Study of planning and operational aspects of tidal in-stream power generation

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STUDY OF PLANNING AND OPERATIONAL ASPECTS OF TIDAL IN-STREAM POWER GENERATION

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A Thesis Submitted To Nanyang Technological University In
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2013

STATEMENT OF ORIGINALITY

I hereby certify that the content of this thesis is the result of work done by me and has not been submitted for higher degree to any other university or institution.

March 2013

Mahda J. Jahromi

To my beloved country

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List of Abbreviations

AR	Auto Regressive
ARMA	Auto Regressive Moving Average
ANFIS	Adaptive Neuro-Fuzzy Inference System
CO-OPS	Center for Operational Oceanographic Products and Services
COF	Cut Off Frequency
CF	Correlation Factor
COV	Co-Variance
cpd	Cycles Per Day
D	Number of Previous Data Values to be Used in a Prediction
EPRI	Electric Power Research Institute
FTDNN	Focused Time Delay Neural Network
F.F	Fitness Function
GA	Genetic Algorithm
GMM	Gaussian Mixture Models
GIC	Grid Imposed Frequency Converter
HAMLS	Harmonic Analysis, Method of Least Squares
HMM	Hidden Markov Model
HPP	Hybrid Power Port
LPF	Low Pass Filter
MPPT	Maximum Power Point Tracking
MLPNN	Multi-Layer Perceptron Neural Network
MAF	Moving Average Filter
MSE	Mean Square Error
NOAA	National Oceanic and Atmospheric Administration
NOS	National Ocean Services
NCOP	National Current Observation Program
NARXNN	Nonlinear Autoregressive with Exogenous inputs Neural Network
N	Number of Neurons in Each Layer of a Neural Network / Number of Non-Linear Units
PCC	Point of Common Coupling
PF	Power Factor
TCS	Tidal Current Speed
TISEC	Tidal In Stream Energy Convertors
U	Eastern-Western Tidal Current Flow Direction
V	Northern-Southern Tidal Current Flow Direction
VSC	Voltage Sourced Converter
VFC	Variable Frequency Converter
var	Variance
Δ	Prediction Horizon (Number of Steps Ahead When Predicting)
ρ	Density
η	Efficiency

Abstract

Even though harnessing marine energy through water wheels can be traced back to the early civilizations, no major step was taken to exploit this vast source of energy until the mid-70s after the first oil crisis. Ever since, many countries have begun to investigate and invest in this vast and predictable source of energy. Nowadays marine energy also referred to as oceanic energy has become a very generic term which spans from the kinetic energy carried by ocean tides and waves to the potential energy of tidal barrages. Out of the aforementioned different types, this study focuses on extracting power from oceanic tidal streams which are generated by the relative motion of the earth and celestial masses interacting via gravitational forces. Based on such generational forces, energy from tidal currents are essentially inexhaustible and as trustworthy as the rising of the sun.

The physics of the energy conversion system of tidal currents is very similar, in principle, to the kinetic energy conversion system in the wind industry; resulting in an inevitable, though superficial, resemblance of many of the existing tidal in-stream energy converting devices to those of the wind turbines. Similar to wind power, the power of a tidal current is proportional to the density of the media and cube of its speed. It is the high density of water which makes the tidal power density impressive, in the sense that at similar sizes and flow speeds one tidal turbine can produce as much power as 1000 wind turbines. Apart from the high power density, based on the astronomical underlying forces, tidal stream energy is also very predictable and independent of terrestrial issues such as fog or lack of wind that can impact the performance of other renewables. High power density and robustness to aesthetic issues makes this source of energy very reliable; a factor which is very important when integrating renewable resources into the conventional networks. Consequently my objective in this study has been to look into the electrical challenges of extracting and transferring of tidal stream power to conventional power networks.

Although the research trend in this area has been gaining momentum in recent years, when searching for similar literature even on a very generic topic such as marine power systems, one still finds very few papers, and even less when the topic is narrowed down to tidal turbines or in-stream generation. This made this study quite challenging at its very early stages since very little technical information was available to start with. In addition, many of the existing papers such as [1-9] merely focused on the overall trends in the field and were basically generic introductions on the topic of oceanic energies. However eventually specific publications helped shaping the study; out of which site feasibility and technology assessment reports of Siddiqui and EPRI [10-13], the work of Hong-da Liu et. al. on connecting methods of intermittent tidal resources to the grid [14] and the review of transmission alternatives of offshore

generation [15] by Alegria, et. al. can be mentioned as a few examples. But probably the most inspiring and influential publications for me were the work of Elghali's group based in the University of Western Brittany (France) who had studied simulation modeling of potential sites for marine current turbines. Although their papers were mostly dedicated to the development of a mechanical model for the turbine blades, their strategy in looking into different parts of a tidal in stream generation system was very helpful[7, 16]. Currently from their latest publications, it seems that they are moving towards designing specified permanent magnet based marine generators [17, 18]. Another paper which I found very inspiring at the beginning of this work was the paper by Biswarup, et. al. of the Imperial College of London who did a thorough investigation on the performance of an Archimedes Wave Swing (AWS) and its effect on the neighboring networks[19]. Although the paper was centered on a wave converting system and not tidal current generators, the author found it very useful due to the many similarities that exist between the two. In other words since the same astronomical generating forces that form tidal currents, shape tidal waves as well, the resulting captured power profiles are similar and so are many of the other electrical challenges.

Despite the limited literature in marine energy systems, when it comes to wind energy and its pertaining technologies, this takes a completely different toll. The wind energy and its affiliated research topics are heading towards maturity and this is very beneficial to marine power systems especially tidal in stream systems which resemble wind turbines both in shape and operational concept. So eventually whenever the required literature in the marine category was missing or was found to be insufficient, the vast literature of the studies in wind energy focusing on similar issues were perceived as inspiring guides and solution to the problem; this is why you will find many of the cited papers in the context to be wind related studies. Out of such, the ones carried out by Reza Iravani of the University of Toronto were most useful particularly his latest book along with A. Yazdani on voltage sourced converters [20].

Chapter 1

Introduction to Tide Fundamentals

1.1 Introduction

Tidal energy is generated by the relative motion of the earth and celestial masses especially the sun and the moon, which interact via gravitational forces. The magnitude of the tides and also the speed of tidal currents at any location is the result of the changing positions of such masses, rotation of the earth, and also the local shape of the sea bed and coastlines. Due to the astronomical nature of the underlying forces, tidal energy is practically inexhaustible and is thus classified as a renewable source of energy. A tidal turbine uses this phenomenon to generate electricity. Numerous techniques have been suggested for extracting marine energy, but the most developed of them can be summarized into the following three main categories:

- *Tidal-Barrages*, which use potential energy (head) difference of sea water
- *Tidal-Stream Energy Converters*, which use the kinetic energy in tidal flows
- *Wave Energy Converters*, which capture the energy in tidal waves

Out of these three, tidal stream power is the focusing point of this study. Similar to wind turbines tidal stream systems make use of the kinetic energy of moving water to power turbines but due to the higher density of water, the available power here is much more significant even at low velocities. Apart from high power density, due to the astronomical underlying forces, tidal energy is also very predictable and independent of terrestrial issues that can impact the performance of other renewables such as lack of wind or fog. High power density and robustness to aesthetic issues thus make tidal energy a very reliable source of energy. Reliability on its own is a very important factor in the integration of most renewable resources into the conventional networks.

Like wind the amount of power in a tidal flow is proportional to the cube of the fluent velocity. So to determine the available tidal flow power in a region over time, the study and forecast of the current speed variations in that region is inevitable. On the other hand, although tidal flows are not as chaotic as wind

speed they too are wavering which means that like wind energy one should not expect a stable flow of power when it comes to tidal currents. This leads to the need for both short and long term prediction of tidal currents so that power system planners and regulators could make detailed schedule plans and set reserve capacities required to compensate for the penetration of such resources.

It should be noted that there have been numerous researches on wind speed prediction like those in [21-24], reviews and comparisons of different prediction methods such as the papers by Lei [25], Costa[26] and Wu [27], or even in other power engineering problems like load estimation [28-34]; but when searching for similar literature on Tidal Current Speed (TCS) prediction, one will find very few papers. Regardless of the prediction techniques or type of data used in these studies, they can all be categorized into two main groups of short and long term predictions. While there can be many definitions for short and long term prediction, the one that is considered in this study, refers to operational and planning concepts.

Although there are articles on long-term prediction like [35-37] that have either used model identification based techniques or models based on deterministic chaos theory, few methods tend to provide accurate long term instantaneous prediction values especially for complex systems or radically varying time series such as that of wind or tidal speed. On the other hand in most of the available literature on time series prediction, the forecasting horizon is only a few steps or a few days, even when it has been declared as long term prediction [38, 39]. This is almost the case for all the studies utilizing artificial intelligent based models. In author's previous publication [40], the performance of different prediction methods of both mathematical (Auto-Regressive models) and soft computing types (Neural and Fuzzy Neural Networks) regarding forecasting of tidal current speed were explored. It was shown that because of the myopic nature of such methods and the problem of error accumulation they cannot predict too far into the future, these techniques will be reviewed again in subsection 1.4. However in this study, it will be shown how these models can be used in conjunction with a conventional tide forecasting technique to produce more consistent and accurate estimations of instantaneous speeds, several years into the future which is the equivalent of predicting several thousand time steps; this is the main contribution of this study. The performance of each technique is tested using the same actual recorded data [41], so that an unbiased comparison could be made.

In order to evaluate the prediction techniques and also to be able to compare their results and performances a %Fit index was used that is briefly described alongside some of the required pre-processing routines, in subsection 1.2. Later on in the study, three categories of estimation routines are

introduced and studied in different sections. In subsection 1.3 the conventional Harmonic Analysis of Least Square (HAMLS) of tidal data is explained and applied, in subsection 1.4 the featured mathematical and soft computing based prediction techniques are put to test and in subsection 1.5 a novel technique is proposed which combines the previous estimation methods. Conclusively in subsection 1.6 it is shown that based on the results obtained, the hybrid technique shows improvement when compared to each of its constituent components.

1.2 Accuracy Index and Data Pre-processing

Unlike tides, currents can follow any of three spatial dimensions – one vertical and two horizontal. In this study, the vertical direction is neglected, so it is assumed that the tidal currents only move horizontally. In addition to that, although simultaneous analysis of water current in two horizontal dimensions is usually possible, to simplify the problem further, the dimensionality of the problem is reduced to one having the largest possible variance, i.e. the principal axis. Figure (1.1) shows the justification of this dimensionality reduction. With each single dot showing where the tip of an individual current vector would lie in a U (East – West), V (North – South) plot it is clear that most of the total variance of the data lies in one axis, named here as Up . This means that elimination of the Vp axis does not alter the analysis significantly [42] and so an eigenvector algorithm is used to mathematically orient the data to the new Up, Vp coordinates and the heading is assigned by convention, i.e. positive coordinate axis represents the new flood (landward) direction and the negative coordinate axis represents the new ebb (seaward) direction. More details of each of such pre-processing can be found in [40, 42].

The data used in this study, are taken from [41] and belong to the Philadelphia station during 2007 to 2009 period. Before the raw data could be used, a number of pre-processing had to be done since as the website itself states in its disclaimer the data are not verified and hence may contain mistakes. So the data are checked, filtered for high frequency hash removal, de-trended and rotated to principal coordinates before any other use. The channel profile of where the recording station is situated is clear, so when going through the UV plot, any data that lies too far from the Up axis is either a sensor misread or result of Eddie currents that do not contribute to power generation (at least not with the current technology) and can be eliminated.

Figure 1.1: Tidal Current Data in a (U,V) Plane

According to [42] and based on experimental results the explained pre-processing does not alter the HAMLS model, meaning that HAMLS model can be fitted on the original data as long as the recording serial time remains accurate; on the other hand although filtration of the data increases the training efficiency of the model free estimators, it is not vital. In practice though, the high frequency distortions do not reach the generator since the mechanical time constants of the turbine itself do not let through any sudden and rapid changes of speed, in other words the mechanical elements act as a low pass filter. However the integrity of the data must not be altered in any pre-processing, this is why a MSE index was used to verify this matter; Figure1.2 shows the actual and processed data of the Philadelphia station during the first week of Jan 2007, the Mean Square Error regarding the two sets of data here was less than 3%.

Figure 1.2: Tidal Speed (U) Before and After Processing

In order to be able to evaluate and compare the performances, a %Fit index is used which is defined as follows.

$$\%Fit = \left(1 - \frac{\|y - \hat{y}\|}{\|y - \bar{y}\|}\right) \times 100 \quad (1.1)$$

Here y represents the actual values, $\hat{y}$ estimated values, and $\bar{y}$ is the mean of y , where y is the current speed flowing in the direction of the principal axis. A fit value of 100% corresponds to a perfect fit while a 0% indicates that the fit is no better than guessing the output to be a constant ($\hat{y} = \bar{y}$). Due to this definition, it is possible for Fit% values to become negative. A negative fit is worse than 0%.

1.3 Harmonic Analysis of Tides

Orbital paths of astronomical bodies that cause tidal flows are very nearly circular, so sinusoidal variations are suitable for tides; consequently tidal patterns can be decomposed into many sinusoids having many fundamental frequencies, corresponding to, many different combinations of the relative motions of the earth and celestial masses. The study of tides by harmonic analysis was begun by Laplace, William Thomson (Lord Kelvin), and George Darwin, while A.T. Doodson extended their work, introducing the Doodson Number notation to organize the hundreds of resulting terms. The method that is used today to analyze a water current time series is commonly known as Harmonic Analysis, Method of Least Squares (HAMLS), which is a modified version of what Doodson, had proposed in his book [43]. This method achieves a progressive reduction in variance by adding harmonic terms with specific astronomical frequencies to a general least squares model. With the astronomical frequencies associated with celestial relative motions, the success of a harmonic model depends entirely on the ocean's response to celestial (gravitational) forcing at these same frequencies, recognizing that oceans are free to respond to local (meteorological) forcing at the same time. The equation for the harmonic model in this instance is:

$$y(t) = y_0 + \sum_{j=1}^m f_j H_j \cos(\omega_j t + \theta_j - \delta_j) = y_0 + \sum_{j=1}^m R_j \cos(\omega_j t - \varphi_j) \quad (1.2)$$

Where:

t = Time in serial hours (serial time is the fractional number of hours past a specified time origin as opposed to 24-hour solar time)

$y(t)$ = Predicted water current speed at time t

y_0 = Mean water current speed

f_j = Lunar node factor for j^{th} constituent*

H_j = Mean amplitude for j^{th} constituent over 18.6-year lunar cycle

ω_j = Frequency of j^{th} constituent

θ_j = Nodal phase for j^{th} constituent*

δ_j = Phase of j^{th} constituent for the time origin in use (Midnight Beginning December 31, 1899)

m = Number of constituents

*For purely solar constituents, $f_j = 1$ and $\theta_j = 0$. (Others are obtained by formula, see [23])

Figure 1.3: HAMLS Model Based on Philadelphia

So basically what HAMLS does is a simple but powerful procedure of obtaining tidal constituent amplitude (R_j) and phase (ϕ_j) angles, the so-called *tidal harmonic constants* needed for tidal modeling and tidal predictions using Equation (1.2). The least squares criterion requires a solution for the harmonic constants that will produce the minimum possible sum of squared differences for a series of observations y_t (actual values) of length n , described mathematically as follows:

$$\text{Minimize} \left(\sum_{t=1}^n [y_t - y(t)]^2 \right) \quad (1.3)$$

For this purpose, Equation (1.2) is rewritten in the equivalent form as:

$$y(t) = A_0 + \sum_{j=1}^m A_j \cos \omega_j t + \sum_{j=1}^m B_j \sin \omega_j t \quad (1.4)$$

Where:

$$A_0 = y_0, \\ R_j = \sqrt{A_j^2 + B_j^2} = f_j H_j \text{ and } \varphi_j = \tan^{-1}\left(\frac{B_j}{A_j}\right) = \delta_j - \theta_j$$

The unknowns (A_0, A_j, B_j) in Equation (1.4) are obtained by solving the general matrix equation for least squares approximations:

$$[C] = M^{-1} \cdot N \quad (1.5)$$

In the above, $[C]$ is a $(2m + 1) \times 1$ vector of unknowns, so that, $[C] = [A_0 A_1 B_1 A_2 B_2 \dots A_m B_m]^T$, with $M = [X]^T [X]$ and $N = [X]^T [Y]$, where:

$$[X] = \begin{bmatrix} 1 & \cos \omega_1 t_1 & \sin \omega_1 t_1 & \dots & \cos \omega_m t_1 & \sin \omega_m t_1 \\ 1 & \cos \omega_1 t_2 & \sin \omega_1 t_2 & \dots & \cos \omega_m t_2 & \sin \omega_m t_2 \\ 1 & \cos \omega_1 t_3 & \sin \omega_1 t_3 & \dots & \cos \omega_m t_3 & \sin \omega_m t_3 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ 1 & \cos \omega_1 t_n & \sin \omega_1 t_n & \dots & \cos \omega_m t_n & \sin \omega_m t_n \end{bmatrix}$$

and $[Y] = [U_1 U_2 U_3 \dots U_n]^T$, which is a vector containing n observations.

Note that, while there is a term representing the mean in Equation (1.2), there is no term representing a linear trend. This is because as mentioned before data prior to analysis are de-trended. Literally this procedure should be called a *water current analysis* (rather than a tidal current analysis) because the measured currents in coastal waterways, gulfs etc. oscillate at both tidal and non-tidal frequencies, including frequencies so low they appear as a mean current or linear trend, but “tidal current analysis” is a well-established phrase that is also used here. As will be explained in later sections the linear trend can be picked up by model free estimators. The objective at this stage is to separate the tidal components so that a *tidal current prediction* can be made with terms that are predictable due to their astronomical nature. In

this section this theory is put to test using actual recorded data from Philadelphia recording station [41]. According to [42] even a 29-day record of water currents, yields enough information to capture the most significant terms in the list of all known constituents. Figure 1.3 shows the HAMLs model used in predicting future current speeds. The model as explained before was calculated based on Philadelphia station data, recorded during Jan 2007. As can be seen the HAMLs model tracks the actual recordings quite well. The following prediction results are all based on the same HAMLs model created from a 31 day (Jan 2007) harmonic analysis of Philadelphia station's TCS data. The prediction has been carried out for the first 1000 data points subsequent to the indicated prediction horizon (Δ); so for example if the prediction horizon is said to be 1 month, then the prediction has been done for the first 1000 data points of March 2007 or if the prediction horizon is said to be 1 year then the first 1000 data points generated by the HAMLs model for Feb 2008 were compared with their actual values in that period. It should be noted that the sampling time for the recorded data as stated by [41] is 6 minutes, so 1000 data points represent 100 hours or a little bit more than 4 days which should be enough to indicate whether the model is forecasting accurately or not.

Figure1.4: Prediction results using the HAMLs model

The performance of a HAMLs model can be improved by including different harmonic constituents. If more influential constituents are considered in the calculation the resulting model would explain a better portion of the actual variations. The two Figures bellow show two periodograms drawn with respect to the number of constituents considered in the earlier HAMLs calculation process. A higher number of

constituents in the analysis usually results in a better model but selection of too many constituents for a very short dataset can result in an unstable model when used for future predictions. As can be seen in Figure 1.5, there is a large peak with energy of $3.5 \times 10^{-3}(J)$ at 3.862 cycles per day (cpd) and two smaller ones at lower frequencies. Consulting the available constituent frequency table [42], the first frequency is close to the quarter diurnal constituents M4 and MS4 while the second falls midway between semidiurnal constituents LAM2 and L2. The second Figure shows the periodograms when those constituents have been taken into account. As can be seen the spike near 4cpd is almost gone but the other two low frequency spikes have not changed at all. This is because no astronomical model can ever explain 100 percent of the variance in current speed data. Low-frequency currents from meteorological forcing and ‘noise’ virtually guarantee it. Predictions from astronomical harmonic constituents need therefore, to be modified to encounter aesthetic issues such as sea floor changes, wind periods (which may blow water into or out of a bay, causing a significant change in current speeds), extreme storm surges or changes to atmospheric pressures which can increase or decrease water levels, etc.

But inclusion of all such phenomena leads to complex and uncertain calculations, resulting in less reliable predictions. This is where other types of predicting methods can come into play. To avoid such complexities one could make use of model-free estimation techniques trained by actual data that have been recorded over a sufficient period of time. Recently, neural networks, fuzzy systems and identification based techniques such as auto-regressive models have been successfully applied to many real world problems. These techniques are referred to as model-free estimators because they build models from input/output data without other knowledge about the system. In the following section some of these models are introduced further and applied to the TCS prediction problem.

Figure 1.5: Residual periodograms using $[O1, K1, N2, M2, S2]$ as constituents

Figure 1.6: Residual periodograms using $[O1, K1, N2, M2, S2, M4, MS4, LAM2, L2]$ as constituents

1.4 Model Free Estimators

There are various prediction techniques. Some of those are specifically designed for specific conditions, while others are more generic. Here six different estimation techniques have been selected based on their recognition in forecasting literature and also their iterative learning nature. The six model-free estimators considered in this study, are Multi Perceptron Neural Network (MLP), Focused Time Delay Neural

Network (FTD), Nonlinear Auto Regressive Neural Network (NARX), Adaptive Neuro-Fuzzy Inference System (ANFIS), linear and non-linear Auto-Regressive Models (AR). These models are divided into two categories: Soft-Computing based techniques which embrace the neural networks and ANFIS and Identification based models which include linear and non-linear auto regressive models. All techniques have been optimized using a Genetic Algorithm which searches for the best combination of factors affecting the performance of these models. In this section a brief introduction of each model precedes the overall results which are given as barographs at the end.

1.4.1 Prediction by Soft Computing Methods

Soft Computing techniques are a set of methods that have been inspired from nature and are used to solve problems that conventionally are very difficult to solve. Neural Networks and ANFIS are two prominent soft computing based learning algorithms that use input and target values. However in most time series prediction scenarios, there is only one set of data, e.g. the recorded tidal current speeds in this study; so the same dataset should be used to provide both the input and target values. In other words future values (Targets) are predicted based on previous data values (Inputs). Regarding soft computing models used in this study, there are many factors that can influence their performance. Apart from the two choices of data sampling formats (see appendix), there are a wide range of parameters like the number of layers and neurons in each layer, number of membership functions, initial conditions, learning factor etc. that can affect their performances. Unfortunately due to the nature of these networks in most cases there is no sound method for selecting such variables; however here a Genetic Algorithm was utilized to adjust the models to their optimized performance. The GA tries different combinations of starting states, network architecture (number of layers and neurons), type of input/target data sampling (see Appendix), number of training samples (D), and where applicable regressors. This process is repeated for each solution in a generation so that new generations are ameliorated compared to their predecessors. This way an optimal solution (here being an effective combination of the model parameters) can be found. Although the chance of GA failing to find a true or global optimum increases in complex or wide search spaces, GA usage however, does guarantee, finding more favorable solutions compared to random or arbitrary assigning of such parameters. The soft computing models used here include ANFIS and several neural networks of static and dynamic nature. Static (feed-forward) networks have no feedback elements and contain no delays; the output is calculated directly from the input through feed-forward connections. In dynamic networks, the output depends not only on the current input to the network, but also on the previous inputs, outputs, or states of the network. Therefore dynamic networks require a different training method and have a different architecture to those of static networks. FTD and NARX are both of dynamic nature while MLP is a static network. Here each model is described further.

1.4.1.1 Multi-Layer Perceptron (MLP)

As the name implies a MLP neural network can have several layers. Each layer has a weight matrix, a bias vector, and an output vector. The layers of a MLP network play different roles. A layer that produces the network output is called an *output layer*. All other layers are called *hidden layers*. The network is repeatedly exposed to a set of training data and based on the resulting outputs, errors are calculated which are used to adjust weights and biases. This will eventually lead to optimum weight and bias values that can mimic any available input-output correlation, given that the network *can* mimic the model. The MLP network is the most commonly used neural network in prediction studies, among which [44-46] are a few examples. Although MLP is itself of static nature, the type of training used in this study, is similar to that of a dynamic network, since previous values are used as inputs and future values are used as target which is similar to a network with a tapped delay input. The difference is in the type of sampling used in the training which in this case is of fixed time steps (Appendix Fig. A1.1). The parameters selected for optimization here were, the number of network layers, number of neurons in each layer, the number of previous data used in prediction (D), and also initial values. Since these parameters are all dependent, the problem cannot be solved as a conventional decoupled multi-variable optimization problem and this is why a specific genotype was developed for the GA optimization.

1.4.1.2 Focused Time Delay (FTD)

A Focused Time Delay Neural Network is a dynamic network which consists of a feed-forward network similar to MLP with a tapped delay line at the input. The dynamic nature of the network makes it very flexible and also efficient in predicting time series. The optimization procedure for this network was similar to that of the MLP network with slight differences in the chromosome design regarding the GA routine. The sampling used for this network was of variable time steps (Appendix Fig. A1.2).

1.4.1.3 NARX

The FTD network is a focused network, with the dynamics only at the input layer and is therefore a feed forward network. The nonlinear autoregressive network with exogenous inputs (NARX) is a recurrent dynamic network, with feedback connections enclosing several layers of the network. The NARX model is based on the linear AR model, which is commonly used in time-series modeling. So this network is basically a combination of a neural network and an autoregressive model. The defining equation for the NARX model is:

$$y(t) = f \left(\begin{matrix} y(t-1), y(t-2), \dots, y(t-n_y) \\ u(t-1), u(t-2), \dots, u(t-n_u) \end{matrix} \right) \quad (1.6)$$

where the next value of the dependent output signal $y(t)$ is regressed on previous values of the output signal, and also the previous values of the independent (exogenous) input signal $u(t)$. One can implement the NARX model by using a feed forward neural network to approximate the function f . This implementation also allows for a vector AR model, where the input and output can be multidimensional. The optimization regarding NARX networks included the number of layers and neurons and also the delay vectors pertaining to its input and output.

1.4.1.4 ANFIS

The acronym ANFIS derives its name from *Adaptive Neuro-Fuzzy Inference System*, and is basically a combination of fuzzy inference systems and iterative learning techniques similar to those used in neural networks. The basic structure of fuzzy inference system, is a model that maps input characteristics to input membership functions, input membership function to rules, rules to a set of output characteristics, output characteristics to output membership functions, and the output membership function to a single-valued output or a decision associated with the output. There are various membership functions with different shapes. The shape of such functions is dependent on a number of parameters which can be treated as variables and need to be adjusted for the system to act in accordance to a predefined behavior. Therefore by initially assigning a number of flexible membership functions and progressively updating their parameters based on the difference of the input and output values (same as training neural networks), the fuzzy inference system can be tuned for different applications such as data modeling or estimation. Using a given input/output data set, ANFIS constructs a fuzzy inference system (FIS) whose membership function parameters are tuned using either a back propagation algorithm alone or in combination with a least squares type of method. This adjustment allows the fuzzy systems to learn from the data it is modeling. ANFIS has also been quite attractive for researchers working on intelligent forecasting models, [46, 47] are two examples of such. The type of membership function used in this study, is a typical Gaussian shaped function which can be described using the following equation.

$$f(x, \delta, c) = e^{\frac{-(x-c)^2}{2\delta^2}} \quad (1.7)$$

Due to the exponential increase of the size of ANFIS's rule-set an upper limit was imposed on D (number of previous data values used to predict future value) and $nMFs$ (Number of Membership Functions). Since the rapid increase in the size of the rule base, limits the number of feasible MFs and D values, there was no need for GA optimization as the search space was small enough for an exhaustive trial and error of different combinations within the assigned limits. The sampling format used here was similar to that of Appendix Fig. A.1.

1.4.2 Prediction By Identification Based Methods

There are many identification based routines such as Auto Regressive Moving Average (ARMA), Box-Jenkins Model [48] (BJM), Hammerstein-Wiener Model [49], etc. Auto Regressive or AR is an identification based method which comes quite handy when searching for mathematical estimation of dynamic systems by the use of measured data. The resulting model can then be used to simulate the output of the system for any given input or more generally system analysis, prediction etc. AR is especially useful for modeling systems that cannot be easily modeled from first principles or specifications, having said that tidal current streams are marked nonlinear systems that are very difficult to analyze and model, hence the whole system is seen as a black-box modeling process, where the measured data determine the model structure. AR modeling is very prominent in wind speed prediction, [46, 50] are three examples of such. There are two types of AR modeling, linear and nonlinear. Here both models have been examined and their results were plotted against results of other prediction methods. In the following a brief introduction of AR models is given. It must be bared in mind that in this case there is no input but rather just one set of data, named $y(t)$ which is recorded TCS data.

1.4.2.1 Linear Auto Regressive (A.R) models

The following linear polynomial model structure is the basis for linear AR modeling:

$$y(t) = \sum_{i=0}^N A_i(q).y(t - \theta_i) + \frac{C(q)}{D(q)} e(t) \quad (1.8)$$

The polynomial A_i , C and D contain the time-shift operator q . N is the total number of inputs, and θ_i is the i^{th} delay that characterizes the delay response time. The variance of the white noise $e(t)$ is assumed to be λ . To estimate a polynomial model, initially the model order should be specified as a set of integers. The general polynomial equation is written in terms of the time-shift operator q . The definition of this time-shift operator is best explained as in the following discrete-time difference equation:

$$y(t) = a_0 y(t) + a_1 y(t - T) + a_2 y(t - 2T) \quad (1.9)$$

Here T is the sampling interval and q^{-1} is a time-shift operator that compactly represents such difference equations using $q^{-1}.y(t) = y(t - T)$, hence:

$$y(t) = a_0 y(t) + a_1. q^{-1} y(t) + a_2. q^{-2}. y(t) = A(q).y(t) \rightarrow A(q) = a_0 + a_1. q^{-1} + a_2. q^{-2} \quad (1.10)$$

The model structures differ by the number of such polynomials that are included in the structure. Thus, different model structures provide varying levels of flexibility for modeling the dynamics and noise characteristics. As will be described shortly the linear AR can be implemented as a subset of the more general non-linear AR modeling.

1.4.2.2 Non-Linear Auto Regressive Models

Nonlinear AR models describe nonlinear structures using a parallel combination of nonlinear and linear blocks. The nonlinear and linear functions are expressed in terms of variables called *regressors*, which can be calculated by performing transformations of the $y(t)$ data based on the specified model order. For example, regressors can be delayed data, such as $y(t - 1)$ or $y(t - 3)$ or even nonlinear functions, such as $\tan(y(t - 1))$ or $y(t - 1)^2$, though in this study, only regressors of delayed data type have been used. The predicted output $\hat{y}$ of a nonlinear model at time t is given by the following general equation:

$$\hat{y} = F(\mathbf{y}) = f(\mathbf{y}) + L^T \cdot \mathbf{y} \quad (1.11)$$

Here $\hat{y}$ is a summation of linear and nonlinear regression functions. f is a nonlinear function which can be approximated by a nonlinearity estimator such as a *binary partition tree*, a *sigmoid* or a *wavelet network*. The following equation provides a general description of f .

$$f(y) = \sum_{k=1}^N \alpha_k \cdot G(\beta_k \cdot (y - \gamma_k)) \quad (1.12)$$

The second term in equation (1.11), describes the linear part which was explained previously. Here L is a vector describing the polynomial coefficients pertaining to each regressor. So the linear AR model can consequently be seen as a subset of the general non-linear model. Figure 1.7 shows how the predicted output of this model is formed.

Figure 1.7: General AR Model

Regarding Equation (1.12) G is the nonlinearity function, N is the number of nonlinearity units, and α, β and γ are the parameters of the nonlinearity estimator. As aforementioned the non-linear portion of the model is selected from a list of 3 pre-defined non-linear mathematical functions. The optimization in this case was on selecting the best non-linearity, regressor set and also the number of nonlinear units that could result in the highest $\%Fit$.

1.5 Comparison of Results

The following bar graph depicts the overall performances of all the aforementioned model free prediction techniques. As can be seen from the results the accuracies are quite high, in some cases almost exact, but only when the prediction horizon is limited to 5-10 steps, which with respect to the sampling intervals of 6 minutes converts to less than an hour ahead prediction. Looking at the results it is evident that the NARX, FTD and non-linear AR models are the best options for short term predictions with NARX having slightly better results.

Figure 1.8: Results of Model Free Predictors (using 1 year of training data)

Figure 1.9: Results of the top three model-free estimators regarding different prediction horizons (Δ)

Figure 1.9, shows the prediction results considering longer prediction horizons. It is seen that the non-linear AR model is more robust to the increase of prediction horizon as its fit percentage line decreases with a slower pace compared to the other two. One should bear in mind that the decreasing rate pertaining to ANFIS, MLP and Linear AR are much higher than that of the depicted models.

The high precision of these models in short term predictions is of merit only in operational objectives and cannot be used for planning purposes; however their ability in learning hidden trends in sequential data is very useful. In the next section it will be shown how this property can be put to use in long term prediction of tidal currents.

1.6 The Hybrid Method

As seen from the results of previous sections, although the HAMLs method is good in long term prediction, since it merely relies on astronomical factors, it lacks the ability of tuning itself to terrestrial effects such as weather and sea bed changes, temperature variations etc. On the other hand while model free predictors are good when it comes to learning unobservable trends, their accuracies drop drastically when predicting more than an hour into the future, in other words they are very myopic. So, in order to have both virtues at the same time a hybrid method is proposed by the author. The best way to describe the hybrid method is to exemplify it using a coarse and fine tuning concept, meaning that the coarse prediction is done by the HAMLs method while model free estimators (NARX, FTD and AR) do the fine tuning. So a short set of TCS data will be used to make a harmonic model using the HAMLs method and future predictions of that model are used as the input to a model free estimator. This way the model free estimator can be trained based on the difference between the HAMLs predictions and actual recorded values. This ultimately results in the model free estimator learning the difference of values which are by and large caused by terrestrial effects. Since most of such effects have rather long time constants with their effects perceivable only in durable time frames, the longer the training (in terms of number of data used) the more accurate the final outcomes will become.

Here the same HAMLs model used in subsection 1.3 was used to produce the inputs to the selected top three model free estimators of subsection 1.4. These models were then trained in two different scenarios using separate sets of data; One being recorded over a 6 months period and the other a year of consecutive TCS data. Figure 1.10 shows the best predictions by the hybrid model in each situation. Evidently in both situations the Hybrid model shows improvement compared to the conventional predictions by the HAMLs method. Details of the last three models can be found in the appendix. Although the improvement may not be profound, one should bear in mind that any improvement in the accuracy of tidal velocities will affect the power density calculations treble since power density is proportional to the cube of fluid velocity. Also from the result one can see the effect of the training data size, since a longer training set, results in a better performance. This is mostly because the secondary predictor (model free predictor) has more time to pick up any concealed trends and so can generalize the input more effectively.

Figure 1.10: Comparison of fit percentages of the Hybrid and HAMLS Models.

1.7 Discussion

In this section, three types of prediction methods are introduced and tested. The conventional HAMLS method was proved to be the most efficient and robust predictor, mimicking tidal current patterns quite effectively. It only requires a few weeks of recorded data and is able to predict well into the future. However HAMLS method is limited to celestial movements and does not consider any terrestrial factors that also influence tidal flows and current speeds. Model free estimators on the other hand have been proven to be very good in learning indefinite trends in time series data, but are very myopic especially when the data under study is chaotic. Nevertheless when limited to a dozen time steps they are almost perfect and so can be very useful for short term predictions and operational purposes. Using both predictors sequentially, a novel predicting technique was proposed that could incorporate both properties simultaneously. Although the training period and the amount of data needed is much more than in the case of merely using HAMLS, the hybrid method was found to produce more accurate results by incorporating both terrestrial and celestial factors that shape and influence tidal currents.

Although the intention here has been to include a vast number of prediction alternatives, obviously the available methods and models are much more than the ones utilized in this study. There are novel estimation techniques such as GMM and HMM that use probabilistic models or individual neural

networks specifically design for predicting time series of different kind, which have not been included in this study, but are recommended for any future investigations. In addition, the HAML method itself can be improved further by employing better models like the one discussed in [51]. Therefore this study should be seen as a preliminary attempt to combine conventional oceanographic methods with novel artificial intelligent based techniques to depict the prominent property of oceanic energy, i.e. its predictability.

1.8 Scope of Work

Since tidal stream power is highly dependent on the current speed, my investigation begins with a comprehensive study of the nature of tides and tidal current speed forecasting techniques. In this part of the study, it is shown how combining conventional prediction methods that rely merely on astronomical circulating periods, with model free estimators such as neural networks can improve tidal speed forecasting, estimating the speed several years into the future. Knowledge of the current speed behavior allows power system planners and regulators to design detailed power schedules and plan reserve capacity settings that are needed in compensating for the penetration of intermittent resources. The study then continues with the design and performance evaluation of a tidal in-stream generation system that utilizes induction generators in conjunction with multilevel voltage sourced converters.

1.9 Organization of the Thesis

Chapter 1 deals with different tidal current forecasting techniques and comparison of their results.

Chapter 2 provides fundamental concepts required for the converter. An overview of the structure of the Voltage-Sourced Converters (VSC) that has been used throughout the study is presented.

Chapter 3 provides insight into how a decoupled active/reactive power control strategy can be achieved and employed in a power port system that transfers variable amounts of tidal power to a utility grid.

Chapter 4 consists of the control system designing.

Chapter 5 provides the design details for current speed forecasting, which is used in Hybrid Power Port.

Chapter 6 deals with the performance evaluation of the entire system.

Chapter 7 draws conclusions to the overall study and makes recommendations for future works and improvements. Finally references are given in chapter 8.

Chapter 2

Overview of Neutral Point Clamped Voltage-Sourced Converter

2.1 Introduction

To connect a variable frequency system, in this case a generator that rotates at different speeds, to a grid that is of constant frequency, the power must be conditioned through a back to back AC/DC/AC converting system. Such a system requires two sets of converters, one at the machine side and another at the grid side, tied together through a DC link. In this study regardless of the control systems that drive each converter or their operating modes, i.e. inverting or rectifying, the structure of the two converters and the components used in them are the same. In other words the two different converter systems differ only in their control systems and not their architecture. Thus I begin section two of my thesis with an overview of the structure of the Voltage-Sourced Converters (VSC) that I have used throughout the study.

Based on the number of different output voltage levels that a VSC provides, they can be categorized into two or multi-level converters. Although the 2-level VSCs are the dominant building blocks for a wide variety of apparatuses in the medium to high power level applications, the switch cells in these converters must bear the whole reverse DC link voltage when they are in the *OFF* state. This problem makes the converter feasibility in the higher voltage/power applications quite limited as the type of switch cells that can withstand such high voltage/power levels are often the state of the art and naturally very expensive. Apart from the cost issue beyond a certain voltage rating, even the best available switches may not be able to meet the voltage requirements. One approach to solving the problem is to break the voltage requirement among a number of switch cells that have lower voltage ratings and are connected in series; however this approach has many problems such as the need for synchronized switching signals and similar snubber systems that provide equal voltage division among the switches. Simultaneous switching in turn requires precise timings, wirings and noise effect minimization. Furthermore no two switch modules are ever identical which calls for pains taking signal tuning. So in general series switch connection is not very desirable. In view of the foregoing issues, multilevel converters provide a better solution when it comes to the high-power industrial and utility applications [52, 53]. There are different types of multilevel converters each with their cons and pros, such as the Cascaded H-Bridge (CHB), the

Flying Capacitor (FC), and the Neutral Point Clamped (NPC) converter. In this study, a 3-level (NPC) converter was preferred since unlike the CHB it does not require a separate DC source and hence avoids a bulky and nonstandard transformer and is also relatively cheaper than the FC type converters [54]. The 3-level Neutral Point Clamped (NPC) converter[55] is a multilevel voltage sourced converter that offers an alternative to the switch voltage rating problem by reducing the switch voltage burden. In this approach, each switch cell withstands only half of the DC voltage; thus, reducing the number of switches to be connected in series. Moreover, compared to the conventional 2-level converters the 3-level NPC provides a 3-phase AC voltage that has a lower Total Harmonic Distortion (THD) level. This chapter introduces the NPC converter which is used extensively throughout this study.

2.2 Circuit Structure

Figure 2.1 shows the schematic diagram of a 3-level NPC. As is illustrated in the Figure, the 3-level NPC is made up of three identical half bridge NPCs, the so called converter legs. The three converter legs are connected in parallel and supplied at the DC side by a divided voltage source. The voltage reference here is the DC side midpoint, which is noted in the Figure as node 0. Once connected to an AC system the converter allows a bidirectional flow of power from the DC-side source to the three-phase AC system. The AC system can be an electric machine, a utility power grid, a passive load or a combination of all.

Figure 2.1: Schematic diagram of a 3-level NPC converter

2.3 Operation Principles

The AC-side terminal voltage of a3-level half-bridge NPC is given by:

$$\bar{V}_t(t) = \frac{V_{DC}}{2} m(t) \quad (2.1)$$

where $\bar{V}_t$ indicates the average value of V_t and $m(t)$ is the modulating index.

Consequently the three phase voltages of the NPC converter depicted in Figure 2.1 can be expressed as:

$$\begin{cases} V_{ta}(t) = m_a(t) \frac{V_{DC}}{2} \\ V_{tb}(t) = m_b(t) \frac{V_{DC}}{2} \\ V_{tc}(t) = m_c(t) \frac{V_{DC}}{2} \end{cases} \quad (2.2)$$

What Equation (2.1) indicates is that the converter AC-side terminal voltage $V_t(t)$ can be controlled by $m(t)$, i.e. the modulation signal. In a voltage source inverter system, the terminal voltages and currents are usually required to be sinusoidal. In order for the inverter AC-side terminal voltages to be balanced, m_{abc} must also be a balanced three-phase sinusoidal function. $m_{abc}(t)$ is usually generated by a closed-loop control system that regulates i_{abc} . In general however, m_{abc} can be expressed in the following form, where $\varepsilon(t)$ holds the phase angle and frequency information and $\hat{m}$ is the modulation index amplitude.

$$\begin{cases} m_a(t) = \hat{m}(t) \cos[\varepsilon(t)] \\ m_b(t) = \hat{m}(t) \cos[\varepsilon(t) - \frac{2\pi}{3}] \\ m_c(t) = \hat{m}(t) \cos[\varepsilon(t) - \frac{4\pi}{3}] \end{cases} \quad (2.3)$$

The NPC converter must be designed so that its AC terminal voltages are able to change amplitude and/or phase angle promptly. Based on Equation (2.2), rapid changes in the amplitude and phase angle of the terminal voltages are possible by controlling the m_{abc} . Consequently like in many other three phase systems vector control strategies (in $\alpha\beta$ or dq frame) are used extensively in converter control systems.

2.3.1 Neutral Point Current

Although the DC-side midpoint in a 2-level converter is normally considered as the reference node for circuit analysis, practically it is neither available nor accessible. However the DC-side midpoint of the 3-level NPC shown in Figure 2.1 is a physical node which can be accessed. In the NPC converter, the DC voltage is equally divided in two and connected to the midpoint node. It can be shown that the midpoint current of a 3-level neutral point clamped converter is given by [20]:

$$i_{np}(t) = -\left(\frac{6\hat{m}\hat{i}}{\pi}\right)\cos\gamma \sum_{h=3,9,15,\dots}^{+\infty} \frac{1}{h} \sin\left(\frac{h\pi}{2}\right) \cos(h\epsilon) + \left(\frac{3\hat{m}\hat{i}}{\pi}\right) \sum_{h=3,9,15,\dots}^{+\infty} \frac{1}{h+2} \sin\left(\frac{h\pi}{2}\right) \cos(h\epsilon + \gamma) \\ + \left(\frac{3\hat{m}\hat{i}}{\pi}\right) \sum_{h=3,9,15,\dots}^{+\infty} \frac{1}{h-2} \sin\left(\frac{h\pi}{2}\right) \cos(h\epsilon - \gamma) \quad (2.4)$$

where $\hat{i}$ indicates the peak output current (phase). As Equation (2.4) indicates, while the neutral point current has no DC component, it includes odd triple-n harmonics, in which the third harmonic is the dominant one. Hence, it can be approximated as:

$$i_{np}(t) \approx -\left(\frac{2\hat{m}\hat{i}}{\pi}\right)\cos\gamma \cos(3\epsilon) i_{np}(t) - \left(\frac{3\hat{m}\hat{i}}{5\pi}\right)\cos(3\epsilon + \gamma) - \left(\frac{3\hat{m}\hat{i}}{\pi}\right)\cos(3\epsilon - \gamma) \\ = \left(\frac{4\hat{m}\hat{i}}{5\pi}\right)[-2\cos\gamma \cos(3\epsilon) - 3\sin\gamma \sin(3\epsilon)] \quad (2.5)$$

Equation (2.5) can be rewritten in a compact form as [20]:

$$i_{np}(t) \approx \left(\frac{4\hat{m}\hat{i}}{5\pi}\right)\sqrt{4 + 5\sin^2\gamma} \cdot \cos(3\epsilon + \zeta) = \left(\frac{4\hat{m}\hat{i}}{5\pi}\right)\sqrt{9 - 5\cos^2\gamma} \cdot \cos(3\epsilon + \zeta) \quad (2.6)$$

where $\zeta = \pi - \tan^{-1}(1.5 \tan \gamma)$.

Equation (2.6) indicates that the size of the midpoint current is governed by the converter operating point. It also implies that i_{np} amplitude is a nonlinear function of the converter power factor as well as being directly proportional to the converter AC terminal voltage. Consequently, the current amplitude is maximized, when the converter operates at zero power factor ($\cos \gamma = 0$) and at minimum when power is exchanged at unity power factor. Based on (2.6), the following inequality estimates the amplitude of the midpoint current:

$$0.51\hat{m}\hat{i} \leq i_{np} \leq 0.76\hat{m}\hat{i} \quad (2.7)$$

What this indicates is in cases where the 3-level NPC is mainly exchanging reactive power with the AC system the size of its midpoint current can reach 76% of the AC-side current. Furthermore even if the converter operates at near unity power factor, the amplitude of the relative midpoint current is still not less than 51%. Therefore, the DC voltage sources must accommodate a relatively large 3rdharmonic midpoint current, which is very important when implementing the 3-level converter design. Compared to the equivalent 2-level converter this can be seen as a disadvantage of the 3-level NPC converter.

2.3.2 Using a Capacitive DC Voltage Divider

Figure 2.2 shows an alternative configuration for delivering the two equal DC voltages required in the 3-level NPC converter. As can be seen from this Figure, the 3-level NPC employs a capacitive voltage divider at its DC side. As shown in Figure 2.3, the configuration of Figure 2.2 can be augmented to constitute a back-to-back converter configuration, permitting a bidirectional power exchange between two AC systems (not shown in the Figure) that are interfaced with the converters.

Figure 2.2: Three level NPC with a capacitive DC voltage divider

Figure 2.3: Back to back converter system configuration

If the two capacitors are nominally identical, the total DC voltage would be equally divided between them. However in reality if no corrective measures are taken, the moment the system starts exchanging power, DC capacitors voltages will drift from their nominal values. The voltage drifting occurs under both steady state and transient conditions. The cause of this phenomenon is the tolerances of the converter components as well as the asymmetries of the switches gating commands. Another issue with this type of a converter is that, due to the large third-harmonic component of the midpoint current, each of the partial DC voltages includes a third-harmonic ripple that causes low-order voltage harmonics at the converter AC terminals. To reduce the harmonic voltage, adequately large DC capacitors must be employed.

2.3.3 Capacitor Voltage Drift Phenomenon

Equations (2.4) and (2.6) indicate that the midpoint current, $i_{np}(t)$, has no DC component. However equations (2.4) and (2.6) were developed based on the assumption that the three half-bridge NPCs that constitute the converter shown in Figure 2.1 are identical and their AC side currents are balanced 3-phase waveforms. Furthermore, it is implicitly assumed that in each cycle of the modulating waveform the switching functions of each leg is perfectly symmetrical. In practice however, tolerances of the circuit components in conjunction with implementation limitations, like sensor offsets, digital truncation errors etc., result in deviation from the ideal condition. Therefore, inevitably, the midpoint current of leg and thus the midpoint current of the converter will include DC components. As small as the DC component may relatively be, once integrated by the capacitors they become accumulated leading to the drifting of the two DC voltages [56, 57].

Numerous methods have been proposed in the literature to prevent the voltage drift phenomenon, such as the annihilation of the i_{np} DC component through an external converter [57, 58]. Thus a compensator processes the voltage difference between the two capacitors and indicates how much current the external converter should inject into the midpoint. Apparently the main shortcoming of this approach is the requirement for additional hardware which not only adds to the system cost it increases the system complexity. An alternative approach proposes to modify the switching functions, in order to nullifying the i_{np} DC component [56, 59]. Although this approach requires a more elaborate control strategy, it provides an economically more feasible and a technologically more elegant solution to the problem. Details of this method are presented as follows.

2.3.4 Balancing the Two Capacitor Voltages

To counteract the DC voltage imbalance a small offset is added to the modulating waveforms of (2.3) as follows,

$$\begin{cases} m_a(t) = m_0 + \hat{m}(t) \cos[\epsilon(t)] \\ m_b(t) = m_0 + \hat{m}(t) \cos[\epsilon(t) - \frac{2\pi}{3}] \\ m_c(t) = m_0 + \hat{m}(t) \cos[\epsilon(t) - \frac{4\pi}{3}] \end{cases} \quad (2.8)$$

$m_0(t)$ is the small offset variable that is added to the sinusoidal components of $m_{abc}(t)$, intentionally introducing an asymmetry in the switching functions so that the time lapse of the period during which the switch pair (Q_{1-1}, Q_{4-1}) is pulse-width modulated will be different from that of the (Q_{1-2}, Q_{4-2}) pair. Consequently, the time interval during which i_{np} is positive will be different from the one during which i_{np} is negative, in other words i_{np} will constitute a DC component. The i_{np} DC component in this case is defined as[20]:

$$i_{np0} = -\frac{6\hat{i} \cos \gamma}{\pi} m_0(t) \quad (2.9)$$

Hence the converter midpoint current is a combination of the third-harmonic component (i_{np3}) formulated by Equation (2.6) and the DC component (i_{np0}) presented in (2.9), Figure 2.4.

Figure 2.4: Partial DC side voltages circuit model representation

So based on Equation (2.9) it can be seen that the i_{np} , DC component is a linear function of a variable gain $(-\frac{6\hat{i}\cos\gamma}{\pi})$ and the switching offset (m_0). Under various power factor and output current ratings the gain can change from zero to high values. Therefore to balance the two DC voltages, a closed-loop system is required to control m_0 under various converter output conditions [56, 58]. It can be shown that by assuming two identical capacitors each having a capacitance of $2C$ (an equivalent capacitance of C):

$$\begin{cases} V_{DC} = \langle V_1 \rangle_0 + \langle V_2 \rangle_0 \\ \frac{d}{dt} (\langle V_1 - V_2 \rangle_0) = \left(\frac{1}{2C} \right) i_{np0} \end{cases} \quad (2.10)$$

According to (2.10), $\langle V_1 - V_2 \rangle_0$, i.e. the DC component of $V_1 - V_2$ is proportional to the integral of i_{np0} . Thus, $\langle V_1 - V_2 \rangle_0$ is a constant value (preferably zero) should i_{np0} be zero. Ideally $i_{np} = 0$ corresponds to $m_0 = 0$, yet, in practice due to inherent asymmetries of the system, m_0 must assume a nonzero value to ensure $i_{np} \equiv 0$. Subsequently, m_0 must be inevitably determined in a closed-loop system that attempts to regulate $\langle V_1 - V_2 \rangle_0$ zero. Based on (2.10), $\langle V_1 - V_2 \rangle_0$ is required as the feedback signal in the closed-loop voltage equalizing process. However, in practice, only $V_1 - V_2$ and not its DC component is available for measurement. To relate $\langle V_1 - V_2 \rangle_0$ to $V_1 - V_2$, V_1 and V_2 are expressed as:

$$\begin{cases} V_1 = \langle V_1 \rangle_0 + \langle V_1 \rangle_3 = \langle V_1 \rangle_0 + \hat{V}_{r3} \sin(3\omega t + \zeta) \\ V_2 = \langle V_2 \rangle_0 + \langle V_2 \rangle_3 = \langle V_2 \rangle_0 - \hat{V}_{r3} \sin(3\omega t + \zeta) \end{cases} \quad (2.11)$$

Subtracting both sides of (2.11), the following is obtained:

$$\langle V_1 - V_2 \rangle_0 = (V_1 - V_2) - 2\hat{V}_{r3} \sin(3\omega t + \zeta) \quad (2.12)$$

As Equation (2.12) indicates, $\langle V_1 - V_2 \rangle_0$ is equal to $V_1 - V_2$ plus a high frequency component which means that $\langle V_1 - V_2 \rangle_0$ can be extracted by passing $V_1 - V_2$ through a low pass filtering. Figure 2.5 shows the closed-loop voltage equalizing scheme in block diagrams. A saturation block is used to limit m_0 to a small value such that $|m_0| \ll m$. Dividing the controller output by $(-i \cos \gamma)$ achieves a feed-forward compensation that makes the loop gain independent of the system operating point. Should a constant gain replace the feed-forward signal, its sign must still be included in the control loop as a multiplicative term. Otherwise, the loop gain assumes opposite signs in rectifying and inverting modes of operation, which results in the instability of the system when the converter moves from one mode to the other.

Figure 2.5: Control block diagram of the partial DC-side voltage balancing system

2.4 Voltage Equalizing Controller

The control block diagram of Figure 2.5 indicates that the control plant is predominantly an integrator that has a variable gain of $-\hat{i} \cos \gamma$. Bearing in mind that $\hat{i}$ is the amplitude of the AC current while γ is the converter AC output power angle. Since both $\hat{i}$ and γ are functions of the converter operating point, $(-\hat{i} \cos \gamma)$ can assume different (+ve or -ve) values. Therefore, if a fixed-structure compensator is employed, although being stable at one operating point, and may become unstable at another. To make the loop gain decoupled from the operating point, the compensator output is divided by $-\hat{i} \cos \gamma$. Since $P(t) = \text{Re} \left\{ \frac{3}{2} \vec{v}(t) \cdot \vec{i}^*(t) \right\}$ the converter AC-side terminals instantaneous real power becomes:

$$P_t(t) = \text{Re} \left(\frac{3}{2} \vec{V}_t(t) \cdot \vec{i}^*(t) \right) = \frac{3}{4} V_{DC} \hat{m} (\hat{i} \cos \gamma) \quad (2.13)$$

Dividing both sides of (2.13) by $(3/4)V_{DC}\hat{m}$, the following is deduced:

$$\hat{i} \cos \gamma = \left(\frac{4}{3V_{DC}\hat{m}} \right) P_t(t) \quad (2.14)$$

Neglected the instantaneous power that is exchanged with the interface reactors, $P_t(t) \approx P_s(t)$ and (2.14) can be rewritten as:

$$\hat{i} \cos \gamma \approx \left(\frac{4}{3V_{DC}\hat{m}} \right) P_s(t) \quad (2.15)$$

According to Equation (2.15) in view of the fact that V_{DC} and $\hat{m}$ are positive quantities, $\hat{i} \cos \gamma$ and $P_s(t)$ have identical signs. In addition, m typically assumes value from a rather narrow range that is 0.7 to 1.0, and V_{DC} is often maintained at a relatively constant value. Thus, $(\hat{i} \cos \gamma)$ is approximately proportional to $P_s(t)$. Equation (2.15) also indicates that the loop gain of the voltage equalizing system illustrated in Figure 2.5 is mainly a function of the real-power that is exchange between the converter and the AC system, assuming opposite signs under rectifying ($P_s(t) \leq 0$) and inverting ($P_s(t) \geq 0$) modes of operation. Figure 2.6 shows the complete control block diagram in which the variable gain $(-\hat{i} \cos \gamma)$ is replaced by a block that reads the sign of P_s .

Figure 2.6: 3-level NPC converter modified partial DC-side voltage equalizing system control block diagram

In the control loop of Figure 2.6, P_s can be seen as a disturbance input as it does not impact $\langle V_1 - V_2 \rangle_0$ in the steady state, even if $H_{Equ}(s)$ is a pure gain; this is due to the control plant including an integral term. In the design process the compensator gain is selected based on the converter rated real power. As can be seen from the Figure the desired difference value between the two DC capacitor voltages is normally set to zero. In addition, as explained before, $V_1 - V_2$ includes a dominant 3rd harmonic component that must be diminished by the low pass filter $F(s)$. This is possible if $F(s)$ is assigned a pair of complex-conjugate zeros at $s = \pm j(3\omega_0)$. Depending on the frame work used P_s can be also be calculated as a function of the, $\alpha\beta$ or dq system variables, i.e.:

$$P_s(t) = \begin{cases} \frac{3}{2} [V_{s\alpha}(t)i_{\alpha}(t) + V_{s\beta}(t)i_{\beta}(t)], & \alpha\beta \text{ Frame} \\ \frac{3}{2} [V_{sd}(t)i_d(t) + V_{sq}(t)i_q(t)], & dq \text{ Frame} \end{cases} \quad (2.16)$$

2.5 Single model for the 3/ and 2/ level converters

A comparison between the conventional 2-level and the 3-level NPC converters shows that the expression for the DC current of the 2-level converter is identical to the expression for the DC component of the DC current of a 3-level NPC converter. The reason for this is that in either converter topologies only the constant component of the converter DC current contributes to the power exchange, bearing in mind that if the capacitor voltages are equal and stable, the DC component of i_{np} in the 3-level NPC is zero. Furthermore, the AC terminal voltage of the 3-level NPC assumes the same form as those of the 2-level converter. “Hence, as long as the terminal voltage/current relationships are concerned, there exists an equivalent 2-level converter for a 3-level NPC [56]. Figures 2.7 and 2.8 illustrate this equivalence property. Figure 2.7 shows a 3-level NPC for which the equivalent 2-level converter is illustrated in Figure 2.8. Even though the internal circuit structures and switching strategies of the two converters are

different, they exhibit identical dynamic behavior if m_{abc} and V_{DC} are selected to be the same. Hence a voltage sourced converter system can employ either the 2-level converter or the 3-level NPC as its AC/DC power processor. However based on the properties mentioned at the beginning of the chapter, only the converter in the latter case, that is the 3-level NPC converter is employed.

Based on the equivalence property, the converter topology (conventional 2-level or 3-level NPC) does not influence the way converter currents and voltages control loop are designed; and since the dynamics of the partial DC-side voltages are decoupled from those of terminal currents/voltages the DC voltage balancing system can be designed independently of the other controllers.

Figure 2.7: Block diagram of an ideal a 3-level NPC Voltage Sourced Converter

Figure 2.8: Block diagram of an ideal 2-level Voltage Sourced Converter

2.6 Concluding Remarks

Neutral point clamped converters are a class of multilevel converters that offer an alternative to reduce or avoid the number of series-connected switches. In the 3-level NPC, each switch cell withstands only half of the DC voltage. Moreover, the 3-level NPC can provide a three-phase AC voltage with a lower THD level. However NPC converters require a capacitive DC voltage divider that needs constant voltage balancing as the partial capacitor DC voltage drift once the converter starts exchanging power. Balanced DC voltages are essential in correct operation of the NPC converter. Consequently a voltage equalizing system was designed based on regular adjustment of the modulation indices resulting in relatively balanced partial DC voltages.

Chapter 3

Grid – Tie Converter Operation and Control

3.1 Introduction

In order to connect and transfer the captured tidal power to the grid, a power port system is needed that monitors the grid status for smooth connection and manages the power transfer without impacting the network. One must bear in mind that the power levels of even one individual tidal turbine can be enough to disrupt the voltage stability and frequency of an electrically remote but energy rich area, that is typical of isolated coasts and many oceanfronts. Consequently this chapter focuses on a class of converter systems referred to as grid imposed converters. On the basis of the unified model presented previously, this chapter presents models and controls for a member of the family of the grid imposed frequency converter systems, namely, the real/reactive-power controller referred to as the power port in dq -frame. A dq -frame vector control is a well-known control strategy that not only reduces the number of required plant controllers in three phase systems from three to two; compared to other vector control strategies like the $\alpha\beta$ -frame, provides a more efficient and independent control of the instantaneous real and reactive powers exchanged between the converter and the AC grid system. Furthermore unlike the $\alpha\beta$ -frame where the control variables and signals are sinusoidal functions of time, dq -vector control of a converter results in control variables that are DC quantities in their steady states. This chapter provides insight into how a decoupled active/reactive power control strategy can be achieved and employed in a power port system that transfers variable amounts of tidal power to a utility grid. In addition, it will be shown how the independent reactive power controller can be utilized in different power factor regulation settings, namely the unified power factor and the voltage regulating schemes.

3.2 Overall Structure

Figure 3.1 shows the schematic diagram of a grid imposed frequency, voltage sourced converter. The converter can represent either a 3-level NPC with a DC side voltage equalizing scheme or a conventional 2-level converter. In either case, the converter is modeled by a lossless power processor including an equivalent DC-bus capacitor, series on-state resistances at the AC side representing conduction losses and a current source that represents the converter switching power losses, as Figure 3.1 shows. The DC side of the converter is interfaced with a DC power source. Each phase of the converter is interfaced with the AC

3.3 Grid Imposed Frequency Converter Control Systems

In case the grid imposed frequency converter of Figure 3.1 is employed as a real/reactive power controller, the objective would be to control the instantaneous real and reactive powers that the converter system exchanges with the AC system, that is, $P_s(t)$ and $Q_s(t)$.

3.3.1 Comparison of the Current and Voltage Mode Controls

Two main methods exist for controlling P_s and Q_s in the converter system of Figure 3.1. The first method that is known as the voltage mode control (Figure 3.2) which although has occasionally been utilized in the industry [63] has been dominantly employed in high voltage/power applications such as in FACTS controllers [64, 65]. Figure 3.2 illustrates that in a voltage controlled converter, the real and reactive power are controlled by the phase angle and the amplitude of the converter AC-side terminal voltage (relative to the PCC voltage) respectively [20]. If the amplitude and phase angle of V_{tabc} are close to those of V_{sabc} , the real and reactive power are almost decoupled and thus two separate controllers can be employed for each one (Fig. 3.2). The voltage mode control is quite simple and has a small number of control loops. However, the main shortcoming of the method is that there is no control loop closed on the converter line current. Consequently, the converter is not protected against overcurrents, and the current may undergo large excursions if faults take place in the AC system or the power commands are rapidly changed. The second approach to the control of the real and reactive power in a voltage sourced converter (Figure 3.1) is referred to as the current mode control. In this approach, the converter line current is closely regulated by a dedicated current control scheme, through the converter AC-side terminal voltage. Here the powers are controlled by the amplitude and the phase angle of the converter line current with respect to the PCC voltage. Consequently the converter is protected against overcurrent conditions. Other advantages of the current mode control include its robustness against variations in the converter or AC systems parameters, higher control precision and superior dynamic performance [66]. Figure 3.3 shows a schematic diagram of the current controlled real/reactive-power controller, where the control is performed in dq -frame. In this mode of control, P_s and Q_s are controlled by the line current components i_d and i_q respectively. Here the feedback and feed-forward signals are initially transformed to the dq -frame and then processed by compensators to produce the control signals in dq -frame. Finally, the control signals are transformed back to the abc -frame and fed to the converter system (Fig. 3.3). As can be seen from the Figure to protect the converter, the reference commands i_{dref} and i_{qref} are limited by corresponding adjustable saturation blocks.

Figure 3.2: Schematic diagram of a voltage controlled, real/reactive power controller

Figure 3.3: Schematic diagram of the current controlled real/reactive power controller in the dq -frame

3.3.2 System Equations and Dynamic Models

Assuming the AC system voltage in the converter system of Figure 3.3 is expressed as:

$$\vec{V}_s(t) = \hat{V}_s e^{j(\omega_0 t + \theta_0)} \equiv \begin{cases} V_{sa}(t) = \hat{V}_s \cos(\omega_0 t + \theta_0) \\ V_{sb}(t) = \hat{V}_s \cos\left(\omega_0 t + \theta_0 - \frac{2\pi}{3}\right) \\ V_{sc}(t) = \hat{V}_s \cos\left(\omega_0 t + \theta_0 - \frac{4\pi}{3}\right) \end{cases} \quad (3.1)$$

in which, $\vec{V}_s(t)$ is the space-phasor equivalent of V_{sabc} , ω_0 is the AC system (source) frequency, $\hat{V}_s$ is the peak value of the line-to-neutral voltage and θ_0 is the source initial phase angle. At the same time dynamics of the converter AC side (Figure 3.3) can be described by the following space-phasor equation:

$$L \frac{d\vec{i}}{dt} = -(R + r_{on})\vec{i} + \vec{V}_t - \vec{V}_s \quad (3.2)$$

By expressing (3.2) in a dq -frame, substituting for $\vec{i} = i_{dq} e^{j\rho}$, $\vec{V}_t = V_{tdq} e^{j\rho}$ and decomposing the result into its real and imaginary components, one deduces:

$$\begin{cases} L \frac{di_d}{dt} = L \left(\frac{d\rho}{dt} \right) i_q - (R + r_{on})i_d + V_{td} - \hat{V}_s \cos(\omega_0 t + \theta_0 - \rho) \\ L \frac{di_q}{dt} = -L \left(\frac{d\rho}{dt} \right) i_d - (R + r_{on})i_q + V_{tq} - \hat{V}_s \sin(\omega_0 t + \theta_0 - \rho) \end{cases} \quad (3.3)$$

The two equations in (3.3) are not in the standard state space format and so the new control variable ω where $\omega = d\rho/dt$ is introduced. Consequently by assuming $\omega(t) \equiv 0$ and also a zero initial condition for ρ , the two equations in (3.3) take the following form:

$$\begin{cases} L \frac{di_d}{dt} = -(R + r_{on})i_d + V_{td} - \hat{V}_s \cos(\omega_0 t + \theta_0) \\ L \frac{di_q}{dt} = -(R + r_{on})i_q + V_{tq} - \hat{V}_s \sin(\omega_0 t + \theta_0) \end{cases} \quad (3.4)$$

At the same time by considering the space phasor $\vec{f}(t) = f_\alpha + jf_\beta$, the dq to $\alpha\beta$ -frame transformation can be defined as:

$$\begin{cases} f_d + jf_q = \hat{f}e^{j(\omega t + \theta_0)} \cdot e^{j\theta_0} \\ \hat{f}e^{j(\omega t + \theta_0)} \equiv \overline{f(t)} \end{cases} \quad (3.5)$$

Substituting for $\overline{V_s}(t) = \widehat{V_s}e^{j(\omega_0 t + \theta_0)}$ in (3.5), results in:

$$\begin{cases} V_{sd} = \widehat{V_s} \cos(\omega_0 t + \theta_0 - \rho) \\ V_{sq} = \widehat{V_s} \sin(\omega_0 t + \theta_0 - \rho) \end{cases} \quad (3.6)$$

Thus (3.3) can be rewritten as:

$$\begin{cases} L \frac{di_d}{dt} = L\omega(t)i_q - (R + r_{on})i_d + V_{td} - V_{sd} \\ L \frac{di_q}{dt} = -L\omega(t)i_d - (R + r_{on})i_q + V_{tq} - V_{sq} \\ \frac{d\rho}{dt} = \omega(t) \end{cases} \quad (3.7)$$

It can also be shown that if $\omega = \omega_0$ and $\rho(t) = \omega_0 t + \theta_0$, the following second order linear system which is excited by a constant input $\widehat{V_s}$ explains the converter, AC side dynamic equations:

$$\begin{cases} L \frac{di_d}{dt} = L\omega_0 i_q - (R + r_{on})i_d + V_{td} - \widehat{V_s} \\ L \frac{di_q}{dt} = -L\omega_0 i_d - (R + r_{on})i_q + V_{tq} \end{cases} \quad (3.8)$$

One must note that if V_{td} and V_{tq} are DC variables, i_d and i_q will also be DC variables in the steady state.

The mechanism to ensure $\rho(t) = \omega_0 t + \theta_0$ is referred to as the Phase Locked Loop (PLL) system. The structure, model, and stabilization of the PLL designed for this study can be found in the Appendix II.

With reference to the real/reactive power controller depicted in Figure 3.3, the real and reactive powers delivered to the AC system at the PCC can be expressed as [67]:

$$\begin{cases} P_s(t) = \frac{3}{2} [V_{sd}(t)i_d(t) + V_{sq}(t)i_q(t)] \\ Q_s(t) = \frac{3}{2} [-V_{sd}(t)i_q(t) + V_{sq}(t)i_d(t)] \end{cases} \quad (3.9)$$

where V_{sd} and V_{sq} are the AC system dq -frame voltage components which cannot be controlled by the converter. If the PLL is properly designed in a steady state, $V_{sq} = 0$ and the above power equations simplify to:

$$\begin{cases} P_s(t) = \frac{3}{2} [V_{sd}(t) i_d(t)] \\ Q_s(t) = -\frac{3}{2} [V_{sd}(t) i_q(t)] \end{cases} \quad (3.10)$$

Hence $P_s(t)$ and $Q_s(t)$ can be controlled by i_d and i_q , respectively. In this case, let:

$$\begin{cases} i_{dref}(t) = \frac{2}{3V_{sd}} P_{sref}(t) \\ i_{qref}(t) = -\frac{2}{3V_{sd}} Q_{sref}(t) \end{cases} \quad (3.11)$$

If the control system provides fast tracking of the references, i.e. $i_d \approx i_{dref}$ and $i_q \approx i_{qref}$, it results in $P_s \approx P_{sref}$ and $Q_s \approx Q_{sref}$, that is, $P_s(t)$ and $Q_s(t)$, can be individually controlled by their respective reference commands. Since V_{sd} in the steady state is a DC variable, i_{dref} and i_{qref} are also DC variables if P_{sref} and Q_{sref} are moderately constant signals. Thus, as mentioned before, unlike the control system in the $\alpha\beta$ -frame that deals with sinusoidal signals, the dq -frame control system deals with DC variables.

3.3.3 Current Mode Control System

The dq -vector control of the real/reactive power controller of Figure 3.3 is based on (3.7). Thus by assuming a steady-state operating condition and substituting for $\omega(t) = \omega_0$, we deduce:

$$\begin{cases} L \frac{di_d}{dt} = L\omega_0 i_q - (R + r_{on})i_d + V_{td} - V_{sd} \\ L \frac{di_q}{dt} = -L\omega_0 i_d - (R + r_{on})i_q + V_{tq} - V_{sq} \end{cases} \quad (3.12)$$

in which, V_{td} and V_{tq} are:

$$\begin{cases} V_{td}(t) = \frac{V_{DC}}{2} m_d(t) \\ V_{tq}(t) = \frac{V_{DC}}{2} m_q(t) \end{cases} \quad (3.13)$$

The two equations in (3.12) represent the converter model in dq -frame. This model is applicable to both the 2-level and the three-level NPC converters. In (3.12), V_{td} and V_{tq} are control inputs, i_d and i_q are state

variables and V_{sd} and V_{sq} are disturbance inputs. Due to the presence of $L\omega_0$ terms in (3.12), dynamics of i_d and i_q are coupled. To decouple the dynamics, m_d and m_q can be determined as:

$$\begin{cases} m_d = \frac{2}{V_{DC}}(u_d - L\omega_0 i_q + V_{sd}) \\ m_q = \frac{2}{V_{DC}}(u_q - L\omega_0 i_d + V_{sq}) \end{cases} \quad (3.14)$$

where u_d and u_q are two new control inputs [68]. Substituting for m_d and m_q in (3.13) from (3.14), and substituting for V_{td} and V_{tq} from the resultant in (3.12), we deduce:

$$\begin{cases} L \frac{di_d}{dt} = -(R + r_{on})i_d + u_d \\ L \frac{di_q}{dt} = -(R + r_{on})i_q + u_q \end{cases} \quad (3.15)$$

The two equations in (3.15) describe two decoupled, linear and first-order systems. Based on (3.15), i_d and i_q can be controlled by u_d and u_q , respectively. Figure 3.4 shows a block representation of the d and q -axis current controllers of the converter system in which u_d and u_q are the outputs of two corresponding compensators. So while the d -axis compensator processes the error signal $e_d = i_{dref} - i_d$ to provide u_d , which based on (3.14), contributes to m_d , the q -axis compensator in a similar loop processes $e_q = i_{qref} - i_q$ and provides u_q that contributes to m_q . The converter then amplifies m_d and m_q by a factor of $V_{DC}/2$ and generates V_{td} and V_{tq} that, in turn, control i_d and i_q based on (3.12).

According to the above-mentioned control process, the simplified control block diagrams of Figure 3.5, are sketched which is equivalent to the control system of Figure 3.4. It should be noted that in the steady state all the control, feed-forward, and feedback signals in the control system of Figure 3.5 are DC quantities. The Figure also indicates that the control plants in both d - and q -axis current-control loops are identical leading to the corresponding compensators being alike.

Figure 3.4: Control block diagram of a current controlled voltage sourced converter system

Figure 3.5: Simplified block diagram of the current controlled converter of Figure3.4

Considering the two control loops, $k_d(s)$ and $k_q(s)$ can be selected to be a basic proportional integral (PI) compensator, enabling the tracking of a DC reference command. Hence:

$$\begin{cases} k_d(s) = k_q(s) = \frac{k_p s + k_i}{s} \\ k_p = \frac{L}{\tau_i} \\ k_i = \frac{R + r_{on}}{\tau_i} \end{cases} \quad (3.16)$$

where k_p and k_i (proportional and integral gains), are selected according to the loop gain and the plant pole. Once k_p and k_i are selected (based on 3.16), the response of $i_d(t)$ to $i_{dref}(t)$ would be based on a first order transfer function with a time constant τ_i that is a design choice. τ_i should be selected small enough to result in a fast current control response but adequately large enough so that $1/\tau_i$, i.e. the bandwidth of the closed-loop control system, is at least 10 times smaller than the switching frequency of the converter (in rad/s). Depending on the requirements of the application and the converter switching frequency, τ_i is typically selected in the range of a few milliseconds [20]. One should also note that R and L in Equation (3.16) represent the total equivalent resistance and inductance that is seen between the converter and the point of common coupling including any transformer or submarine cable impedances.

3.3.4 Selecting the DC Bus Voltage Level

According to [20], the DC-bus voltage of the real/reactive-power controller must satisfy the following criteria:

$$\begin{cases} V_{DC} \geq 2(\hat{V}_t) & PWM \\ V_{DC} \geq 1.74(\hat{V}_t) & PWM \text{ with 3rd - harmonic injection} \end{cases} \quad (3.17)$$

Consequently $\hat{V}_t$ must properly be evaluated under the worst case operating condition. Since the converter system controls P_s and Q_s , $\hat{V}_t$ should also be expressed in terms of P_s and Q_s . Under the assumptions that V_{sd} is constant $V_{sq} = 0$ and $(R + r_{on}) \approx 0$, it can be shown that:

$$\begin{cases} V_{td} = \left(\frac{2L}{3V_{sd}}\right) \frac{dP_s}{dt} + \left(\frac{2L\omega_0}{2V_{sd}}\right) Q_s + V_{sd} \\ V_{td} = -\left(\frac{2L}{3V_{sd}}\right) \frac{dQ_s}{dt} + \left(\frac{2L\omega_0}{2V_{sd}}\right) P_s \end{cases} \quad (3.18)$$

The amplitude of the AC side terminal voltage can accordingly be calculated as:

$$\hat{V}_t = \sqrt{V_{td}^2 + V_{tq}^2} = \hat{m} \frac{V_{DC}}{2} \quad (3.19)$$

If the conventional PWM technique is employed, $\hat{m}$ can assume values up to unity, whereas with the PWM with third-harmonic injection, $\hat{m}$ can become as large as 1.15. To calculate the $\hat{V}_t$ maximum a worst case scenario is considered where both the active and reactive powers are subjected to simultaneous step changes. In such a case the terminal voltage change can be calculated according to the following equation:

$$\begin{cases} V_{td}(t_0^+) = V_{sd} + \left(\frac{2L\omega_0}{3V_{sd}}\right) Q_{s0} + \left(\frac{2L}{3\tau_i V_{sd}}\right) \Delta P_s \\ V_{tq}(t_0^+) = \left(\frac{2L\omega_0}{3V_{sd}}\right) P_{s0} - \left(\frac{2L}{3\tau_i V_{sd}}\right) \Delta Q_s \end{cases} \quad (3.20)$$

Based on the steady-state power flow and values of ΔP_s and ΔQ_s , $V_{td}(t_0^+)$ and $V_{tq}(t_0^+)$ can be estimated and AC-side terminal voltage maximum value, $\hat{V}_t(t_0^+)$ is then calculated from (3.19). Finally, based on the same equation and according to the PWM strategy adopted, the minimum required DC bus voltage is calculated.

3.4 Active and Reactive Power Controllers

Figure 3.6: Schematic diagram of the controlled active/reactive power port

The objective of the real (P) and reactive (Q) power controller is to control the power flow that the grid-tie converter exchanges with the AC system. As can be seen from Figure 3.6, for simplicity, the turbine-generator set along with its AC/DC variable frequency converter is considered as a black box that exchanges a time-varying power $P_{Gen}(t)$ with the DC side of the system. A capacitor link is used to maintain the DC voltage, implying that the DC bus voltage is not imposed and so needs to be regulated. Also note that unlike before the utility network is not ideal and so a step-up transformer is employed to connect the converter system to a non-stiff grid which is presented by its equivalent Thevenin model. In view of that the converter control system is designed to be capable of stable bidirectional power exchange between the DC bus and the AC network.

3.4.1 Active Power Control and DC Bus Voltage Regulation

Previously the model and control system of the real/reactive power controller were presented. Practically in the real/reactive power controller, the converter DC-bus voltage is not impressed by an ideal voltage source and so needs to be continuously regulated. As mentioned before the power source which in this case is a variable speed tidal turbine-generator set, is assumed to exchange a time varying power ($P_{Gen}(t)$) with the DC bus. Therefore the converter system of Figure 3.6 should enable a bidirectional power exchange between the power source seen here as black box and the AC system. The main control objective for the active power controller is to regulate the DC-bus voltage V_{DC} while the power is being transferred to the utility. As Figure 3.6 illustrates, the kernel of the power port is the real/reactive power controller of Figure 3.3 in which P_s and Q_s can be independently controlled. Consequently two separate controllers are considered for the active and reactive power regulation. To regulate the DC-bus voltage, a feedback mechanism compares V_{DC} with its reference command and adjusts the P_s value accordingly, such that the net power exchanged with the DC-bus capacitor is kept zero. The reactive power controller on the other hand regulates Q_s according to the required reactive power philosophy. In many applications, Q_s is regulated at zero, leading to the converter operating at unity power factor; however as will be seen in later sections, Q_s can also be controlled in a closed-loop mechanism to regulate the PCC voltage.

3.4.1.1 Model of the Active Power Controller

Figure 3.7 shows inside details of the active power controller depicted in Figure 3.6.

Figure 3.7: Control block diagram of the active power controller

The main control requirement of this active power controller is to regulate the DC-bus voltage (V_{DC}). As can be seen from the Figure, V_{DC}^2 is selected as the input for DC voltage regulation rather than V_{DC} . The reason for using the squared value lies in the power balance equation of the system. Here, P_t represents the terminal power (Fig. 3.6) and P_{loss} represents the power loss of the converter.

$$P_{Gen} - P_{loss} - \frac{d}{dt} \left(\frac{1}{2} C V_{DC}^2 \right) = P_t \Rightarrow \left(\frac{C}{2} \right) \frac{dV_{DC}^2}{dt} = P_{Gen} - P_{loss} - P_t \quad (3.21)$$

According to (3.21) the system can be seen as a dynamic model in which V_{DC}^2 is the state variable, P_t is the control input while P_{Gen} and P_{loss} are the disturbances. By expressing the control input (P_t) in terms of P_s and Q_s and some mathematical manipulations, it can be shown that the dynamics of V_{DC}^2 is described by the following equation:

$$\frac{dV_{DC}^2}{dt} = \frac{2}{C} P_{Gen} - \frac{2}{C} P_{loss} - \frac{2}{C} \left[P_s + \left(\frac{2LP_s}{3V_{sd}^2} \right) \frac{dP_s}{dt} \right] + \frac{2}{C} \left[\left(\frac{2LQ_s}{3V_{sd}^2} \right) \frac{dQ_s}{dt} \right] \quad (3.22)$$

Based on (3.22), V_{DC}^2 is the output, P_s is the control input, and P_{Gen} , P_{loss} , and Q_s are the disturbance inputs. As shown in Figure 3.7, V_{DC}^2 is compared with V_{DCref}^2 , and the error signal is processed by the compensator $H_P(s)$, leading to the command P_{sref} being issued for the real-power controller. The real-power controller, in turn, regulates P_s at P_{sref} . To design the active power controller we need the system open loop transfer function which entails the transfer functions of the compensator ($H_P(s)$) itself as well as the power controller and the DC bus voltage dynamics. Assuming the transfer function $G_p(s) =$

$P_s(s)/P_{sref}(s)$ that describes the power controller dynamics, it is noticed that by presuming a constant V_{sd} and multiplying both sides of Equation (3.23) by $\left(\frac{3}{2}\right)V_{sd}$:

$$I_d(s) = G_i(s)I_{dref}(s) \quad (3.23)$$

$G_p(s)$ can be defined as (3.24).

$$\begin{cases} P_s(s) = G_i(s)P_{sref}(s) \\ \therefore G_p(s) = G_i(s) = \frac{1}{\tau_i s + 1} \end{cases} \quad (3.24)$$

On the other hand although Equation (3.22) describes the DC bus dynamics, in which presence of the $P_s \frac{dP_s}{dt}$ and $Q_s \frac{dQ_s}{dt}$ terms make it a nonlinear system. The following equation describes the linearized plant with $\tilde{V}_s$ substituted by V_{sd} .

$$\frac{d\tilde{V}_{DC}^2}{dt} = \frac{2}{C}\tilde{P}_{ext} - \frac{2}{C}\left[\tilde{P}_s + \left(\frac{2LP_{s0}}{3V_{sd}^2}\right)\frac{d\tilde{P}_s}{dt}\right] + \frac{2}{C}\left[\left(\frac{2LQ_{s0}}{3V_{sd}^2}\right)\frac{d\tilde{Q}_s}{dt}\right] \quad (3.25)$$

The superscripts “~” and “0” in Equation (3.25) represent small-signal perturbations and steady-state values of the variables respectively. By applying Laplace transform to (3.25), the DC bus dynamic transfer function can be deduced as:

$$\begin{cases} G_v(s) = \frac{\tilde{V}_{DC}^2}{\tilde{P}_s} = -\left(\frac{2}{C}\right)\frac{\tau s + 1}{s} \\ \tau = \frac{2LP_{s0}}{3V_{sd}^2} = \frac{2LP_{Gen0}}{3V_{sd}^2} \end{cases} \quad (3.26)$$

What Equation (3.26) implies is that τ is proportional to the steady-state real-power flow P_{Gen0} . Thus, if P_{Gen0} is small, τ is insignificant and the plant predominantly becomes an integrator. As P_{Gen} increases, τ becomes larger and results in the phase of $G_v(s)$ shifting. While in the inverting mode of operation P_{Gen0} and τ are both positive which adds to the phase of $G_v(s)$. However, in the rectifying mode of operation, where P_{Gen0} is negative, τ is negative and results in the phase of $G_v(s)$ reducing; the phase drops further as the absolute value of P_{Gen0} becomes larger. According to (3.26), the plant zero is given by $z = -1/\tau$. Consequently a negative τ corresponds to a zero on the right-half plane. As a result, the controlled DC-voltage power port can be seen as a non-minimum-phase system in the rectifying mode of operation [69]. The non-minimum-phase property has a negative impact on the system stability which needs accounting for in the control design process.

3.4.1.2 Designing the Active Power Controller $H_p(s)$

Figure 3.8 shows the block diagram of the DC-bus voltage controller for the hybrid power port of Figure 3.6. As the Figure illustrates the closed loop system is composed of the compensator $H_p(s)$, the real power controller $G_p(s)$ as well as the control plant $G_v(s)$, described by (3.26). The reason for multiplying $H_p(s)$ by -1 is to compensate for the negative sign of $G_v(s)$. The $H_p(s)$, should include an integral term followed by a lead transfer function to compensates for the plant phase lag, ensuring an adequate phase margin at the gain crossover frequency. According to (3.26), $G_v(s)$ will have the largest phase lag when P_{Gen} is at its rated negative value. Consequently adequate phase margin should be guaranteed at this operating point, so that the closed-loop system remains stable for other operating points.

Figure 3.8: DC bus voltage controller block diagram based on the linearized model

When designing the $H_p(s)$, the gain crossover frequency (ω_c), should be selected to be adequately smaller than $G_p(s)$ bandwidth, such that $G_p(j\omega_c) \approx 1 + j0$. $H_p(s)$ is then designed under the worst-case operating condition with an adequately large phase margin. The root-locus design method will be used to design the controller. The advantage of this method is that performance indices, such as the maximum overshoot or the settling time, are related to the pole/zero loci in a straightforward manner and can be readily taken into account during the design process.

3.4.2 Reactive Power Regulation and PCC Bus Voltage Control

Decoupled P and Q controllers allow for the ability of different reactive power control strategies to be implemented. Reactive power can be controlled to be variable, constant or even zero, leading to the system operating in a variable, constant or unified power factor mode. Although most renewable generating devices are made to operate under a constant or unit power factor scheme, assigning a small portion of the grid tie converter, power capacity for reactive power control results in the regulation of the PCC bus voltage in spite of any utility grid weaknesses. Hence the presented reactive power control methodology is most useful in cases where the system is interfaced with a weak AC network.

Figure 3.9: Injected powers to the PCC and Thevenin equivalent model of the grid system

Figure 3.9 shows the Thevenin equivalent circuit of an electric grid and the interfacing PCC bus which separates the Tidal In-Stream Energy Converter (TISEC) system from the network. The Thevenin model consists of its Thevenin impedance, i.e. $Z_{Th} = R_g + jX_g$, and an ideal voltage source which is shown here as an infinite bus of constant voltage ($E \angle 0^\circ$). The stability of a network can be defined by its Short Circuit Level ($SCL = \frac{V_{Grid}^2}{|Z_{Th}|}$); so the higher the impedance or alternatively the lower the short circuit level, the more vulnerable is the network to voltage variability. In this study, voltage variability refers to the deviation of the PCC bus voltage from that of the utility grid ($E \angle 0$). In steady state the PCC and grid voltage difference can be determined by solving the following two equations:

$$\begin{cases} V_{PCC}^2 - V_{Grid}V_{PCC}\cos\theta - (R_gP_s + X_gQ_s) = 0 \\ V_{Grid} \cdot V_{PCC} \cdot \sin\theta + R_gQ_s - X_gP_s = 0 \end{cases} \quad (3.27)$$

with θ being the PCC bus voltage angle with respect to the utility network. In cases where θ is small, the voltage deviation simplifies to the following approximate equation:

$$\Delta V = V_{PCC} - V_{Grid} = \frac{P_s \cdot R_g + Q_s \cdot X_g}{V_{PCC}} \quad (3.28)$$

where, ΔV is the voltage deviation at the PCC; P_s and Q_s are the injected active and reactive powers. Consequently it can be seen that ΔV depends on the injected active and reactive powers as well as the grid Thevenin equivalent impedance (overlooking any line or transformer impedances). It should be emphasized that eq. (3.28) only yields the steady-state value of the voltage variation, assuming over the interval of interest, no network control action has been initiated within the grid to vary Z_{Th} or V_{Grid} .

Since tidal currents change speed over time, the active power that a TISEC injects to the PCC is variable. The variable P injection results in the PCC voltage changing over time, a common power quality problem that is worsened by the level of active power transfer. However as Equation (3.28) shows the voltage deviation is also a function of Q , and since in most power networks $X_g \gg R_g$, the influence of the reactive power on PCC bus voltage is much higher than that of the active power. As a result by managing the amount of reactive power that is injected to the PCC bus, the voltage deviation caused by active power transfer can be counteracted. As mentioned before, although separate series/parallel reactive power compensations have been proposed in the past for other renewables that employ an auxiliary STATCOM or DVR module [60-62], marine current repetitive and predictable profiles provide opportunities for novel PCC bus voltage regulation schemes that counteract fluctuations, through assigning part of the grid-tie converter capacity to reactive power adjustments. In other words the long term forecasting techniques developed in chapter one can be used to meticulously assign a fixed/variable portion of the power port capacity to reactive power control. Model development as well as technical insight and analysis of this strategy is another contribution of this study.

If there is no limit on the amount of reactive power injection, the desirable result for PCC voltage regulation would be $\Delta V = 0$; however, converters can only operate, with Q limited within a specific range that is determined based on the vector difference of their nominal rating and the TISEC generated active power (assuming negligible losses). Letting S stand for the rated capacity of the converter and P for its generated active power, then the reactive power capability of the grid side converter is bounded by:

$$-\sqrt{S^2 - P^2} \leq Q \leq +\sqrt{S^2 - P^2} \quad (3.29)$$

Hence the percentage of the converter power capacity assigned to reactive power management should be selected according to the system active power levels. In practice depending on the size of the power port compared to the generator(s) this percentage can be made variable or fixed. In this study, a constant portion that is equivalent to 25% of the converter apparent power rating is assigned for reactive power control, leaving a good 97% of the converter capacity for the active power transfer. As mentioned before since the X/R ratio of most networks is typically 5 - 15, the effect of Q on ΔV is much higher than that of P , and thus even a relatively small portion of the converter power rating can suffice the Q requirement for ΔV regulation.

3.4.2.1 PCC Bus Voltage Dynamic Models

As is depicted in Figure 3.6, the TISEC system is connected to the power network through a submarine line and a step up transformer. The network itself is modeled by its Thevenin equivalent. The schematic diagram of Figure 3.10 shows a simplified picture of the same systems depicted in Figure 3.6. As can be seen in this Figure the impedance of the transformer is summed up with that of the line and the PCC is assumed to supply a cluster of loads. Depending on the reactive power management strategy the reactive-power reference Q_{sref} , can be set to:

- Zero, under unit power factor scheme
- A constant factor of P_{sref} for constant leading or lagging power factors
- Be controlled through a closed-loop mechanism enabling PCC bus voltage regulation.

Figure 3.10: A schematic diagram of the power port system and load

With reference to Figure 3.10, the electrical nodes where the three phases of the converter are connected to the AC system through the series impedances create the Point of Common Coupling (PCC). Here the PCC voltage is labeled as V_{sabc} . According to Figure 3.10 the converter synchronization signals, i.e. the PLL inputs are obtained from the PCC. Although in the final design no direct load will be imposed on the TISEC, here a generic modeling will be performed, thus the system not only interchanges power with the utility grid it has a direct load coupling on its PCC bus.

To stop the large V_{sabc} voltage notches caused by the modulated waveforms of the converter AC side terminal voltages (V_{tabc}) and the relatively large L_g practically a three-phase series RLC filter needs to be connected in parallel with the power port at the PCC bus. If this problem is not addressed properly load voltage as well as the PLL feedback signals (V_{sd} and V_{sq}) can become highly distorted. Furthermore each RLC branch in the filter needs to be tuned to the dominant pulse-width modulation (PWM) side-band harmonic, but exhibit a large impedance at the grid frequency. This way, the converter current harmonics flow through the RLC filter and so will not penetrate the grid. One must note that since the grid is assumed to be non-stiff, depending on load conditions the magnitude and the phase angle of V_{sabc} can be different from that of V_{gabc} . Consequently any load switching incidents can result in abrupt and large deviations of V_{sabc} . In this section, V_{sabc} regulation through closed-loop Q_s adjustments, is discussed under both large and small signal models of the PCC voltage dynamics

a) PCC Voltage Dynamics under The Large Signal Model

The reactive power controller of Figure 3.10 is designed to regulate the PCC bus voltage (V_{sabc}), in the presence of the load current I_{Labc} , and transfer of the active power P_s . The regulation process takes place by controlling i_{abc} , i.e. the three phase currents entering the PCC bus. Assuming a negligible network resistance (i.e. $X_g \gg R_g$), the following equations hold:

$$\begin{cases} V_{sa} = L_g \frac{di_{ga}}{dt} + V_{ga} + V_{null} \\ V_{sb} = L_g \frac{di_{gb}}{dt} + V_{gb} + V_{null} \\ V_{sc} = L_g \frac{di_{gc}}{dt} + V_{gc} + V_{null} \end{cases} \quad (3.30)$$

and

$$\begin{cases} i_{ga} = i_a - i_{La} \\ i_{gb} = i_b - i_{Lb} \\ i_{gc} = i_c - i_{Lc} \end{cases} \quad (3.31)$$

where V_{null} is the AC system neutral point voltage with respect to the converter virtual or actual DC bus middle point. The AC system Thevenin voltage can be written as:

$$\begin{cases} V_{ga} = \hat{V}_g \cos(\omega_0 t + \theta_0) \\ V_{gb} = \hat{V}_g \cos(\omega_0 t + \theta_0 - \frac{2\pi}{3}) \\ V_{gc} = \hat{V}_g \cos(\omega_0 t + \theta_0 - \frac{4\pi}{3}) \end{cases} \quad (3.33)$$

where $\widehat{V}_g$ is the amplitude of the phase voltage, ω_0 is the frequency of the AC system, and θ_0 is the V_{gabc} initial phase angle. Since the grid-tie converter is controlled in dq -frame which is synchronized to the angle ρ , it can be shown that:

$$\begin{cases} V_{sd} = L_g \frac{di_{gd}}{dt} - L_g \omega i_{gq} + \widehat{V}_g \cos(\omega_0 t + \theta_0 - \rho) \\ V_{sq} = L_g \frac{di_{gq}}{dt} - L_g \omega i_{gd} + \widehat{V}_g \sin(\omega_0 t + \theta_0 - \rho) \end{cases} \quad (3.33)$$

where $\omega = d\rho/dt$ and is controlled by the PLL (Appendix), according to:

$$\frac{d\rho}{dt} = \omega(t) = H(p) \cdot V_{sq}(t) \quad (3.34)$$

The $p = d(\cdot)/dt$ in (3.34) represents the differentiation operator and the PLL compensator transfer function is denoted by $H(s)$. Thus, $H(p) \cdot f(t)$, in which $f(t)$ is an arbitrary function of time, represents the $H(s)$ zero-state response to the input $f(t)$. As explained in the Appendix, the PLL compensator includes an integral term which results in $\omega(t)$ assuming a nonzero steady-state value as V_{sq} settles at zero. Equations (3.30)–(3.34) represent a dynamic system for which i_d and i_q are the control inputs, $i_{Ld}i_{Lq}$ are the disturbance inputs and V_{sd} is the output. Due to the presence of the terms $\widehat{V}_g \cos(\omega_0 t + \theta_0 - \rho)$ and $\widehat{V}_g \sin(\omega_0 t + \theta_0 - \rho)$, the system is nonlinear. Furthermore the converter frequency (ω) is an operating point dependent dynamic variable. Substituting for V_{sq} , from (3.33), in (3.34) can help clarifying this point further:

$$\omega = \frac{d\rho}{dt} = L_g H(p) \left(\frac{di_{gq}}{dt} + \omega i_{gd} \right) + \widehat{V}_g \sin(\omega_0 t + \theta_0 - \rho) \quad (3.35)$$

What Equation (3.35) indicates is that dynamic responses of ρ and ω , as well as their natural transient components that correspond to $i_{gd} = i_{gq} = 0$, include forced components which are functions of i_{gd} and i_{gq} . This is in contrast to the case of a stiff grid where ρ and ω responses merely include natural transient components, and so the PLL dynamics are independent of the rest of the system and the operating point. Hence in a stiff grid scenario by the time the PLL reaches the steady state, $\rho = \omega_0 t + \theta_0$ and $\omega = \omega_0$.

b) PCC Bus Voltage Dynamics under The Small Signal Model

PCC voltage small-signal dynamics can be derived by linearizing (3.34)–(3.35) around a steady-state operating point. By introducing the following perturbed variables:

$$\left\{ \begin{array}{l} V_{sd} = V_{sd0} + \widetilde{V}_{sd} \\ V_{sq} = 0 + \widetilde{V}_{sq} \\ i_{gd} = i_{gd0} + \widetilde{i}_{gd} \\ \omega_0 t + \theta_0 - \rho = -(\rho_0 + \tilde{\rho}) \Rightarrow \frac{d\rho}{dt} = \omega_0 + \frac{d\tilde{\rho}}{dt} \end{array} \right. \quad (3.36)$$

with some mathematical manipulation and substituting for $\widehat{V}_g \cos \rho_0$ and $\widehat{V}_g \sin \rho_0$ the following equation set can be obtained:

$$\left\{ \begin{array}{l} \widetilde{V}_{sd} = L_g s \widetilde{I}_{gd}(s) - L_g \omega_0 \widetilde{I}_{gq}(s) - L_g (i_{gq0} s + \omega_0 i_{gd0}) \tilde{\rho}(s) \\ \widetilde{V}_{sq} = L_g s \widetilde{I}_{gq}(s) + L_g \omega_0 \widetilde{I}_{gd}(s) + [L_g i_{gd0} s - (V_{sd0} + L_g \omega_0 i_{gq0})] \tilde{\rho}(s) \\ \tilde{\rho}(s) = \frac{H(s)}{s} \widetilde{V}_{sq}(s) \end{array} \right. \quad (3.37)$$

The Laplace domain equations in (3.37), describes a linear system which is the small signal equivalent of the system described previously by (3.33) and (3.34). Expressing $V_{sd}(s)$ dynamics in terms of $\widetilde{I}_{gd}(s)$ and $\widetilde{I}_{gq}(s)$, first $\widetilde{V}_{sq}$ can be eliminated between (3.37) second and third equations and then from the resultant equation $\tilde{\rho}$ can be substituted in (3.37) first equation. The result is:

$$\widetilde{V}_{sd}(s) = G_d(s) \widetilde{I}_{gd}(s) + G_q(s) \widetilde{I}_{gq}(s) \quad (3.38)$$

where $G_d(s)$ and $G_q(s)$ are two individual linear transfer functions with parameters that are functions of i_{gd0} and i_{gq0} . The small signal current equations for the system of Figure 3.10 can be written as:

$$\left\{ \begin{array}{l} \widetilde{I}_{gd}(s) = \widetilde{I}_d(s) - \widetilde{I}_{Ld}(s) \\ \widetilde{I}_{gq}(s) = \widetilde{I}_q(s) - \widetilde{I}_{Lq}(s) \end{array} \right. \quad (3.39)$$

Substituting for $\widetilde{I}_{gd}$ and $\widetilde{I}_{gq}$, from (3.39) in (3.38), the following is derived:

$$V_{sd}(s) = G_d(s) \cdot \widetilde{I}_d(s) - [G_d(s) \cdot \widetilde{I}_{Ld}(s) + G_q(s) \cdot \widetilde{I}_{Lq}(s)] + G_q(s) \cdot \widetilde{I}_q(s) \quad (3.40)$$

Equation (3.40) describes a dynamic system with $\widetilde{I}_q$ as the control input and $\widetilde{V}_{sd}$ as the output. The inputs I_{Ld}, I_{Lq} and $\widetilde{I}_d$ are, generally functions of $\widetilde{V}_{sd}$ and $\widetilde{V}_{sq}$ and the external generator power and so cannot be called disturbances. Moreover they also include the harmonic filters dynamics. If I_{Labc} could be measured, or the load had an identified model, the impact of, $\widetilde{I}_{Ld}$ and $\widetilde{I}_{Lq}$ on $\widetilde{V}_{sd}$ could be mitigated through appropriate feed-forward compensation techniques. However, in practice these conditions are satisfied occasionally. Consequently the stability of the control system is delegated to a robust design of the compensator(s) and load dynamics are commonly ignored [20].

Figure 3.11: PCC bus voltage regulator, control block diagram

The control block diagram of the power port, PCC bus voltage regulator is depicted in Figure 3.11 which illustrates how the compensator $H_Q(s)$, processes $V_{sd_ref} - V_{sd}$ and provides Q_{s_ref} . Assuming, $V_{sd} \approx V_{sd0}$, Q_{s_ref} is multiplied by $-2/(3V_{sd0})$ to provide I_{q_ref} . From there, I_q can track I_{q_ref} , based on a closed-loop transfer function $G_i(s)$. As mentioned before, dq -frame current controller parameters can be selected such that $G_i(s)$ functions as a first-order transfer function having an arbitrarily small time constant. It is common for nested control structures to have an adequately lower voltage control closed-loop bandwidth than that of $G_i(s)$, so that $G_i(s)$ is approximated by a unity gain in the process.

c) PCC Voltage Dynamics under the Approximate Model

If $\rho = d\rho/dt = 0$ or in other words transient excursions of ρ and ω are ignored in (3.33), then the following simplified dynamic model can be derived:

$$\widetilde{V}_{sd} \cong L_g \frac{d\widetilde{i}_{gd}}{dt} - L_g \omega_0 \widetilde{i}_{gq} \quad (3.41)$$

Since ρ and $\omega = d\rho/dt$ are functions of i_{gd} and i_{gq} , (3.41) provides a sufficiently accurate description of the PCC bus voltage dynamics provided that i_{gd} and i_{gq} do not change rapidly. Formally, this requires the PCC bus voltage control loop to be adequately slower than the d and q -axis closed-loop current controllers. In addition, i_{Ld} and i_{Lq} must have reasonably low rates of variability. Replacing i_{gd} and i_{gq} from (3.39) in (3.41), we obtain:

$$\widetilde{V}_{sd} \cong L_g \left(\frac{d\widetilde{i}_d}{dt} - \frac{d\widetilde{i}_{Ld}}{dt} \right) - L_g \omega_0 (\widetilde{i}_q - \widetilde{i}_{Lq}) \quad (3.42)$$

Expressing equation (3.42) in the Laplace domain yields:

$$\widetilde{V}_{sd} \cong L_g s \widetilde{I}_d(s) - [L_g s \widetilde{I}_{Ld}(s) - L_g \omega_0 \widetilde{I}_{Lq}(s)] - L_g \omega_0 \widetilde{I}_q(s) \quad (3.43)$$

By comparing (3.43) with (3.40) it is concluded that based on the approximate model $G_d(s)$ and $G_q(s)$ are:

$$\begin{cases} G_d(s) \cong L_g s \\ G_q(s) \cong -L_g \omega_0 \end{cases} \quad (3.44)$$

So if the simplified model of (3.43) is considered as $H_Q(s)$ design basis, then the control block diagram of Figure 3.10 can be altered to that depicted in Figure 3.12.

Figure 3.12: PCC voltage regulator control block diagram, based on the approximate model

3.4.2.2 Designing the Reactive Power Controller $H_Q(s)$

The PCC voltage controller, that is, $H_Q(s)$, is designed based on the block diagram of Figure 3.11. As discussed previously, the plant transfer function $G_q(s)$ can be derived by linearizing the system nonlinear equations. However, as long as the PLL dynamics are ignored, the block diagram of Figure 3.12 can be employed instead as it results in a simpler and lower order $G_q(s)$ model. Based on the control loop model of Figure 3.12, in cases where there is no direct load exerted on the PCC bus the load effect is cancelled ($I_L = 0$) and the system would turn into the classic case of general SISO servo problem, Figure 3.13.

Figure 3.13: Control block diagram of the general SISO servo problem

where:

$$\begin{cases} K(s) = \frac{B(s)}{A(s)} = H_Q(s) \\ G(s) = \frac{D(s)}{N(s)} = \frac{2L_q\omega_0}{3V_{sd0}} \cdot G_i(s) = \frac{2L_q\omega_0}{3V_{sd0}} \times \frac{1}{\tau_i s + 1} \\ d = I_{Ld} \cdot L_g s \end{cases} \quad (3.45)$$

In this case the following classical servo control design procedure is employed:

- The plant coprime polynomial fraction is obtained and formulated as $G(s) = N(s)/D(s)$.
- From specified disturbance and reference types, the internal model, i.e. $W(s)$ is determined
- $\tilde{K} = BA^{-1}$ is designed such that $P_c = \tilde{A}DW + \tilde{B}N$ is a stable polynomial.
- The controller is formed as $K = \frac{\tilde{K}}{W}$.

Since the power reference inputs i.e. (P_{sref}, Q_{sref}) are assumed to change in a stepwise manner (under worst case scenario), $d = \mathcal{L}[a_0 u(t)] \times (L_g s) = \frac{a_0}{s} \times (L_g s) = a_0 L_g$ which is a constant and so $W(s)$ turns out to be a constant as well. Therefore by inspecting the control loop it is seen the plant transfer function is a pure gain and $H_Q(s)$ in its simplest form can be a basic proportional integral (PI) compensator.

The PCC voltage control system closed-loop band width is usually selected to be adequately smaller than that of the closed-loop current controller $G_i(s)$, that is $(1/\tau_i)$. Furthermore parameters of the $H_Q(s)$ compensator can then be selected based on the phase margin and bandwidth requirements.

3.5 Grid Connection Procedure and Supervisory Controller

While the power port system described previously is designed to operate in a grid-imposed frequency mode once connected to the utility network, in case the grid encounters any problems, the system requires being able to separate and reconnect to the network. In other words in addition to the grid-tie mode the system needs to be operable in an autonomous operating mode as well. Unlike before in this mode the operating frequency is not imposed by the grid, and is controlled according to a reference command that takes into account local load as well as reconnection synchronization requirements. The same argument holds true for PCC voltage regulation in cases the system is required to supply a local load.

As Figure 3.14 shows the connection to the AC system is made possible through a main switch that is usually installed at the LV side. This way, the entity including the TISEC system and any local loads may operate in either the grid connected or the autonomous mode should any disturbances be detected in the grid system. From here onward and throughout the rest of the study, the grid-side converter whether operating under the grid or input reference frequency, along with its DC link capacitors, will be referred to as the Hybrid Power Port (HPP).

Figure 3.14: Schematic diagram of the HPP system in controlled frequency operation mode

Even though the wavering nature of tidal currents results in a variable output power, which unless supported by some kind of a storage or power leveling system, cannot be used to supply any local loads; the HPP control system in this study, will be designed generically so that supply of dedicated loads under the islanded condition is also made possible without any requirement of prior load model knowledge. This is achieved through the use of a decoupling feed-forward compensation strategy that was first recommended by Amirnaser Yazdani in [70]. Hence as long as a constant flow of power is provided, any load or clusters of loads are protected from grid distortions through autonomous operation of the TISEC system. At the same time a simple supervisory system is designed for grid dis/re-connections.

3.5.1 System Model and Equations

Figure 3.14 shows the schematic diagram of the controlled-frequency converter system which as can be seen employs the same current-controlled converter kernel as the grid imposed frequency system. In case a local load is to be supplied, each phase of the converter AC side terminals will need to be interfaced with a RLC filter. The control objective in this case is to regulate the amplitude and frequency of the load voltage (V_{abc_L}) in the presence of disturbances in the load current (I_{abc_L}). The series RL branch accompanied by a shunt capacitor C_f form the RLC filter. In other words C_f provides a low-impedance path for switching current harmonics generated by the voltage sourced converter, preventing them from penetrating into the load. Since in this study, controlled frequency system compensators are developed with no specific assumptions regarding the load dynamics or configuration, avoiding C_f , leads to the load voltage harmonic distortion, to be significantly dependent on the load impedance at switching frequency and its harmonics. Employing C_f also ensure that the RL branch is terminated to a node with some degree of voltage support. As can be seen from the Figure unlike the previous grid-tie operating mode where the PLL provided the functioning frequency here the VCO input, i.e. ω , can be changed to control the operating frequency of the system. The load voltage dynamics is however different as the capacitor dynamics also needs to be taken into consideration. Noting that the current controllers are similar to what was explained before, with reference to Figure 3.14, the following state space equations are used to describe the dynamics of the load voltage.

$$\begin{cases} C_f \frac{dV_{sa}}{dt} = i_a - i_{aL} \\ C_f \frac{dV_{sb}}{dt} = i_b - i_{bL} \\ C_f \frac{dV_{sc}}{dt} = i_c - i_{cL} \end{cases} \quad (3.46)$$

By expressing the three phase equations in dq -frame components, it then follows that:

$$\begin{cases} C_f \frac{dV_{sd}}{dt} = C_f(\omega V_{sq}) + i_d - i_{dL} \\ C_f \frac{dV_{sq}}{dt} = C_f(\omega V_{sd}) + i_q - i_{qL} \\ \frac{d\rho}{dt} = \omega(t) \end{cases} \quad (3.47)$$

Hence the following block diagram replaces the one previously depicted in Figure 3.4. One should note that the load currents can be easily provided by means of measuring devices.

Figure 3.15: Block diagram of the load voltage dynamics with the presence of C_f

3.5.2 Load voltage control under autonomous operating mode

As mentioned before the second control objective in the controlled frequency mode of operation is to regulate the amplitude of the load voltage. As Figure 3.14 illustrates, like before the control is exercised in a dq -frame defined by the angle ρ and the angular velocity $\omega = d\rho/dt$. One must note that, regulation of the amplitude of V_{sabc} , i.e. $\hat{V}_s = \sqrt{V_{sd}^2 + V_{sq}^2}$, is equivalent to ensuring that $\hat{V}_s$ tip always resides on a circle.

However, depending on the values of V_{sd} and V_{sq} , $\hat{V}_s$ can assume different angles with respect to the d -axis. A similar routine to the one adopted for the PLL is also employed here where $(V_{sd}, V_{sq}) = (\hat{V}_{sn}, 0)$, with $\hat{V}_{sn}$ being the nominal value of $\hat{V}_s$. Hence similar to the case of the grid imposed frequency mode, V_{sq} is forced to zero. The control process then can be based on the dynamic model illustrated in Figure 3.15. In other words V_{sd} and V_{sq} are controlled by i_{dref} and i_{qref} . Equation set 3.47) indicates that V_{sd} and V_{sq} are coupled and so the system of Figure 3.15 is a multiple input multiple output system. Figure

3.16 illustrates a possible control structure for the HPP system operating in the controlled frequency mode. This topology largely overcomes the difficulties related to the load dynamics and affiliated interdependencies. Figure 3.16 shows how the V_{sd} and V_{sq} coupling is eliminated by the decoupling feed-forward compensation [70]. The feed-forward compensation employed here is similar to the one utilized previously to decouple d and q currents in the current-controlled voltage sourced converter and makes it possible to control V_{sd} by i_{dref} and V_{sq} by i_{qref} . Figure 3.16 also shows that measures of i_{dL} and i_{qL} are added to i_{dref} and i_{qref} , respectively; enabling the compensated system to perform under all load conditions almost the same way as the system without the feed-forward compensation would perform under no-load conditions.

With reference to Figure 3.16, i_{dref} and i_{qref} are determined by the following equation set:

$$\begin{cases} i_{dref} = u_d - C_f(\omega V_{sq}) + i_{dL} \\ i_{qref} = u_q - C_f(\omega V_{sd}) + i_{qL} \end{cases} \quad (3.48)$$

in which u_d and u_q are two new control inputs. Consequently based on (3.23) and (3.24), i_d and i_q respond to i_{dref} and i_{qref} as:

$$\begin{cases} I_d(s) = G_i(s)U_d(s) - C_f G_i(s) \mathcal{L}[\omega V_{sq}] + G_i(s)I_{dL}(s) \\ I_q(s) = G_i(s)U_q(s) - C_f G_i(s) \mathcal{L}[\omega V_{sd}] + G_i(s)I_{qL}(s) \end{cases} \quad (3.50)$$

Figure 3.16: Control block diagram of the HPP system operating in the controlled frequency mode

The letter $\mathcal{L}$ denotes the Laplace transform operator. Substitution of $i_d(s)$ and $i_q(s)$ from (3.49) in the Laplace transforms of (3.47) yields:

$$\begin{cases} s \cdot C_f V_{sd}(s) = G_i(s) U_d(s) + C_f(1 - G_i(s)) \cdot \mathcal{L}[\omega V_{sq}] - (1 - G_i(s)) \cdot I_{dL}(s) \\ s \cdot C_f V_{sq}(s) = G_i(s) U_q(s) - C_f(1 - G_i(s)) \cdot \mathcal{L}[\omega V_{sd}] - (1 - G_i(s)) \cdot I_{qL}(s) \end{cases} \quad (3.51)$$

On the other hand the transfer function $G_i(s) = 1/(\tau_i s + 1)$ has a DC gain of one, and so $[1 - G_i(s)]$ has a zero DC gain. Thus, as long as τ_i is small, $[1 - G_i(s)]$ would be trivial over a relatively wide range of frequencies and so can be approximated by zero. Therefore:

$$\begin{cases} \frac{V_{sd}(s)}{U_d(s)} \approx G_i(s) \cdot \frac{1}{s C_f} \\ \frac{V_{sq}(s)}{U_q(s)} \approx G_i(s) \cdot \frac{1}{s C_f} \end{cases} \quad (3.52)$$

which represent two, linear, decoupled systems with u_d and u_q as inputs and V_{sd} and V_{sq} as respective outputs.

3.5.2.1 Controlled Frequency Mode Compensator Design

The simplified closed-loop block diagram of Figure 3.16 is depicted in Figure 3.17. As this Figure represents two decoupled single-input-single-output control loops it is more insightful for the compensator design.

Figure 3.17: Simplified control block diagram of the controlled frequency mode voltage regulation

Each of the control loops shown in Figure 3.17 includes a real pole at $s = -1/\tau_i$, in addition to one pole at $s = 0$. The simplest compensator for this system that fulfills a zero steady-state error in addition to fast regulation is a basic PI compensator. Thus:

$$K(s) = k_0 \cdot \frac{s + z}{s} \quad (3.52)$$

and so the loop gain turns out to be:

$$l(s) = \frac{k_0}{\tau_i C_f} \left(\frac{s + z}{s + \tau_i^{-1}} \right) \cdot \frac{1}{s^2} \quad (3.53)$$

At low frequencies, $\angle l(j\omega) \approx -180^\circ$, which is due to the double pole at $s = 0$. If $z < \tau_i^{-1}$, the loop angle first increases until it reaches a maximum of δ_m at a certain frequency that is ω_m . For frequencies greater than ω_m , the loop angle drops and asymptotically approaches -180 . The following two equations give the maximum and its affiliated frequency in terms of τ_i and z .

$$\begin{cases} \delta_m = \sin^{-1} \left(\frac{1 - \tau_i z}{1 + \tau_i z} \right) \\ \omega_m = \sqrt{z \tau_i^{-1}} \end{cases} \quad (3.54)$$

If the gain crossover frequency i.e. ω_c is assigned as ω_m , then δ_m becomes the phase margin of the system as long as the compensator proportional gain, k_0 , satisfies $|l(j\omega_c)| = |l(j\omega_m)| = 1$. Yielding in:

$$k_0 = C_f \omega_c \quad (3.55)$$

The foregoing method of compensator design is suitable for a loop gain that has two poles at $s = 0$ (including the PI compensator pole) and one real pole. This method is a well-established routine referred to in the literature as the method of *symmetrical optimum* [71]. Based on this method, the resultant closed-loop system will be of the third order. It can be shown that the closed-loop system always has one real pole at $s = -\omega_c$ and two other complex-conjugate poles that are located on a circle with a radius of ω_c , although the exact locations of the two poles on the circle depend on the selected phase margin which is usually selected to be in the range 30° to 75° .

3.5.3 Grid connection/disconnection procedure

The schematic diagram of Figures 3.18 and 3.19 demonstrate how the grid connection/disconnection takes place. For the system of Figure 3.14, the autonomous-mode occurs once:

- The dq current references, i.e. i_{dref} and i_{qref} are obtained from the dq -axis compensators of the voltage regulation scheme presented in 3.5.2
- The VCO input is switched from the PLL output to the desired frequency, i.e. ω_0
- The grid connecting switch is commanded to open

Figure 3.18: Mode command generation process in the supervisory controller

Once the grid system returns to its normal operating condition, the HPP system may be reconnected allowing for operation in the grid imposed frequency mode, thus the PCC voltage retrieves its nominal qualities. However the HPP system must be resynchronized to the grid voltage, i.e. V_{gabc} before the main switch is reclosed. Although in the controlled frequency mode, system frequency is regulated at the nominal AC system frequency, when the AC system is restored a phase error is likely to exist between V_{sabc} and V_{gabc} . As Figure 3.18 shows, the AC system restoration is detected by comparing the grid voltage amplitude, i.e. V_{gabc} , with a threshold value, that in this case is set to be 90% of its corresponding nominal value. Hence, once $\hat{V}_g$ is larger than the threshold, the AC system is considered to

The diagram illustrates the architecture of a Hybrid Power Port (HPP) capable of both autonomous and grid-connected operation. The HPP block receives reference currents i_{qref} and i_{dref} , and a power factor ρ . It outputs three-phase currents i_{abc} through a line impedance Z_{line} to a main switch. The main switch is controlled by a mode command (0 for autonomous, 1 for grid-connected) and a PLL. The PLL receives grid voltage V_{gabc} and outputs a phase-locked signal. The main switch is connected to a transformer (Yg, Delta) and a grid through a series impedance $Z_{Th} \approx \omega_0 L_g$. The grid voltage is $V_{Grid} \approx 0$. The HPP also includes a local load and a capacitor C_f . The main switch is controlled by a mode command (0 for autonomous, 1 for grid-connected) and a PLL. The PLL receives grid voltage V_{gabc} and outputs a phase-locked signal. The main switch is connected to a transformer (Yg, Delta) and a grid through a series impedance $Z_{Th} \approx \omega_0 L_g$. The grid voltage is $V_{Grid} \approx 0$.

In order to stop multiple transitions between the two modes $\widehat{V}_g$ is passed through a low-pass filter (LPF) followed by a hysteresis block. Figure 3.18 also shows that once V_{gq} and $\omega(t)$ become adequately close to their steady-state values (respectively, zero and ω_0) the synchronization scheme changes the *mode command* from 0 to 1, resulting in the closure of the main switch and the generation of i_{dref} and i_{qref} in proportion to the required real and reactive powers. Thereafter, the converter continues its operation in the grid imposed frequency mode.

3.6 Concluding Remarks

In this chapter the 3-level NPC converter of chapter 2 was utilized in a grid connected power port system. An independent active and reactive power handling system was established through the use of a PLL and dq -frame current controllers. It was shown how the phase angle and the amplitude of the converter line current (with respect to the PCC voltage) could be changed by controlling the converter dq -frame modulation indices. This way, not only the converter is protected against overcurrent conditions, active and reactive powers are managed with superior dynamic performances. Decoupled active and reactive powers controllability in turn allows for a controlled DC voltage power port mechanism with various power factor regulation capabilities, namely the unified power factor scheme where variable amounts of active power can be interchanged with the utility grid without any reactive power demand as well as a voltage regulatory scheme where any PCC bus voltage variability is counteracted through the converter reactive power handling strategy.

Furthermore it was shown how in cases where the utility grid encounters a problem, the system separates from the network and reconnects once the fault is cleared. The network status monitoring is performed through a supervisory controlling mechanism that allows the power port system to detect grid complications as well as enforcing resynchronization criteria for reconnection. A main switch connects and disconnects the converter from the network according to the supervisory controller generated *mode command* signal which can change the system setting from grid connected to autonomous. By entering the autonomous operation mode the system terminal voltage and frequency are controlled internally; allowing supply of dedicated loads under an islanded condition. A decoupling feed-forward compensation strategy was employed to allow independent operation of the system from load status. The power port system along with its supervisory controller was named the Hybrid Power Port (HPP).

Usage of the HPP reactive power capacity for PCC bus voltage regulation as well as the technical insight into the PCC bus voltage regulatory system and its modeling were the main contributions of the author in this chapter.

Chapter 4

Variable Frequency Converter and Maximum Power Point Tracking

4.1 Introduction

In this chapter variable frequency voltage sourced converters are introduced. Variable Frequency Converters (VFCs) make the kernel of the electromechanical energy conversion systems where the converter is the interface between the DC bus and the AC stator terminals of an electric machine. The machine frequency and voltage are controlled by the converter such that the torque could be controlled while the flux is kept at its nominal value. Control the torque consecutively allows for the control of the machine speed and power. In variable speed systems, the converter operates in a variable frequency mode to allow for various speed values, and thus out of different control strategies only dq -frame renders the job effectively as in this strategy controller gains are independent of the system operating frequency. This chapter starts with presentation of asymmetrical three-phase induction machine model and the description of the asynchronous machine vector control and continues with different variable speed generation topologies and maximum power tracking methods.

4.2 Variable Frequency Converter Structure

The schematic diagram of a variable frequency voltage sourced converter is presented in Figure4.1. The Variable Frequency Converter (VFC) system depicted inFigure4.1 employs a 3-level NPC converter that is directly connected to the machine stator. Due to the machine relatively large inductance practically no interfacing inductors are required for damping the current harmonics that are caused by the switching voltage harmonics of the converter[20]. In this study, the VFC is connected through a DC link to the hybrid power port that establishes a bidirectional power flow path between the mechanical prime mover and the utility network. As mentioned in chapter 2, the converter can be modeled by an ideal voltage sourced converter and a current source representing the switching power losses. In this case the three resistors representing the switch conduction losses can be considered as part of the machine stator resistance.

Figure 4.1: Schematic diagram of a variable frequency converter for controlling an asynchronous tidal generator

As Figure 4.1 illustrates the VFC system in this study, is controlled in a dq -frame since the variable-frequency nature of the wave forms complicates the compensator design if other frames such as $\alpha\beta$ are used. The control inputs of the VFC system are the torque command, (T_{eref}) and the reference flux ($Flux_{ref}$). As will be seen the reference flux is set to the machine rated value and normally is kept unchanged. The torque command on the other hand is set to be the output of another control loop which can actively control, T_{eref} to effectively track the turbine maximum power point, the so called Maximum Power Point Tracking (MPPT) scenario.

4.3 Control of VFC system

Although the constant frequency power port system of the previous section was designed to be able to establish a bidirectional power flow while maintaining the DC capacitor voltage regulated, here the converter is controlled to be capable of establishing a bidirectional power flow at variable frequencies assuming the DC voltage is sustained. Hence the control objectives here are managing the machine torque

and flux. According to [20] the electrical and mechanical equations governing the dynamics of a symmetrical three phase AC machine, are:

$$\left\{ \begin{array}{l} \frac{d\vec{\lambda}_s}{dt} = \vec{V}_s - R_s \vec{I}_s \\ \frac{d\vec{\lambda}_r}{dt} = \vec{V}_r - R_r \vec{I}_r \\ \vec{\lambda}_s = L_m[(1 + \sigma_s)\vec{I}_s + e^{j\theta_r}\vec{I}_r] \\ \vec{\lambda}_r = L_m[(1 + \sigma_r)\vec{I}_r + e^{-j\theta_r}\vec{I}_s] \end{array} \right. \quad (4.1)$$

$$\left\{ \begin{array}{l} J \frac{d\omega_r}{dt} = T_m - T_e \\ \frac{d\theta_r}{dt} = \omega_r \\ T_e = \left(\frac{3}{2}L_m\right) \text{Im}\left(e^{-j\theta_r}\vec{I}_s\vec{I}_r^*\right) \end{array} \right. \quad (4.2)$$

Where the stator and rotor leakage factors σ_s and σ_r are defined in Equation (4.3) with L_s , L_r and L_m representing the stator, rotor and magnetizing inductances respectively:

$$\left\{ \begin{array}{l} \sigma_s = \frac{L_s}{L_m} - 1 \\ \sigma_r = \frac{L_r}{L_m} - 1 \end{array} \right. \quad (4.3)$$

In a squirrel cage or a wound rotor asynchronous machine whose rotor terminals are short circuited $\vec{V}_r = 0$, and $\vec{I}_r$ is not measurable. According to [20] by introducing a fictitious space phasor current and defining it as: $\vec{I}_m \triangleq (1 + \sigma_r)e^{j\theta_r}\vec{I}_r + \vec{I}_s$, it can be shown that the machine electric torque simplifies to:

$$T_e = \frac{3}{2} \left(\frac{L_m}{1 + \sigma_r} \right) \hat{I}_m I_q \quad (4.4)$$

What Equation (4.4) imposes is that if $\hat{I}_m$ (which basically corresponds to the machine magnetizing flux) is maintained at a constant value, the machine electric torque becomes a linear function of the q -axis component of the stator current I_q . Hence $\hat{I}_m$ is regulated at the machines nominal magnetizing current corresponding to the rated voltage and $\omega_r = \omega_0$, i.e.:

$$\begin{cases} \hat{I}_{mref} = \sqrt{\frac{2}{3}} \frac{V_{s-nom}}{2\pi f_0 R_s \tau_s} = Constant \\ \tau_s = \frac{(1 + \sigma_s)L_m}{R_s} \end{cases} \quad (4.5)$$

It can also be shown that:

$$\tau_r \frac{\hat{I}_m}{dt} + j(\omega - \omega_r) \tau_r \hat{I}_m = -\hat{I}_m + \vec{I}_s e^{-j\rho} \quad (4.6)$$

where:

$$\begin{cases} \tau_r = \frac{(1 + \sigma_r)L_m}{R_r} \\ \omega_r = \frac{d\theta_r}{dt} \\ \omega = \frac{d\rho}{dt} \end{cases} \quad (4.7)$$

By decomposing Equation (4.6) into its real and imaginary components the following equations are obtained that form the basis of a SCIG rotor-flux observer [67].

$$\begin{cases} \tau_r \frac{\hat{I}_m}{dt} = -\hat{I}_m + Re(\vec{I}_s e^{-j\rho}) = -\hat{I}_m + I_d \\ \omega = \omega_r + \frac{Imag(\vec{I}_s e^{-j\rho})}{\tau_r \hat{I}_{mr}} = \omega_r + \frac{I_q}{\tau_r \hat{I}_m} \end{cases} \quad (4.8)$$

Figure 4.3 shows the realization of the equation set (4.8) in a rotor-flux observer model that is employed in this study. This configuration is generally referred to as the current model flux observer. For the most part, regulation of $\hat{I}_{mr}$ is only exercised either during the system startup process, or when I_{sq} is zero which means that the machine is at standstill. This way, flux establishment is ensured before any torque demand.

Figure 4.2: Schematic diagram of the rotor-flux observer for SCIG

The task of the flux observer presented in Figure 4.2 is to take I_d , I_q and ω_r as inputs and supply $\hat{I}_m$, ρ and ω as outputs. As can be seen from the Figure, the I_d and I_q components are themselves products of an $abc2dq$ transformer block which requires the provided angle ρ_2 in a feedback loop; ρ_2 is also required for the VFC system PWM signal generator. To decouple the dynamics of I_d and I_q , the output ω is utilized in a feed-forward compensation strategy while the third product of the system, i.e. $\hat{I}_m$ is used as a feedback signal for flux regulation. As the system presented in Figure 4.2 requires the machine speed, i.e. ω_r as an input, a speed sensor or an encoder is also required for this topology to work. There are other flux observer systems available in the literature, that are constructed based on an alternative approach in which ω_r is not required; instead stator voltages and currents (V_{dq} and I_{dq}) are used to generate ρ_2 and ω . Each type of flux observer has its own cons and pros; further details of such systems as well as comparisons studies can be accessed in [67, 72, 73].

4.3.1 Rotor Field Coordinates, Vector Control

Figure 4.3 depicts the block diagram of the scheme employed in rotor-field coordinates for vector control of the SCIG. As the Figure illustrates, the d -axis current control loop reference signal, that is $I_{d_{ref}}$, is generated by the flux controller $H_\lambda(s)$ which takes the error signal $\hat{I}_{m_{ref}} - \hat{I}_m$ as input. Since the machine flux does not need changing $\hat{I}_{m_{ref}}$ is a constant value set according to the machine flux nominal value; the $\hat{I}_m$ signal on the other hand is obtained from the flux observer of Figure 4.2. $H_\lambda(s)$, can be as simple as a basic PI controller as long as it's tuned for fast closed loop response. Although, as a result of the large rotor time constant, $I_{d_{ref}}$ and I_d may undergo significant transient excursions if the flux regulating loop is fast, this does not overload the machine or converter, as under normal circumstances flux regulation only takes place at startup, when $I_q = 0$. Alternatively if the speed of the closed-loop response is not an issue, I_d can settle at the steady state value of $\hat{I}_m$ in about $5\tau_r$, given $I_{d_{ref}} = \hat{I}_m$.

The machine torque control loop is also depicted in Figure 4.3. According to Equation (4.4), the machine torque is $\frac{3}{2} \left(\frac{L_m}{1+\sigma_r} \right) \hat{I}_m$ multiplied by I_q . Consequently the reference value for $I_{q_{ref}}$, is determined by dividing the T_{eref} by $\frac{3}{2} \left(\frac{L_m}{1+\sigma_r} \right) \hat{I}_m$. It should be noted that T_{eref} in our design is the output of another compensator which aims at regulating the generator speed, i.e. ω_r ; hence the torque control loop is closed and the system is robust to model uncertainties. An oversight of the system in Figure 4.3 shows that the I_d and I_q currents are controlled by the reference commands generated by the flux and torque compensators respectively. Therefore this is a two-input-two-output system that is inherently nonlinear

and consequently hard to control. However as seen previously in similar situations, the system can be transformed into two decoupled, single-input–single-output linear time-invariant subsystems through the use of proper feed-forward compensation techniques. In such a case each of the resultant systems can be characterized by a first-order transfer function $G_i(s)$ that relates I_d to I_{dref} and I_q to I_{qref} (Fig. 4.3). Consequently $G_i(s)$ parameters can be tuned for faster tracking of the system reference inputs.

Figure 4.3: Control block diagram of the vector-controlled SCIG in rotor field coordinates

4.3.2 Controlling the Machine dq -Frame Currents

Although the variable frequency converter only controls the stator voltages, as seen in the previous subsection, the machine flux and torque are controlled, through the d and q axis currents respectively. Therefore, mathematical expressions must be developed to relate the stator dq -frame currents to the voltages. The stator terminal voltage and current are related based on (4.1). Substituting for I_r and λ_s in the stator flux equations results in the following equation:

$$L_m \frac{d}{dt} \left[\frac{(1 + \sigma_s)(1 + \sigma_r) - 1}{1 + \sigma_r} \vec{I}_s + \frac{1}{1 + \sigma_r} \hat{I}_m e^{j\rho} \right] = \vec{V}_s - R_s \vec{I}_s \quad (4.9)$$

By defining:

$$\begin{cases} \sigma = 1 - \frac{1}{(1 + \sigma_r)(1 + \sigma_s)} \\ \tau_s = \frac{L_m(1 + \sigma_s)}{R_s} \end{cases} \quad (4.10)$$

And decoupling the resultant complex equation into its real and imaginary components, the following equations are deduced:

$$\begin{cases} \left(\sigma \tau_s \frac{dI_d}{dt} + I_d \right) = \sigma \tau_s \omega I_d - (1 - \sigma) \tau_s \frac{d\hat{I}_m}{dt} + \frac{1}{R_s} V_d \\ \left(\sigma \tau_s \frac{dI_q}{dt} + I_q \right) = -\sigma \tau_s \omega I_d - (1 - \sigma) \tau_s \omega \hat{I}_m + \frac{1}{R_s} V_q \end{cases} \quad (4.11)$$

The two equations in (4.11) represent a nonlinear system for which I_d and I_q are both state variables and outputs and V_d and V_q are the inputs. According to (4.11), dynamics of I_d and I_q are coupled. Furthermore in consideration of the fact that ω and $\hat{I}_m$ are both functions of I_d and I_q , due to the presence of terms like ωI_d , ωI_q , and $\omega \hat{I}_m$ the system is nonlinear. By defining the two new control inputs, u_d and u_q this nonlinearity can be avoided in the control system [20]:

$$\begin{bmatrix} u_d = \sigma \tau_s \omega I_q - (1 - \sigma) \tau_s \frac{d\hat{I}_m}{dt} + \frac{1}{R_s} V_d \\ u_q = -\sigma \tau_s \omega I_d - (1 - \sigma) \tau_s \omega \hat{I}_m + \frac{1}{R_s} V_q \end{bmatrix} \Rightarrow \begin{cases} \left(\sigma \tau_s \frac{dI_d}{dt} + I_d \right) = u_d \\ \left(\sigma \tau_s \frac{dI_q}{dt} + I_q \right) = u_q \end{cases} \quad (4.12)$$

The two equations in (4.12) represent two first order and decoupled subsystems that have unity DC gains. As Figure 4.4 illustrates, while I_d and I_q are controlled by u_d and u_q respectively, they in turn are obtained from two corresponding PI compensators that processes $I_{d_{ref}} - I_d$ and $I_{q_{ref}} - I_q$. Parameters of the compensators are determined based on the block diagrams of Figure 4.4.

Figure 4.4: Control block diagram of the induction machine stator current controls, based on Equation (4.12)

By assigning the PI controller parameters as shown in (4.13):

$$k(s) = \frac{k_p s + k_i}{s}, \begin{cases} k_p = \frac{\sigma \tau_s}{\tau_i} \\ k_i = \frac{1}{\tau_i} \end{cases} \quad (4.13)$$

the current transfer functions can be found to be:

$$\begin{cases} \frac{I_d(s)}{I_{dref}(s)} = G_i(s) = \frac{1}{\tau_i s + 1} \\ \frac{I_q(s)}{I_{qref}(s)} = G_i(s) = \frac{1}{\tau_i s + 1} \end{cases} \quad (4.14)$$

The time constant τ_i is a designer's choice. The actual control signals V_d and V_q are calculated based on (4.12), as:

$$\begin{aligned} V_d &= R_s \left[u_d - \sigma \tau_s \omega I_q - (1 - \sigma) \tau_s \frac{d\hat{I}_m}{dt} \right] \\ V_q &= R_s \left[u_q - \sigma \tau_s \omega I_d - (1 - \sigma) \tau_s \omega \hat{I}_m \right] \end{aligned} \quad (4.15)$$

Mathematically selecting the compensator parameters according to (4.14) will result in elimination of the plant pole at $p = -1/\sigma \tau_s$ by the PI compensator zero. However in practice most of the machine parameters are measured subject to tolerances. In addition, they vary with changes in the operating conditions, e.g. L_m and σ are obtained from measurements, that come with errors, and R_s undergoes

significant changes as machine temperature varies. Although such changes may result in a weak or incomplete pole cancellation, designing the compensators based on the machine nameplate parameters often renders the compensations effective, and the first-order transfer functions of (4.14) can still be used to characterize the resultant closed-loop subsystems [20].

Based on the methodology explained above the complete dq -frame current-control implementation illustrated in Figure 4.3 can be explained as follows:

While the system receives the command references $I_{d_{ref}}$ and $I_{q_{ref}}$ from the flux and torque compensators of Figure 5.4, $\hat{I}_m$ and ω are taken from the flux observer of Figure 4.2. Subsequently, V_d and V_q are calculated based on (4.15). To compensate for the converter voltage conversion gain, m'_d and m'_q are calculated by dividing V_d and V_q by $V_{DC}/2$. The resultant values are then delivered to the PWM signal generator. Since, the machine is interfaced with the VFC via a three-wire connection; the third-harmonic injected PWM is adopted to permit a lower DC-bus voltage. Figure 4.3 also shows that to avoid differentiation and its associated noise-amplifying characteristics the controller implementation is slightly modified with respect to (4.15) as the term $(1 - \sigma)\tau_s d\hat{I}_m/dt$ is omitted from the V_d expression. Since, except during the start-up process, $\hat{I}_m$ is regulated at a constant value the average of $d\hat{I}_m/dt$ is zero and so the simplification poses no issue.

4.4 Variable Speed Generation Systems

Previously the variable frequency converter system was described. In this section it will be demonstrated that a grid connected variable speed generation system can be implemented by a back to back connection of a variable frequency converter (VFC) and a hybrid power port (HPP). As a result while the VFC system controls the turbine generator, the HPP provides the grid interface. Furthermore based on the type of current speed variation several Maximum Power Point Tracking (MPPT) methods are introduced; namely the *Constant K*, *Variable K* and *Imposed Delay* routines. While the *constant K* method is widely being employed in most variable generation systems, it is shown that once the flow speed variations become too rapid this conventional method does not do a good job in tracking the maximum power point. Consequently the *Variable K* and *Imposed delay* methods are proposed which enhance the MPPT by boosting the system acceleration/deceleration.

4.4.1 Different Variable Speed Generation Topologies

Three dominant types of variable speed generation systems are illustrated in Figures 4.5(a)–(c). They include:

- Type *a* (Figure 4.5(a)) that employs an induction machine in which the frequency and rotor speed are adjusted by a full scale power electronic converter. The same converter also allows the transfer of real power from the variable frequency generator to the constant frequency utility network.
- Type *b* (Figure 4.5(b)) in which the variable speed system is based on a doubly fed induction machine that utilizes a cheaper power-electronic converter system. Although the machine synchronous speed is directly dictated by the grid frequency, the rotor speed is designed to be controlled through the rotor connected power converter. The same converter permits a bidirectional power exchange between the machine rotor and the utility network.
- Type *c* (Figure 4.5(c)). The principle of operation in this topology is conceptually the same as that of type *a*, with the exception that here synchronous or permanent magnet machines are employed instead. The full scale converter system in this topology adjusts the frequency of the stator circuit allowing the rotor speed to be variable. As the Figure illustrates in case a machine with high number of poles is used the gearbox can be eliminated.

The system that is employed and analyzed in this study is of type (a) which means that an asynchronous machine is employed along with a fixed ratio gearbox. The full scale AC/DC/AC converter on the other hand, is realized by utilizing the hybrid power port system of chapters 3 along with the variable frequency converter introduced earlier. The two back to back voltage sourced converters interfacing the generator-turbine system and the AC utility network.

Figure 4.5: Basic schematic diagrams of three main categories of variable speed generation systems

4.5 Turbine model characteristics

Operation of a conventional horizontal axis turbine can be characterized by its mechanical power, given by Equation (4.16)

$$P_{Turbine} = \frac{1}{2} \rho A V^3 C_p(\lambda, \beta) \quad (4.16)$$

where ρ , is the mass density of the medium in kg/m^3 , $A = \pi R^2$ is the turbine swept area

in m^2 with R representing the turbine radius in m , and V is the flow speed in m/s . $C_p(\lambda, \beta)$ is called the capacity factor or power capturing efficiency. This factor is limited to $16/27 \cong 59\%$ by the well-known Betz law. For wind turbines, C_p have typical values in the range of 25-30% while for marine turbines, C_p is estimated to be in the range of 35-50% [74]. β is the pitch angle of the turbine blades in degrees and λ is the tip-speed ratio defined as:

$$\lambda = \text{Tip Speed Ratio} = \frac{R \cdot \omega_{tur}}{V} \quad (4.17)$$

where ω_{tur} is the turbine angular velocity in rad/s . Equation (4.17) indicates that λ is the ratio of the turbine blade tip tangential speed to the speed of the incoming flow. Consequently what Equation (4.16) infers is that, the turbine power is the product of the available power in the stream ($0.5\rho AV^3$) and the capacity factor $C_p(\lambda, \beta)$ which determines how much of the available power can be extracted by the turbine. The former term cannot be controlled however, the latter, that is, $C_p(\lambda, \beta)$ can be manipulated through its constituent terms λ and β . According to (4.17), λ is a function of V and ω_{tur} , consequently the control of C_p indeed boils down to the control of ω_{tur} and β . Although there are some analytical expressions available for C_p that defines it as a static function [75, 76], in practice each turbine has a unique $C_p(\lambda, \beta)$ characteristic which is a dynamic, highly nonlinear, function of λ and β . Figure 4.6 depicts a typical $C_p(\lambda, \beta)$ versus λ curve, drawn for five different values of pitch angle ($\beta = 0$ to $\beta = 20^\circ$). The characteristics of the C_p curve of Figure 4.6 is taken from the wind turbine characteristics provided in [77] with some modifications to allow for an arbitrary selection of the λ_{opt} and C_{pmax} values. The following equation defines $C_p(\lambda, \beta)$ and Figure 4.4 shows the Simulink implemented model.

$$\left\{ \begin{array}{l} C_p(\lambda, \beta) = c_1 \left(\frac{c_2}{\lambda_i} - c_3\beta - c_4 \right) e^{\frac{c_5}{\lambda_i}} + c_6\lambda \\ \frac{1}{\lambda_i} = \frac{1}{\lambda + 0.08\beta} - \frac{0.035}{\beta^3 + 1} \\ c_{1-6} = [0.5176 \quad 116 \quad 0.4 \quad 5 \quad 21 \quad 0.0068] \end{array} \right. \quad (4.18)$$

Figure 4.6: Capacity factor vs. tip speed ratio, characteristic curve of the employed turbine for different values of β

Figure 4.7: Simulation model of the turbine described by Equation (4.18) with adjustable tip speed ratio

The differences between the constant and variable speed systems can be explained further on the basis of the properties of the capacity factor (C_p). In a constant-speed system, λ cannot be controlled as ω_{tur} is constant and so λ changes with V which leads to C_p not being able to maintain its maximum value persistently; resulting in the nonoptimum operation of the turbine over a wide range of flow speeds. In a variable speed system however as long as ω_{tur} is adjusted in proportion to V , λ can be kept constant and equal to λ_{opt} thus maximizing C_p . Consequently for any flow speed, since C_p assumes its maximum value the turbine power is maximized.

The C_p value that determines the turbine power output can be controlled by changing any or both of its constituent variables, i.e. λ and β . In this study, β is not controlled and is maintained at zero. However

practically in most variable speed systems, β is actively controlled limiting the turbine power in case the power exceeds its rated value, and is set to its maximum value, e.g., 90° , to stop the power generation under extreme conditions. The tip speed ratio on the other hand is adjusted to be equal to its optimum value which according to the $\beta = 0$ curve (Figure 4.6) equals to 5. According to Figure 4.6, $C_p(\lambda = 0) \approx 0$, but as λ increases C_p also increases until it reaches the peak value at $\lambda = \lambda_{opt}$; thereafter, a further increase in λ results in the drop in the C_p value. It should be noted that the peak value of C_p , that is, $C_{pmax} = C_p(\lambda_{opt})$, is the largest when β is zero; as β is increased, the peak value of C_p drops, and so does the turbine power P_{tur} . The C_{pmax} value for $(\lambda = 5, \beta = 0)$ is arbitrary set to be 50%. To simplify the analysis of a variable speed system where a gearbox is involved, as instructed in [20] a per unit system is employed in which:

- P_b = machine nominal power
- ω_b = machine nominal electrical frequency ; Thus turbine base speed in case a gearbox is employed = ω_b/N , (where N is the gearbox ratio).
- $T_b = P_b/\omega_b$.

In this method the machine number of poles can be accounted for as part of the gearbox ratio. Based on this per-unit system, the drive train is assumed to obey a single-mass model and the turbine per-unit speed is equal to that of the machine rotor speed, i.e. $\omega_{tur-pu} = \omega_{r-pu}$. Consequently ω_{r-pu} and ω_{tur-pu} may be used interchangeably. The under study turbine per unit power speed characteristic is depicted in Figure 4.8 for different flow speeds, assuming $\beta = 0$. Each characteristic curve in Figure 4.7 corresponds to a particular flow speed resulting in the affiliated turbine power P_{tur-pu} becoming a function of only the rotor speed ω_{r-pu} . According to Figure 4.8, for a given flow speed the turbine power is insignificant at small rotor speeds, however, similar to the C_p curve, the power increases as the rotor speeds up, until the power peak value corresponding C_{p-Max} is reached. This power peak takes place at a rotor speed that corresponds to λ_{opt} ($\lambda_{opt} = 5$ for the turbine characterized by Figure. 4.6). Thereafter any further increase in ω_{r-pu} results in the continuous reduction of the turbine power until the power curve crosses the zero axis at a relatively high rotor speed. The Figure also shows that as the flow speed increases the rotor speed corresponding to the maximum power shifts toward a higher value which is expected since for a constant β and λ_{opt} to keep the tip speed ratio constant in a faster flow the rotor speed must increase as well. The dashed line that joins the peak power points in the Figure is called the “maximum power point trajectory”. In the next section, it will be demonstrated that this curve is proportional to the rotor speed cubed.

Figure 4.8: Turbine power output plotted against its angular velocity for various incoming flow speeds

4.6 Maximum Power Extraction Methods

Apart from the three distinguished tidal flow patterns, i.e., diurnal, semi-diurnal and composite currents, based on the length of their repetitive cycles, tidal currents can be divided into two main categories namely, the channel type which is the dominant pattern in natural channels and estuaries with consistent and predictable [78] speed variations and the costal or near-shore flow type which has a much shorter period and so changes speed and direction more rapidly, Figure 4.9.

Figure 4.9: Different types of tidal stream speed profiles

Actual recorded current speed profiles illustrated in the following Figures, confirm the aforementioned characteristics. Figure 4.10 depicts the current speed profile that was recorded at Martinez-Amorco station in November 2010. This station is located in the strait that connects the San Pablo bay in California to the bays of Suisun and Grizzly. As can be seen from the Figure once the current builds up, the tidal stream pattern in that channel is quite consistent with minor speed variations. On the other hand Figure 4.11 shows the speed profile of a harbor near the Ship Island in Mississippi; apparently although the general tidal flow pattern is still observable, the instantaneous speed variations have increased.

Figure 4.10: A channel type current speed profile, recorded at Martinez-Amorco station, 1/11/2010 [87]

Figure 4.11: A near-shore type current speed profile, recorded at Day Marker station, 7/11/2010 [87]

The changing speed calls for a changing generator speed if the tip speed ratio is to be regulated at its optimum value, thus providing the maximum turbine power regardless of the current speed change. This procedure is usually referred to as the Maximum Power Point Tracking (MPPT). There are various MPPT methods available for wind and solar power systems such as the hill-climbing search control, power

signal feedback or the use of fuzzy logic controllers [79-81]. However no articles in the literature show MPPT methods being examined for marine in-stream turbines. In this section, initially the conventional MPPT method which is widely used in variable speed generation systems is introduced and analyzed when the tidal turbine experiences a composite tidal flow. Consequently it is shown how the method defects if employed under boisterous speed profiles that are typical of near shore regions. This conventional method is then considered as the benchmark in the analysis of other MPPT techniques. Based on the characteristics of the resource and in order to overcome the MPPT problem in near-shore regions, two MPPT strategies are proposed by the author referred to as the *variable K* and *imposed delays* techniques which provide a much faster speed tracking through boosting the generator acceleration. These two methods are then compared to the conventional MPPT routine, (referred to as the *constant K* method) in different scenarios. A detailed case study is presented in later chapters to demonstrate, the advantages of the proposed methods when used under the tidal flow patterns of coastal regions. This should be seen as a timely contribution to the emerging field of marine current power generation.

4.6.1 Conventional maximum power point tracking

As mentioned before, variable speed systems are desired to output their maximum possible electrical power, regardless of the speed. Based on the static $C_p(\lambda, \beta = 0)$ characteristics curve shown in Figure 4.6, in order to maximize the turbine power, regardless of the tidal stream speed changes, λ must be regulated at λ_{opt} such that $C_p(\lambda = \lambda_{opt}, \beta = 0) = C_{pmax}$, where C_{pmax} is the peak value of C_p at λ_{opt} . Consequently, based on (4.17):

$$V = \frac{R \cdot \omega_{tur_{opt}}}{\lambda_{opt}} \quad (4.19)$$

where $\omega_{tur_{opt}}$ is the turbine speed at which $\lambda = \lambda_{opt}$. Replacing V from (4.19) in (4.16), we obtain:

$$P_{tur_{opt}} = \left(\frac{0.5 \rho A R^3 C_{p-max}}{\lambda_{opt}^3} \right) \omega_{tur_{opt}}^3 \quad (4.20)$$

Furthermore dividing both sides of (4.20) by $\omega_{tur_{opt}}$, the optimal turbine torque is deduced as:

$$T_{tur_{opt}} = \left(\frac{0.5 \rho A R^3 C_{p-max}}{\lambda_{opt}^3} \right) \omega_{tur_{opt}}^2 \quad (4.21)$$

Expressing equations (4.20) and (4.21) in terms of per-unit values yields:

$$\begin{cases} P_{tur-opt_{pu}} = K_{opt} \cdot \omega_{r-opt_{pu}}^3 \\ T_{tur-opt_{pu}} = K_{opt} \cdot \omega_{r-opt_{pu}}^2 \end{cases} \quad (4.22)$$

where $\omega_{tur-opt_{pu}}$ is replaced by $\omega_{r-opt_{pu}}$, and the constant K_{opt} is

$$K_{opt} = \frac{0.5 \rho A R^3 \omega_b^3 C_{pmax}}{N^3 P_b \lambda_{opt}^3} \quad (4.23)$$

Equation (4.22) infers that under the variable-speed regime, the maximum turbine power within reach is proportional to the cube of the turbine speed. Equation (4.22) also shows that assuming a constant λ , the turbine torque must be proportional to the square of the turbine speed [20]. Thus, to operate on the maximum power point the control system must force the turbine torque to change in proportion to the rotor speed squared. Moreover the system must ensure that the turbine speed changes according to the flow speed such that $\lambda = \lambda_{opt}$. These two objectives can be realized if the following relationship is imposed on the machine [82]:

$$T_{e-pu} = -K_{opt} \times \omega_{r-pu}^2 \quad (4.24)$$

where T_{e-pu} is the machine electrical torque in per units. The rationale behind this is easily explained according to the machine mechanical dynamics described by:

$$2H \frac{d\omega_{r-pu}}{dt} = \sum T = T_{tur-pu} + T_{e-pu} = T_{tur-pu} - (K_{opt} \cdot \omega_{r-pu}^2) \quad (4.25)$$

where H is the inertia constant in seconds, defined as:

$$H = \frac{\left(\frac{1}{2} J \omega_b^2\right)}{P_b} \quad (4.26)$$

In this equation, J represents the moment of inertia in $kg \cdot m^2$, P_b is the power base in Watts and ω_b is the base angular speed in rad/s . Dropping the p.u. subscript, Figure 4.10 presents a graphical visualization of (4.25) in which the torque speed characteristic of the turbine is superimposed on the curve describing $-T_e = K_{opt} \cdot \omega_r^2$. As Figure 4.10 indicates if, $\omega_r < \omega_{r_{opt}}$, then $T_{tur} > K_{opt} \omega_r^2$. Thus according to (4.25), $d\omega_r/dt$ is positive leading to the turbine-generator set accelerating and ω_r increases. In a similar fashion if, $\omega_r > \omega_{r_{opt}}$, then $T_{tur} < K_{opt} \omega_r^2$, $d\omega_r/dt$ is negative, and ω_r decreases. Consequently the

optimum operating point is stable and a steady state at which $d\omega_r/dt = 0$ and ω_r remains constant at $\omega_{r_{opt}}$ can be reached. At this point based on (4.25) the turbine torque equals $T_{tur} = K_{opt} \cdot \omega_{r_{opt}}^2$. This outcome agrees with (4.22), which was developed based on the assumption of a constant λ and a maximized power.

The machine electrical power under this strategy can be obtained by multiplying both sides of (4.24) by ω_r , in view of the machine efficiency. Thus:

$$P_e = -K_{opt} \omega_r^3 \times \eta \quad (4.27)$$

The turbine power speed characteristic superimposed on the $-P_e = K_{opt} \cdot \omega_r^3$ curve is illustrated in the Figure 4.12. As Figure 4.13 shows, for any flow speed, the two curves intersection point shows the maximum power corresponding to that particular speed. Despite sudden tidal flow speed fluctuations that due to the inertia of the drive train, ω_r assumes smoother variations which means that the machine power formulated by (4.27) is also considerably smoother than the turbine power. Although practically a steady-state condition is rarely reached, the foregoing methodology can enable electrical power maximization in an average bases.

Figure 4.12: Graphical illustration of the conventional MPPT strategy in which machine electrical torque T_e changes in proportion to ω_r^2

Figure 4.13: Graphical illustration of the conventional MPPT strategy in the steady state, when the machine and turbine torques are balanced and the optimum speed is reached.

4.6.2 Enhanced Maximum Power Point Tracking With Acceleration Boost

As was seen in the previous section a maximum power extraction can be assured if the machine electric torque is commanded to change proportionally to the square of its rotor speed, i.e.:

$$\left\{ \begin{array}{l} J \frac{d\omega_r}{dt} = T_m - T_e \\ T_{e_{ref}} = -K \cdot \omega_r^2 \\ K = K_{opt} = \frac{0.5\rho AR^3 \omega_b^3 C_{p_{max}}}{N^3 P_b \lambda_{opt}^3} \end{array} \right. \quad (4.28)$$

where ω_b and P_b are the base rotor speed (nominal electrical frequency of the machine), and power (nominal power of the electrical machine) respectively. N is the gearbox ratio and R is the turbine radius. Despite the simplicity and robustness of the conventional MPPT method where the machine electrical torque (T_e) is commanded to change in proportion to ω_r^2 , adequate time is required for the mechanical system to respond to any sharp flow speed changes, and the turbine-generator speed reaching

its optimum value. The limited dynamic response of this method makes it ineffective when subjected to rapid speed variations, such as what a near shore tidal turbine experiences. To overcome this problem, it is proposed for the control system to be modified so that the machine is enabled to accelerate/decelerate faster.

It is observed from (4.28) that the machine acceleration ($d\omega_r/dt$) is a function of J (the moment of inertia which is a virtue of the machine) and the difference between the turbine mechanical torque and the electrical torque of the machine, i.e. $(T_m - T_e)$. The turbine mechanical torque (T_m) is a speed dependent variable, i.e. $T_m = f(\omega_r)$, however taking into account the much larger mechanical time constants compared to the electrical ones, the turbine mechanical torque can be assumed constant at the instant the machine electrical torque (T_e) is commanded to change through the variable frequency converter. Hence the machine acceleration can be controlled by altering the electrical torque of the machine, i.e. T_e . In situations where the flow speed changes occur in shorter time intervals it is proposed for the electrical torque to be controlled so that the electro-mechanical torque difference $(T_m - T_e)$, is made greater, causing a more rapid generator-turbine speed change. In other words by commanding T_e to change to values that are greater/smaller than what the conventional MPPT method proposes, the machine is enabled to accelerate/decelerate faster, causing a more rigorous tracking of the flow speed variations. This in turn results in a better capacity factor (C_p) profile and thus a more effective power capture. It must be noted that in practice the acceleration/deceleration boost should be designed according to the system electrical and mechanical limitations to avoid damages. In this study, however to demonstrate the change acceleration boost makes on the capacity factor, no mechanical limits were imposed and so the acceleration/deceleration were set to take place freely at instances of abrupt flow speed variations, decreasing exponentially to zero as the turbine speed reaches its optimum value. The acceleration boost can be implemented in two ways. Initially it is proposed for the K coefficient to be regulated according to the readings from a flow meter. Here a PI controller monitors the difference between the actual and optimum rotor speeds and regulates the K value (4.28) accordingly. This method is accurate and robust however requires an auxiliary flow meter that is a demerit as it brings new complexities and decreases the system reliability. Consequently in the second approach to the problem the need for an auxiliary flow meter is eliminated. As will be seen later on, in this his strategy, the acceleration is increased by introducing a variable time delay that controls the electro-mechanical torque difference, i.e. $T_m - T_e$. The performance of both techniques will be put to test in the coming chapter.

4.6.2.1 Variable K Method

As was explained previously the operating principal behind the conventional MPPT is that if $\omega_r < \omega_{r-opt}$, then $T_m > K_{opt}\omega_r^2$. Consequently, $d\omega_r/dt$ which is proportional to $T_m - T_e$ becomes positive leading to the turbine-generator acceleration, increasing ω_r . On the other hand, if $\omega_r > \omega_{r-opt}$, then $T_m < K_{opt}\omega_r^2$, $d\omega_r/dt$ is negative, and so ω_r decreases. Once the generator speed reaches the optimum value, i.e. $\omega_r = \omega_{r-opt}$ the change in speed stops as $d\omega_r/dt = T_m - T_e = 0$. At this point $T_m = T_e = K_{opt}\omega_{r-opt}^2$. Thus, by commanding T_e to settle at a smaller/greater value than $K_{opt}\omega_r^2$, a greater acceleration/deceleration can be created respectively. In other words by commanding T_e to change to values that are greater/smaller than what the conventional MPPT method proposes, the gap between T_m and T_e is made greater causing a boost in the machine acceleration/deceleration at the instants of flow speed changes.

Figure 4.14: K regulator in acceleration boost MPPT method

Figure 4.14 depicts the proposed variable K regulating system. This controller is designed to operate at the K_{opt} value under normal conditions while adjusting the K value when the tip speed ratio (λ) deviates from its optimal value. As can be seen from the Figure a PI controller is employed to regulate the initial K_{opt} value when the generator speed (ω_r) deviates from its optimum value ($\omega_{r-opt} = \frac{N\lambda}{R} \cdot V$). The threshold block stops unnecessary fluctuations in the K once the speed reaches its optimal value. One must note that an advantage of regulating K instead of a direct torque or current regulation is the considerable reduction in the PI compensator gains.

4.6.2.2 Imposed Time Delay Method

After observing the implemented conventional and variable K , MPPT methods for some time, it was seen that if at any stage the T_{eref} command was delayed the generator would accelerate/decelerate according to the changes happening in T_m until a point where the system would come to a complete standstill or become unstable. However if the delay was made short the system could reach its optimum speed much faster than the conventional MPPT method without becoming unstable. Thus the imposed delay MPPT

was designed in a way such that the T_{eref} command would experience short transient delays at instances where the change in speed was significant. However an acceleration monitoring device replaces the speed meter. The acceleration monitor could initiate the delays once the acceleration passes a predefined threshold. This way the acceleration boost technique could be repeated without the need for an auxiliary flow meter, since most encoders do provide the acceleration as well as the rotation speed and angle. The same encoder that is required for generator speed monitoring for the dq - vector control implementation can be used to monitor the system acceleration as well. The following schematic diagram show the T_{eref} command is being generated in the proposed MMPT method.

Figure 4.15: T_{eref} generating system under the imposed variable time delay method

One must note that though the maximum delay value can be selected arbitrary, the threshold value and the delay clearance time interval are two sensitive values that need to be selected meticulously. Hence while a small threshold results in the machine inappropriate acceleration/deceleration, a high threshold changes the system from its acceleration boost mode into the conventional MPPT type. Furthermore a long delay time can make the system unstable. So while the threshold can be set by checking the system acceleration when operated in the conventional mode, the delay clearance time should be assigned in view of the magnitude of the speed variation and the mechanical limitations of the system. It was found that it is much better to use a smaller delay that lasts longer rather than a bigger delay value that is short.

4.7 Concluding Remarks

In this chapter initially the control mechanism of a variable frequency converter suitable for an induction machine was introduced. Similar to previous chapters a dq -frame based control strategy was employed, enabling separate control of the machine flux and electrical torque in the rotor-field coordinates. Moreover different variable speed generation topologies were explained. It was shown that the system employed in this study, utilizes a full scale converter that acts as the interface between the variable speed/frequency generator and the constant frequency grid. The generator can be selected to be either a SCIG or WRIG coupled to the turbine through a gearbox. Furthermore the concept of maximum power point tracking was explained for propeller type turbines in which the extracted power is a function of the tip speed ratio (λ) and pitch angle (β). It was shown that the major obstacle in tracking speed variations that are rapid, such as those that coastal areas experience, is the time period that it takes the turbine to reach the optimal speed. Thus two new MPPT routines (*Variable K* and *Imposed Delay*) were introduced and compared to a conventional routine (*Constant K*) that is widely practiced in variable speed generation systems. It was seen that the two proposed alternative MPPT methods overcome the shortcoming of the conventional routine when it comes to spontaneous speed variations. However the (*Variable K*) routine had the disadvantage of requiring current speed measurements which if not practically impossible, is technically very challenging to implement, given that the turbine blades do not experience the same current speed at different locations. Hence a third method (*Imposed Delay*) was introduced in which the fast dynamic response of the *Variable K* method was to some extent maintained without the need of actual flow speed measurements. This method relies on a variable imposed delay system that widens the gap between the mechanical and electrical torque at instances of sudden speed change.

Distinction between the channel type and near-shore tidal currents and proposition of accelerator boosting techniques for a more abrupt maximum power point tracking, i.e. *Variable K* and *Imposed Delay* MPPT methods are the main contributions of the author in this chapter.

Chapter 5

Designing Details of the HPP and VFC Controllers

5.1 Introduction

This chapter begins with using the predictive models developed in chapter one for current speed forecasting which will be used in sizing a Hybrid Power Port (HPP). The chapter then continues with detail design descriptions of the HPP and variable frequency converter control systems described in chapters 3 and 4. Consequently based on actual tidal current speed recordings an example HPP system is designed and tested in two scenarios which differ according to their type of reactive power control strategy. Furthermore a variable frequency converter is designed based on the parameters of a Wound Rotor Induction Machine (WRIM). The variable frequency converter is then tested under a flywheel condition in which the machine is controlled under motoring and generating scenarios.

One must note that the combination of the HPP system and the variable frequency converter results in a full scale converter that is employed in the final chapter to evaluate the performance of a multi-machine TISEC under different tidal current variations.

5.2 Site Assessment and HPP Sizing

As was explained in section I of the study, depending on the geographical situation, and the bathymetry of the region, tidal stream velocity can vary in one of three different patterns, the diurnal, semi-diurnal and the mixed current profiles. Out of the aforementioned types the most prevalent tidal flow is the semi-diurnal currents with two floods and two recessions in each lunar day [83]. Figure 5.1 is an example of such a flow showing how the current changes speed and direction in one day. This recording is performed at the Cape May Inlet Entrance at NJ Coast, in Mid Oct 2000 which will be considered as the case example in this section.

As can be seen from the 180° jumps on the direction plot in Figure 5.1, the flow overturns just about every 6 hours. In spite of the change in direction the available power per square meter of this flow pattern is depicted in Figure 5.2. Injection of such power profiles to a non-stiff network which is typical of the electrically remote but energy rich areas of most estuaries and ocean fronts, can lead to bus voltage

variability, a common power quality problem. Thus regardless of the power capturing method employed the main objective in this section is to:

- Maintain a stabilized DC voltage while a variable amount of active/reactive power is being transferred through the power port to the utility network.
- Maintaining the voltage quality of the PCC bus during the power transfer.
- Enabling the system for smooth grid connection/disconnection and autonomous operation

Figure 5.01: Cape May Station, Current Speed Data. 17 Oct 2000[41]

Figure 5.02: Cape May Station Power Density Profile. 17 Oct 2000

Although PCC voltage regulation for other renewables has been dealt with by series/parallel reactive power compensations in the past, most alternatives proposed in the literature employ an auxiliary STATCOM or DVR module [60-62]. As was explained in Section II, marine current, repetitive and predictable profiles however provide opportunities for novel Point of Common Coupling (PCC) voltage regulation schemes that utilize part of the grid-tie converter capacity for reactive power control. Here the prediction algorithms developed in chapter one, are employed for site power density assessment leading to the sizing of the required hybrid power port. A 29 day study of the stream behavior at Cape May station [84] is used to develop the HAMLS model. As mentioned before this time span is enough to capture the most influential astronomical constituent frequencies affecting the tidal flows in that area. The model is then used in conjunction with a NARX neural network to form the combined predicting model which is then trained using 8 months of recorded current speed data. Using the trained hybrid prediction model the current speeds of the region is later estimated over a 5 year planning period. The following surface speed histogram shows the current speed distribution of the site under study during a five years planning period.

Figure5.03: Estimated surface current speed histogram

Based on this Figure, for the TISEC systems to be able to capture the available stream power, at least 97% of the times, they should be designed to withstand speeds of up to 2.15 (m/s). Accordingly assuming a TISEC with 3 turbines interconnected at the DC side, each having a turbine radius of 10m and a total conversion efficiency of 50%, the required power capacity of the HPP is calculated to be

approximately 2.34 MVA. Therefore the power capacity of the HPP in this study is rounded up and assumed to be 2.5 MVA. As mentioned in Section 2, part of this capacity will be used for reactive power regulation.

5.3 Designing Details of The Hybrid Power Port (HPP)

Figure 5.4, shows the simplified schematic diagram of the complete system under study. As can be seen from the diagram a full scale AC/DC/AC converter is used to connect two individual or a cluster of turbine-generator sets to the power grid through a submarine line and a step-up transformer. In the following sections the designing factors pertaining to different parts of the HPP system, i.e. the PLL, 3rd harmonic injector and all corresponding controllers will be described in detail.

Figure 5.04: Schematic diagram of the real/reactive (hybrid) power port system and its grid connection, for the sake of clarity the soft starter module required for DC capacitor initial charging is not included

5.3.1 Designing Details of the Phase Lock Loop (PLL)

Considering the system depicted in Figure 5.4 with $\omega_0 = 2\pi \times 60 \text{ (rad/s)}$ and $\hat{V}_s = 11000 \times \sqrt{2/3} = 8.98 \text{ KV}$, the PLL compensator $H_{PLL}(s)$ is designed as explained in the Appendix. The transfer function $H_{PLL}(s)$ includes a pole at $s = 0$ and two conjugate zeros at $s = \pm j2\omega_0$. Furthermore, two real

poles are included at $s = -2\omega_0$, ensuring the loop gain magnitude decreases with a slope of -40 (dB/dec) after $\omega > 2\omega_0$. Therefore:

$$H_{PLL}(s) = \left(\frac{h}{\widehat{V_{sn}}} \right) \frac{s^2 + (2\omega_0)^2}{s(s + 2\omega_0)^2} P(s) \quad (5.1)$$

Here $\widehat{V_{sn}}$ is $\widehat{V_s}$ nominal value and $P(s)$ is a transfer function that has no zeros at $s = 0$. Based on the control block diagram of Figure App.3 (Appendix II), the loop gain is formulated as:

$$l(s) = h \cdot \frac{s^2 + (2\omega_0)^2}{s^2(s + 2\omega_0)^2} \cdot P(s) \quad (5.2)$$

Assuming a gain crossover frequency of $\omega_c = 200 \text{ (rad/s)}$ and a phase margin of 60° , if $h \cdot P(s) = 1$, the phase angle of $l(s)$ is calculated as:

```
>> w0 = 2*pi*60;
>> L = tf([1 0 (2*w0)^2],[1 4*w0 (2*w0)^2 00]);
>> bode(L);
>> [Mag, Phase] = bode(L, 200)
Mag =
    2.1713e-005
Phase =
   -209.7121
```

A lead compensator provides the required phase advance to the loop gain. Here the phase margin is chosen to be 90° , and so $P(s)$ is selected to be two cascade lead filters each providing a phase margin of $\sigma_m = 45^\circ$ at $\omega_c = 200 \text{ (rad/s)}$:

$$\begin{cases} P(s) = \left(\frac{s + (p/\alpha)}{s + p} \right)^2 \\ p = \omega_c \sqrt{\alpha} \\ \alpha = \frac{1 + \sin \sigma_m}{1 - \sin \sigma_m} \end{cases} \quad (5.3)$$

Hence:

$$\sigma_m = 45^\circ \Rightarrow F(s) = \left(\frac{s + 83}{s + 482} \right)^2$$

$$\Rightarrow l(s) = \frac{h \cdot (s^2 + 568516)(s^2 + 166s + 6889)}{s^2(s^2 + 1508s + 568516)(s^2 + 964s + 232324)} \quad (5.4)$$

```
>> L1 = tf([1 0 (2*w0)^2],[1 4*w0 (2*w0)^2 0 0]);
```

```
>> F = tf([1 83],[1 482])^2;
```

```
>> L2 = L1*F;
```

```
>> [Mag , Phase] = bode (L2 , 200)
```

Mag =

3.7386e-006

Phase =

240.1409

Figure 5.05: PLL open loop Bode plot, i.e. $l(s)$ with $h = 1$

Hence h and consequently $h/\widehat{V}_{sn}$ are calculated as:

$$|l(j200)| = 1 \Rightarrow h \times 3.7386 \times 10^{-6} \Rightarrow h = 2.675 \times 10^5$$

$$\widehat{V}_{sn} = 8.98 \text{ KV} \Rightarrow \frac{h}{\widehat{V}_{sn}} = 29.78 \approx 30 \quad (5.5)$$

Therefore the final compensator has the following transfer function:

$$H_{PLL}(s) = \frac{30 (s^2 + 568516)(s^2 + 166s + 6889)}{s(s^2 + 1508s + 568516)(s^2 + 964s + 232324)} \left(\frac{\text{rad/s}}{V} \right) \quad (5.6)$$

The Bode diagram of $l(j\omega)$ based on the final compensator design of (5.6) is plotted in Figure 5.6. Apparently, although for $\omega \ll \omega_c = 200$, the loop magnitude i.e. $|l(j\omega)|$ drops with a slope of -40 dB/dec , near ω_c it reduces to about -20 dB/dec with its angle i.e. $\angle l(j\omega)$ rising to about 240° which corresponds to a phase margin of 60° at $\omega = \omega_c$. As Figure 5.6 illustrates for $\omega > \omega_c$ the loop magnitude $|l(j\omega)|$ again continues to drop with a slope of -40 dB/dec . This is a favorable characteristic as V_{sq} AC components are reduced in case of V_{sabc} experiences any harmonic distortion. Particularly at $\omega = 6\omega_0$, $|l(j\omega)| \approx -30 \text{ dB}$. In cases where a step-up transformer is used to elevate the voltage the PLL circuit must be modified according to the transformer specifications. Figure 5.7 shows the required modifications compared to the original system of Figure App.4 (Appendix II). As can be seen ρ needs to be advanced according to the vector group of the transformer (e.g. $+\pi/6$ for $Yd1$) while V_{sdq} is calculated by multiplying the output of the $abc \Rightarrow dq$ block by $1/N_t$, where N_t is the transformer voltage (turn) ratio.

Figure 5.06: Bode plot of the compensated open loop PLL, i.e. $l(s)$ with $h = 30$

Figure 5.08: Designed PLL startup response under normal conditions.

Figure 5.09: Response of the designed PLL to disturbances in grid frequency.

According to Equation (5.1) $H(s)$ is normalized so is independent of $\widehat{V}_{sn}$. Therefore, the PLL is robust to any balanced voltage changes as well, however as will be demonstrated unbalanced voltage changes can cause ripple in the frequency and dq voltages. Figure 5.10 illustrates the way the PLL responds to a sudden V_{sabc} voltage imbalance.

Figure 5.10: Designed PLL response to a sudden voltage imbalance in the AC system

As can be observed from the plots of Figure 5.10, in the beginning the PLL is in a steady state. At $t = 0.05s$, the three phase AC system voltage, that is V_{sabc} becomes unbalanced such that a 0.2 p.u. negative sequence fundamental voltage harmonic component is introduced to the system. At $t = 0.15s$, V_{sabc} goes back to its pre-disturbance balanced condition. Figure 5.10 illustrates the response to the voltage distortion and to maintain the DC component of V_{sq} at zero, the $H_{PLL}(s)$ transiently changes ω . The Figure also shows that due to the V_{sabc} negative-sequence component, both V_{sq} and V_{sd} voltages include a 120 Hz sinusoidal ripple that is not completely annihilated by the $H_{PLL}(s)$. However the $H_{PLL}(s)$ suppression of the ripple is enough to keep ω and ρ essentially free of distortion. But this ripple can still impact the performance of other controllers as they require a zero V_{sq} and a fairly constant V_{sd} to operate correctly. So subject to the size and complexity of the voltage imbalance the system is vulnerable when it comes to faults in the upstream AC network.

It's worth mentioning that there are many papers in the literature that focus on design alternatives of fault tolerant systems, here however the adapted strategy is to disconnect the TISEC once the grid encounters such problems and entre an autonomous operating condition that will be described later in the chapter.

5.3.2 Designing Details of the Active Power Controllers

As was explained in Section 2, the active power controller is designed not only to allow for the active power flow in and out of the power port but also to keep the DC bus voltage regulated. The controller is designed based on the block diagram presented in Figure 3.8. In Figure 3.8, $G_v(s)$ is a function of the operating point (Equation 3.26) Therefore, $H_p(s)$ is designed for the worst case scenario which corresponds to the rectification mode operating point, that when P_{Gen} is negative.

As was mentioned in chapter three, τ_i in Equation (3.24) is a design choice and so has been selected here to be $\tau_i = 1$ (ms), hence:

$$\frac{P_s(s)}{P_{sref}(s)} = G_p(s) = G_i(s) = \frac{1}{\tau_i s + 1} = \frac{1000}{s + 1000} \quad (5.7)$$

Based on Equation set (3.16):

$$k_d(s) = k_q(s) = \frac{k_p s + k_i}{s}, \quad \begin{cases} k_p = \frac{L_{eq}}{\tau_i} = \frac{250 \times 10^{-6}}{10^{-3}} = 0.25 \\ k_i = \frac{R_{eq}}{\tau_i} = \frac{6 \times 10^{-3}}{10^{-3}} = 6 \end{cases} \quad (5.8)$$

Equation 5.7 shows that $G_p(s)$ bandwidth is 1000(rad/s). Consequently to avoid excessive phase lag in the control loop of Figure 3.8, ω_c is selected to be considerably smaller than $G_p(s)$ bandwidth. In this study, it's assumed for ω_c to be 200 rad/s.

According to Figure 3.8, the loop gain is:

$$l(s) = -H_p(s).G_p(s).G_v(s) \quad (5.9)$$

Based on (3.26):

$$\begin{cases} G_v(s) = \frac{\widetilde{V_{DC}}^2(s)}{\widetilde{P_s}(s)} = -\left(\frac{2}{C}\right) \frac{\tau s + 1}{s} \\ \tau = \frac{2L_{eq}P_{Gen0}}{3V_{sd}^2} = \frac{2 \times 250 \times (\pm 2.5)}{3 \times \left(480 \times \sqrt{\frac{2}{3}}\right)^2} = \pm 1.063 \text{ (ms)} \end{cases} \quad (5.10)$$

Inclusion of an integral term in $H_P(s)$ ensures a zero steady state error. So $H_P(s)$ is assumed to be:

$$H_P(s) = D(s) \frac{k_0}{s} \quad (5.11)$$

where k_0 is a constant gain, and $D(s)$ is a transfer function without any zeros at $s = 0$. Substituting for $G_v(s)$ and $H_P(s)$ in (5.11), respectively, the following is obtained:

$$l(s) = D(s) \cdot k_0 \left(\frac{2}{C} \right) \frac{\tau s + 1}{s^2(0.001s + 1)} \quad (5.12)$$

Considering the same $\omega_c = 200$, as before and selecting a capacitor size of $C = 9625 \mu F$, if $D(s) = 1$, then $k_0 = 192.05$, yields $|l(j200)| \cong 1$

```
>>L = tf( [0.220207.79],[0.001 1 0 0]);
>> [Mag, Phase] = bode(L,200)

Mag =
0.0052
Phase =
-179.3540
>> 1/Mag
ans =
192.0558
```

Therefore, the loop gain without any compensation becomes:

$$l_{un-cmp}(s) = 39907 \frac{\tau s + 1}{s^2(0.001s + 1)} \quad (5.13)$$

Figure 5.11 illustrates the magnitude and phase plots of the uncompensated loop gain, for three critical situations where $P_{Gen0} = 2.5 MW$, $P_{Gen0} = 0$, and $P_{Gen0} = -2.5 MW$. Figure 5.11 shows that all three operating points have similar uncompensated loop gain magnitudes, i.e. $|l_{un-cmp}(j200)| = 1$. The loop angles on the other hand have different values, namely: $\angle l_{un-cmp}(j200)$ is -179.3° , -191.3° , and $+156.7 \equiv -203.3$, corresponding to $P_{Gen0} = +2.5$, 0 , and $-2.5 MW$, respectively.

```

>>L1 = tf ( [39907.7*1.063e-3 39907.7] , [0.001 1 0 0] ); % PGen0 = +2.5MW
>>L0 = tf ( [0 39907.7] , [0.001 1 0 0] ); % PGen0 = 0 MW
>>L2 = tf ( [-39907.7*1.063e-3 39907.7] , [0.001 1 0 0] ); % PGen0 = -2.5MW

>> [Mag , Phase] = bode(L1,200)
Mag =
    1.0002
Phase =
   -179.3076

>> [Mag , Phase] = bode(L0,200)
Mag =
    0.9783
Phase =
   -191.3099

>> [Mag , Phase] = bode(L2,200)
Mag =
    1.0002
Phase =
   156.6877

```

Therefore, it is seen that the closed-loop system is poorly stable for $P_{Gen0} = +2.5 \text{ MW}$, and unstable for $P_{Gen0} = 0$ and $P_{Gen0} = -2.5 \text{ MW}$. To ensure a stable closed-loop system under all operating points, $\angle l_{unc-mp}(j200)$ is corrected by letting $D(s)$ in (5.12) to be the lead filter:

$$D(s) = n_0 \frac{s + (\frac{p}{\alpha})}{s + p} \quad (5.14)$$

where, p represents its pole, n_0 is the filter gain and $\alpha (> 1)$ is a real constant. The phase provided by this lead compensator is given by:

$$\sigma_m = \sin^{-1} \left(\frac{\alpha - 1}{\alpha + 1} \right) \quad \text{at} \quad \omega_m = \frac{p}{\sqrt{\alpha}} (= \omega_c) \quad (5.15)$$

A phase margin of, 45° is selected for the worst case scenario, i.e. $P_{Gen0} = -2.5 \text{ MW}$, hence $\angle D(j200)$ is required to be $|-204 - (-180 + 45)| = 69^\circ$. Solving for α, p and n_0 , with $\sigma_m = 69^\circ$, $\omega_m = 200 \text{ (rad/s)}$ and $|D(j200)| = 1$, the following is obtained:

$$D(s) = 5.43 \frac{s + 29.11}{s + 1079.07} \quad (5.16)$$

```
>>L = tf ( [1 29.11] , [1 1079.07]);
>> [Mag , Phase] = bode(L,200)

Mag =
    0.1842

>> 1/Mag
ans =
    5.4300
```

Substituting for $D(s)$ in (5.11) and (5.12), from (5.16), $l(s)$ can be rewritten as (5.17), where the subscript “*cmp*” stand for compensated:

$$l_{cmp}(s) = 216695 \times \left(\frac{s + 29.11}{s + 1079.07} \right) \times \left(\frac{\tau s + 1}{s^2(0.001s + 1)} \right) \quad (5.17)$$

and so $H_P(s)$ would be:

$$H_P(s) = 1042.83 \times \frac{s + 29.11}{s(s + 1079.07)} [\Omega^{-1}] \quad (5.18)$$

Figure 5.11 illustrates Bode plots of the compensated as well as the uncompensated loop gains. As before the graphs are plotted for $P_{Gen0} = +2.5$, $P_{Gen0} = 0$, and $P_{Gen0} = -2.5$ MW. Figure 5.11 illustrates that $|l_{cmp}(j200)| = 1$ for all three operating points. In addition, $\angle l_{cmp}(j200)$ are shown to be equal to -108.09° , -120.09° , and -132.09° , corresponding to $P_{Gen0} = +2.5$, 0 , and -2.5 MW, respectively. Hence under all three operating points the closed loop system is stable with phase margins that range from 45° to 81° .

Figure 5.011: Bode plots of the compensated and uncompensated open loop gains

```

>>% PGen0 = +2.5MW
>>L1 = tf( 216695 * [ 1 29.11 ] , [1 1079.07]) * tf( [1.063e-3 1] , [0.001 1 0 0] );
>>
>>% PGen0 = 0MW
>>L0 = tf( 216695 * [ 1 29.11 ] , [1 1079.07]) * tf( [0 1] , [0.001 1 0 0] );
>>
>>% PGen0 = -2.5MW
>>L2 = tf( 216695 * [ 1 29.11 ] , [1 1079.07]) * tf( [-1.063e-3 1] , [0.001 1 0 0] );
>>
>>[Mag , Phase] = bode(L1,200)
Mag =
1.0002
Phase =
-108.0891
>>[Mag , Phase] = bode(L0,200)
Mag =
0.9783
Phase =
-120.0915
>>[Mag , Phase] = bode(L2,200)
Mag =
1.0002
Phase =
227.9061

```

5.3.2.1 Designing Details of the DC Voltage Equalizing System

Considering the voltage converter system of Figure 5.4 that utilizes the 3-level NPC converter of Figure 2.1 and adopts the 3rd harmonic injection PWM strategy and given the following system parameters:

$$\begin{cases} P_{s0} = 2.5MW \\ 2C = 19250 \mu F \\ V_{DC} = 2500 V \\ \hat{V}_t \approx 391 V \\ \hat{m} = \frac{2\hat{V}_t}{V_{DC}} \approx 0.31 \\ \omega_0 = 377 \text{ rad/s} \end{cases}$$

the plant transfer function, in the closed-loop system of Figure 2.6, becomes:

$$G(s) = \left(\frac{4}{\pi C V_{DC} \hat{m}} \right) \frac{1}{s} = \frac{0.169}{s} = \frac{169}{s} [(kA)^{-1}] \quad (5.19)$$

Let $F(s)$ be:

$$F(s) = \frac{s^2 + (3\omega_0)^2}{(s + 3\omega_0)^2} = \frac{s^2 + 1131^2}{s^2 + 2262s + 1131^2} \quad (5.20)$$

Consequently the third-harmonic component of capacitor DC voltages i.e. $V_1 - V_2$ is suppressed, $F(s)$ has a DC gain that equals one, and the loop gain decreases for $\omega > 3\omega_0$. If the compensator is assumed to be a pure gain the voltage equalizer system, that is $K(s) = k$ and $|P_s| = 2.5 \text{ MW}$, then the loop gain would become:

$$l(s) = K(s).G(s).F(s).|P_s| = (423k) \frac{s^2 + 1131^2}{s(s^2 + 2262s + 1131^2)} \quad (5.21)$$

Figure 7.12 depicts the corresponding bode diagram for this open loop system.

```
>> L = tf ( 423 * [1 0 1131^2] , [1 2262 1131^2 0] );          % Pso = +2.5MW
>> bode(L,{10,1e5});
>> [Mag , Phase ] = bode(L,200);
>> 1/Mag
ans =
0.5033
```

Figure 5.012: Bode plot of the open loop gain with $k = 1$

For $\omega \leq \omega_0/10$, remains more or less constant at -90° , equivalent to a phase margin of 90° . For $\omega > \omega_0/10$, $\angle l(j\omega)$ drops with a slope of nearly $-90^\circ/\text{dec}$, which is mostly due to the two real pole present at $\omega = \omega_0$. Consequently if the phase margin is required to be 70° , then the loop gain crossover frequency ω_c , needs to be larger than $\omega_0/10$, that is $\omega_c = 200 \text{ rad/s}$. Substituting for $\omega_c = 200 \text{ rad/s}$ in equation $|l(j\omega_c)| = 1$, we find $k = 0.503 \text{ (kV)}^{-1}$, for which the exact phase margin value turns out to be equal to 69.9° . The Bode diagram of the new loop gain is plotted in Figure 5.13 in which the approximate bandwidth of the closed-loop system is seen to be about $\omega_b = 400 \text{ (rad/s)}$.

Figure 5.13: Bode plot of the open loop gain with $k = 0.503 \times 10^{-3}$

5.3.2.2 Testing the Active Power Controller and DC Voltage Equalizer

The power port converter system of Figure 5.14 is configured so that initially $P_{Gen} = 0$, converter switching signals are blocked, and the controllers are inactive. Since the focus is on DC voltage regulation, the network is assumed to be stiff and so the LV side of the transformer is modeled as an ideal three phase voltage source, leading to $V_{sabc} = 480 \angle 0$. The DC side capacitors are charged through the anti-parallel diodes of converter switch cells and a soft starting resistor (not shown in the Figure) of $R_{ss} = 0.5 \Omega$, to about 800V. At $t = 0.2s$, switching signals are unblocked, the soft starting resistor is bypassed and all controllers are activated.

Figure 5.014: Schematic diagram of the real/reactive (hybrid) power port system and its grid connection

At this stage V_{DCref} is linearly increased to its nominal value of 2500V. As can be seen from Figure 5.15 in order to move V_{DC} up to its reference value, the controller commands a negative P_{sref} to import real power from the AC system to the DC side. At about $t = 0.35s$, V_{DC} is regulated at its reference value of 2500V. After that P_{sref} and P assume small values corresponding to the converter power loss. At $t = 0.5s$, P_{Gen} changes stepwise from 0 to 2.0MW; this causes a small over-shoot in V_{DC} . The compensator reacts to this disturbance and increases P_{ref} to bring V_{DC} back to 2500V. At $t = 1.25s$, P_{Gen} changes stepwise from 2.0 to -1.5MW, consequently, V_{DC} undergoes an under-shoot until the compensator reacts and reduces P_{sref} . At $t = 2.0s$, P_{Gen} changes from -1.5 to 0MW, consequently, V_{DC} undergoes another over-shoot until the compensator reacts and increases P_{sref} . It should be noted that the pattern of the under-shoot at $t = 1.25s$ is different from that of the overshoot at $t = 2.0s$ due to the different phase margins (and frequency responses) for these two operating points.

On the middle plot, the transferred reactive power is plotted. The reactive power reference signal (Q_{sref}) is commanded to change stepwise at $t(sec) = [0.75, 1.0, 1.5, 1.75, 2.0, 2.25]$ to $Q_{sref}(MVar) = [1.0, -1.0, 1.0, 0.0, 1.0, 0.0]$. These changes are designed so that all PQ combinations are tested. As can be seen from the results, despite the dramatic changes in the active power, the H_P controller manages to regulate the DC voltage quite effectively and so the V_{DC} level is extensively maintained regardless of

the P and Q changes. At the same time it is observed that, P_s and Q_s have very little influence on each other and hence can be controlled quite independently. This is mainly due the correct operation of the PLL which keeps its V_q output zero at all times. Figure 5.6 shows the PLL generated dq voltage outputs as well as its trace profile of the grid frequency. As can be seen from the Figure after the short transient period at the beginning for the rest of the simulation time V_q is kept at zero while V_d tracks the grid voltage. To show the performance of the DC voltage equalizer system, the DC voltage difference (ΔV_{DC}) as well as the different modulation index constituents are depicted in Figure 5.17. As can be seen from (ΔV_{DC}) plot in this Figure the voltage equalizer renders the capacitor voltage balancing quite effectively, and so the ΔV_{DC} doesn't extend beyond 50V which is less than 2% of the nominal voltage (2500V).

Figure 5.15: Dynamic performance of the HPP in controlling the DC voltage

Figure 5.016: Grid three phase voltages as well as the PLL response

Figure 5.017: Capacitors DC voltage difference, modulating signals and the corrective modulation offset signals

Other plots in Figure 5.17 show the modulating signals (m_{abc}), the DC offset value (m_0) added to the modulating signals in order to balance the two partial DC voltages and the 3rd harmonic injected modulating signals ($m_{aug-abc} = m_{abc} + 3^{rd} Hrm$) while the active and reactive power profiles shown in Figure 5.15 were being transferred to the grid. The m_0 plot in Figure 5.17 shows how the closed loop controller of Figure 2.6 regulates this offset value to maintain a balanced V_1 and V_2 voltage values. The $\Delta V_{DC} = V_1 - V_2$ plot shows that although the difference between the two DC voltages is limited their difference changes according to the operating condition. This is mainly due to the 3rd harmonic component of the converter midpoint current, which increases as the converter terminal power factor decreases; causing the third-harmonic ripples of the two voltages and their difference ΔV_{DC} to increase when the 3-level NPC starts to exchange reactive power.

5.3.3 Reactive Power Controllers

As was explained in chapter three, the compensator in the PCC voltage control loop can be designed according to the simplified model in which the PLL, harmonic filter and load dynamics are ignored. Consequently, the high order transfer function describing the plant model is approximated by a constant gain (Figure 3.12). In this study, the PCC bus is assumed to be without any direct load connections and so is assumed to be merely as an interface between the converter and the grid network. Therefore based on the block diagram of Figure 3.12, the open loop transfer function is:

$$l(s) = H_Q(s) \cdot \frac{2}{3V_{sd0}} G_i(s) \cdot \underbrace{(L_g \omega_0)}_{X_s} \quad (5.22)$$

Since the network considered for this study is an 11 KV medium voltage network with a short circuit level of 5 MVA and a $X/R = 8$. Hence:

$$\begin{cases} V_{sd0} = 8.98 \text{ KV} \\ X_s = 24.01 \Omega \end{cases} \quad (5.23)$$

As mentioned before, $\tau_i = 1.0 \text{ ms}$ and so $G_i(s)$ is:

$$G_i(s) = \frac{1}{\tau_i s + 1} = \frac{1000}{s + 1000} \quad (5.24)$$

In view of (5.22) – (5.24), the compensator of the reactive power controller is designed to have the following transfer function:

$$H_Q(s) = \frac{5000}{s} [KA] \quad (5.25)$$

This results in an $\omega_c = 89 \text{ (rad/s)}$, and a phase margin of about 85° . It should be noted that the crossover frequency of $\omega_c = 89 \text{ (rad/s)}$ is approximately 11 times smaller than $G_i(s)$ bandwidth, i.e. $1/\tau_i = 1000 \text{ (rad/s)}$, which means that the loop gain is essentially an integrator and the closed-loop transfer function for the most part is a first order with a time constant that is equal to $1/\omega_c \cong 11 \text{ ms}$. By selecting a relatively large phase margin, the closed loop system stability in the presence of unmodeled dynamics and parameter uncertainties can to some degree be assured. Bearing in mind that the variable grid inductance is not precisely known and the dynamics of the PLL, filters and any local loads present were ignored. The plot of the resultant closed-loop system is shown in Figure 5.18. As will be seen later, this system is able to regulate the PCC voltage quite fast and efficiently.

Although the input to the reactive power controller presented in Figure 3.12 was V_{sdref} , to realize a more pragmatic approach here the input is selected to be the difference between the *rms* values of the PCC and grid positive sequence voltages, i.e. $\left(\frac{\sqrt{3}}{3} |V_a + aV_b + a^2V_c|\right)$, in which V_a, V_b and V_c represent the phase to ground voltages, Figure 5.4. One should note that since the PCC bus voltage regulation is performed with respect to V_{Grid} , this reference voltage needsto be obtained from a dispatching center.

Figure 5.018: Bode diagram of the closed-loop reactive power control system

5.3.3.1 Testing the Reactive Power Controlling System

To examine the proposed reactive power controller $H_Q(s)$, two scenarios are examined where the converter has to transfer a certain amount of active power to the grid, under initially a leading and then a controlled power factor regime. The changes in the active power are designed to resemble that of a semi-diurnal tidal stream. In the first example the active power transfer takes place under a reactive power injection strategy in which Q_s varies in proportion to P_s , resulting in a leading but constant power factor. One must note that the leading power factor scheme is quite common in the wind industry and is performed to achieve satisfactory voltage profiles in the vicinity of the wind farms [62]. In the second scenario the reactive power is actively controlled to maintain the PCC bus voltage at its rated value. In both scenarios the active power is transferred to a non-stiff network through underwater cables and a step-up transformer, Figure 5.19. Since the focus here is the effect that active and reactive power injection have on the PCC bus voltage the electric network is assumed to be quite weak. Thus although the standard short circuit level of an 11 KV network is 200 MW[85], the network employed in this study, is assumed to have a short circuit level of only 5 MW, indicating a much weaker network than the average. As will be seen this intentionally assigned small short circuit level can better highlight the effectiveness of the proposed HPP system even under operation in such exceptionally weak networks.

Figure 5.19: Schematic diagram of the real/reactive (hybrid) power port system and its grid connection

Case 1- Voltage Deviation Under A Fixed Power Factor Scheme

In this example active power transfer takes place under a constant power factor that is very near unity, i.e. ($PF = 0.99$ lead). Since it is common to express a generating systems reactive power status in terms of the power factor, the required reactive power to achieve this power factor can be defined as a function of the active power by:

$$\begin{cases} Q_s = k \cdot P_s \\ k = \frac{\sqrt{1 - PF^2}}{PF} \end{cases} \quad (5.26)$$

For a leading but very near unity power factor of $PF = 0.99$, $k = +0.142$; so in this case Q_{sref} is set to be 14.2% of P_{sref} and thus Q_s changes according to P_s . Figure 20 shows P_s , Q_s and V_{DC} plots in this scenario. The initial starting condition for the power port system is designed to be the same in both cases, where initially $P_{Gen} = 0$, converter switching signals are blocked, and the controllers are inactive, during this period the DC-side capacitors are charged from zero to about 700V via the antiparallel diodes of the switch cells and through a soft starting resistor. The soft starter limits the capacitor inrush current and is automatically bypassed at $t = 0.5s$ once the converter switching signals are unblocked and controllers are activated. Consecutively, at this stage V_{DCref} is raised linearly from 700 to its nominal value of 2500V. To increase the DC voltage the active power controller, i.e. $H_p(s)$ commands a negative P_{sref} , importing real power from the AC system to the converter DC side. At about $t = 0.7s$, V_{DC} is regulated at 2500V. After this initialization stage, as long as the input power to the converter (i.e. P_{Gen}) is zero, P_{sref} and P_s assume small values that correspond to the system power losses.

At $t = 1.0s$, P_{Gen} starts to increase reaching its maximum of 1.6MW at $t = 1.75s$, after this the input power decreases to nearly zero but rises again at $t = 2.5s$. In the second drift the input power reaches a maximum value of 2.0 MW at $t = 3.25s$. As expected the plot of the reactive power is also seen to follow the same trend as the active power profile, which is due to the fixed leading power factor setting. In general the power plots of Figure 5.20 show that the grid transferred powers follow their references quite effectively and a well maintained V_{DC} profile regardless of the changes happening in the active and reactive power profiles. However the PCC bus voltage profile and the ΔV plot are not very promising. As can be seen from Figure 5.21 the PCC bus voltage increases at the time of power injection intervals by 10% which is enough to be considered as a power quality problem known as voltage swell.

Figure 5.020: Active (P_s) and reactive (Q_s) powers transferred to the grid as well as the DC voltage under the fixed power factor scheme. PF=0.99

Figure 5.021: Voltage deviation $\left(\% \Delta V = 100 \times \frac{V_{PCC} - V_{Grid}}{V_{Grid}}\right)$ under the fixed power factor scheme. PF=0.99

Case 2- Voltage Deviation under a Voltage Regulation Scheme

This scenario shows the performance of the same system as in the previous example, with the reactive power controller $H_Q(s)$ enabled. Figures 5.22 and 5.23 show the power plots as well as the DC and AC voltage profiles. It is seen that unlike before by using a small portion of the converter power capability for reactive power control the ΔV values here are damped to less than 2%. As can be seen from the plots,

while the active power profile is the same as before, Q_s is actively controlled to compensate for the voltage changes that are imposed by the active power injection. It is also observed that unlike the previous case, here Q_s and P_s power profiles moves in opposite directions, which infers the way H_Q controller manages the voltage distortions that are caused by the injection or draw of active power.

Figure 5.022: Active (P_s) and reactive (Q_s) powers transferred to the grid as well as the DC voltage under the Voltage Regulating (VR) scheme

Figure 5.023: Voltage deviation, $\left(\% \Delta V = 100 \times \frac{V_{PCC} - V_{Grid}}{V_{Grid}}\right)$ under the Voltage Regulating (VR) scheme.

The two cases above, clearly demonstrate the effectiveness of the proposed active voltage regulation by the grid-tie converter and the successful operation of the reactive power controller H_Q .

5.3.4 Designing the Autonomous Mode, Controllers

Considering the HPP system of Figure 4.6 in which:

$$\left\{ \begin{array}{l} C_f = 2500 \mu F \\ L = 50 \mu H \\ R + r_{on} = 3 m\Omega \\ V_{DC} = 2500 \\ f_s = 1680 Hz \\ \tau_i = 1.0 ms \\ \omega_0 = 377 rad/s \\ (V_{sdref}, V_{sqref}) = (400, 0) \\ RL Load connected as Yn \begin{cases} R_{Load} = 83 \Omega \\ L_{Load} = 137 \mu H \end{cases} \end{array} \right. \quad (5.27)$$

Although the final system will not supply any local loads, to verify the control designs of Section II, as can be seen from the list of parameters in (5.27), a local RL load is employed. While the supervisory control scheme is the same as what was explained in chapter four, the voltage regulation loop compensator is designed according to the provided system parameters and a phase margin of 55° , thus base on (4.9), one finds $z = 100 (rad/s)$. It then follows from (4.9) and (4.10) that $\omega_c = 316 (rad/s)$ and $k_0 = 0.8$. Thus:

$$K(s) = 0.8 \left(\frac{s + 100}{s} \right) \quad (5.28)$$

Figure 5.34 illustrates the loop gain Bode plots of the off-grid mode compensator. As Figure 5.24 illustrates, the phase margin at $\omega_c = 316 (rad/s)$ is about 55° .

Figure 5.24: Open loop frequency response of the voltage regulator in autonomous operating mode

```
>>L1 = tf ( 0.8 * [1 100] , [1 0] ) * tf( [1000] , [1 1000]) * tf( [1] , [2500e-6 0]);
>>bode( L1 , {10,1e5} );
>>margin(L1)
```

5.3.4.1 Testing the HPP under Autonomous Operating Mode

Considering the system of Figure 5.25 with the parameters provided in (5.27) and (5.28), three scenarios are designed in which the HPP is tested for appropriate autonomous mode voltage and frequency management, correct and smooth grid connection and finally grid fault detection and separation. Since the focus of this part is primarily on the HPP supervisory control, the network is assumed ideal having a Thevenin equivalent voltage of $480 V_{rms}$ which means that no step-up transformer is required. The DC side voltage is also assumed to be ideal inferring that it is imposed through an ideal DC voltage source.

Figure 5.25: Schematic diagram of the HPP and its supervisory controlling system for grid connected and autonomous operation modes. The grid in this case is assumed to be ideal

Scenario 1

This scenario is designed so that the system is tested during its autonomous operation. Here initially the converter switching signals are blocked, all controllers are inactive and the *Mode Command* is kept at zero, resulting in the main switch remaining open and the HPP, *dq*-reference currents being governed

by $(V_{sd_{ref}}, V_{sq_{ref}}) = (391, 0)$. At the same time the operating frequency is determined through the frequency reference that is kept constant at $\omega_0 = 377 \text{ (rad/s)}$. Initially the system is started without any loads connected. At $t = 0.01\text{s}$, switching signals are unblocked, controllers are enabled and references are ramped to their predefined values. It is seen that by $t = 0.05\text{s}$ the V_{sd} voltage reaches its reference value of 391V. To test the voltage controller, at $t = 0.15\text{s}$, $V_{sd_{ref}}$ is commanded to increase to 550V, and back to 391V at $t = 0.25\text{s}$. Figure 5.26 shows the system response to this command. To demonstrate the system independence from the load dynamics, as was stated in chapter 4, at $t = 0.35\text{s}$ the load is switched on which although causes a sudden drop in the V_{sdq} voltages the system manages to regulate them back to their reference values and so by $t = 0.40\text{s}$, $(V_{sd}, V_{sq}) = (391\text{V}, 0)$. At $t = 0.45\text{s}$ the load is dropped off which as can be seen from the three phase and dq -axis currents, results in an even greater transient. Despite this the results demonstrate that the system is able to damp these transients in less than three 60Hz cycles. Hence while islanded from the network, the HPP is able to regulate and control the voltage and frequency regardless of the presence or type of any local loads.

Figure 5.26: Dynamic response of the HPP in autonomous operation mode to changes in reference voltage commands as well as sudden load shifts

Scenario 2

This scenario is designed so that the system is tested during initial grid connection. Similar to the previous case initially the converter switching signals are blocked, all controllers are inactive and the main switch is open inferring operation of the HPP system in the autonomous (controlled frequency) mod for which $(V_{sd_{ref}}, V_{sq_{ref}}) = (391V, 0)$ and $\omega_0 = 377 \text{ (rad/s)}$. At $t = 0.01s$, switching signals are unblocked, controllers are enabled and references are ramped to their predefined values. It is seen that by $t = 0.05s$ the load voltage reaches its reference value of 391V, however the system still remains in the autonomous mode due to the transient state of the PLL frequency. By the time the PLL regulates V_{gq} to zero and its track of the grid frequency reaches steady state, the VCO frequency input is switched from the constant reference $\omega_s = \omega_0 = 377$ to ω_g , hence the system frequency is increases by $\Delta\omega$ to reach the grid frequency. As the phase (V_a) voltage plots demonstrate, at this stage onwards the load and grid voltages become synchronized. One must note that at this stage the main switch is still open. Once all grid connection criteria (Figure 4.5) are met the “Grid Connection” flag goes up and consequently a short while later, i.e. at $t = 0.175s$, the main switch is closed and grid connection is established. At the same time the HPP supervisory control system switches the operation mode from *controlled frequency* to *grid-imposed frequency* mode which means that $i_{dq_{ref}}$ are now defined by the reference $P&Q$ powers.

Figure 5.027: HPP responses, voltage and frequency values prior and after the initial grid connection

Furthermore to this scenario Figure 5.28 is provided that depicts the HPP and the load dq -currents, i.e. i_{dq} and i_{Ldq} . As can be seen from this figure while prior to grid connection the HPP is providing most of the required load current, once the grid connection is established, although the load current stays constant the HPP current output reduces to zero. This is due to the fact that after the grid connection the system operation mode is changed and so the HPP is controlled with regards to the active and reactive power references, i.e. (P_{sref}, Q_{sref}) , which are both zero. From this point onward the required load current is provided by the utility network. Another property worth mentioning is the removal of the high frequency components which are perceivable in the HPP output current but not on the load current. As explained in chapter 4 the presence of C_f in combination with the line RL components creates a low pass filter that helps removing the high frequency hash.

Figure 5.028: Load and HPP dq -currents, in autonomous and grid connected operation modes

Scenario 3

This scenario is designed so that the system is tested for load voltage protection during grid faults, such as significant drops in its magnitude. Here the system starts while being under load while the rest of the initial conditions are similar to what was explained before. Hence by $t = 0.05$ s the load voltage reaches its reference value of 391V, and by the time all grid connection criteria are met the “Grid Connection”

flag goes up. Consequently a short while later, i.e. at $t = 0.175s$, the main switch is closed and grid connection is established, Figure 5.29. At this stage the system is working in the *grid-imposed frequency mode*. At $t = 0.30s$ the grid undergoes 30% voltage sag that is sensed by the HPP supervisory controller and a short while later, the mode command is switched to *autonomous mode operation*. The time delay between fault detection and mode command shift is predefined in the supervisory controlling system and is enforced to prevent the breaker from multiple switching. As can be seen from the results while in the autonomous mode, the voltage controller is able to regulate the load voltage in less than one cycle; thus V_s is restored to the reference value of 391V at $t = 0.33s$. At $t = 0.40s$ the grid recovers and a short while later once all grid connection criteria are met again, the main switch is reclosed.

Figure 5.29: HPP responses, voltage and frequency values before and after the grid voltage disturbance

5.4 Design Details of the Variable Frequency Converter

Here the variable frequency converter introduced in chapter four and its affiliated compensators are designed and tested with respect to the machines parameters. The machine considered for this study is a Wound-Rotor, Asynchronous Generator (WRAG) whose rotor terminals are short circuited. The generator is then considered to be connected to the tidal turbine through a fixed ratio gearbox. Depending

on the length of the DC bus and status of the neutral points (Figure 2.3), DC voltage balancing can be performed by either one or each of the converters individually. In this study, it is assumed that neutral points are not connected and so each NPC converter performs DC voltage balancing independently. Furthermore, in this section instead of P_s, P_{DC} will be used in the DC voltage equalizing system, Figure 2.6. Figure 5.40 depicts the asynchronous generator – turbine system along with its Variable Frequency Converter (VFC).

Figure 5.30: Variable frequency converter and controlling system topology

Table 5.1 lists the characteristics of the asynchronous machine employed in this study. A time constant of $\tau_i = 3.0 \text{ ms}$ is selected for the current controllers shown in Figure 5.4, hence based on Equation (5.13) and the machine parameters given above, the following parameters are obtained:

$$\begin{cases} k_p = \frac{\sigma \tau_s}{\tau_i} = 13.6 \\ k_i = \frac{1}{\tau_i} = 333 \text{ s}^{-1} \end{cases} \quad (5.29)$$

Table 5.1: Asynchronous Machine Parameters [20]

<i>Nominal Power</i>	2250 hp (1.678 MW)
<i>Nominal Voltage</i> (V_{sn})	2300 V (line to line rms)
<i>Nominal Frequency</i> (ω_0)	377 rad/s
R_s	29 m Ω
R_r	22 m Ω
L_m	34.6 mH
L_s	35.2 mH
L_r	35.2 mH
σ_s	0.0173
σ_r	0.0173
σ	0.0337
τ_s	1.213 s
τ_r	1.6 s

The flux controller transfer function (Figure 5.3) is selected as:

$$H_\lambda(s) = \frac{180(s + 0.625)}{s} \quad (5.30)$$

which is based on canceling the most unstable pole pertaining to the machine dynamics and also the internal-model principle. Based on Equation 5.5, and machine parameters, $\hat{I}_{m_{ref}}$ is set to its nominal value of 141A.

5.4.1 Variable Frequency Converter Testing

Here the variable frequency converter is tested for both motor and generator operation capability. Hence the test is designed so that the asynchronous machine is initially energized without any torque demand. After the machine flux build up stage, the electrical torque is controlled so that the machine starts off as a free accelerating motor, hence the converter commands a temporary positive torque pulse that is enough to steadily increase the rotor speed. A short while after T_e reference is decreased back to zero and so the machine stops accelerating, maintain a steady speed. At this stage there is enough momentum built in the rotor for regeneration. Thus the converter is commanded to reverse the electrical torque reference, leading to the extraction of the machine kinetic energy; in other words a flywheel system is formed. Figure 5.31 illustrate the response of the machine to the variable frequency converter torque and flux references.

As can be seen from the plots of the Figure, initially the machine is stalled; all controllers are disabled, and the PWM switching signals are blocked. At $t = 0.1s$, the controllers are enabled and the switching signals are released. Thus, to regulate $\hat{I}_m$, the flux compensator steps up I_{dref} , that is rapidly tracked by I_d . Figure 5.31 also shows that I_{dref} is limited to 1500A which is to protect the machine and the converter from over-current. The plots show that while $I_d = 1500A$, $\hat{I}_m$ rises toward its reference value at a fairly low speed which is due to the large rotor time constant. At $t = 0.263s$, $\hat{I}_m$ reaches its reference value, and thus the flux controller reduces I_{dref} which is tracked by I_d . Consequently eventually $\hat{I}_m$, I_{dref} , and I_d become equal (=141A) in the steady state.

Figure 5.031: Response of the WRIG variable-frequency converter system to free acceleration reference commands.

Since T_{eref} (and thus I_{qref}) is kept zero while the machine flux is building up, the machine torque T_e and rotor speed remain zero until $t = 0.5s$ at which T_{eref} is increased step wise to 40KNm. The plots of Figure 5.31 show how T_e and I_q track their reference values. These results also confirm that similar to T_e , I_q also changes as a first-order exponential function; reaching its steady-state value in $5\tau_i$ that is about 15 ms.

Figure 5.032: Waveforms of the stator three phase voltages and currents

Figure 5.033: Waveforms of the stator and rotor frequencies (ω_s , ω_r) as well as the dq -frame synchronization angle

Figures 5.32 and 5.33 illustrate the generators terminal voltage (V_{abc}), currents (I_{abc}), the output frequency of the converter system (ω_s) and the angle for abc to dq -frame transformation (ρ_2). Figure 5.33 illustrates that from $t = 0.1$ to $0.5s$, while ω_r and I_q are both zero, ω_s is also zero and ρ remains unchanged. Therefore as illustrated by Figure 5.32, the modulating signal and the stator current are DC waveforms. One must note that since the machine is being operated without any load or prime mover $T_L = 0$. Consequently once T_e becomes nonzero, the non-zero torque difference causes the machine to start rotating and so ω_r increases linearly with time. As can be seen from Figure 5.33, the slope at which the speed increases, i.e. from $t = 0.5$ to $0.6s$ is twice that of its deceleration at $t = 0.8$ to $1.0s$. The reason for this is the magnitude of the T_{eref} signal at these two intervals which as can be seen from the T_e plot in Figure 5.31 has halved from $t = 0.8$ to $1.0s$. Hence it takes almost twice as much time for the speed to decrease to zero.

The I_{abc} plot in Figure 5.32 also shows that from $t = 0.1$ to $0.4s$, $I_b = I_c = -I_a/2$ which is consistent with $I_q = 0$, since $I_d + jI_q = \vec{I}e^{-j\rho_2}$, $\vec{I} = (2/3)[I_a + e^{j(2\pi/3)}I_b + e^{j(4\pi/3)}I_c]$ and $\rho_2 = 0$. It is interesting to note that, from $t = 0.1$ to $0.5s$ the converter maintains essentially DC currents in the stator windings to regulate the flux. As Figure 5.33 shows, from $t = 0.5s$ onward as I_q and ω_r start to rise, ω_s increases accordingly, noting that since $I_q > 0$, ω_s is slightly larger than ω_r , as understood from Equation (5.8).

As Figure 5.33 illustrates, ρ_2 and its slope increase as ω_s increases; same goes for the frequency of the modulating signal and the stator current. From the same Figure it can be observed that after $t = 0.5s$ the amplitude of I_{abc} is constant whenever the dq -axis currents are both constant. Another point worth mentioning is the change of three phase current sequence that occurs from $t = 0.8s$ onwards. This is due to the machine generating power (generator action) opposed to the previous motoring action that consumed power. The average machine electrical power, $P_e = T_e \cdot \omega_r$, and the power transferred to the DC link, i.e. $P_{DC} = I_{DC}V_{DC}$ are depicted in Figure 5.34. As can be seen from the Figure from $t = 0.1$ to $0.5s$, $P_e = 0$, while P_{DC} assumes a small negative value indicating power consumption. Over that period of time P_e is zero since $T_e = 0$, and P_{DC} is nonzero since $I_{DC} \neq 0$; thus, power is drawn from the DC side to make up for machine flux build up. After $t = 0.5s$, when ω_r increases linearly (Figure 5.33) and T_e is constant, P_e and P_{DC} increase linearly. Due to the resistive losses in the machine stator and rotor, P_{DC} is slightly larger than P_e . These results show that the VFC is working properly and machine torque control can be utilized to manage the power exchange between the mechanical source interfaced with the machine and the entity interfaced with the converter DC side.

Figure 5.034: Converter DC power (P_{DC}) profile as well as the machine electrical power (P_e)

5.5 Concluding Remarks

In this chapter the predictive models developed in chapter one for current speed forecasting was used for sizing a Hybrid Power Port (HPP). Moreover detail design descriptions of the HPP, variable frequency converter control systems described in chapters 3 and 4 were given. The controllers required for an example HPP system was then designed and tested in two scenarios which differ according to their type of reactive power control strategy. Similarly the controllers required for a variable frequency converter was designed based on the parameters of a Wound Rotor Induction Machine (WRIM). The variable frequency converter was then tested under a fly-wheeling condition in which the machine was controlled under motoring and generating scenarios.

The combination of the HPP system and the variable frequency converter results in a full scale converter that is employed in the final chapter to evaluate the performance of a multi-machine TISEC under different tidal current variations.

Performance Evaluation of the Complete System

In this section the complete TISEC system is examined under a tidal shaped speed profile. The TISEC system is composed of three separate turbine generators which are connected at the DC level and share the same power port, Figure 6.1. In practice depending on the span of the installation site and the distance between the machines, the DC voltage varies at different parts of the system, causing formation of reverse or inner-circulating currents which in turn result in the increase of the system losses as well as the risk of facing machines turning into motors. Although the use of dispersed and large capacitors that are supported by the natural distributed capacitance of the submarine line can reduce such voltage differences, utilization of interfacing diode rectifiers (equipped with bypassing systems) for generator, DC line connection is also recommended. Further insight into this issue is outside the scope of this study and is recommended by the author for future studies. Here however the DC voltage is assumed to be consistent throughout the DC line.

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The machines and their variable frequency converters are assumed identical; however each machine utilizes a different MPPT system. Therefore while one of the generators employs the conventional *Constant K* method the other two will utilize the *Variable K* and *Imposed Delay* MPPT techniques. To make the comparisons easier, it is assumed for the three turbines to experience the same tidal flow velocities. Two tidal flow speed variation are considered in this study, that mimic the speed variations of natural channel (with slow and smooth speed variations) and near shore (with very rapid and spontaneous) flow speed profiles. Figure 6.2 shows the two tidal stream speed profiles employed in this study. As can be seen from the Figure, the two speed profiles have similar shapes and resemble a mixed semi-diurnal tidal profile having two high and two low tides of different sizes. Although in reality such speed variations take place under much longer time periods, that is, one lunar day for channel type and a few minutes for near shore flows, due to the software limitation the time spans are scaled down to 100 and 10 seconds respectively. Yet still as will be seen from the obtained results, these time periods are enough to illustrate the advantages and shortcomings of each of the aforementioned MPPT routines and the performance of the complete TISEC when employed in channel or near shore regions.

Figure 6.2: Near-shore and channel type tidal stream speed profiles employed in the study

The utility network in this study is an 11 KV grid with a short circuit level of just 5MW. According to [85] the standard short circuit level of an 11KV network should be 200MW, indicating that the network

under study here is much weaker than the average 11KV network. This small short circuit level is intentionally assumed to illustrate the effectiveness of the proposed reactive power control system even under operation in exceptionally weak grids. Other system characteristics such as machine, transformer or converter parameters are summarized in table 6.1. Based on the speed of the tidal current variations the TISEC system is tested for the power capturing efficiency as well as the performance of the HPP.

Table 6.1: Complete system and controller parameters of Figure 6.1

<u>Grid</u> Voltage = 11 Kv Frequency = 60 Hz SCL = 5MVA , $X/R = 8$	<u>Line Parameters</u> $L = 50 \mu\text{H}$ $R = 2 \text{ m}\Omega$ $\omega_0 = 377 \text{ rad/s}$
<u>Transformer</u> $Y_g/\Delta 1$ $S_n = 2.5 \text{ MVA}$, $f = 60 \text{ Hz}$ LV = 480 (Ph – Ph rms) HV = 11KV (Ph – Ph rms) $R_1 = R_2 = 0.0325 \text{ pu}$ $L_1 = L_2 = 0.818 \text{ pu}$ $R_m = L_m = 500 \text{ pu}$	<u>Converters</u> Type: 3 Phase, NPC, 3 Level $r_{\text{on}} = 1 \text{ m}\Omega$ Forward Threshold Voltage $V_d = 1 \text{ V}$ Carrier Frequency = 1680 Hz Modulation type: PWM + 3rd Harmonic Injection $C = 9625 \mu\text{F}$, $C_f =$ $V_{\text{dc,ref}} = 2500 \text{ V}$
<u>Turbine & Gear Box</u> Radius = 10 m $C_{p-\text{Max}} = 0.5$ at $\beta = 0$ $\lambda_{\text{opt}} = 5$ $\tau = 0.5 \text{ s}$ Gear Ratio (N) = 20 $\eta_{GB} = 96\%$ (Avg)	<u>Machines</u> Voltage = 2.3 KV , Power = 2250 hp $f_0 = 60 \text{ Hz}$ $R_s = 29 \text{ m}\Omega$, $R_r = 22 \text{ m}\Omega$ $L_s = 35.2 \text{ mH}$, $L_m = 34.6 \text{ mH}$, $L_r = 35.2 \text{ mH}$ $\tau_s = 1.213 \text{ s}$, $\tau_r = 1.6 \text{ s}$ $\sigma = 0.03384$ $J = 63.87 \text{ Kg. m}^2$, No of Poles = 4
<u>HPP Controllers (Grid Imposed Frequency Mode)</u> $H_P = 1043 \frac{s + 29}{s^2 + 1079s}$ $H_Q = \frac{5000}{s}$, $F(s) = \frac{1}{.0005s + 1}$ $K_d(s) = K_q(s) = \frac{(0.25s + 6)}{s}$ $G_{\text{PLL}}(s) = \frac{30(s^2 + 568520)(s^2 + 166s + 6890)}{s(s^2 + 1505s + 568520)(s^2 + 962s + 232325)}$	<u>HPP Controllers (Autonomous Mode)</u> $\tau_i = 1.0 \text{ ms}$ $\omega_0 = 377 \text{ (rad/s)}$ $(V_{sd\text{ref}}, V_{sq\text{ref}}) = (391, 0)$ $K(s) = 0.8 \left(\frac{s + 100}{s} \right)$
<u>Machine side Controllers (VFC)</u> $H_\lambda = 186 \frac{s + 0.625}{s}$ $G_i(s) = \frac{I_{dq}(s)}{I_{dq\text{ref}}(s)} = \frac{(13.6s + 333)}{s}$ $PI = \frac{50s + 25}{s}$	<u>DC Voltage Equalizer</u> $\text{LPF} = \frac{s^2 + (3\omega_0)^2}{(s + 3\omega_0)^2}$ $\omega_0 = 377 \text{ (rad/s)}$ $k(s) = 0.503 \times 10^{-3} \text{ (V}^{-1}\text{)}$

6.2 Initializations

The system depicted in Figure 6.1 starts in the autonomous operation mode. While disconnected from the grid the DC capacitors are assumed to have been charged via soft starting resistors and a power source to their nominal value, providing a total DC voltage of 2500V. During this period the converter switching signals are blocked, all controllers are inactive and the *Mode Command* is kept at zero, resulting in the main switch remaining open and the HPP, dq -reference currents as well as the operating frequency being governed by $(V_{sd_{ref}}, V_{sq_{ref}}) = (391, 0)$, and $\omega_0 = 377$ (rad/s) respectively. In practice all such values can be obtained online through a local dispatching center. At $t = 0.01s$, switching signals are unblocked, controllers are enabled and references are ramped to their predefined values. Figure 6.3 illustrates that by $t = 0.07s$ the V_{sd-LV} voltage reaches its reference value. By the time the PLL regulates V_{sq} (shown in the legend as V_{sq-HV}) to zero and its track of the grid frequency reaches steady state, the VCO frequency input is switched from the constant reference $\omega_s = \omega_0 = 377$ to ω_g , hence system frequency is decreased by $\Delta\omega$ towards the grid frequency until the two become equal, bearing in mind that the main switch is still open. Once all grid connection criteria (Figure 4.5) are met the “Grid Connection” flag goes up and consequently a short while later, i.e. at $t = 0.16s$, the main switch is closed and grid connection is established. Thus the HPP supervisory control system switches the operation mode from *controlled frequency* to *grid-imposed frequency* mode which means that the HPP $i_{dq_{ref}}$ are now defined by the reference active and reactive powers. One must note that the VCO output in this case is increased by $\pi/6$ to compensate for the transformer 30° phase delay. As the phase (V_a) voltage plot demonstrates, at this stage onwards the LV and PCC voltages are locked with a fixed 30° phase difference.

As the V_{dq} plot shows once the main switch closes the LV side q -axis voltage (V_{sq-LV}), jumps from zero to 200V. This is due to the shift in current reference commands ($i_{dq_{ref}}$) which are now the HPP transferred active and reactive powers. Since at this stage (Q_{sref}) is zero, converter neither produce nor intakes reactive power and so C_f produced reactive power is transferred to the grid causing the net reactive power to become positive and this is reflected on the V_{sq-LV} plot. One must note that since at this stage V_{sd-LV} is more or less the same as before, the low voltage peak value, i.e. $\widehat{V}_s$ slightly increases.

At $t = 0.25s$, $V_{sq-HV} = 0$, which means that the PLL has reached a steady state and the HPP-grid connection has successfully been established.

Figure 6.3: Voltage and frequency plots before and after the initial HPP-Grid connection

Once the grid connection is stable, the machine-side converters can be activated. Hence, at $t = 0.3s, 0.5s$ and $0.7s$ respectively the three variable frequency converter switching signals are released. The reason for not releasing all machine converter switching signals all at once is to limit the total energization current demand. Consequently to regulate, the machine fluxes ($\hat{I}_m$), the three flux compensators, step up their $I_{d_{ref}}$ sequentially, which are rapidly tracked by their d -axis stator currents (I_d). For protection purposes the machine stator dq -currents are limited to 1500A. As Figure 6.4 illustrates, while each generator $I_d = 1500A$, its magnetization current, i.e. $\hat{I}_m$ rises towards the reference value which for all machines is set to be 141A; once the set point is reached flux controllers reduce $I_{d_{ref}}$ down maintaining them at the value of 141A. The power required for energization of the machines is obtained from the grid, thus causing the transferred power (P_s) as well as the DC power (P_{DC}) to be negative. The negative DC powers in the P_{DC} plot of Figure 6.4 indicate that the direction of the power flow is from the grid side to the machine (AC) side. Hence during the energization period the three machines consume power. Up to this stage all three turbines were stalled (using pitch and/or mechanical break systems), thus no torque demand was exerted on the machines and $\omega_{r1} = \omega_{r2} = \omega_{r3} = 0$. This can also be seen from the I_q measurement plots which as can be seen are fluctuating round the zero value. At $t = 1.0s$ all turbine breaks are realized and the system is allowed to follow the tidal flow pattern.

Figure 6.04: Generators 1 to 3, energization plots, and the DC bus power (P_{dc}). A negative P_{dc} indicates power consumption and vice versa

6.3 Maximum Power Point Tracking (MPPT)

At this stage the TISEC is ready for power generation. Since the HPP can control the output power factor the C_f can be disconnected and so at $t = 1.01s$ the three phase capacitor bank is disconnected. From this point onward 2 scenarios are examined where each of the three generators and their corresponding MPPT method are tested, first in a natural channel where the speed variation is slow and secondly with respect to the near shore speed variations where the tidal speed changes value and direction much more rapidly.

Scenario I: Channel Type Speed Profiles (Slow Speed Variations)

In this scenario the three machine face a slow changing tidal speed profile that resembles the natural channel type tidal flows. Figure 6.2a depicts the type of speed variation that each of the three turbines face in this scenario. Plots of Figures 6.5-6.7 depict how the SCIG stator dq currents are controlled during various stages of the tidal flow for the *Constant K*, *Variable K* and *Imposed Delay* MPPT methods. As can be seen from the Figures after the initial flux buildup, I_d and hence all three machine fluxes stay fairly constant throughout the rest of the simulation, regardless of the changes happening in their speed or I_q . The three Figures also show that while I_d currents do not differ much, I_q in the *Variable K* scheme undergoes much higher fluctuations, especially near the peak values. The fluctuations are the result of the torque controller ($T_{e_{ref}}$) going to extremes for keeping the C_p value at maximum. This is a disadvantage of this MPPT method compared to the other two.

Figure 6.5: Stator dq -frame currents (I_{dq}), machine magnetization current (I_m) and electrical torque (T_e) plots for the first generator that employs the *constant* KMPPT method, operating under the natural channel type speed variations

Figure 6.06: Stator dq -frame currents (I_{dq}), machine magnetization current (I_m) and electrical torque (T_e) plots for the second generator that employs the *variable* KMPPT method, operating under the natural channel type speed variations

Figure 6.7: Stator dq -frame currents (I_{dq}), machine magnetization current (I_m) and electrical torque (T_e) plots for the third generator that employs the *Imposed Delay* MPPT method, operating under the natural channel type speed profile

Figure 6.08: ω_r , C_p and Acceleration plots for the three generators operating under the natural channel type speed variations

Figure 6.8 shows the three generator rotor speeds plotted against the changing flow speed, as well as the performance coefficients (C_p) and the accelerations. As can be seen from the C_p plots, apart from the startup and places where the current speed decrease to very low values, the three generator performances do not vary much and so the three MPPT methods lead to capturing of comparable amounts of power when employed in sites with slow speed variations. However the *Variable K* method is able to produce the highest accelerations and so the third generator has the shortest startup as well as the best C_p profile. The *Imposed Delay* and *Constant K* methods take the second and third places

Scenario II: Coastal Type Speed Profiles (Rapid Speed Variations)

Plots of Figures 6.9-6.11 show the three machines I_{dq} variations under the three MPPT methods. As can be seen from the Figures, similar to the natural channel flow scenario, after the initial flux buildup, I_d and hence all three machine fluxes stay fairly constant throughout the rest of the simulation, regardless of the changes happening in their speed or I_q . Although like before $T_{e_{ref}}$ and consequently (T_e) and I_q in the *Variable K* scheme undergoes much higher fluctuations, the extra control effort is much more fruitful in this scenario as the C_p profiles of Figure 6.12 illustrate.

Figure 6.9: Stator dq -frame currents (I_{dq}), Magnetization current (I_m) and machine electrical torque (T_e) plots for the first generator that employs the *constant KMPPT* method, operating under the coastal type speed variations

Figure 6.10: Stator dq -frame currents (I_{dq}), Magnetization current (I_m) and machine electrical torque (T_e) plots for the first generator that employs the *Variable* KMPPT method, operating under the coastal type speed variations

Figure 6.11: Stator dq -frame currents (I_{dq}), Magnetization current (I_m) and machine electrical torque (T_e) plots for the first generator that employs the *Imposed Delay* MPPT method, operating under the coastal type speed variations

Figure 6.12: Rotor speed(ω_r), Capacity factor (C_p) and Acceleration plots for the three generators operating under the coastal type speed variations

As can be seen from the C_p and Acceleration plots in Figure 6.12, the *Variable K* MPPT method manages to maintain the capacity factor at maximum even during sudden speed drops where the other two methods show much weakness. In addition, it is seen that the *Constant K* MPPT method takes much longer to reach the optimum speed especially during the startup and instances where the speed is rapidly dropping and picking up again. These instances are very typical of near-shore currents as they change direction quite often. The acceleration plots also show that the *Variable K* and the *Imposed Delay* MPPT methods produce the highest accelerations which are cogent with their ability to track the rapid speed variations and maintaining the optimum C_p value. The differences in the C_p profiles translate into differences in the amount of power that each MPPT method can extract from the available power in the stream. To demonstrate this more effectively the power that each of the three different MPPT schemes extracts (P_e) under the channel type and near-shore situations is plotted against the total extractable power as well as the power transferred to the grid (P_s), Figures 6.13 and 6.14. For the sake of clarity a moving average filter has been employed to smooth the extracted and transferred powers. One must notice that unlike P_s , P_e graphs do not exhibit the initial negative power pertaining to the capacitor charging and machine flux buildup, which is basically due to the way this power is defined; here P_e is calculated as the product of the electrical torque and the machine rotational speed, i.e. $P_e = T_e \omega_r$, therefore it does not encounter machine I_d variations that cause the negative startup power as reflected in P_s .

Figure 6.13: Extracted (P_e) and Transferred Powers (P_s) plotted against the Extractable Power (P_{Ext}) under a slowly changing flow speed profile. The performances of the three MPPT routines are almost identical

Plot of Figure 6.13 shows that the performances of the three MPPT routines are almost identical, with minute superiority of the *Variable K* method during the startup and saddle regions where either the speed is quite low or alternatively the flow direction is changing. Be that as it may, the three MPPT methods result in very different extracted profiles once employed in near-shore flow patterns. In other words the benefit of the *Controlled K* and *Imposed Delay* MPPT methods will become more distinct once tidal speed changes happen more rigorously. Figure 6.14 shows the extracted powers under each MPPT method plotted against the total extractable power when the tidal speed changes 10 times faster than the previous case, i.e. same as the near-shore speed profile of Figure 6.2. As Figure 6.4 illustrates, even though there are 4 drifts in the stream pattern (Figure 6.2), the *Constant K* MPPT scheme misses most of the power in the first drift, however *Variable K* and *Imposed Delay* routines are capable of capturing a significant portion of the power available in the first stream. The cause of this poor startup mainly lies in the high inertia of the machine, turbine and gearbox, preventing a prompt follow-up of the tidal flow speed profile. Consequently, while the machine is picking up speed; much of the available energy in the first drift is lost. Although the startup problem also exists in places with channel type flow profiles, due to the slower speed variations, the machines have sufficient time to reach the optimal speed.

Figure 6.14: Extracted (P_e) and Transferred Powers (P_s) plotted against the Extractable Power (P_{Ext}) under a rapidly changing tidal speed profile

Apart from the startup for the rest of the simulation the superiority of the *Controlled K* scheme under rapid speed variations is moderately noticeable, yet still the dependency of the method on current speed monitoring makes it practically challenging if not impossible. The *Imposed Delay* method on the other hand, though not being able to deliver as much power as the *Variable K* method, performs much better than the conventional *Constant K* MPPT method without any dependency on the current speed and so is a much more practical approach to the problem of maximum power capturing of near shore flow profiles. Despite this all three methods are less efficient in near-shore flow patterns as opposed to the channel type indicating that it's much more difficult to capture the power in instant drifts and rigorous tidal flows that change speed and direction quickly.

While the difference between the extracted (P_e) and transferred (P_s) power profiles depict the system electrical losses, the difference between P_s and the total extractable power profile shows the total losses of the system including both the electrical and mechanical losses. Hence considering the results of the previous Figures, it's easy to see that once the system is running most of the losses in the system are electrical rather than mechanical, which is mostly due to the low operating voltage of the converters

compared to their handling power range which results in an increase of their conduction losses and losses in the submarine cables. Although this problem can be overcome by increasing the operating voltages the need for HVDC systems or employment of an underwater transformer adds to the complexity of the and overall cost of the system.

To measure the efficiency of each MPPT method and also the overall efficiency the surface area under the P_{Ext} , P_e and P_s power plots corresponding to the extractable, captured and transferred energy sums are calculated as:

$$\begin{cases} \eta_{Total} = \frac{\int P_s(t) dt}{3 \times \int P_{Ext}(t) dt} \times 100 \\ \eta_{Gi} = \frac{\int P_{ei}(t) dt}{\int P_{Ext}(t) dt} \times 100 \end{cases} \quad (6.1)$$

$$\begin{cases} P_s(t) = \frac{3}{2} [V_{sd}(t) \cdot I_{sd}(t) + V_{sq}(t) \cdot I_{sq}(t)] \\ P_e = T_e \times \omega_r \\ P_{Ext}(t) = \frac{1}{2} \rho A V^3 C_{pmax} \end{cases} \quad (6.2)$$

where η_{Total} represents the overall efficiency and η_{Gi} represents the individual machine efficiencies with respect to their MPPT technique. Since here P_s is calculated with respect to the PCC bus voltage and current measurements (Figure 6.1), the calculated efficiency accounts for the mechanical losses as well as the losses pertaining to the converters, submarine cable and the step-up transformer. Table 6.2 summarizes the different efficiency percentages under near-shore and channel type speed profiles.

What these results infer is that while the three methods are basically indifferent (with the *Constant K* method having the advantage of avoiding an excessive flow meter and auxiliary control systems), when used in channels or water gates where the tidal currents undergo very slow speed changes of significant magnitude, the *Controlled K* and *Imposed Delay* scheme perform much more efficiently when it comes to handling rapid speed changes and thus are more suitable for bidirectional tidal generators intended for coastal and near-shore tidal farms where the time gap between tidal flood and ebbing is narrow, and the speed changes are more spontaneous. In such cases the change in power capturing efficiency reaches 20% which considering the magnitude of the tidal power density and planning periods, translate into millions of dollars. In addition, while the *Variable K* MPPT method results in a more or less higher capturing efficiency the *Imposed Delay* alternative has the advantage of not requiring current speed measurements and so can be easily implemented in real case scenarios.

Table 6.2: Efficiency percentages under different speed profiles

6.4 PCC and DC bus Voltage Regulation

The performance of the HPP in terms of efficiently regulating the DC voltage as well as the PCC bus Voltage Regulating (V.R) system is depicted in Figures 6.15-6.18. As can be seen from the results of Figures 6.15 and 6.16, PCC voltage regulation through reactive power control was satisfactorily performed under both the channel and near-shore type flows. Thus no matter how abrupt the captured tidal flows and consequently active power changes are, by employing part of the HPP converter power capacity for reactive power regulation even in networks that are much weaker than the standard networks, the PCC bus voltage can be regulated effectively (with respect to V_{Grid}) through the HPP injecting or absorbing reactive power. In this study, in view of the grid nominal voltage, the proposed reactive power control has resulted in an average voltage deviation of less than 0.5%, regardless of the type tidal flow speed variation. The results however show that the near unity, i.e. $PF = 0.99$, scheme causes the PCC bus voltage to fluctuate according to the variations in the captured power. Under this scheme the PCC bus voltages undergoes voltage swells as high as 7-8% which is enough to offset transformer tap positions as well as other utility system settings that can lead to many problems in the local network.

Figure 6.15: PCC voltage deviations $\% \Delta V = 100 \left(\frac{V_{PCC} - V_{Grid}}{V_{Grid}} \right)$ plotted against P_s and Q_s , under the *Constant Power Factor* and *PCC Voltage Regulating* schemes. Channel type tidal flow speed variations

Figure 6.16: PCC voltage deviations $\% \Delta V = 100 \left(\frac{V_{PCC} - V_{Grid}}{V_{Grid}} \right)$ plotted against P_s and Q_s , under the *Constant Power Factor* and *PCC Voltage Regulating* schemes. Near-shore tidal flow speed variations

Figure 6.17: Performance of the HPP, DC voltage management under various power factor schemes. Channel type flow speed variations

Figure 6.18: Performance of the HPP, DC voltage management under various power factor schemes. Near-shore type flow speed variations

Figures 6.17 and 6.18 give further details of the system variables in this study. The two Figures show the DC voltage as well as the DC capacitor voltage differences, i.e. $\Delta V_{DC} = V_1 - V_2$, under each of the two scenarios where the HPP power factor is fixed and variable, while the tidal flow speed was changing gradually (Figure 6.17) and rapidly (Figure 6.18). As is illustrated from the results of these two Figures, regardless of the HPP power factor (reactive power) regulation scheme and the type of tidal speed variations the HPP controller manages the active power transfer in such a way that maintains a very steady DC voltage. In addition, the DC voltage equalizer keeps the DC voltage divided evenly between the two capacitors, by modifying the converter modulation index every now and then. Incorrect operation of this system would lead to an unbalanced DC voltage among the two capacitors and consequent malfunction of the 3-level NPC converter. These results show that the DC voltage variations as well as the DC capacitor voltage differences (i.e. $V_1 - V_2$) are satisfactorily limited to less than 1.6% of the nominal DC link voltage (2500V), under all circumstances.

6.5 Concluding Remarks

As was illustrated before, while channel type tidal currents have very gradual and consistent speed variations that also happen to be very predictable [78], speed changes in costal currents are more spontaneous. Furthermore, regardless of the type of streams, due to their variable nature, output from tidal generators is very unsteady. Thus extraction and integration of tidal power into the conventional electric networks bring about many challenges such as the need for maximum power extraction and smooth power transfer without degradation of the voltage quality. In this chapter, different proposed solution designs were tested for both near-shore and channel type tidal turbines. It was shown how widening the gap between electrical and mechanical torques can increase the turbine acceleration/deceleration which in turn leads to a better tracking of the tidal current speed profile and thus increases the power capturing efficiency. Moreover it was shown how the tidal speed forecasting method can be used to size a hybrid power port. Thus a HPP was sized and its performance was put to test under various power factor regimes. It was seen that the designed HPP not only does a good job in synchronizing with the grid, which allows for smooth connection and disconnection, according to its control settings it also capable of maintains either a unity power factor or a steady PCC bus voltage during power transfers. In other words by assigning a small portion of the grid-tie converter power capacity for reactive power control, the HPP can either regulate the PCC bus voltage or maintain a constant power factor while transferring the captured tidal power to the grid.

Chapter 7

Conclusions and Recommendations

7.1 Conclusions

Marine or oceanic energy refers to the energy carried by ocean tides, waves, salinity, and ocean temperature differences. Among the aforementioned types, this study focused on the oceanic tidal stream power which began to draw attention in the mid-70s after the first oil crisis. The periodic rise and fall of the water along ocean coasts and their rhythmic daily patterns are familiar phenomena which are caused for the most part by the relative positions of the moon and Earth. Tidal currents occur in conjunction with the rise and fall of the tide. In other words the vertical motion of the tides along with the rotation of the earth causes the water to move horizontally, creating currents. Such recurrent cycles of current buildup along with the high density of water, make tidal stream power a renewable resource with extraordinary energy density when compared to other renewables.

Consequently development of this emerging technology presents unprecedented opportunities to increase power extraction and efficiency that are yet to be explored. The lunar semidiurnal tide, with two high and two low waters each day is the principal world tide. The amount of displaced water in a lunar cycle each day is tremendous, however it happens very gradually over a course of several hours. Consequently in the open ocean where the water enjoys a vast basin the tidal currents caused by the rise and fall of water, have lower velocities. In addition to the lunar tides at these so called near-shore areas, stream patterns are also influenced by propagation and reflection of tidal waves. As the wave propagates into shallower water, its wave speed decreases and the energy contained between crests is compressed both into a smaller depth and a shorter wavelength; thus the tide height and the tidal stream strength must increase accordingly. Once reaching the shore the water returns (usually at an angle) to the sea causing reverse currents. This phenomenon can also be explained through the Bernoulli potential and kinetic energy conservation equations and the continuity of mass principle [86]. One must note that this water flooding and ebbing occurs more frequently than the lunar tides, imposing regular and rapid speed variations on near-shore and coastal currents. On the other hand in water straits, estuaries and natural channels where water streams are restricted in a much narrower space, the current speed variations are essentially shaped by the lunar tides with wave propagation having less effect. In these areas tidal streams are of much higher speed and rectilinear, meaning they flow back and forth in a somewhat straight line, changing speed and direction much more gradually than the coastal flows. Consequently most of the existing tidal turbines are

designed and installed in channel type areas where a broad strait connects two areas of the sea. While speed variations intensify as the distance to the shore decreases, coastal flows on the whole are apparently weaker than the channel type streams. This fact raises a fundamental question of whether it's feasible to exploit such tidal currents in the first place. The current trend in the tidal industry can also be used as a bear witness to support the argument, since nearly all tidal power companies find natural channels and water straits to be more attractive for installing their turbines. Be that as it may, the concept of offshore wind power and the advantages that co-generation brings with itself, substantiates the need for a closer study of near shore currents and ways of increasing the productivity of near-shore tidal turbines. In addition to that the same technology can also be applied to other devices that experience similar boisterous conditions.

Regardless of the type of tidal streams, due to their variable nature output from tidal generators are highly unsteady. Thus extraction and integration of tidal power into the conventional electric networks bring about many challenges such as the need for maximum power extraction and smooth power transfer without degradation of the voltage quality. In this study, different solution designs were proposed for both near-shore and channel type tidal turbines. It was shown how widening the gap between electrical and mechanical torques can increase the turbine acceleration/deceleration which in turn leads to a better tracking of the tidal current speed profile and thus increases the power capturing efficiency.

Different tidal speed forecasting techniques were also explored to increase the accuracy and reliability of speed estimations that ultimately reveal the amount of available power in the region under study. The long term estimation models were then used to design a hybrid power port that not only allows for smooth grid connection/disconnections; it maintains the PCC bus voltage during various stages of the power transfer both in autonomous and grid-connected scenarios. Through an example it was shown how the long term tidal speed forecasts can be used in determining the power capacity of the hybrid power port as well as the required amount of reactive power production. Several simulation examples as well as comparison studies were included to illustrate the effectiveness of all the proposed systems under different circumstances.

7.2 Recommendations

Although the initial objectives were met, the study has opened many new frontiers that need further investigation. Like the study itself, the accomplishments can be divided into three parts that are briefly reviewed here. In chapter one a conventional tidal speed forecasting system was enhanced with model free estimation techniques that increased the prediction accuracy by bringing in, any hidden trends influencing the tidal flow into the forecasting procedure. In other words the conventional system that merely relied on celestial movements and astronomical periods was enhanced in such a way to include terrestrial effects leading to more realistic forecasting of the tides. This was a timely contribution to the conventional maritime studies and was cited now and again in many subsequent papers. With that being said it's also worth mentioning that although many of the benchmarking model free prediction methods were employed and examined, the available methods and models are much more than the ones utilized in this study. There are novel estimation techniques such as the Hidden Markov Model (HMM) that use probabilistic modeling or individual neural networks that are specifically designed and tuned for predicting time series of different kind. These models were not included in this study but can and are highly recommended for any future investigations. In addition to that, the HAMLs method itself is based on a mean least square protocol that can be replaced and improved further by employing better mathematical alternatives.

In the converter design chapter four main contributions were made that as mentioned before were the maximum power point tracking system, that was designed for near-shore tidal turbines as well as the integrated or hybrid power port structure that enabled the grid side inverter to actively regulate the point of common coupling bus voltage in autonomous and grid-connected modes of operation. Since the HPP is designed to maintain a steady DC link voltage, the concept of energy storage can be introduced to the current research in future studies. It follows that under such circumstance, one must make sure that battery charging only takes place during power generation so that the system does not intake the grid power for charging the batteries.

The hybrid power port in this study was designed to be robust to voltage sags and small frequency changes; however the PLL would not function properly under an unbalanced and/or highly distorted grid voltage profile, leading to the HPP system deficiency or alternatively change of its operation mode under such circumstances. Enhancement of the HPP control system with fault tolerant algorithms could be a big step into commercialization of the design since in most revised grid interfacing standards it is advised for the renewable resources to stay connected even during grid faults and variations of substantial magnitude. Here again, introduction of storage systems can help prevent loss of energy should the grid become highly

distorted, forcing the system into an autonomous mode. Furthermore though the HPP was designed to regulate the PCC bus voltage in both autonomous and grid connected modes, there are improvements that need to be done before employing the system in real size applications. As an example, the autonomous-state voltage regulation system unties itself from any prior knowledge of the load model by adapting a decoupling feed-forward compensation strategy that uses real-time load current measurements. Therefore should any measuring device adapted in a real case scenario fail, the system will simply not deliver. Consequently a redundant watch dog alternative routine needs to be implemented in commercial systems in which a load forecasting system estimates and provides the load current values under a failed or malfunctioning measuring device situation. The grid connected mode on the other hand ignores the network load status altogether and although the load effect is analyzed and provided in the control loop, no local loads were assumed when designing and testing the voltage regulator. A good point to start any future study would be to evaluate and analyze the effect of different load models on the HPP voltage control system and the affiliated complexities.

The machine side converter unit in this study was designed to regulate the machine flux and electrical torque independently. The decoupling process was made possible through a conventional flux observer unit that intakes the stator current as well the rotor angular velocity. Consequently a submergible encoder unit is required for the system to work. Such a device increases the system vulnerability and is not very amenable in practical applications. A different approach is to estimate the rotor position through the use of stator voltage and current waveforms. However if the system is to operate correctly the designing of the affiliated control system would become more complicated as the phase delays caused by the cable and other variable interferences caused by the ongoing changes in the subsea environment need to be actively monitored and controlled. Thus a sensor-less method that could extract the necessary information for estimation of the rotor position from distorted voltage and current measurements is a very good topic for any future studies; and last but certainly not least although a new maximum power point tracking system was designed to be suitable for near shore turbines that experience rapid speed changes and thus require acceleration boost for efficient tracking of the current speed, the number of MPPT alternatives explored in the study were limited. Now that a variable speed system with independent flux and torque regulating loops is available many of the existing MPPT techniques proposed in other industries such as wind and solar can be adapted or examined for tidal turbine applications, and so could again be seen as a good starting point in any future study. That being said, it's also worth mentioning that the distinction between natural channel and near shore tidal patterns mentioned for the first time in this study can also be explored further through joint ventures and interdisciplinary research that takes the nature of tidal flows into account when designing a maximum power capturing system.

Chapter 8

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APPENDIX I

Tidal Speed Prediction

In time-series prediction, the known values of the time series up to the point in time, say, t , must be used to predict the value at some point in the future, say, $t + P$. A typical method for this type of prediction is to create a mapping from D sample points, sampled every Δ units in time, where Δ can be either fixed or varying. So the input dataset would be a D -dimensional vector of more or less one of the following forms; with I representing the input, T target, and x the time series under study.

$$I_1(t) = [x(t - (D - 1)\Delta), \dots, x(t - \Delta), x(t)]$$

$$T(t) = x(t + \Delta)$$

Figure A1.1: Fixed Δ

$$I_2(t) = [x(t - \Delta_D), \dots, x(t - \Delta_1), x(t)]$$

$$T(t) = x(t + P)$$

Figure A1.2: Variable Δ

Here P (prediction horizon) is equal to Δ when the input is formatted as in I_1 (Figure A1.1) and is an independent variable when the input is arranged as in I_2 (Figure A1.2).

The following set of tables describes the architecture of the model free estimators that have been used at different stages of this study. The number of layers and neurons are presented by the numbers enclosed in a bracket so that for example a neural network with two layers having 10 and 5 neurons respectively is displayed as [10 5]. Where the inputs have been sampled as in Figure (A1.2), D values are presented in brackets and where the input has been as in Figure (A1.1), D values are presented as a constant. So for example if D is expressed as $D = [1: 3, 7, 9]$ this means that the input vector has been:

$$I(t) = [x(t-1), x(t-2), x(t-3), x(t-7), x(t-9)]$$

Or if D is a constant like $D = 3$, it simply implies that the input and target vectors have been:

$$I(t) = [x(t-2\Delta), x(t-\Delta), x(t)] \text{ and } T(t) = x(t+\Delta)$$

A similar explanation holds for the regressor sets of non-linear auto regressive estimators while linear ARs have been quoted with a subscript number that indicates their order. In both tables N represents the number of non-linear units.

Table A1.1: Description of the Model Free Estimators used in Section I

	Predicti on Horizon (In Steps)*	Size of Training Dataset = 1 Week	Size of Training Dataset = 1 Year
MLP	$\Delta = 5$	[15], $D = 10$ <i>Fit = %90.67</i>	[15], $D = 10$ <i>Fit = %93.63</i>
	$\Delta = 10$	[15 10 5], $D = 10$ <i>Fit = %31.47</i>	[10 10], $D = 15$ <i>Fit = %35.73</i>
FTD	$\Delta = 5$	[6], $D = [1:9\ 12:15\ 21\ 28]$ <i>Fit = %99.38</i>	[5 1], $D = [1:10\ 12:15\ 19]$ <i>Fit = %99.59</i>
	$\Delta = 10$	[10 10 5], $D = [1:13\ 17\ 23]$ <i>Fit = %91.64</i>	[10 10], $D = [1:8\ 10:13\ 28]$ <i>Fit = %94.19</i>
NARX	$\Delta = 5$	[10], $D = [1:13\ 19:21\ 26\ 29]$ <i>Fit = %98.46</i>	[5 3], $D = [1:10\ 12:15\ 19]$ <i>Fit = %99.75</i>
	$\Delta = 10$	[10 5], $D = [1:10\ 18:22\ 25]$ <i>Fit = %92.61</i>	[10 6], $D = [1:6\ 12:16\ 25\ 29]$ <i>Fit = %95.56</i>
ANFIS	$\Delta = 5$	$nMFs = [3]$, $D = 3$ <i>Fit = %78.44</i>	$nMFs = [3]$, $D = 5$ <i>Fit = %87.94</i>
	$\Delta = 10$	$nMFs = [3]$, $D = 2$ <i>Fit = %33.47</i>	$nMFs = [3]$, $D = 3$ <i>Fit = %36.03</i>
Linear AR	$\Delta = 5$	$AR_{(25)}$, <i>Fit%</i> = 23.58	$AR_{(30)}$, <i>Fit%</i> = 30.88
	$\Delta = 10$	$AR_{(25)}$, <i>Fit%</i> = 2.27	$AR_{(30)}$, <i>Fit%</i> = 8.743
Non-Linear AR	$\Delta = 5$	Sigmoid-Net, $N=12$ Regressors = [5:8,10,12,13,16,18,19,23:25,27,29] <i>Fit%</i> = 96.73	Sigmoid-Net, $N=9$ Regressors = [1:4,6,9:15,17,23,25,27,29] <i>Fit%</i> = 99.18
	$\Delta = 10$	Wave-Net, $N=4$ Regressors = [1:4,6,8,10,12,18,19,24,25,27,30] <i>Fit%</i> = 88.26	Sigmoid-Net, $N=10$ Regressors = [1:3,5:7,12,14:16,19,21,23,26,27,30] <i>Fit%</i> = 90.39

* Since each time-step corresponds to 6 minutes, $\Delta = 5$ denotes 30 minutes and $\Delta = 10$ is one hour of prediction

Table A1.2: Description of the Model Free Estimators used in the Hybrid Model of Section I

	Predicti on Horizon	Size of Training Dataset = 6 Months	Size of Training Dataset = 1 Year
NARX	$\Delta = 1$ Year	[10 4], <i>Delay Vectors</i> = [15 15] <i>Fit</i> = %89.18	[15 6], $D = [20\ 20]$ <i>Fit</i> = %93.01
AR	$\Delta = 2$ Years	Sigmoid-Net, $N=10$ Regressors = [1:10,12:17,19,24:26,28,30] <i>Fit</i> = %83.36	Wave-Net, $N=20$ Regressors = [1:9,12:21,23,25,28:30] <i>Fit</i> = %87.11
	$\Delta = 3$ Years	Sigmoid-Net, $N=15$ Regressors = [1:11,12,16,19,24:27,29,30] <i>Fit</i> = %80.67	Sigmoid-Net, $N=25$ Regressors = [1:8,12:16,18,20:22,24,26,28:30] <i>Fit</i> = %82.64

APPENDIX II

Phase Locked Loop (PLL)

$$\begin{cases} V_{sd} = \hat{V}_s \cos(\omega_0 t + \theta_0 - \rho) \\ V_{sq} = \hat{V}_s \sin(\omega_0 t + \theta_0 - \rho) \end{cases} \quad (A2.1)$$

Taking into account the equation set (A2.1), Equation 3.7 of chapter 3 can be rewritten as:

$$\begin{cases} L \frac{di}{dt} = L\omega(t)i_q - (R + r_{on})i_d + V_{td} - V_{sd} \\ L \frac{di}{dt} = -L\omega(t)i_d - (R + r_{on})i_q + V_{tq} - V_{sq} \\ \frac{d\rho}{dt} = \omega(t) \end{cases} \quad (A2.2)$$

Since $V_{sq} = \hat{V}_s \sin(\omega_0 t + \theta_0 - \rho)$, $\rho(t) = \omega_0 t + \theta_0$ corresponds to $V_{sq} = 0$. Hence, a mechanism is required to regulate V_{sq} at zero. This can be achieved according to the following feedback law:

$$\omega(t) = H(p) \cdot V_{sq} \quad (A2.3)$$

In which $H(p)$ is a linear transfer function and $p = d(\cdot)/dt$ is the differentiation operator. Substituting for V_{sq} from Equation (A2.2) in (A2.3), and substituting for ω from Equation (A2.3) in (A2.2), the following is deduced:

$$\frac{d\rho}{dt} = H(p) \hat{V}_s \sin(\omega_0 t + \theta_0 - \rho) \quad (A2.4)$$

Equation (A2.4) represents a nonlinear dynamic system, referred to as the Phase Locked Loop (PLL) [88, 89]. The main function of the PLL is to regulate ρ at $\omega_0 t + \theta_0$; despite this, in view of its nonlinear characteristic, the PLL can exhibit unsatisfactory behavior under certain conditions. For example, if the PLL starts from an initial condition in which $\rho(0) = 0$ and $\omega(0) = 0$, then the term $H(p) \hat{V}_s \sin(\omega_0 t + \theta_0 - \rho)$ in (A2.4) would be a sinusoidal function of time with the frequency ω_0 . Consequently, if $H(s)$ has a lowpass frequency response, the right-hand side of Equation (A2.4) and also $d\rho/dt$ would demonstrate small sinusoidal perturbations around zero, the PLL falls in a limit cycle, and ρ does not track $\omega_0 t + \theta_0$. Therefore to prevent such limit cycles from happening, the aforementioned control law is modified as:

$$\omega(t) = H(p)V_{sq}(t), \quad \omega(0) = \omega_0 \text{ and } \omega_{Min} \leq \omega \leq \omega_{Max} \quad (A2.5)$$

where $\omega(0) = \omega_0$ and $\omega(t)$ is limited to ω_{Min} and ω_{Max} , selected to be close to ω_0 ; defining a narrow range of $\omega(t)$ variations. The range of variations should be selected sufficiently wide to permit excursions of $\omega(t)$ during transients. If the PLL tracks $\omega_0 t + \theta_0$, the term $\omega_0 t + \theta_0 - \rho$ would be close to zero and so, $\sin(\omega_0 t + \theta_0 - \rho) \approx (\omega_0 t + \theta_0 - \rho)$. Consequently Equation (A2.4) can be simplified to:

$$\frac{d\rho}{dt} = H(p)\hat{V}_s(\omega_0 t + \theta_0 - \rho) \quad (A2.6)$$

Equation (A2.6) represents a classical feedback control loop in which $H(s) \cdot \hat{V}_s$ is the transfer function $\omega_0 t + \theta_0$ is the reference input and ρ is the output of the effective compensator (Figure A2.1). Based on the methodology presented above Figure A2.2 illustrates a schematic diagram of the PLL in which the PLL transforms V_{sabc} to V_{sdq} and adjusts the rotational speed of the dq -frame (ω) such that in the steady state V_{sq} is forced to zero resulting in $\rho = \omega_0 t + \theta_0$ and $V_{sd} = \hat{V}_s$. The integrator in Figure A2.2 is realized by means of a voltage-controlled oscillator (VCO) which can be regarded as a settable integrator whose output (ρ) resets to zero whenever it reaches 2π .

Figure A2.1: Control block diagram of the PLL

Figure A2.2: Complete control block diagram of the PLL

PLL dynamic performance is mostly influenced by $H(s)$. Inclusion of two integrators in the loop gain can ensure a negligible steady state error and so $H(s)$ is designed to have at least one pole at $s = 0$. The other $H(s)$ poles and zeros are determined according to the closed-loop band width of the PLL and system phase and gain margins. Sinusoidal components in V_{sd} and V_{sq} that are usually caused by unbalanced or harmonically distorted three-phase voltages are evenly important when designing the PLL. Between the different AC components of V_{sq} , the $2\omega_0$ constituent is more important[20]. Accordingly to attenuate the double-frequency component of V_{sq} , one pair zeros at $s = \pm j2\omega_0$, are included in $H(s)$. Using this method, the PLL closed-loop bandwidth can be selected to be arbitrarily large.

List of Publications

Journals and Transactions

- M. J. Jahromi, A. I. Maswood, and K. J. Tseng, "*Design and Evaluation of a Tidal In-Stream Generator Power Port*," Systems Journal, IEEE, vol. PP, pp. 1-1, 2013.
- M. J. Jahromi, A. I. Maswood, and K.J. Tseng, "*Design and Evaluation of a New Converter Control Strategy for Near Shore Tidal Turbines*," Industrial Electronics, IEEE Transactions on, vol. PP, pp. 1-1, 2012.
- M. J. Jahromi, A. I. Maswood, and K.J. Tseng, "*Long Term Prediction of Tidal Currents*," Systems Journal, IEEE, vol. 5, pp. 146-155, 2011.

International Conferences

- Mahda J. Jahromi, A. I. Maswood, K.J. Tseng, Bicky Bhangu, "*Tidal In-Stream Generation with Improved PCC Voltage Profile*", IEEE-PES Innovative Smart Grid Technologies Conference, Washington, D.C. January 17, 2012
- Mahda J. Jahromi, A.I. Maswood, and K.J. Tseng, Bicky Bhangu, "*Maximum Power Extraction of Tidal Streams*", IEEE - PES Innovative Smart Grid Technologies Conference, Washington, D.C., January 17, 2012
- Mahda J. Jahromi, A.I. Maswood, and K.J. Tseng, "*Comparison of Different Techniques for Short Term Prediction of Tidal Current Speeds*", IEEE Power & Energy Society General Meeting, Minneapolis, Minnesota, USA, July 25-29, 2010
- Mahda J. Jahromi, M.H. Haque and K.J. Tseng, "*Siting and sizing of distributed generation using three new indices*", IASTED International Conference on Solar Energy, SOE, Phuket, March 2009