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Nanyang Technological University
School of Electrical and Electronic Engineering

**ADVANCED CONTROL FOR LARGE SCALE
NON-CONVENTIONAL POWER SYSTEMS**

Wu Si

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Summary

This thesis concerns the applications of advanced controls for large scale non-conventional power systems. The so-called non-conventional power system is defined as the power system with grid-connected renewable power generation sources in this thesis. Among all kinds of renewable power generation sources, the installed capacity of wind power is the largest with the growing over 25% per year. Thus, it is reasonable and significant to set wind power as a representative of the renewable power generation sources in the non-conventional power system. In this work, researches are concentrated on the self-control of wind turbine power generation system in non-conventional power systems and the global control of the entire non-conventional power system.

From a Wind Energy Conversion System's (WECS) point of view, it always desirable to capture as much power as possible. This depends on both that the WECS is able to obtain accurate and real-time wind information and efficient and accurate controls are employed in the WECS control systems. For example, for distributed WECS or grid-connected WECS in the case of normal operation of the power system, WECSs main concern is its own control, including wind turbine Maximum Power Point Tracking (MPPT), real-time wind speed estimation, turbine pitch system control and sensorless control of WECS, etc.

In the controls of variable-speed variable-pitch Wind Turbine Power Generation System (WTPGS), the turbine shaft rotational speed should be controlled optimally with respect to the variable wind speed in order to achieve MPP-T. A precise real-time wind speed estimation method and sensorless control for variable-speed variable-pitch WTPGS are proposed in this thesis. The wind speed estimation is realized by an Extreme Learning Machine (ELM) based nonlinear

input-output mapping neural network. A specific design characteristic of the wind turbine is used for improving the mapping accuracy while considering the control of turbine pitch angle system. The proposed wind speed estimation is established by using the information of the wind mechanical torque and the turbine rotational speed which are estimated by a two-loop linear observer. The estimated wind speed is then used to determine the optimal rotational speed and pitch angle commands.

In the wind generator control case, including rotational speed control and current control, in order to obtain high control performance and system robustness, the advanced controller must achieve fast dynamic response during the transient state and robustness against parametric variations and unstructured dynamics. An Optimal Reset Control (ORC) scheme based on linear principle is proposed for the generators control. The major advantages of the ORC method over the traditional control methods are as follows: 1) to reduce the overshoot of the step response without degrading other specifications; 2) to suppress controllers saturation effectively; 3) to accelerate the tracking speed with less error. With these advantages, the ORC has also been used in the design of uncertainty observer. Due to the fast convergence of the ORC uncertainty observers, feed-forward compensations of uncertainty terms can be employed on-line in the wind generator control loops to eliminate the uncertainty effects. The ORC control scheme can achieve fast dynamic response during the transient process and robustness against uncertainties in the generators controls.

The proposed methods are tested on a wind turbine installed with Permanent Magnet Synchronous Generator (PMSG), through both simulations in Matlab/Simulink software and experiments on a PMSG WECS setup. The effectiveness of the proposed methods is verified by performance comparisons with the existing methods.

However, with considering the grid integration, the connection of large wind turbines to the grid has a large impact on grid stability. Basically, the grid codes require wind turbines to have an operational behavior more similar to that of conventional generation capacity and take more responsibility in network operation. The status of wind turbines is thus changing from being simple energy

sources to having power plant status. This means that WECS have to share some of the duties carried out by the conventional power plants.

Modern power systems are large scale and complex. They consist of a collection of several independent and spatially distributed dynamic subsystems with grid-connected wind power plants interconnected together. The conventional controllers may not dominate the global controls of the power system with grid-connected WTPGS and cannot prompt the cooperation of the Synchronous Generator (SG) and the WTPGS which may negatively affect the global stability. In this thesis, works mainly focus on the control designs for power system transient stability enhancement with voltage regulations. A robust backstepping control scheme is proposed for the SG which can both achieve transient control and voltage regulation within one control structure. When a large disturbance occurs, the proposed backstepping law is employed to define variable electrical power output objectives of the SG which can correct the actions of the conventional voltage controllers. Since the power angle and frequency controls of SG are decoupled from the power and terminal voltage controls with the backstepping controls, one control structure for both the transient stabilizing control and voltage regulation is realized.

Moreover, the proposed robust back-stepping control scheme provides a method of coordination and control of the traditional power system and the WTPGS. The power deviation between the electrical power output objective of the SG and its real output power can be offered as the reference for the wind turbine captured power. When the power system is subjected to a severe disturbance, the rotational speed and pitch angle commands can be obtained from the power deviation through the mapping characteristic of the wind turbine in the transient process which can dominate the WTPGS cooperating with the global system and helping to enhance the transient stability and voltage regulation capacity of the global system.

The proposed controller design technique has been tested on Single Machine Infinity Bus (SMIB) power system with grid-connected WPTGS. The applications of the proposed controllers to multi-machine power system with grid-connected WPTGS have also been shown. Their performances are compared with those of

other controllers and evaluated using different cases. Results show the effectiveness of the proposed control strategies.

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List of Symbols

SMIB Power System

$\delta(t)$	the power angle of the generator, in radian;
$\omega(t)$	the relative speed of the generator, in radian/second;
$P_m(t)$	the mechanical input power, in p.u.;
$P_e(t)$	the active electrical power delivered by the generator, in p.u.;
ω_0	the synchronous machine speed, in radian/second; $\omega_0 = 2\pi f_0$
D	the per unit damping constant, in p.u.;
H	the per unit inertia constant, in second;
$E'_q(t)$	the transient EMF in the quadrature axis of the generator, in p.u.;
$E_q(t)$	the EMF in the quadrature axis, in p.u.;
$E_f(t)$	the equivalent EMF in the excitation coil, in p.u.;
T'_{d0}	the direct axis transient short circuit time constant, in second;
$I_f(t)$	the excitation current, in p.u.;
$I_q(t)$	the quadrature axis current, in p.u.;

$Q(t)$	the reactive power, in p.u.;
$V_t(t)$	the generator terminal voltage, in p.u.;
k_c	the gain of the excitation amplifier, in p.u.;
$u_f(t)$	the input of the SCR amplifier of the generator, in p.u.;
x_{ds}	$= x_T + x_L + x_d$;
x'_{ds}	$= x_T + x_L + x'_d$;
x_T	the reactance of the transformer, in p.u.;
x_d	the direct axis reactance of the generator, in p.u.;
x'_d	the direct axis transient reactance of the generator, in p.u.;
x_L	the reactance of the transmission line, in p.u.;
x_s	$= x_T + x_L$;
x_{ad}	the mutual reactance between the excitation coil and the stator coil, in p.u.;
V_s	the infinite bus voltage, in p.u.;

Multi-machine Power System

$\delta_i(t)$	power angle of the i th generator, in radian;
$\omega_i(t)$	rotor speed of the i th generator, in radian/second;
$P_{mi}(t)$	mechanical input power of the i th generator, in p.u.;
$P_{ei}(t)$	active electrical power of the i th generator, in p.u.;
ω_0	synchronous machine speed, in radian/second; $\omega_0 = 2\pi f_0$
D_i	per unit damping constant of the i th generator, in p.u.;

H_i	per unit inertia constant of the i th generator, in second;
$E'_{qi}(t)$	transient EMF in the quadrature axis of the i th generator, in p.u.;
$E_{qi}(t)$	EMF in the quadrature axis of the i th generator, in p.u.;
$E_{fi}(t)$	equivalent EMF in the excitation coil of the i th generator, in p.u.;
T'_{doi}	direct axis transient short circuit time constant of the i th generator, in second;
$I_{fi}(t)$	excitation current of the i th generator, in p.u.;
$I_{qi}(t)$	quadrature axis current of the i th generator, in p.u.;
$I_{di}(t)$	direct axis current of the i th generator, in p.u.;
$Q_{ei}(t)$	reactive electrical power of the i th generator, in p.u.;
$V_{ti}(t)$	generator terminal voltage of the i th generator, in p.u.;
k_{ci}	gain of the excitation amplifier of the i th generator, in p.u.;
$u_{fi}(t)$	input of the SCR amplifier of the generator of the i th generator, in p.u.;
B_{ij}	admittance between the i th generator and j th bus, in p.u.;
$\delta_{ij}(t)$	phase angle difference between $E'_q(t)$ and the j th bus, in radian;
x_{Ti}	reactance of the transformer of the i th generator, in p.u.;

x_{di}	direct axis reactance of the i th generator, in p.u.;
x'_{di}	direct axis transient reactance of the i th generator, in p.u.;
x_{adi}	mutual reactance between the excitation coil and the stator coil of the i th generator, in p.u.;

Acronyms

AC:	Alternating Current
DC:	Direct Current
DFL:	Direct Feedback Linearization
DSP:	Digital Signal Processor
ELM:	Extreme Learning Machine
FACTS:	Flexible Alternating Current Transmission Systems
HAWT:	Horizontal Axis Wind Turbine
IGBT:	Insulated Gate Bipolar Transistor
LQR:	Linear Quadratic Regulation
MPPT:	Maximum Power Point Tracking
ORC:	Optimal Reset Control
ORL:	Optimal Reset Law
PCC:	Point of Common Coupling
PI:	Proportional-Integral
PID:	Proportional-Integral-Derivative
PMS:	Permanent Magnet Synchronous
PMSG:	Permanent Magnet Synchronous Generator
PMSM:	Permanent Magnet Synchronous Motor

PT: Park Transformation
PWM: Pulse Width Modulation
RBF: Radial Basis Function
SG: Synchronous Generator
SMC: Sliding Mode Control
SMES: Superconductive Magnetic Energy Storage
SVM: Space Vector Modulation
TSR: Tip Speed Ratio
VSVP: Variable-Speed Variable-Pitch
WECS: Wind Energy Conversion System
WTPGS: Wind Turbine Power Generation System

Introduction

1.1 Power System

The function of an electric power system is to convert energy from one of the naturally available forms to the electrical form and to transport it to the points of consumption. The generation, transmission and distribution of electrical power are the three main tasks, or primary functions of a power system. Conventional electric power systems vary in size and structural components. However, they all have the same basic characteristics [1]:

- are comprised of three-phase AC systems operating essentially at constant voltage. Generation and transmission facilities invariably use three-phase equipment.
- use synchronous machines for generation of electricity. Prime movers convert the primary sources of energy (fossil, nuclear and hydraulic) to mechanical energy that is, in turn, converted to electrical energy by synchronous generators.
- transmit power over significant distances to consumers spread over a wide area.

Figure 1.1 illustrates a sense of the overall utility generation, transmission, and distribution power system. Electric power is produced at power generation and

transmitted to consumers through a complex network of individual components, including transmission lines and transformers.

Power system operates normally in a balanced state: the electrical energy produced and consumed throughout the whole system should be equal at each moment in time. Consequently, the system operation relies on the control of the system power generation and its adaptation to the fluctuations of the uncertain system load.

Problems of Conventional Power System

Alternative energy such as renewable energy is gradually receiving attentions because of:

1. The conventional power generators which are driven by fossil fuels based prime movers are extracted from the available natural reserves. However, these fossil fuels are not infinite and will be exhausted in the near future. This dictates the gradual incorporation of sustainable energy sources in the power system generation.
2. The adverse environmental impacts of the operation of fossil-fuel-fired power plants, such as the global climate change and the greenhouse effect caused by the greenhouse gases emissions. The Kyoto protocol required a collective reduction by 5.2% in greenhouse gas emissions among 160 industrialized nations. This, among other reasons, is one of the primary driving forces behind an increase in renewable energy generation globally.

Figure 1.1. Conventional power generation, transmission, and distribution system

1.2 Non-conventional Power System

As discussed in the previous section, the development and implementation of alternative energy such as renewable energy is encouraged by environmental, social and political forces. In this thesis, the power system with the mentioned grid-connected alternative power generation sources is called non-conventional power system. The alternative energy sources mainly comprise the renewable energy sources like wind energy, solar energy and etc.. The rise of the renewable energy sources is supported by the advancements in supporting technologies like power-electronic converters and controllers. The basic structures of the conventional power system and the non-conventional power system are presented in Figure 1.2 and Figure 1.3, respectively. Some brief introductions of the renewable energy sources are presented in the following.

Figure 1.2. Conventional power system

Wind Energy

Wind has been utilized as a source of power for thousands of years for such tasks as propelling sailing ships, grinding grain, pumping water, and powering factory machinery. Wind power generation is realized by extracting kinetic energy from the wind passing through the blades of wind turbine. Then the power generators (including asynchronous and synchronous generators) can be driven by the wind turbine for power generation.

Wind energy is one of the most mature of the various renewable energy technologies and has recently gained much favor all over the world. The total amount of economically extractable power available from the wind is considerably more than present human power use from all sources [2][3].

Solar Energy

Solar power generation systems convert sunlight directly into electricity [4] which is usually achieved by photovoltaic (PV) solar panels. A PV cell consists of two or more semiconductor layers of specific physical properties. These layers are arranged in such a way that when the PV cell is exposed to sunlight, the photons cause the electrons to move in one direction (crossing the junctions of the layers) and a direct current (DC) is generated. Currently, PV energy cost is still high.

There are other kinds of the so-called alternative energy sources like biomass, tidal energy, wave energy and geothermal energy. Since the total capacity proportion of these energy sources in the entire power system is still small, so the work in this thesis does not cover the content of these generation sources.

Figure 1.3. Non-conventional power system

1.2.1 Control of Non-conventional Power System

In non-conventional power system, the power system control scheme should fully takes into account the coordination and cooperation of the conventional centralized power generators and the renewable energy sources as shown in Figure 1.4. Basically, different from the conventional centralized power generators, the renewable energy sources have the following two characteristics:

- **Instability and uncertainty.** The mentioned renewable sources such as wind and sunlight are relatively irregular in occurrence. From the renewable energy sources' point of view, it is desirable to capture as much power as possible. However, on the perspective of power market and planning, theoretically, the power grid requires stable power supplies which do not negatively affect global stability and reliability of the power systems. So the self-controls of the renewable power plants should overcome the instability and uncertainty of the sources.
- **Small capacity and flexible controllability.** Compared to the conventional centralized power generations, the capacity of the renewable power plants are relatively small. They can be expected to have satisfactory power control capability similar to that of conventional generation capacity and take more responsibility in network operation. Due to flexible system structures, the renewable power plants shall act as conventional power plants providing wide range of controls. For example, during a system fault, the controls of the renewable power plants may contribute more to the fast recovery of the entire non-conventional power systems.

1.2.2 Grid-connected Wind Energy Conversion System

Currently, due to the increased penetration of the wind power into the electrical grid, the power system becomes more vulnerable and dependent on the wind energy production. The wind power generation system has some unique characteristics as compared to conventional fossil fuel generation systems:

1. The capacity of single wind turbine power generator is relatively small. Wind turbine installations are distributively over a geographical area much larger than typical fossil fuel plant. They can be quite remote from load centers.
2. Since the source of energy is wind, the production of electrical power from wind power generation system is intermittent by nature.
3. Conventional power generation plants all employ synchronous electrical generators. In contrast, wind power generation technologies utilize a variety of different types of electrical generators including induction generators and permanent magnet synchronous generators.

It is vital for the electric power industry to have a concise source of information that defines the distinctive characteristics of wind generation and how its impact on system performance is to be assessed through proper modeling and analysis. As a new grid-integrated power generation system, the reliability, stability and dynamic behavior of controls of the wind power generation systems are really great challenges for modern industrial engineers.

In this thesis, first of all, a coordination control structure will be built for the entire non-conventional power system. After that, wind turbine power generation system, the current most important renewable energy, will be employed in the research of non-conventional power system as a representative of the renewable

Figure 1.4. Control block diagram for non-conventional power system

power sources. Wind power will act as the only renewable power source existing in the non-conventional power system to be studied. Both the self-control of wind power generation system and the coordination control of global non-conventional power system with grid-integrated wind power generation system will be explored in details.

1.3 Motivation

Since that, in this thesis, the wind power is the unique renewable power source existing in the non-conventional power system, all the following discussions are carried out about the wind energy conversion system.

Related to the interconnection of WTPGS and the utility grid, the control objectives of the WTPGS should be considered from both sides: the WTPGS themselves and the power grid. From a WTPGS's point of view, it is desirable to capture as much power as possible. For example, in the case of normal operation of the grid-integrated WTPGS in power system (steady-state), WTPGS's main concern is its own control, including wind turbine Maximum Power Point Tracking (MPPT), real-time wind speed estimation, turbine pitch system control and sensorless control of WTPGS, etc.

For the WTPGS, on one hand, it is necessary to get accurate wind speed information. Accurate real-time wind information is great significance for the control of WTPGS and power system planning and operation. According to the aerodynamic model of wind turbine, the turbine extracted power is a cubic function with respect to the wind speed. If the wind speed is immediately considered as a constant, the captured power is positively proportional to a power coefficient which is represented with respect to the wind turbine Tip Speed Ratio (TSR) and the turbine pitch angle. In other words, as long as the real-time wind speed information is obtained, the optimal control commands of the turbine rotor speed and the pitch angle can be determined. Normally, wind speed anemometers are employed for the wind speed estimation.

On the other hand, high effective controllers for the power generator of WTPGS should be proposed to make the responses of the generator as fast as possible. As

electrical machine, in order to achieve satisfactory control performance, parametric variations, system uncertainties and unstructured dynamics should be overcome by the generator control of WTPGS. Since the current control performances directly relate to the generator performance, it is necessary to design high performance current controllers for the power generator of WTPGS. To obtain high control performance and system robustness, the advanced current controller must achieve fast dynamic response during the transient state, lower current ripple in the steady state and robustness against parametric variations and unstructured dynamics. Moreover, in variable-speed WTPGS, the turbine shaft rotational speed should be controlled optimally with respect to the variable wind speed in order to achieve Maximum Power Point Tracking (MPPT). Assume the real-time time-varying wind speed can be accurately estimated, the optimal rotor speed reference can be obtained easily which is rapidly variable following the variations of wind speed. Thus, WTPGS generator speed controller must be highly accurate and efficient in order to ensure fast and accurate speed control response.

However, on the perspective of power market and planning, theoretically, the power grid desires a wind farm to support stable power supply while does not negatively affect the global stability and reliability of the power systems. Even during a system fault, the control of the wind power plants cannot only guarantee the stability and ensure enough capabilities for their own low voltage ride through, but can contribute to the fast recovery of the entire power systems.

For the power grid side, considering the grid integration of WTPGS to the traditional power system, robust and efficient entire controllers should be employed in the power systems in particular consideration of high degree of nonlinearities, uncertainties and interconnections of synchronous generators (SG) and WTPGS.

Power system is a highly nonlinear uncertain system by nature. It may operate over a wide operating range. Electric power system is a complex network of individual components, including synchronous generators, renewable power sources, transformers, buses, switching devices, transmission lines and Flexible Alternating Current Transmission Systems (FACTS) devices. Since all the components are inherently nonlinear and can operate from a normal condition to an abnormal one, power system could operate over wide regions. In a traditional

power system, the system uncertainty are mainly caused by variations of load and network structures except the unmodeled dynamics and parametric uncertainties. Load level and dynamic behaviors change continuously due to small load events or abruptly due to large event [1][5]. The structure of electrical power network and its parameters are susceptible to change from time to time by severe disturbance, repair and failure and so on. Therefore, the controller design for modern power system should consider various nonlinearities and uncertainties over wide operating range. When the power system consists of a collection of several independent and spatially distributed dynamic subsystems with grid-connected wind power plants interconnected together. The power system global controllers should dominate the global controls of the power system with grid-connected WTPGS, simultaneously, can prompt the cooperation of the SGs and the WTPGS which may significantly affect the global stability.

Different control objective control objectives and actions must be employed under different operating ranges of power system. This thesis mainly focus on the two of most important control problems of power system which are real power/frequency regulation and reactive power/voltage regulation, respectively. However, the two control objectives sometimes are conflicting with each other when the system uncertainty exists. Due to that the transient stability and voltage regulation are both important properties of power system controls, the controllers were always designed separately to deal with the two problems. If two individual robust controllers are designed respectively for different objectives on different subsystems of the power system, the global power system stability cannot easily be guaranteed in theory. Thus, above all, the power system control should guarantee the system stability theoretically.

Moreover, the power system control scheme should fully takes into account the coordination and cooperation of the traditional power system and the WTPGS. The connection of large wind turbines to the grid has a large impact on grid stability, especially during transient period. It requires WTPGS to have satisfactory power control capability more similar to that of conventional generation capacity and more responsibility in network operation. The power control capability means mainly that the wind turbines have to share some of the duties

carried out traditionally by conventional power plants, such as regulating active and reactive power and performing frequency and voltage control on the grid. Wind farms connected at the transmission level shall act as a conventional power plant providing a wide range of controlling the output power based on transmission system operator's demands and also participation in primary and secondary control [6][7]. When the power system is subjected to a severe disturbance, how can the control of WTPGS both guarantee the stability of itself and make the WTPGS cooperate well with the global power system and helping to enhance the transient stability and voltage regulation capacity of the global system at the same time is an interesting research topic.

The conventional control block diagram of traditional power system with grid-connected wind energy conversion system is shown in Figure 1.5. In this thesis, the controls of the transmission system including the FACTs won't be discussed, because they were adequately studied in the previous literatures [8][9][10][11].

Figure 1.5. Control block diagram of power system with grid-integrated WECS

1.4 Literature review

This section reviews some literatures which address associated researches with this thesis.

Reset Control

The reset control was firstly proposed by Clegg [12] to overcome the limitations of linear control. This reset controller consists of an integrator and a reset law which resets the output of the integrator to zero when its input vanishes. The reset control is helpful in reducing the controller windup caused by integration. However, due to the fixed traditional reset law, performance improvement is limited. For example, before the first zero-crossing of tracking error, the traditional reset mechanism is not active, thus the rise time cannot be reduced and overshoot is in general inevitable because of inertia of physical elements. Actually, performance improvement does not come from blind resettings [13] but from the interaction between the baseline system and the reset mechanism. In [14], in order for that the reset mechanism cooperates better with the baseline system, a feedback was introduced in the reset law.

It has been proven that Reset control can achieve some control specifications beyond the ability of ordinary linear control and realize much better sensor noise suppression without degrading disturbance rejection or losing margins [15][16]. These advantages make reset control an effective and significant technique for transient performance improvement. Actually, reset control can be more general, for instance, the reset time and the amount of reset can both be designed so that the reset law and the baseline system can cooperate better with each other. The Optimal Reset Control (ORC) design propose an Optimal Reset Law (ORL) to dominate the reset time and amount of reset according to a previous defined performance index. The design of the reset law aims to minimize the performance index. The ORC problem is transferred to a Linear Quadratic Regulation (LQR) problem, so its solution can be derived by solving a Riccati equation. The ORC can retain fast response, shorten settling time and improve steady-state stability. Obviously, ORC is a kind of nonlinear control scheme. By transferring

the ORL design problem into Linear Quadratic Regulation (LQR) problem, the entire design process of ORC is based on linear theory [17]. Some specifications can be achieved and much better sensor noise suppression without degrading disturbance rejection or losing margins can be realized by ORC. Furthermore, the system stability is guaranteed inherently according to the ORC principle.

It is worth pointing out that there are many other techniques with similar design ideas as the ORC. For example, impulsive control [18] and sliding mode control (SMC) [19], etc. Reset control system is also known as a special impulsive system which is closely related to the optimal impulsive system [18]. Both the reset time and the reset value could be arbitrarily determined in the case of impulsive control [20][21]. If the reset value is fixed and the reset instant need to be determined, it belongs to the pulse width modulation control. On the other hand, the proposed reset control operates with fixed reset instant, and the reset value need to be determined. Since the purposed ORC is designed based on linear theory, it is superior than other impulsive controls for its simple design process and easy industrial application. SMC is also a widely used advanced control scheme. However, it is always dependent on complex design process, even hard to avoid chattering. In addition, SMC cannot be easily applied to multi-input/output systems as the ORC.

The ORC is designed and applied to meet the needs from the industry. Due to its linear design principle, the ORC is relatively easy. On the other hand, ORC is a kind of nonlinear control scheme which can achieve some specifications beyond the ability of linear controller. Recently, reset control design application has been expanded to hard disk drive servo systems [15][16]. However, the previous works are all based on the ORC design for a constant reference system mainly for the improvement of step transient response. More often the tracking reference for the system is time-varying and unpredictable.

PMS Machine Controls

Permanent Magnet Synchronous (PMS) machine has been widely used in industry. It has several advantages such as high power density, high torque to inertia ratio and relatively high efficiency [22][23]. Due to the good performance

of PMS machine, it has also attracted more and more attention of researchers in current, speed and position controls drive systems. In order to achieve satisfactory control performance, the field-orientation control is usually employed to eliminate its non-linearity and coupling [24].

[25][26][27] have shown some advanced and robust current controllers design methods. According to the results shown in these literatures, good current control performances have been achieved. Since the current control performances directly relate to the drive performance, it is necessary to design high performance current controllers for PMS machine. As the earliest applied control method, Proportional-Integral (PI) control technique was regarded as an useful scheme in PMS machine current control [28]. With its simple structure, good stability, high reliability and easy parameters adjustment, PI control technique is still one of the main technology in industrial control. The design of perfect performed PI controllers are highly dependent on precise mathematical model of PMS machine. However, uncertainties of PMS machine caused by parametric variations, modeling errors and external load torque disturbances are negative influences for the precise modeling. When too big or high-frequency uncertainties exist in PMS machine, the current control performances may not be ensured by conventional PI schemes, so as the stability.

To obtain high control performance and system robustness, the advanced current controller must achieve fast dynamic response during the transient state, lower current ripple in the steady state and robustness against parametric variations and unstructured dynamics. The predictive current control is proposed in [29]. Although it gives an ideal response for the basic modeling, its anti-uncertainties and anti-disturbances capabilities are limited. Moreover, the predictive control requires too complicated design processes. Sliding Mode Control (SMC) is also widely used [30]. The limitation of SMC is that the control gains are difficult to be designed when the PMS machine includes uncertain parameters. Furthermore, robust current controllers are proposed in [27][31].

Actually, robust and anti-uncertainty capabilities are the key points of PMS machine current control. And real-time estimations of the disturbance and uncertainty are proven to be effective for solving these problems by feed-forward

compensations [32][33]. [34][35][36] show some methods to estimate the armature inductances of PMS machine. However, the uncertainties caused by the variations of the stator-winding resistance and magnet's flux linkage are not considered. [37][38][39] propose some adaptive estimation schemes for parameters. The time-varying parameters were treated as extra states in the PMS machine. Nevertheless, the estimation stability, errors and speed are still problems which should be solved. In [27][31], the PMS machine robust current control are achieved by observing the uncertainties through adaptive internal models. However, the used steepest descent method which may decrease the estimation and response speed even affect the control performance. Although some effective disturbance observers for PMS machine were proposed in [32][33][40], they may not suitable for industrial systems due to the complicated computations and control algorithms. Additionally, adaptive and real-time tuning controls [41] are also explored against the uncertainties. Note that the design of these methods are always too complex and hard to be extended to different systems. In [30], sliding mode based uncertainty estimators are proposed in the SMC loops. However, it is hard to give the proper parameters of SMC and estimators which makes its extension application inconvenient.

The speed controllers design of PMS machine drive systems have also been investigated. The adaptive control [42], self-tuning control [43][44], nonlinear control [45][46][47], neural network control [48] and fuzzy control [49][50] are considered for the speed controllers design, respectively. These proposed methods have more or less realized the improvements of performance of PMS machine speed control. The conventional controllers for PMS machine speed control are proportional plus integral (PI). The PMS machine are commonly used for variable speed machine drive systems to produce high-efficiency, high-reliability, and low-cost drive performance. Most of the PMS machine PI speed control parameters are tuned using symmetrical optimum method [51], pole placement method [52] or even trial-and-error method.

From the application perspective, PMS machine speed controllers are designed not only to focus on the robust and anti-uncertainty performances, but also to meet the following three requirements: 1) to track the time-varying speed ref-

erence rapidly and accurately; 2) to reduce mechanical and electrical stresses during transient period; 3) to satisfy anti-windup characteristic of controllers. Essentially, it is a problem of the PMS machine dynamic speed control. A hybrid fuzzy-PI controller is proposed in [53]. As aforementioned, the transient performance of PI are limited and affected by load disturbances and speed variations. The fuzzy controller is employed in [53] mainly to improve the transient performance. Evidently, fuzzy logic and the combination switching functions will make the controller design complicated. Furthermore, the time-varying speed reference isn't considered in [53]. The load and torque in the power train of the AC electric machines during transient period should be reduced to protect the electric drive system [54]. In [55], some anti-windup strategies for PI speed controllers of PMS machine are proposed. There is an application of anti-windup scheme used for the speed controller of PMS machine [56]. These strategies are all realized by adding extra loops into the conventional PI. But no analysis was given about the effects on stability of the additional loops.

Moreover, design robust and high-performance position controllers for PMS machine drive system is one of the major topics for PMS machine applications. Kinds of position controllers design methods have been proposed in literatures, such as optimal controller [57], precise position controller with disturbance observer [40][39], new robust controller [58], fuzzy optimization techniques [59][41] and adaptive controller [60][61]. These papers have shown their advantages respectively in the position control.

As a multi-loop control system, set the normal triple loops system as an example, PMS machine position controller design processes are complicated even require repeated trial and error verification. The reason is that the design of the controllers parameters should consider the response speeds, frequency bandwidth of the interaction and coordination among each loop of regulators. Theoretically, the more loops the control system has, the more difficult the controllers design will be. Furthermore, each of the three loops are based on feedback control. It means that more cumulative measurement errors and noises will be taken into the control system in the triple loops control system than in double loops and single loop control systems. The measurement errors and noises will cause the

position ripple even make the steady-state stability worse.

Originally, three loops Proportional-Integral (PI) controller is designed for PMS machine position control. However, the influences of uncertainties of PMS machine caused by parametric variations and external load torque disturbances must be inevitably considered in precise position control. The design of perfect performed PI or PID controllers are high dependent on precise mathematical model of PMS machine. Then the good position control performance may not be ensured by conventional PI or PID control schemes. Several nonlinear controllers were proposed to deal with the problems of conventional PI or PID. For instance, SMC, fuzzy optimization control and adaptive control are applied in [62][63], [59][41] and [60][61]. Although the position control performances have been significantly improved by these control schemes, the controller structures are always hard to be arranged in order to guarantee the whole system stability.

Wind Speed Estimation and WTPGS Sensorless Control

Recently, wind farms prefer to employ DFIG wind energy conversion system, which has several advantages when compared to fixed-speed wind energy conversion system [64][65][66][67]. These advantages, including speed control, reduced flicker, and four-quadrant active and reactive power capabilities, are primarily achieved via control of a rotor side converter, which is typically rated at around 30%-35% of the generator rating for a given rotor speed variation range of 25%.

The wind power generation system installed with PMSG has received increasing attention [68][69][70][71]. The PMSG can provide high-efficiency and high-reliability power generation, since there is no need for external excitation and no copper losses in the rotor circuits. In addition, PMSG WECS with power electronic converters can be easily controlled. It is worth pointing out, as a permanent magnet synchronous machine, the PMSG is small in size with high power density, which reduces the cost and weight of wind turbines. Moreover, this variable-speed wind turbine concept with its gearless construction is also an appropriate solution for offshore wind farms, as it requires low maintenance, reduced losses and costs, and at the same time has high efficiency and good controllability [72][73].

In the operating wind speed range, in order to achieve the MPPT of wind turbine, the turbine shaft rotational speed should be adjusted optimally with respect to the variable wind speed. Such turbine rotor speed control should base on the real-time information of wind speed [74]. Normally, wind speed anemometers are employed for the wind speed estimation. However, high cost of precise anemometer limits the extensive use of this equipment. For example, the surrounded installed anemometers cannot provide adequate and accurate wind speed information for every wind turbine in wind farms [75] and precise anemometer doesn't deserve to be used for a distributed small WTPGS.

Recently, the mechanical sensorless MPPT controls have been reported in literatures. Some of the literatures propose turbine maximum power extraction controls independent of wind speed information. It can avoid the difficulties of wind speed measurements or estimations. For instance, a fuzzy-logic-based control is given in [76]. The fuzzy control adaptively performs an online incremental/decremental searcher to drive the WTPGS rotational speed until the extracted power maintains at its maximum condition. However, this method associates with the system's dynamics to some extent. If the wind speed is in sharply changing, the relative slow search speed of this method may not make the maximum power extraction realized opportunely. Some other literatures show their sensorless MPPT control methods based on wind speed on-line estimations [77][78]. The lookup table can be used as one estimation method [78] however it requires too much memory space to ensure the estimation accuracy and may result in a time-consuming search for the solution. The real-time calculation method of the nonlinear function roots [77] was used to solve the inverse function problem for estimating the real-time wind speed. However, it may result in complex calculations and high computational burden for the computer, therefore, reducing system performance. Alternatively, a novel algorithm based on the theory of support-vector regression (SVR) for wind-speed estimation in wind power generation systems is proposed in [79]. After the off-line training, a specified model is obtained to determine the wind speed online from the instantaneous turbine information. The SVR model is too slow due to the large number of support vectors generated. Especially, it is much slower than feed-forward neu-

ral networks. The high performance of using SVR to make real-time estimation cannot be guaranteed [80].

Artificial neural networks (ANNs) are used to implement nonlinear time-varying input-output mapping instead of the lookup table and SVR based mapping in some literatures [75][81]. It has been proven that ANN methods can overcome the drawbacks of the previous mapping methods. [75] propose a ANN based wind speed estimation method for a direct-drive small WTPGS. This method achieves fast and smooth wind speed estimation based on an ideal WTPGS. However, the power loss of the wind turbine is neglected which sometimes cannot be ignored. Moreover, and the method cannot be applied on variable-pitch WTPGS. In [81], Gaussian radial basis function network (GRBFN) wind speed estimation based sensorless output maximization control is proposed with considering the pitch activation and the power loss. Although the estimation and control performances are both good, the performance of this method can still be improved to some extent. For example, intensive sampling data of the three training inputs always include too much data. In order to finish the training, the number of the data should be decreased which will reduce the regression accuracy of the ANN theoretically. Furthermore, this method is not robust against the air density variations and no test is given when the pitch angle system is activated. The accuracy of ANN based wind speed estimation can be improved mainly by two general ways. One is using advanced training algorithm to improve the estimation accuracy. The other is simplifying the structure of ANN as much as possible which can accelerate the estimation speed and improve the estimation accuracy as well [80].

ELM is a perfect algorithm [82][83][84][85][86]. It takes lots of advantages such as fast training speed, high training accuracy, easy design process and no manual tuning. Extreme Learning Machine (ELM) ANN based input-output mapping can be used for the WTPGS wind speed on-line estimation and sensorless controls. In the WTPGS controls, the commonly used pitch angle control strategy is the classical proportional-integral-derivative (PID) control [87][88]. It is always designed based on a linearized wind turbine model around a specific operating point [89]. So its global robust fast response and stability may not be ensured.

The well trained ELM ANN controller could replace the conventional controller with the advantage of increasing execution speed precisely.

Power System Transient Stabilization Control and Voltage Regulation ■

Power systems are large and nonlinear systems. The transient stability of a power system is its ability to maintain synchronism when subjected to a severe transient disturbance. And the voltage regulation problem means that power system should regulate the voltage quality to its prefault steady condition after the severe transient disturbance. The transient stability and voltage regulation are both important properties of power system controls when subjected to a severe transient disturbance [5][1].

Due to that the conventional controllers are designed based on approximately linearized models, such as power system stabilizers, they may only deal with local stability around an operating point. Nonlinear control theory is applied to cancel the power system nonlinearity which can alleviate the uncertainties caused by operating point variations and improve the system stability and control performance as well. As mentioned in [90][91][92][93], feedback linearization approaches are successfully used for design of transient controllers and voltage regulators in power systems. In [94], using the direct feedback linearization (DFL) technique, nonlinear DFL excitation controllers were proposed which achieve transient stability improvements. However, the voltage recovery may not be realized in such transient stabilizing controls, because the power system structure may considerably vary from its pre-fault one. Since the power quality is an important property of power supply, the voltage feedbacks are introduced in the DFL transient controllers' design to achieve the post-fault voltage regulations in [95][96]. Although the post-fault voltage regulation problems are realized by such voltage controllers, they are only effective around the normal operating point. When a large disturbance occurs, the power systems may not be survived by the voltage controllers.

Obviously, different control objectives should be concerned for the power system control in different operating region. In the power system transient stabilizing

control, on one hand, transient controller should be considered more when the operating point is far away from its pre-fault one. It ensures that the power system doesn't lose synchronism when a large disturbance occurs. On the other hand, voltage controller is needed when the system operates around the operating region to realize the voltage regulation problem. [97][98][99] proposed coordinated controllers to deal with this problem. By given switching laws, the activations of the transient controller and the voltage regulator can be switched and cooperated with each other. However, the switching strategy may cause a discontinuity of the system behavior. In [100][101], global controllers were employed to coordinate the transient stabilizer and voltage regulator. Soft switching membership functions are used instead of the switching laws which makes the actions of controllers more smooth and robust. The simulations shown have demonstrated the better performance of global controllers. Unfortunately, it seems that the global control performance is too dependent on the performance of the soft switching membership function since an inappropriate one may cause the loss of synchronism of the SG. Actually, the system stability of the global control cannot be ensured in theory.

There are still some problems existing in the power systems global control. For example, two individual robust controllers should be designed respectively for different subsystems of the power system in order for transient stabilizing control and voltage regulation. Theoretically, the global stability cannot be guaranteed by each of the two controllers unless the appropriate switching membership function can be found. To some extent, it can only rely on coincidence. What's more, The design processes are inevitably complicated, especially in large, multi-machine interconnected power systems. The other is when other power regulation devices are simultaneously installed in the system for the same purposes of transient stability enhancement and voltage regulation [102][103], the global controlled power system and the power regulation device cannot cooperate well with each other. For instant, the energy storage systems and the renewable power sources, etc. The global controller should be redesigned to ensure the stability of the new power system. Furthermore, there is still no theory proposed for finding the optimal controller in this condition.

WTPGS Stabilization Control

During a grid fault, the original power system and the grid-connected WTPGS will impact each other inevitably and complicatedly [104]. The transient power system controls needs mainly that the wind turbines have to share some of the duties carried out traditionally by conventional power plants, such as regulating active and reactive power and performing frequency and voltage control on the grid. For the WTPGS controls, pitch controller for wind turbine is always employed in the WTPGS transient stabilization control during the power system transient process [5][105]. The main purpose of the pitch controller for a wind turbine is to maintain WTPGS output power at its rated value when the wind speed goes beyond the rating [106]. However, it can control the wind turbine rotor speed by adjusting the wind mechanical torque with respect to the PCC bus frequency which indirectly provides a method for WTPGS stability enhancement. The commonly used pitch angle control strategy is always designed based on a linearized wind turbine model around a specific operating point [87][88][89]. So its global robustness and fast response may not be the best for the highly nonlinear wind turbine system [107]. Alternatively, some terminal installed energy storage systems are employed in the WTPGS to improve wind generator stability, such as the super-conducting magnetic energy storage (SMES) unit [108] [109][102][110] and dynamic reactive compensation [111]. However, high cost of the SMES limits the extensive installation of this device [112]. More often, it prefers that the stability enhancement of WTPGS can be realized by its self control. Overall, these existing methods are based on the WTPGS information locally which may lack coordination and integration with the rest of the power system.

1.5 Objectives

The main research objectives of this thesis are represented in the Figure 1.6.

PMS machine has been used in wind turbine power generation system as permanent magnet synchronous generator and pitch positioning servo motor. Advanced controls for PMS machines is the first object of research in this thesis. In

Figure 1.6. Thesis objectives

order to achieve satisfactory control performance, it is necessary to design high performance controllers for current, speed and position controls of PMS machine systems.

Since the current control performances directly relate to the dynamic performance of PMS machine, it is necessary to design high performance current controllers for PMS machine. Actually, robust and anti-uncertainty capabilities are the key points of PMS machine current control. To obtain high control performance and system robustness, the advanced PMS machine current controller must achieve fast dynamic response during the transient state, lower current ripple in the steady state and robustness against parametric variations and unstructured dynamics.

Design robust and high-performance position controllers for PMS machine drive system is also one of the major topics for PMS machine applications. Some of the industrial positioning system is based on PMS servo machine which can be applied on machine tools, wind turbine pitch system and etc.. The influences of uncertainties of PMS machine caused by parametric variations and external load torque disturbances must be inevitably considered in precise position control. The design of perfect performed PI or PID controllers are high dependent on precise mathematical model of PMSM.

In the case of speed control for PMS machine, the controller should not only overcome the robust speed control for PMS drive systems, but also ensure high-performance speed control when the PMS machine acts as a power generator such as being used in WTPGS installed with Permanent Magnet Synchronous Generator (PMSG). From the application perspective, PMS machine speed controllers are designed not only to focus on the robust and anti-uncertainty performances, but also to meet the following three requirements: 1) to track the time-varying speed reference rapidly and accurately; 2) to reduce mechanical and electrical stresses during transient period; 3) to satisfy anti-windup characteristic of controllers.

In the wind speed estimation and sensorless control for WTPGS, in order to achieve the MPPT of wind turbine, the turbine shaft rotational speed should be adjusted optimally with respect to the variable wind speed. Such turbine rotor speed control command should base on the real-time information of wind speed. The accurate real-time wind speed for individual wind turbine always cannot be obtained by the wind speed anemometers, especially in big wind farms containing several distributed wind turbines. It is significant to propose effective, accurate, fast and robust wind speed estimation method for each wind turbine locally without wind speed sensor. Then the optimal rotational speed command and pitch angle command both can be easily calculated.

The transient stability and voltage regulation are both important properties of power system controls when the power system subjected to a severe transient disturbance. However, the two control objectives sometimes are conflicting with each other when the system uncertainty exists. The controller for both the transient stabilization control and voltage regulation of synchronous generator should be coordinated in one control structure without undermining the stability of the entire power system. When other power regulation devices, for instant, the energy storage systems, are simultaneously installed in the power system for the same purposes of transient stability enhancement and voltage regulation, the controller should make the traditional power system and the power regulation device cooperate well with each other.

In a power system with grid-connected WTPGS, it is important to analyze the

transient stability and voltage regulation problems of both the original SG and the grid-connected WTPGS. The conventional controllers for SG and WTPGS are always designed locally and separately. Due to the lacks of integration and coordination, whether the individual controllers can provide good entire system control performances in a complex power system (including SG and WTPSG) is questionable. The controllers should make the SG and WTPGS cooperate well with each other, especially in the condition when severe disturbances occur in the complex power system.

1.6 Main contributions of the thesis

The main contributions of this thesis are summarized in this section which mainly concerns the following two parts (Fig. 1.7):

- Self-control of wind turbine power generation system. In order to achieve optimal self-control performance of WTPGS, some advanced controls for the electric machine in the WTPGS, the MPPT of the WTPGS, the real-time wind speed estimation and the sensorless control scheme of the WTPGS are proposed in the self-control part.
- Coordination control of the global non-conventional power system. A coordination control scheme based on power system backstepping control is proposed for both of the conventional centralized power generators and the grid-integrated WTPGS in the non-conventional power system.

1.6.1 Self-control of Wind Turbine Power Generation System

Robust Control of Permanent Magnet Synchronous Machine

Real-time estimations by observer of the disturbance and uncertainty are proven to be effective for solving the robust and anti-uncertainty problems by

feed-forward compensations in the PMS machine controls. Linear two pieces cascaded coupled uncertainties observer is proposed to estimate the uncertainties of PMS machine caused by parametric variations and external load torque disturbances. With feed-forward control technique, the uncertainties can be compensated in control loop. Since the observer can easily be ensured relatively fast convergent by the design, the PMS machine whole system control stability is guaranteed.

As a nonlinear control technique, the ORC method is proposed to unify the controller and the observer designs together. The ORC can retain fast response, shorten settling time and improve steady-state stability. Actually, performance improvement comes from the interaction between the baseline system and the reset mechanism [16]. The ORC propose an ORL to dominate the reset amount according to a previous defined performance index. The ORC problem can be

Figure 1.7. Main contributions of the thesis

transferred to a regular LQR problem. Obviously, although the joint ORC scheme is designed based on linear principle, it can achieve some control specifications beyond the ability of ordinary linear control and realize much better system robustness and sensor noise suppression without degrading disturbance rejection or losing margins [14].

Since it is designed based on linear theory, the ORC is superior than some other advanced controls for its simple design process and easy application. With the application of the ORC, the two loops uncertainties observer can achieve fast and dynamic convergent to the real-time uncertainties. The decoupling terms for uncertainty feed-forward compensations can operate more effectively in the control loops. Then the ORC is applied on the baseline PI controllers which are designed on the basic model of PMSG machine.

Optimal Reset Control of PMSG WTPGS

The discrete-time ORC design proposed in this thesis is an advanced reset control. The proposed discrete-time ORC fits for time-varying reference and can be easily applied in industrial discrete-time control system. Through merging the uncertainty and time-varying of reference into the ORL, the proposed ORC law can overcome perfect tracking problem effectively and this ORC law is redesigned of a discrete-time form.

The discrete-time ORC is applied on the generator of WTPGS to improve the transient and dynamic responses. Take the PMSG WTPGS as an example, the baseline controllers for PMSG WTPGS are PI. The PMSG are commonly used for small variable-speed wind turbines to produce high-efficiency, high-reliability, and low-cost wind power generation. The control of the generator side 3-phase converter is always based on the d, q double axis dynamic model [51]. The required d, q components of the rectifier voltage vector are derived from two PI inner-loop current controllers. In order to achieve the MPPT of wind turbine, an outer-loop PI shaft speed controller should be employed in the q -axis component. Similarly, an outer-loop PI for minimum power losses should be employed in the d -axis component.

A PMSG can be regarded as an inverse Permanent Magnet Synchronous Motor (PMSM). Due to the intermittent characteristic, wind speed may go up and down around the cut-in wind speed frequently. Therefore, WTPGS will probably start up and stop frequently. The load and torque in the power train of the AC electric motor during startup should be reduced to protect the electric drive system [54]. The ORC reduces the mechanical stress on the motor and shaft, as well as the electrodynamic stresses on the attached power cables and electrical distribution network, extending the lifespan of the system beyond the ability of PI. In [55], some anti-windup strategies for PI speed controllers of PMSM are proposed. There is an application of anti-windup scheme used for the speed controller of PMSG [56]. These strategies are all realized by adding extra loops into the conventional PI. But no analysis was given about the effects on stability of the additional loops. The ORC can solve the controller windup problem itself without affecting the stability. Moreover, according to the wind speed variation, wind turbine rotational speed should be adjusted to make the tip speed ratio maintained at the optimal value. Then the maximum power coefficient can be achieved for MPPT. Essentially, this is a problem of the PMSG/PMSM dynamic speed tracking. A hybrid fuzzy-PI controller is proposed in [53]. As aforementioned, the transient performance of PI are limited and affected by load disturbances and speed variations. The fuzzy controller is employed in [53] mainly to improve the transient performance. Evidently, fuzzy logic and the combination switching functions will make the controller design complicated. The ORC can realize the same performance as the hybrid fuzzy-PI based on a relatively simple linear design process. Furthermore, the time-varying speed reference isn't considered in [53].

ELM Wind Speed Estimation and WTPGS Sensorless Control

Artificial neural networks (ANNs) are used to implement nonlinear time-varying input-output mapping instead of the lookup table and SVR based mapping in some literatures [75][81]. It has been proven that ANN methods can overcome the drawbacks of the previous mapping methods. The accuracy of ANN based wind speed estimation can be improved mainly by two general ways. One

is using advanced training algorithm to improve the estimation accuracy. The other is simplifying the structure of ANN as much as possible which can accelerate the estimation speed and improve the estimation accuracy as well. In this thesis, ELM ANN based input-output mapping of a specific design characteristic of the wind turbine is proposed for the WTPGS sensorless control with wind speed on-line estimation. ELM is a perfect algorithm. It takes lots of advantages such as fast training speed, high training accuracy, easy design process and no manual tuning. By observing the information of the wind turbine, the two real-time inputs of ELM ANN can be calculated. This process not only reduces the number of training inputs to reduce the ANN structure and increase the estimation accuracy, but makes the ELM mapping independent of the air density. Thus the wind speed estimation in the entire operating region can be achieved effectively and robustly and the maximum turbine power extraction control is dominated from this estimated wind speed. Additionally, when the generated power exceeds its rated value, an ELM based pitch angle controller puts into operation to limit the captured power. The commonly used pitch angle control strategy is the classical proportional-integral-derivative (PID) control. It is always designed based on a linearized wind turbine model around a specific operating point. So its global robust fast response and stability may not be ensured. The well trained ELM ANN controller could replace the conventional controller with the advantage of increasing execution speed precisely.

1.6.2 Coordination Control of the Global Non-conventional Power System

Backstepping Control Scheme for Non-conventional Power System

In this thesis, a backstepping control scheme is proposed to deal with the transient stabilization control and voltage regulation problems for power system. The backstepping technique [113] is employed to integrate the transient stabilization control and voltage regulation of SG in one control structure. Overall, this method is based on the best pursuit of system input-output power balance in considering the entire power system as a whole. Above all, the backstepping

technique can guarantee the system stability theoretically. From the power point of view, the power angle and frequency controls of each SG are decoupled from its electrical power and terminal voltage controls. By a defined error index of the power angle and frequency for each SG, a variable electrical power objective can be proposed to dominate a correction law for the existing voltage regulator. Thus, both the transient stabilization control and voltage regulation of each S-G will be achieved at the same time within one backstepping control structure. Obviously, the backstepping control structures for each of the SG are robust and decentralized. Moreover, the proposed power objective index can offer a power reference for the power regulation device. And the device may work with its decentralized controller to track the power reference which provides good coordination and controls between the original power system and the power regulation device. A super-conducting magnetic energy storage (SMES) unit is employed as the power regulation device for the test in this thesis. The proposed backstepping control scheme is implemented in the simulations on both a single-machine infinite bus power system and a multi-machine power system with respect to different transient faults.

With the proposed backstepping control scheme, a coordination control scheme is proposed for the WTPGS in a power system with grid-connected WTPGS. In such a power system, it is important to analyze the transient stability and voltage regulation problems of both the original SG and the grid-connected WTPGS. With the variable electrical power objective and the SG real electrical output power of the backstepping controlled original SG system, a reasonable output power reference for WTPGS can be determined. According to the WTPGS modeling, it is easy to derive the corresponding real-time turbine captured power reference. Subsequently, the problem goes to make the WTPGS tracking the turbine captured power reference as fast as possible by its decentralized rotor speed controller and pitch controller under the condition of ensuring the wind turbine stability. According to the torque characteristic and torque coefficient curve of wind turbine, one can easily determine the stable operation region of the wind turbine [106]. In this thesis, the artificial neural networks (ANNs) input-output mapping method is implemented to build dynamic searching laws for searching the rotor

speed reference and pitch angle reference within the stable operation region, respectively. Obviously, wind turbine stability can be guaranteed by the proposed control scheme. The proposed control scheme is also verified in the simulations on both a single-machine infinite bus power system with a WTPGS installed at the Point of Common Coupling (PCC) and a multi-machine power system with a WTPGS installed at one of the PCC of one SG with respect to different transient faults.

1.7 Organization of the thesis

There are eight chapters in this thesis which is organized as follows:

Chapter 1 introduces the motivation of this research, some related literature reviews about this topic, the objectives and the major contributions of this thesis.

Chapter 2 presents the mathematical modelings of PMS machine systems, including PMSM drive system and PMSG, the wind turbine power generation system, SMIB power system and multi-machine power system, respectively. The basic concept of the DFL technique and its application on both the SMIB power system and multi-machine power system. These models are utilized to design and analyze PMS machine controllers, wind speed estimator and sensorless controller of wind turbine power generation system and power generator excitation controllers in all the chapters in the next.

Chapter 3 gives the concepts of continuous-time ORC mainly for the improvement of control system step transient response. For the system with time-varying and unpredictable tracking reference, the discrete-time ORC design is proposed in this chapter. The proposed ORC fits for time-varying reference and can be easily applied in industrial discrete-time control system.

Chapter 4 presents the following: 1) advanced position and current controls with uncertainties estimation based on ORC principle for PMSM drive systems; 2) discrete-time ORC speed controller for PMSG system with time-varying reference to reduce the overshoot of the step response without degrading other specifications, suppress controller's saturation effectively and accelerate the tracking speed with less error.

In Chapter 5, the PMSG WTPGS MPPT controller is designed based on the discrete-time ORC principle. In fact, it is a special application of the discrete-time ORC speed controller for PMSG system on the PMSG WTPGS. Afterward, ELM based input-output mapping of a specific design characteristic of the wind turbine is proposed for the WTPGS wind speed real-time estimation. By the estimated wind speed, the sensorless maximum turbine power extraction control and pitch control are both introduced. An ELM based pitch controller is used to replace the conventional pitch controller as well.

Chapter 6 proposes the backstepping control design procedure for the power system transient stabilization control and voltage regulation. The related control technique is able to solve the power system nonlinearities, uncertainties and interconnection problems among each power component. It also provides a coordination method of different devices with different controllers for the same power system transient stabilization control and voltage regulation control tasks. The proposed backstepping control is applied on the conventional SMIB power system, the SMIB power system with PCC installed SMES unit and multi-machine power system, respectively. When considering a multi-machine power system, the system uncertainties and interactions are taken into account in the design of decentralized controllers for every SG excitation.

In Chapter 7, the backstepping control scheme illustrated in the previous chapter is proposed to deal with the transient stabilization control and voltage regulation problems in non-conventional power systems with PCC bus integrated WTPGS. Above all, the backstepping technique can guarantee the system stability theoretically. Subsequently, the backstepping control scheme can transfer the transient stabilization control and voltage regulation problems of WTPGS to making the rotational speed and pitch controls of the WTPGS more active by its decentralized rotor speed controller and pitch controller under the condition of ensuring the wind turbine stability. It provides an integrated and coordinated control structure between SGs and WTPGS. Overall, controller designs for the WTPGS is based on the best pursuit of system input-output power balance in considering the entire non-conventional power system as a whole which should consider the on-line information of all the SGs during the transient period.

Chapter 8 concludes the thesis and recommends tasks for future research.

System Modeling

2.1 Introduction

This thesis aims at developing advanced control schemes for WECS and large scale power systems with wind power generation sources. Since WECS and large scale power systems are all nonlinear and complex systems, it is important to develop proper mathematical models of the systems before analysis and the systematical controller designs.

In the modeling of WECS, there are mainly two parts which should be considered. One is the wind turbine mechanical dynamics, and the other is the electrical part of the generator which includes the activities of power electronic devices. This project takes the wind turbine system installed with PMSG as a representative of the WECS, and studies the controls of WECS in-depth. Duo to its excellent performance, PMSG WECS has received much interest and considerable attention of researchers and is becoming the trend of development in the future wind power generation system. Through research accumulations, the modeling of the PMSG WECS has reached a consensus.

As known, power systems are large scale, highly nonlinear and interconnected dynamic systems. In order to represent the entire dynamics of a power system, very high order and high nonlinear model should be employed. However, it is unfeasible and unnecessary for power system control designs and simulations.

Thus, in this study, approximated simple power system models are employed which preserve the primary transient characteristics. The modeling of SMIB power system and multi-machine power system will be introduced in this chapter. The transformer magnetizing branch is ignored and only the leakage reactance is represented. The AC transmission lines are represented by equivalent π circuit with charging capacitance at both sides. To reduce transmission network, the electrical loads are further represented by constant impedance [1]. Three phase balanced systems are assumed and represented on a per-phase basis.

The following sections of this chapter are arranged as: the modelings of PM-S machine systems, including PMSM drive system and PMSG, are discussed in Section 2.2; the wind turbine system model is described in Section 2.3; the mathematical models of SMIB power system and multi-machine power system are introduced in Section 2.4 and Section 2.5, respectively; in Section 2.6, DFL technique is employed to linearize both the SMIB power system and multi-machine power system.

2.2 Permanent Magnet Synchronous (PMS) Machine

A PMS Machine is an electrical machine where the magnetic field of the rotor is produced by a permanent magnet instead of a coil. It is known as synchronous machine because the speed of the rotor must always match the stator frequency [114]. The advantages of PMS Machines are:

1. do not require additional DC supplies for producing the rotor magnetic fields;
2. avoid the use of slip rings, hence are simpler and maintenance free;
3. do not require condensers for maintaining the power factor as the induction machines.

Since PMSG WECS is used as the main object of study for WECS and PMSG can be regarded as power inverse system of PMSM, controls of both the two

electrical machines (PMSM and PMSG) have been studied in this thesis. Their mathematical models are also given respectively.

2.2.1 PMS Motor (PMSM) Drive System

In this subsection, the electrical and mechanical models of PMSM are shown. According to Park Transformation (PT), transfer abc 3-phase voltages and currents to $dq0$ systems. The 0 term represents imbalances in the a, b, c phase.

The PT transformation matrix can be described as,

$$T_{abc} = \frac{2}{3} \begin{bmatrix} \cos \theta_m & \cos(\theta_m - \frac{2\pi}{3}) & \cos(\theta_m + \frac{2\pi}{3}) \\ \sin \theta_m & \sin(\theta_m - \frac{2\pi}{3}) & \sin(\theta_m + \frac{2\pi}{3}) \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{bmatrix}. \quad (2.1)$$

So obtain the voltages and currents in $dq0$ systems,

$$\begin{aligned} v_{dq0} &= [T_{abc}] v_{abc} \\ i_{dq0} &= [T_{abc}] i_{abc}. \end{aligned} \quad (2.2)$$

The electrical dynamic model of PMSM in its rotating synchronous reference frame is given as [22][23]

$$\begin{aligned} v_{sd} &= R_m i_{sd} + L_m \frac{di_{sd}}{dt} - L_m \omega i_{sq} \\ v_{sq} &= R_m i_{sq} + L_m \frac{di_{sq}}{dt} + L_m \omega i_{sd} + \omega \psi \end{aligned} \quad (2.3)$$

where R_m is the stator resistance. ψ is the magnet flux and ω is the motor electric speed. i_{sd} and i_{sq} are the d -axis and the q -axis currents respectively, and v_{sd} and v_{sq} are the d -axis and the q -axis voltages. For simplification, the inductances of d, q axes are set equal $L_{sq} = L_{sd} = L_m$. If the pole pair number is p , then $\omega = p\omega_r$. ω_r expresses the rotor speed.

The electromagnetic torque is given as

$$T_e = \frac{3}{2} p (\psi i_{sq} - (L_{sd} - L_{sq}) i_{sd}) = \frac{3}{2} p \psi i_{sq}. \quad (2.4)$$

According to (2.4), the electromagnetic torque is only relevant to the quadrature current component.

The torque dynamic equations of PMSM mechanical model can be described as an one mass lumped system,

$$\begin{aligned}\dot{\theta}_r &= \omega_r \\ J_m \dot{\omega}_r &= -D_m \omega_r + T_e - T_L\end{aligned}\tag{2.5}$$

where T_L is the external load torque on the rotor in the opposite direction of the electromagnetic torque. θ_r is the rotor position. D_m and J_m are the damping constant and the moment of inertia, respectively [40][39].

2.2.2 PMS Generator (PMSG)

The PMSG is an a, b, c phase AC electric system. Transferring it to DC system by the park transformation as shown in Subsection 2.2.1. The dqo system dynamic model of a PMSG in its rotating synchronous reference frame is [51][115]

$$\begin{aligned}v_{sd} &= -R_s i_{sd} - L_s \frac{di_{sd}}{dt} + L_s \omega i_{sq} \\ v_{sq} &= -R_s i_{sq} - L_s \frac{di_{sq}}{dt} - L_s \omega i_{sd} + \omega \psi\end{aligned}\tag{2.6}$$

where R_s is the generator resistance. ψ is the magnet flux and ω is the generator electric speed. The inductances of d, q axes are equal $L_{sq} = L_{sd} = L_s$, If the pole pair number is p , then $\omega = p\omega_g$. And ω_g expresses the rotor speed.

The electromagnetic torque is given as

$$T_{em} = \frac{T_g}{p} = \frac{3}{2}p(\psi i_{sq} - (L_d - L_q)i_{sd}) = \frac{3}{2}p\psi i_{sq}.\tag{2.7}$$

According to (2.7), the electromagnetic torque is only relevant to the quadrature current component. The torque dynamic equation on the shaft can be de-

scribed as an one mass lumped system [116],

$$J_g \frac{d\omega_g}{dt} = -K_g \omega_r + T_m - T_g. \quad (2.8)$$

where T_g is the electrical torque to the PMSG shaft on the opposite direction of mechanical torque T_m , K_g and J_g are damping constant and the moment of inertia, respectively. The generated power is given as,

$$P_g = T_g \omega_g. \quad (2.9)$$

2.3 Wind Turbine System

A wind turbine is a device that converts kinetic energy from the wind into mechanical energy. If the mechanical energy is used to produce electricity, the device may be called a wind generator or wind charger. Developed for over a millennium, today's wind turbines are manufactured in a range of vertical and horizontal axis types. The smallest turbines are used for applications such as battery charging or auxiliary power on sailing boats; while large grid-connected arrays of turbines are becoming an increasingly large source of commercial electric power.

Wind turbines can rotate about either a horizontal or a vertical axis, the former being both older and more common. Horizontal-axis wind turbines (HAWT) have the main rotor shaft and electrical generator at the top of a tower, and must be pointed into the wind. Small turbines are pointed by a simple wind vane, while large turbines generally use a wind sensor coupled with a servo motor. Most have a gearbox, which turns the slow rotation of the blades into a quicker rotation that is more suitable to drive an electrical generator. In this thesis, the eyes are only focused on the modeling and applications of HAWT systems.

2.3.1 Aerodynamic Modeling of Wind Turbine

According to Betz theory, the extracted power P_w by the wind turbine from the wind is a cubic function of wind speed [117],

$$\begin{aligned} P_w &= \frac{1}{2} \pi R^2 \rho C_p(\lambda, \beta) V^3 \\ \lambda &= \frac{\omega_m R}{V} \\ P_w &= T_w \omega_m \\ T_w &= \frac{1}{2} \pi R^3 \rho V^2 \frac{C_p(\lambda)}{\lambda} \end{aligned} \quad (2.10)$$

where ρ is the air density (kg/m^3), R is the radius of the blade (m), V is the wind speed (m/s), C_p expresses the power coefficient which is a function of tip speed ratio λ and pitch angle β . ω_m is the mechanical angular velocity of the turbine shaft, T_w is the wind torque on the shaft.

If the pitch system isn't considered, thus, $\beta = 0$. Then the power coefficient C_p is a function versus tip speed ratio λ [118],

$$C_p = \sum_{i=0}^6 a_i \lambda^{i+1} \quad (2.11)$$

Assume that the drive train to a sufficient level of accuracy can be described by two inertias interconnected by a spring and damper with viscous friction on each inertia. The external forces to this two-mass system are the aerodynamic torque T_w on the low speed shaft and generator reaction torque T_g on the high speed shaft. The motion equations are given by [119]

$$\begin{aligned} J_m \dot{\omega}_m &= T_w - B_m \omega_m - \mu(\omega_m - \omega_g) - K\theta_\Delta \\ J_g \dot{\omega}_g &= -T_g - B_g \omega_g + \mu(\omega_m - \omega_g) + K\theta_\Delta \\ \dot{\theta}_\Delta &= \omega_m - \omega_g \end{aligned} \quad (2.12)$$

where ω_m and ω_g are the turbine and generator rotational speed, respectively; θ_Δ is the position deviation between the two shafts. J_m and J_g are the moment

of inertias of the turbine and the generator, respectively; B_m and B_g are the mechanical damping coefficients of the turbine and the generator, respectively; μ is the damping coefficient of the flexible coupling between the two masses and K is the shaft stiffness.

Alternatively, for some small wind turbine or direct drive WTPGS, the torque dynamic equation on the shaft can be described as one-mass lumped systems [116],

$$J_t \frac{d\omega_g}{dt} = -D_t \omega_g + T_w - T_g \quad (2.13)$$

where D_t expresses the damping constant of turbine and J_t is the moment of inertia of the lumped system.

2.3.2 Pitch System of Wind Turbine

According to (2.10), if the pitch system is considered in the aerodynamic modeling of wind turbine, the power coefficient $C_p(\lambda, \beta)$ can be expressed with respect to the tip speed ratio (TSR) λ and the pitch angle β [120],

$$\begin{aligned} C_p(\lambda, \beta) &= 0.73 \left(\frac{151}{\lambda_i} - 0.58\beta - 0.002\beta^{2.14} - 13.2 \right) e^{-\frac{18.4}{\lambda_i}} \\ \frac{1}{\lambda_i} &= \frac{1}{\lambda - 0.02\beta} - \frac{0.003}{1 + \beta^3}. \end{aligned} \quad (2.14)$$

When the wind speed and extracted power increase above the rating, the pitch control is activated to maintain the mechanical extracted power at the rated value.

The conventional pitch control system of WTPGS is shown in Figure 2.1. The pitch system time constant is given as T_{servo} [121][122].

2.4 Single-Machine Infinite-Bus (SMIB) Power System

The classical third order dynamic generator model has been commonly used for the design of excitation controller. The high order model with many nonlinearities is not considered in the controller design but in the simulations. The classical third order dynamical model of the SMIB power system can be described as [100],

Mechanical Equations

$$\dot{\delta}(t) = \omega(t) \quad (2.15)$$

$$\dot{\omega}(t) = -\frac{D}{2H}\omega(t) + \frac{\omega_0}{2H}(P_m(t) - P_e(t)) \quad (2.16)$$

Generator Electrical Dynamics

$$\dot{E}'_q(t) = (E_f(t) - E_q(t))\frac{1}{T'_{d0}} \quad (2.17)$$

Electrical Equations

Figure 2.1. WTPGS pitch control system

$$E_q(t) = \frac{x_{ds}}{x'_{ds}} E'_q(t) - \frac{x_d - x'_d}{x'_{ds}} V_s \cos \delta(t) \quad (2.18)$$

$$E_f(t) = k_c u_f(t) \quad (2.19)$$

$$P_e(t) = \frac{V_s E_q(t)}{x_{ds}} \sin \delta(t) \quad (2.20)$$

$$I_q(t) = \frac{V_s}{x_{ds}} \sin \delta(t) = \frac{P_e}{x_{ad} I_f(t)} \quad (2.21)$$

$$Q(t) = \frac{V_s}{x_{ds}} E_q(t) \cos \delta(t) - \frac{V_s^2}{x_{ds}} \quad (2.22)$$

$$E_q(t) = x_{ad} I_f(t) \quad (2.23)$$

$$V_t(t) = \frac{1}{x_{ds}} [x_s^2 E_q^2(t) + V_s^2 x_d^2 + 2x_s x_d x_{ds} P_e(t) \cot \delta(t)]^{\frac{1}{2}} \quad (2.24)$$

2.5 Multi-Machine Power System

Multi-machine power system is a typical large-scale interconnected nonlinear system, where an N machine power system consists of N independent dynamic subsystems interconnected together. Thus the N machine system is reduced to an electrical network with N internal machine buses. The classical synchronous model is used in the multi-machine power system [123].

The important equations in the analysis of synchronous machine are the rotational inertia equations describing the effect of unbalance between the electromagnetic torque and the mechanical torque of the machine. The motion of the rotor of the i th ($i = 1, 2, \dots, n$) machine is given by the swing equation:

Mechanical Equations

$$\dot{\delta}_i(t) = \omega_i(t) - \omega_0 \quad (2.25)$$

$$\dot{\omega}_i(t) = -\frac{D_i}{2H_i} (\omega_i(t) - \omega_0) + \frac{\omega_0}{2H_i} (P_{mi}(t) - P_{ei}(t)) \quad (2.26)$$

To obtain a time solution for the rotor angle, the expression of the mechanical and the electrical power is needed to develop. However, the complete dynamics

of generator is very complicated, including the dynamics of stator windings, field windings, damping windings, exciter and governor. So the most controller designs use an approximate linearized one-axis generator model. This simplification requires the following assumptions [1]:

1. The mechanical power input remains constant during the period of the transient.
2. Damping or asynchronous power is negligible.
3. The synchronous machine can be represented (electrically) by a voltage source behind a transient reactance.
4. The mechanical angle of the synchronous machine rotor coincides with the electrical phase angle of the voltage behind transient reactance.
5. The only electrical transient considered in the study is the varying $E'_q(t)$ and $E'_d = 0$.
6. The transient saliency is ignored, i.e., $X'_d = X'_q$.
7. The static exciter is simply modeled by a constant gain and its excitation limits as the time-constant of the exciter is relatively short.
8. Assume the exciter is to be supplied from a compensated supply so that its excitation function is not impaired during the fault period.
9. Load are replaced by constant shunt admittances.
10. Network is reduced to the machine internal buses by eliminating all physical buses of the system.

Under these assumptions, the electrical dynamics of the synchronous machine with generator excitation and multi-machine network are represented as:

Electrical Dynamics of Synchronous Generator

$$\dot{E}'_{qi}(t) = (E_{fi}(t) - E_{qi}(t)) \frac{1}{T'_{d0i}} \quad (2.27)$$

Electrical Equations

$$E_{qi}(t) = E'_{qi}(t) - (x_{di} - x'_{di})I_{di}(t) \quad (2.28)$$

$$E_{fi}(t) = k_{ci}u_{fi}(t) \quad (2.29)$$

$$P_{ei}(t) = \sum_{j=1}^n E'_{qi}(t)E'_{qj}(t)B_{ij} \sin \delta_{ij}(t) \quad (2.30)$$

$$Q_{ei}(t) = - \sum_{j=1}^n E'_{qi}(t)E'_{qj}(t)B_{ij} \cos \delta_{ij}(t) \quad (2.31)$$

$$I_{di}(t) = - \sum_{j=1}^n E'_{qj}(t)B_{ij} \cos \delta_{ij}(t) \quad (2.32)$$

$$I_{qi}(t) = \sum_{j=1}^n E'_{qj}(t)B_{ij} \sin \delta_{ij}(t) \quad (2.33)$$

$$E_{qi}(t) = x_{adi}I_{fi}(t) \quad (2.34)$$

$$V_{tqi}(t) = E'_{qi}(t) - x'_{di}I_{di}(t) \quad (2.35)$$

$$V_{tdi}(t) = x'_{di}I_{qi}(t) \quad (2.36)$$

$$V_{ti}(t) = [V_{tdi}^2(t) + V_{tqi}^2(t)]^{\frac{1}{2}} \quad (2.37)$$

2.6 Direct Feedback Linearization (DFL)

Due to that the conventional controllers for power systems are designed based on approximately linearized models, such as power system stabilizers, they may only deal with local stability around an operating point. Nonlinear control theory is employed to cancel the power system nonlinearity which can alleviate the uncertainties caused by operating point variations and improve the system stability and control performance as well.

Feedback linearization which can transform the nonlinear power system to be linear over a wide range has been widely used in power system excitation controller design. Its main advantage over approximate linearization depends on avoiding the local nature without ignoring the nonlinearities and interconnections of the nonlinear power systems. Direct feedback linearization approaches [124][125] are successfully used for design of power system controllers. As a more feasible method which preserves physical states than the geometric algorithm version, the DFL is briefly introduced in this section.

2.6.1 Application of DFL Technique on SMIB Power System

The application of DFL technique on a SMIB power system [100] shown in Section 2.4 is given in this subsection .

Since $E'_q(t)$ is physically unmeasurable, differentiate (2.18) and (2.20) to eliminate $E'_q(t)$,

$$\begin{aligned}\dot{P}_e(t) &= \frac{V_s}{x_{ds}} \sin \delta(t) \dot{E}_q(t) + \frac{V_s}{x_{ds}} \cos \delta(t) E_q(t) \omega(t) \\ \dot{E}_q(t) &= \frac{1}{T'_{d0}} [k_c u_f(t) - E_q(t)] + \frac{x_d - x'_d}{x'_d} V_s \sin \delta(t) \omega(t).\end{aligned}$$

Thus,

$$\begin{aligned}\dot{P}_e(t) &= -\frac{1}{T'} P_e(t) + \frac{1}{T'} \left\{ \frac{V_s}{x_{ds}} \sin \delta(t) [k_c u_f(t) + T'_{d0} (x_d - x'_d) \frac{V_s}{x'_{ds}} \sin \delta(t) \omega(t)] \right. \\ &\quad \left. + T'_{d0} \frac{V_s}{x_{ds}} \cos \delta(t) E_q(t) \omega(t) \right\}.\end{aligned}\tag{2.38}$$

By DFL transformation, (2.15), (2.16) and (2.17) yields

$$\begin{aligned}\dot{\delta}(t) &= \omega(t) \\ \dot{\omega}(t) &= -\frac{D}{2H} \omega(t) - \frac{\omega_0}{2H} \Delta P_e(t) \\ \Delta \dot{P}_e(t) &= -\frac{1}{T'} \Delta P_e(t) + \frac{1}{T'} v_f(t)\end{aligned}\tag{2.39}$$

where $\Delta P_e(t) = P_e(t) - P_m(t)$, $T' = \frac{x'_{ds}}{x_{ds}} T'_{d0}$ and

$$\begin{aligned} v_f(t) = & \frac{V_s}{x_{ds}} \sin \delta(t) [k_c u_f(t) + T'_{d0} (x_d - x'_d) \frac{V_s}{x_{ds}} \sin \delta(t) \omega(t)] \\ & + T'_{d0} \frac{V_s}{x_{ds}} E_q \cos \delta(t) \omega(t) - P_m(t). \end{aligned} \quad (2.40)$$

Note that $v_f(t)$ is the new virtual input for the linearized system (2.39). The mapping (2.40) from $u_f(t)$ to $v_f(t)$ is invertible, except when $\sin \delta(t) = 0$ (which is not in the normal operating region of the generator).

The real excitation control input $u_f(t)$ can be obtained as,

$$\begin{aligned} u_f(t) = & \frac{x_{ds}}{k_c V_s \sin \delta(t)} [v_f(t) - T'_{d0} \frac{V_s}{x_{ds}} E_q(t) \cos \delta(t) \omega(t) + P_m(t)] \\ & - T'_{d0} (x_d - x'_d) \frac{V_s}{k_c x_{ds}} \sin \delta(t) \omega(t)]. \end{aligned} \quad (2.41)$$

The physical limit of the excitation voltage is taken as

$$-3 \leq k_c u_f(t) \leq 6.$$

According to (2.24), the generator terminal voltage $V_t(t)$ is a nonlinear function with respect to $\delta(t)$, $P_e(t)$ and the system structure parameters. If some change is made to the system structure during the fault, $V_t(t)$ will reach another post-fault equilibrium point if $\delta(t)$ and $P_e(t)$ are forced to go back to their pre-fault steady states. In other words, if the post-fault power system structure is different from the pre-fault one, the terminal voltage $V_t(t)$, power angle $\delta(t)$ and electrical power $P_e(t)$ cannot recover back to their own pre-fault states at the same time. In order to avoid the undesirable abnormal voltage condition in practice, a voltage regulator should be employed in the excitation control to force the post-fault voltage back to its nominal value, especially in post-transient period. Power angle $\delta(t)$ and electrical power $P_e(t)$ controls should be relaxed to achieve the new post-fault equilibrium point.

A conventional linear robust voltage controller is given in [95], rewrite (2.24)

gives

$$\begin{aligned}
V_t(t) &= \frac{1}{x'_{ds}} \sqrt{E_q'^2(t)x_s^2 + V_s^2x_d'^2 + 2x_sx_d'E_q'(t)V_s \cos \delta(t)} \\
&= \sqrt{\frac{P_e^2(t)x_s^2}{V_s^2 \sin^2 \delta(t)} + \frac{V_s^2x_d'^2}{x_d'^2} + \frac{2x_sx_d'P_e(t) \cos \delta(t)}{x_d' \sin \delta(t)}}. \tag{2.42}
\end{aligned}$$

Differentiating the $V_t(t)$ equation gives

$$\begin{aligned}
\dot{V}_t(t) &= \left(-\frac{P_e^2(t)x_s^2 \cos \delta(t)}{V_t(t)V_s^2 \sin^3 \delta(t)} - \frac{x_sx_d'P_e(t)}{V_t(t)x_d' \sin^2 \delta(t)} \right) \dot{\delta}(t) \\
&\quad + \left(\frac{P_e(t)x_s^2}{V_t(t)V_s^2 \sin^2 \delta(t)} + \frac{x_sx_d' \cos \delta(t)}{V_t(t)x_d' \sin \delta(t)} \right) \dot{P}_e(t) \\
&= f_1(t)\dot{\delta}(t) + f_2(t)\dot{P}_e(t) \tag{2.43}
\end{aligned}$$

where

$$\begin{aligned}
f_1(t) &= -\frac{P_e^2(t)x_s^2 \cos \delta(t)}{V_t(t)V_s^2 \sin^3 \delta(t)} - \frac{x_sx_d'P_e(t)}{V_t(t)x_d' \sin^2 \delta(t)} \\
f_2(t) &= \frac{P_e(t)x_s^2}{V_t(t)V_s^2 \sin^2 \delta(t)} + \frac{x_sx_d' \cos \delta(t)}{V_t(t)x_d' \sin \delta(t)}.
\end{aligned}$$

Employing the DFL compensating law and substituting (2.39) in (2.43),

$$\dot{V}_t(t) = f_1(t)\omega(t) - \frac{f_2(t)}{T'_{d0}} \Delta P_e(t) + \frac{f_2(t)}{T'_{d0}} v_f(t). \tag{2.44}$$

Selecting $[\Delta V_t(t), \Delta \omega(t), \Delta P_e(t)]^T$ as the system state vector, a new linearized subsystem can be developed to use for design of voltage controller [100]. Since $f_1(t)$ and $f_2(t)$ are dependent on the operating conditions, their bounds can be found within a certain operating region.

2.6.2 Application of DFL Technique on Multi-Machine Power System

The application of DFL technique on a multi-machine power system shown in Section 2.5 is given in this subsection.

Since $E'_{qi}(t)$ is physically unmeasurable, differentiate (2.30) to eliminate $E'_{qi}(t)$,

$$\begin{aligned}\dot{P}_{ei}(t) &= \sum_{j=1}^n \dot{E}'_{qi}(t) E'_{qj}(t) B_{ij} \sin \delta_{ij}(t) + \sum_{j=1}^n E'_{qi}(t) \dot{E}'_{qj}(t) B_{ij} \sin \delta_{ij}(t) \\ &\quad + \sum_{j=1}^n E'_{qi}(t) E'_{qj}(t) B_{ij} \cos \delta_{ij}(t) (\omega_i(t) - \omega_j(t)) \\ &\quad + \sum_{j=1}^n E'_{qi}(t) E'_{qj}(t) \dot{B}_{ij} \sin \delta_{ij}(t)\end{aligned}$$

Let $\Delta P_{ei}(t) = P_{ei}(t) - P_{mi}$, thus,

$$\begin{aligned}\Delta \dot{P}_{ei}(t) &= -\frac{1}{T'_{d0i}} \Delta P_{ei}(t) + \frac{1}{T'_{d0i}} [k_{ci} I_{qi}(t) u_{fi}(t) + (x_{di} - x'_{di}) I_{di}(t) I_{qi}(t) \\ &\quad - T'_{d0i} Q_{ei}(t) \Delta \omega_i(t) - P_{mi}] + \sum_{j=1}^n E'_{qi}(t) [\dot{E}'_{qj}(t) B_{ij} + E'_{qj}(t) \dot{B}_{ij}] \sin \delta_{ij}(t) \\ &\quad - \sum_{j=1}^n E'_{qi}(t) E'_{qj}(t) B_{ij} \cos \delta_{ij}(t) \Delta \omega_j(t) \\ &= -\frac{1}{T'_{d0i}} \Delta P_{ei}(t) + \frac{1}{T'_{d0i}} v_{fi}(t) \\ &\quad + \sum_{j=1}^n E'_{qi}(t) [\dot{E}'_{qj}(t) B_{ij} + E'_{qj}(t) \dot{B}_{ij}] \sin \delta_{ij}(t) \\ &\quad - \sum_{j=1}^n E'_{qi}(t) E'_{qj}(t) B_{ij} \cos \delta_{ij}(t) \Delta \omega_j(t)\end{aligned}\tag{2.45}$$

where $v_{fi} = k_{ci} I_{qi}(t) u_{fi}(t) + (x_{di} - x'_{di}) I_{di}(t) I_{qi}(t) - T'_{d0i} Q_{ei}(t) \Delta \omega_i(t) - P_{mi}$.

The real excitation control input $u_{fi}(t)$ can be obtained as,

$$u_{fi}(t) = \frac{1}{k_{ci}I_{qi}(t)}[v_{fi}(t) + P_{mi} - (x_{di} - x'_{di})I_{di}(t)I_{qi}(t) + T'_{d0i}Q_{ei}(t)\Delta\omega_i(t)]. \quad (2.46)$$

The physical limit of the excitation voltage is taken as

$$-6 \leq k_{ci}u_{fi}(t) \leq 6.$$

By DFL compensation, (2.25), (2.26) and (2.27) yields

$$\begin{aligned} \dot{\delta}_i(t) &= \Delta\omega_i(t) \\ \Delta\dot{\omega}_i(t) &= -\frac{D_i}{2H_i}\Delta\omega_i(t) - \frac{\omega_0}{2H_i}\Delta P_{ei}(t) \\ \Delta\dot{P}_{ei}(t) &= -\frac{1}{T'_{d0i}}\Delta P_{ei}(t) + \frac{1}{T'_{d0i}}v_{fi}(t) \\ &\quad + \sum_{j=1}^n E'_{qi}(t)[\dot{E}'_{qj}(t)B_{ij} + E'_{qj}(t)\dot{B}_{ij}] \sin \delta_{ij}(t) \\ &\quad - \sum_{j=1}^n E'_{qi}(t)E'_{qj}(t)B_{ij} \cos \delta_{ij}(t)\Delta\omega_j(t). \end{aligned} \quad (2.47)$$

Note that $v_{fi}(t)$ is the new virtual input of the excitation loop for the linearized system (2.47). The mapping of $u_{fi}(t)$ to $v_{fi}(t)$ are invertible, except when $I_{qi}(t) = 0$ (which are not in the normal operating region of the generator). The DFL linearized model of the generators are independent of the operating point of the system. Since $\Delta\omega_i(t)$ and $Q_{ei}(t)$ are locally measurable, $I_{di}(t)$ and $I_{qi}(t)$ can be calculated from the three phase current of the i th generator.

After the DFL compensation, the generator models are linear with respect to the new control inputs $v_{fi}(t)$ in the steady state of multi-machine power system. However, the DFL compensating law still contains nonlinearities due to the effects of remote dynamics of the j th and i th generator, such as switching of transmission lines when a fault occurs in the power system, the system parameter, B_{ij} will change considerably. In moment of switching breakers under large disturbances, the derivative of susceptance $\dot{B}_{ij}$ changes dramatically and it cannot be bounded.

However, it remains zero when structure of transmission lines does not change.

The generator terminal voltages $V_{ti}(t)$ are easy to be measured and the equilibrium points for $V_{ti}(t)$ are well defined. The operating points of the given third-order synchronous generators can be defined uniquely by $V_{ti}(t)$, $\omega_i(t)$ and $P_{ei}(t)$. Furthermore, voltage regulation can be expected by the feedback controls involving $V_{ti}(t)$. For permanent fault, in order to avoid the undesirable abnormal voltage condition in practice, voltage regulators should be employed in the excitation controls to force the post-fault voltage back to the nominal values. Power angle and electrical power controls should be released to achieve the new post-fault equilibrium points the same as in SMIB power system shown in section 2.6.1.

If some change is made to the system structure during the fault, $V_{ti}(t)$ will reach other post-fault equilibrium points even if $\delta_i(t)$ and $P_{ei}(t)$ are forced to go back to their pre-fault steady values. In order to avoid the undesirable abnormal voltage conditions in practice, voltage regulators should be employed in the excitation controls for each synchronous generator to force the post-fault voltages back to their nominal values, especially in post-transient period.

Rewrite (2.37) gives

$$V_{ti}(t) = \left\{ [E'_{qi}(t) + x'_{di} \sum_{j=1}^n E'_{qj}(t) B_{ij} \cos \delta_{ij}(t)]^2 + [x'_{di} \sum_{j=1}^n E'_{qj}(t) B_{ij} \sin \delta_{ij}(t)]^2 \right\}^{\frac{1}{2}}. \quad (2.48)$$

To simplify the analysis, the following assumption should be made. As in [96], considering the i th generator, $E'_{qj}(t)$ are constant and neglecting the effects of $\omega_j(t)$, where $j = 1, 2, \dots, n$. By differentiating the $V_{ti}(t)$ equation gives

$$\dot{V}_{ti}(t) = \frac{\partial V_{ti}(t)}{\partial \delta_i(t)} \dot{\delta}_i(t) + \frac{\partial V_{ti}(t)}{\partial E'_{qi}(t)} \dot{E}'_{qi}(t) = f_{1i}(t) \Delta \omega_i(t) + f_{2i}(t) \dot{P}_{ei}(t)$$

where

$$\begin{aligned} f_{1i}(t) &= -\frac{(1 + x'_{di}B_{ij})[-E'_{qi}(t)B_{ij} - Q_{ei}(t)]V_{tqi}(t)}{V_{ti}(t)I_{qi}(t)} - \frac{x'_{di}(1 + x'_{di}B_{ij})P_{ei}(t)}{V_{ti}(t)} \\ f_{2i}(t) &= \frac{(1 + x'_{di}B_{ij})V_{tqi}(t)}{V_{ti}(t)I_{qi}(t)}. \end{aligned}$$

Employing the DFL compensating law and substituting (2.47) in (2.49),

$$\dot{V}_{ti}(t) = f_{1i}(t)\Delta\omega_i(t) - \frac{f_{2i}(t)}{T'_{d0i}}\Delta P_{ei}(t) + \frac{f_{2i}(t)}{T'_{d0i}}v_{fi}(t). \quad (2.49)$$

Selecting $[\Delta V_{ti}(t), \Delta\omega_i(t), \Delta P_{ei}(t)]^T$ as the system state vector of the i th generator, the DFL-compensated model can be developed. Since $f_{1i}(t)$ and $f_{2i}(t)$ are dependent on the operating conditions, their bounds can be found within certain operating regions.

2.7 Summary

This chapter presented the mathematical models of the systems involved in this thesis. The dynamic models of PMS machines are given in Section 2.2 including PMSM and PMSG, respectively. The dynamic model of PMSM is employed in Section 4.2 and Section 4.3 to investigate the advanced controllers for PMSM position and current controls with considering the system parametric uncertainties and unmodeled dynamics, respectively. For the PMSG, the given model is introduced as the generator of WTPGS in this thesis. The advanced MPPT and sensorless controls for WTPGS are proposed in Chapter 5 based on PMSG WTPGS. The modeling of the wind turbine system introduced in Section 2.3.1 including the aerodynamic part and the pitch system is used for all the control designs of wind turbine system in this thesis. The power system models are chosen in consideration of excitation controller design and model accuracy. The dynamic models of synchronous generators are simplified as classical third order systems under certain assumptions. By using the original physical variables of the power system models, DFL is employed in the power systems to transform

the nonlinear power system model into linear model directly. DFL can linearize the power systems in their global state space or in large enough region of state space. In the next chapter, the optimal reset control theory will be introduced in both continuous-time and discrete-time forms in details.

Optimal Reset Control (ORC)

3.1 Introduction

The reset control was firstly proposed by Clegg to overcome the limitations of linear control [12]. This reset controller consists of an integrator and a reset law which resets the output of the integrator to zero when its input vanishes. The reset control is helpful in reducing the controller windup caused by integration. However, due to the fixed traditional reset law, performance improvement is limited. For example, before the first zero-crossing of tracking error, the traditional reset mechanism is not active, thus the rise time cannot be reduced and overshoot is in general inevitable because of inertia of physical elements. Actually, performance improvement does not come from blind resettings but from the interaction between the baseline system and the reset mechanism [13]. In order for that the reset mechanism cooperates better with the baseline system, a feedback was introduced in the reset law. It has been proved that reset control can achieve some control specifications beyond the ability of ordinary linear control and realize much better sensor noise suppression without degrading disturbance rejection or losing margins. These advantages make reset control an effective and significant technique for transient performance improvement.

Actually, reset control can be more general, for instance, the reset time and the amount of reset can both be designed so that the reset law and the baseline

system can cooperate better with each other. The ORC design propose an ORL to dominate the reset time and amount of reset according to a previous defined performance index [14]. The design of the reset law aims to minimize the performance index. The ORC problem is transferred to a regular LQR problem, so its solution can be derived by solving a Riccati equation. The ORC is proposed to retain fast response and short settling time and to improve steady-state stability. Additionally, the overshoot will be totally eliminated. ORC are applied on baseline controllers in this section.

Design of linear controllers such as PI are always difficult to take into account all the control task requirements. Smaller damping factor will make the response faster but leads to larger overshoot as well. Larger bandwidth can achieve faster response and shorter settling time but results in worse steady-state stability. The ORC is designed and applied to meet the needs from the industry. Due to its linear design principle, the ORC is relatively easy. On the other hand, ORC is a kind of nonlinear control scheme which can achieve some specifications beyond the ability of linear controller. Moreover, the ORC can solve the controller windup problem itself without affecting the stability. It is worth pointing out that there are many other techniques with similar design ideas as the ORC. For example, impulsive control [18] and SMC [19], etc. Reset control system is also known as a special impulsive system which is closely related to the optimal impulsive system. Both the reset time and the reset value could be arbitrarily determined in the case of impulsive control. If the reset value is fixed and the reset instant need to be determined, it belongs to the pulse width modulation control. On the other hand, the proposed reset control operates with fixed reset instant, and the reset value need to be determined. Since the purposed ORC is designed based on linear theory, it is superior than other impulsive controls for its simple design process and easy industrial application. SMC is also a widely used advanced control scheme. However, it is always dependent on complex design process, even hard to avoid chattering. In addition, SMC cannot be easily applied to multi-input/output systems as the proposed ORC.

A typical reset control system is depicted in Figure 3.1. In the following section, the existing continuous-time ORC theory will be introduced. Afterward-

s, the discrete-time ORC principle will be proposed for time-varying reference tracking problems.

Figure 3.1. Block diagram of a reset control system

3.2 Existing Optimal Reset Control Theory

Firstly, the ORC theory was proposed in a continuous-time control system for tracking of constant reference. The basic theory of the existing continuous-time optimal reset control will be briefly introduced in this section [17].

Consider a linear control system

$$\begin{aligned}\dot{x} &= Ax + Bu \\ y &= Cx\end{aligned}\tag{3.1}$$

where $x \in \mathbb{R}^n$ is the state of the plant, $u \in \mathbb{R}^m$ and $y \in \mathbb{R}^p$ are the control input and the system output, respectively. A , B and C are constant system matrices with compatible dimensions. The following shows a general form of a reset controller,

$$\begin{cases} u = Me + Hz + Gx, \\ \dot{z} = Ee + Dz, & t \neq t_k, \\ z(t^+) = \rho_k(x, z, r), & t = t_k, \end{cases}\tag{3.2}$$

where $z \in \mathbb{R}^q$ is the state of the controller. r is the reference signal which is assumed to be constant. $\rho_k(x, z, r)$ is the reset value at time instant t_k which is a designed value using optimal reset law. D , E , M , G and H are all constant matrices with compatible dimensions. $e = r - y$ is the tracking error and the reset time period is defined as $\Delta t_k = t_{k+1} - t_k$.

Then combining the plant (3.1) and the reset controller (3.2) yields the state space equations,

$$\begin{cases} \dot{\bar{x}} = \bar{A}\bar{x} + \bar{B}r, & t \neq t_k \\ z(t^+) = \rho_k(\bar{x}, r), & t = t_k \\ y = \bar{C}\bar{x}, \end{cases} \quad (3.3)$$

where $\bar{x} = (x^T \ z^T)^T$, $\bar{C} = (C, 0_{p \times q})$,

$$\bar{A} = \begin{pmatrix} A + BG - BMC & BH \\ -EC & D \end{pmatrix}, \quad \bar{B} = \begin{pmatrix} BM \\ E \end{pmatrix} \quad (3.4)$$

Set an assumption for the steady state of the system (3.1).

Assumption 1. For any $r \in \mathbb{R}^p$, there exists $\bar{x}_r = (x_r^T, z_r^T)^T \in \mathbb{R}^{n+m}$, such that

$$\begin{cases} \bar{C}\bar{x}_r = 0 \\ \bar{A}\bar{x}_r + \bar{B}r = 0. \end{cases} \quad (3.5)$$

Based on this assumption, the steady-state control input u_r is given by

$$u_r = Gx_r + Hz_r. \quad (3.6)$$

Define a cost function J_k in each reset time interval $(t_k, t_{k+1}]$,

$$J_k = \int_{t_k}^{t_{k+1}} [e^T(s)Q_k e(s) + (u(s) - u_r)^T R_k (u(s) - u_r)] ds \quad (3.7)$$

where Q_k and R_k are positive semi-definite matrices.

Problem 1. The optimal reset law design problem means design reset value ρ_k which makes the system asymptotically stable and meanwhile minimize the cumu-

lative cost function $J(\infty)$,

$$J(\infty) = \sum_{i=0}^{\infty} J_i. \quad (3.8)$$

The ORL can be designed by transferring the ORL problems to linear quadratic regulation problems. Firstly, make a coordinate transformation as follows,

$$\begin{cases} \xi_x = x - x_r \\ \xi_z = z - z_r \end{cases} \quad (3.9)$$

where x_r and z_r are defined as the steady-state values. Thus derive the system on the steady state point from (3.3),

$$\begin{cases} \dot{\xi} = \bar{A}\xi, & t \neq t_k \\ \xi_z(t_k^+) = \bar{\rho}_k, & t = t_k, \\ e = -\bar{C}\xi, \end{cases} \quad (3.10)$$

where $\xi = (\xi_x, \xi_z)^T$ and

$$\bar{\rho}_k = \rho_k - z_r. \quad (3.11)$$

Meanwhile,

$$u - u_r = (G - MC)\xi_x + H\xi_z, \quad (3.12)$$

$$\xi(t_{k+1}) = e^{\bar{A}\Delta t_k} \xi(t_k^+). \quad (3.13)$$

Thus

$$\begin{aligned} J_k &= \int_{t_k}^{t_{k+1}} [\xi_x^T(s) C^T Q_k C \xi_x(s) \\ &\quad + ((G - MC)\xi_x + H\xi_z)^T R_k ((G - MC)\xi_x + H\xi_z)] ds \\ &= \xi^T(t_k^+) \Theta_k \xi(t_k^+). \end{aligned} \quad (3.14)$$

Therefore, the cumulative cost function can be described as

$$J(\infty) = \sum_{i=0}^{\infty} \xi^T(t_k^+) \Theta_k \xi(t_k^+) \quad (3.15)$$

where

$$\begin{aligned}
\Theta_k &= \int_0^{\Delta t_k} e^{\bar{A}^T s} \Omega_k e^{\bar{A} s} ds, \\
\Omega_k &= \begin{pmatrix} \Omega_k^{11} & \Omega_k^{12} \\ \Omega_k^{12T} & \Omega_k^{22} \end{pmatrix}, \\
\Omega_k^{11} &= C^T Q_k C + (-MC)^T R_k (-MC), \\
\Omega_k^{12} &= (-MC)^T R_k H, \\
\Omega_k^{22} &= H^T R_k H
\end{aligned} \tag{3.16}$$

System (3.10) can be written as

$$\begin{pmatrix} \xi_x(t_{k+1}) \\ \xi_z(t_{k+1}) \end{pmatrix} = e^{\bar{A}\Delta t_k} \begin{pmatrix} \xi_x(t_k) \\ \bar{\rho}_k \end{pmatrix}. \tag{3.17}$$

Partition $e^{\bar{A}\Delta t_k}$ as follows,

$$e^{\bar{A}\Delta t_k} = \begin{pmatrix} \Gamma_A(k) & \Gamma_B(k) \\ * & * \end{pmatrix}. \tag{3.18}$$

So it is easy to get,

$$\xi_x(t_{k+1}) = \Gamma_A(k)\xi_x(t_k) + \Gamma_B(k)\bar{\rho}_k. \tag{3.19}$$

Problem 2. *The problem is to design a control sequence $\bar{\rho}_k$, $k = 1, 2, \dots$, of considering LQR problems of system (3.19) which both ensures the original system asymptotically stable and makes the cumulative cost function index minimized,*

$$J(\infty) = \sum_{k=0}^{\infty} \begin{pmatrix} \xi_x(k) \\ \bar{\rho}(k) \end{pmatrix}^T \Theta_k (\xi_x(k)^T \quad \bar{\rho}(k)^T)^T. \tag{3.20}$$

In this problem, it assume that the reset time period Δt_k and the index weight matrices Q_k , R_k are constant.

Proposition 1. *Suppose that there exist a positive number $\varepsilon > 0$ such that*

$$\lambda_{\min}(\Gamma_B^T \Gamma_B) \geq \varepsilon > 0, \forall k. \tag{3.21}$$

Then by rearrangement and transformations on system (3.10), $\Gamma_A(k)$ and $\Gamma_B(k)$ are ensured constant and of full rank as well.

Proposition 2. *Suppose that $(\Gamma_A(k), \Gamma_B(k))$ is controllable so that*

$$\lambda_{\min}(\Omega_k) \geq \varepsilon > 0, \forall k. \quad (3.22)$$

Then the ORL Problem 1 is equivalent to LQR Problem 2.

In order to achieve the above equivalence, we only need to prove that system (3.19) is asymptotically stable if and only if system (3.10) is asymptotically stable. In the following, we assume that system (3.19) is asymptotically stable. Then

$$\lim_{k \rightarrow \infty} \xi_x(k) = 0. \quad (3.23)$$

$$\lim_{k \rightarrow \infty} \Gamma_B \bar{\rho}(k) = 0. \quad (3.24)$$

According to (3.24),

$$\lim_{k \rightarrow \infty} \xi_z(k) = \lim_{k \rightarrow \infty} \bar{\rho}(k) = 0. \quad (3.25)$$

That means

$$\lim_{k \rightarrow \infty} \xi(k) = 0. \quad (3.26)$$

Thus the system (3.10) is asymptotically stable. Since Problem 1 and Problem 2 are proved equivalent, it is simple to derive the Optimal Reset Law by solving the corresponding LQR problem. Thus the optimal solutions can directly be obtained by solving some Riccati equations. Partition Θ_k ,

$$\Theta_k = \begin{pmatrix} \bar{Q} & \bar{T} \\ \bar{T}^T & \bar{R} \end{pmatrix}. \quad (3.27)$$

$$\Theta_k = \sum_{i=0}^{n-1} [(\bar{A}^T)^i (\bar{C}^T Q_{k,i} \bar{C} + N^T R_{k,i} N) (\bar{A})^i]. \quad (3.28)$$

If (Γ_A, Γ_B) is controllable, the optimal reset law which stabilizes the previous plant (3.10) (go back to the original system (3.1), this means that the output y asymptotically tracks the time-varying reference r) and meanwhile minimizes the

cumulative cost function $J(\infty)$ is given by

$$\rho^*(k) = -K(x(k) - x_r) + z_r. \quad (3.29)$$

K is determined by

$$K = (\Gamma_B^T S \Gamma_B + \bar{R})^{-1} (\Gamma_B^T S \Gamma_A + \bar{T}^T). \quad (3.30)$$

S here is the solution of the Riccati equation

$$S = \tilde{Q} + \bar{\Gamma}_A^T S \bar{\Gamma}_A - \bar{\Gamma}_A^T S \bar{\Gamma}_B (\Gamma_B^T S \Gamma_B + \bar{R})^{-1} \Gamma_B^T S \bar{\Gamma}_A \quad (3.31)$$

where $\bar{R}$ and $\tilde{Q}$ are positive definite.

$$\begin{aligned} \bar{\Gamma}_A &= \Gamma_A - \Gamma_B \bar{R}^{-1} \bar{T}^T \\ \tilde{Q} &= \bar{Q} - \bar{T} \bar{R}^{-1} \bar{T}^T. \end{aligned} \quad (3.32)$$

Furthermore, the cost function $J(\infty)$ minimum is given by

$$J^*(\infty) = (x(0) - x_r)^T S (x(0) - x_r). \quad (3.33)$$

3.3 Discrete-time Optimal Reset Control

The previous works are all based on the ORC design for a constant reference system mainly for the improvement of step transient response. More often the tracking references are time-varying and unpredictable.

The discrete-time ORC design proposed in this section is an advanced reset control. The majority of industrial control systems depend on computer control. And the most applications are feedback controls which based on sampling of the states. The proposed ORC is fit for time-varying reference and can be easily applied in industrial discrete-time control system. Through merging the uncertainty and time-varying of reference into the ORL, the proposed ORC law which is redesigned in discrete-time form can overcome perfect tracking problem

effectively.

Set a continuous linear dynamic plant as,

$$\begin{aligned}\dot{x} &= Ax + Bu \\ y &= Cx\end{aligned}\tag{3.34}$$

where $x \in \mathbb{R}^n$ is the state of the plant, $u \in \mathbb{R}^m$ and $y \in \mathbb{R}^p$ are the control input and the system output respectively. A , B and C are constant system matrices with compatible dimensions.

Set an assumption about the steady state of the open-loop system (3.34).

Assumption 2. *For any time-varying reference $r(t) \in \mathbb{R}^p$ (abbreviated as r in the rest of this section), there exists a system steady-state x_r and a controller steady-state input u_r such that*

$$\begin{cases} r = Cx_r \\ \dot{x}_r = Ax_r + Bu_r. \end{cases}\tag{3.35}$$

Based on this assumption, the steady-state control input u_r is given by

$$u_r = (B^T B)^{-1} B^T \dot{x}_r - (B^T B)^{-1} B^T A x_r.\tag{3.36}$$

Additionally, in order to support this assumption, u should be a baseline controller which both makes the system stable and convergent to x_r .

Discretize this continuous linear plant (3.34), thus zero-order-hold equivalent description of (3.34) with the sampling time T_s is given by [126]

$$\begin{cases} x(i+1) = A_d x(i) + B_d u(i) \\ y(i) = C_d x(i) \end{cases}\tag{3.37}$$

where $A_d = e^{AT_s}$, $B_d = \int_0^{T_s} e^{A(T_s-\tau)} B d\tau$, $C_d = C$.

Denoted by T_r the constant reset time period. In each T_r , there are n sampling intervals, i.e., $T_r = nT_s$. At any time instant $t = kT_r + iT_s$ with $i = 0, 1, 2, \dots, n-1$. $x(k, i)$, $y(k, i)$ and $u(k, i)$ are the system states, output and control input

respectively.

Introduce in a time-varying output tracking reference $r(k, i) \in \mathbb{R}^p$ for the system (3.37) at $t = kT_r + iT_s$. The system steady-state for $t = kT_r + iT_s$ should be $x_r(k, i)$ and the steady-state control input $u_r(k, i)$ is given by

$$u_r(k, i) = (B_d^T B_d)^{-1} B_d^T (x_r(k, i) - x_r(k, i - 1)) \frac{1}{T_s} - (B_d^T B_d)^{-1} B_d^T A_d x_r(k, i). \quad (3.38)$$

Remark 1. *r is an uncertain time-varying reference to be tracked. After discretization of the system with sampling period T_s , it defaults that r shares the same sampling period T_s with the system. Then the tracking problem essentially becomes discrete-time tracking to each sampling point of reference $r(k, i)$. So we make an assumption that r maintains constant during each sampling interval $[(i - 1)T_s, iT_s]$. If r cannot keep constant during $[(i - 1)T_s, iT_s]$, the sampling for r with T_s will cause the loss of its unignorable high-frequency parts. Definitely, the tracking performance cannot be ensured due to the low sampling rate and the system sampling period should be decreased. Briefly, it comes to the conclusion that system sampling rate should be positive proportional to the content of high-frequency in the reference signal.*

A general discrete-time reset controller is represented as,

$$\begin{cases} u(k, i) = Me(k, i) + Hz(k, i) \\ z(k, i) = Ee(k, i - 1) + Dz(k, i - 1) \\ z(k, 0) = \rho_k(x(k - 1, n - 1), r(k - 1, n - 1), r(k, 0)) \end{cases} \quad (3.39)$$

where $u(k, i)$ means the zero-order-hold control input of (3.37) during time instant $t = kT_r + iT_s$ and $t = kT_r + (i + 1)T_s$. $z(k, i) \in \mathbb{R}^q$ is the state of the controller. $\rho_k(x(k - 1, n - 1), r(k - 1, n - 1), r(k, 0))$ is the designed reset value using ORL at time instant $t = kT_r$. D , E , M and H are all constant matrices with compatible dimensions. $e(k, i) = r(k, i) - y(k, i)$ is the tracking error.

Denoting $z_r(k, i)$ as the steady-state of $z(k, i)$. From (3.38), $z_r(k, i)$ can be

expressed as,

$$\begin{aligned} u_r(k, i) &= H z_r(k, i) \\ z_r(k, i) &= (H^T H)^{-1} H^T (B_d^T B_d)^{-1} B_d^T (x_r(k, i) - x_r(k, i-1)) \frac{1}{T_s} \\ &\quad - (H^T H)^{-1} H^T (B_d^T B_d)^{-1} B_d^T A_d x_r(k, i). \end{aligned} \quad (3.40)$$

Then combining the plant (3.37) and the reset controller (3.39) yields the discrete-time state space equations,

$$\begin{cases} \bar{x}(k, i) = \bar{A} \bar{x}(k, i-1) + \bar{B} r(k, i-1) \\ z(k, 0) = \rho_k(x(k-1, n-1), r(k-1, n-1), r(k, 0)) \\ \bar{y}(k, i) = \bar{C} \bar{x}(k, i) \end{cases} \quad (3.41)$$

where $\bar{x}(k, i) = (x(k, i)^T \ z(k, i)^T)^T$, $\bar{C} = (C_d, 0_{p \times q})$,

$$\bar{A} = \begin{pmatrix} A_d - B_d M C_d & B_d H \\ -E C_d & D \end{pmatrix}, \quad \bar{B} = \begin{pmatrix} B_d M \\ E \end{pmatrix}. \quad (3.42)$$

The steady-state of the closed-loop system (3.41) is $(x_r(k, i)^T, \ z_r(k, i)^T)^T$.

Define a cost function $J(k, i)$ at each time instant $t = kT_r + iT_s$,

$$J(k, i) := e^T(k, i) Q_{k,i} e(k, i) + (u(k, i) - u_r(k, i))^T R_{k,i} (u(k, i) - u_r(k, i)). \quad (3.43)$$

where $Q_{k,i}$ and $R_{k,i}$ are positive semi-definite weight matrices.

Problem 3. *The ORL design problem means design reset value ρ_k which makes the system (3.41) asymptotically stable with respect to $(x_r(k, i)^T, \ z_r(k, i)^T)^T$ and meanwhile minimize the cumulative cost function $J(\infty)$,*

$$J(\infty) := \sum_{k=0}^{\infty} \sum_{i=0}^{n-1} J(k, i). \quad (3.44)$$

The ORL is designed by transferring the ORC problems to the LQR problems.

Make a coordinate transformation as,

$$\begin{cases} \xi_x(k, i) = x(k, i) - x_r(k, i) \\ \xi_z(k, i) = z(k, i) - z_r(k, i) \end{cases} \quad (3.45)$$

Thus derive the equivalent system on the steady state point from (3.41),

$$\begin{cases} \xi(k, i+1) = \bar{A}\xi(k, i) \\ \xi_z(k, 0) = \bar{\rho}(k, 0) \\ e(k, i) = -\bar{C}\xi(k, i) \end{cases} \quad (3.46)$$

where $\xi(k, i) = (\xi_x^T(k, i) \ \xi_z^T(k, i))^T$ and

$$\bar{\rho}(k, 0) = \rho(k, 0) - z_r(k, 0). \quad (3.47)$$

The discrete-time form of Assumption 2 is given by

$$r(k, i) = y_r(k, i) = C_d x_r(k, i) \quad (3.48)$$

and,

$$\begin{aligned} & u(k, i) - u_r(k, i) \\ &= M(r(k, i) - C_d x(k, i)) + Hz(k, i) - Hz_r(k, i) \\ &= -MC_d x(k, i) + Mr(k, i) + H(z(k, i) - z_r(k, i)) \\ &= -MC_d(x(k, i) - x_r(k, i)) + H(z(k, i) - z_r(k, i)) \\ &= -MC_d \xi_x(k, i) + H \xi_z(k, i) \\ &= [-MC_d \ H] \xi(k, i). \end{aligned} \quad (3.49)$$

The cumulative cost function can be simplified as,

$$J(\infty) = \sum_{k=0}^{\infty} \left(\sum_{i=0}^{n-1} J(k, i) \right)$$

$$\begin{aligned}
&= \sum_{k=0}^{\infty} \left(\sum_{i=0}^{n-1} (e^T(k, i) Q_{k,i} e(k, i) \right. \\
&\quad \left. + (u(k, i) - u_r(k, i))^T R_{k,i} (u(k, i) - u_r(k, i))) \right) \\
&= \sum_{k=0}^{\infty} \left(\sum_{i=0}^{n-1} (\xi^T(k, i) \bar{C}^T Q_{k,i} \bar{C} \xi(k, i) \right. \\
&\quad \left. + \xi^T(k, i) [-MC_d \ H]^T R_{k,i} [-MC_d \ H] \xi(k, i)) \right) \\
&= \sum_{k=0}^{\infty} \left(\sum_{i=0}^{n-1} (\xi^T(k, i) (\bar{C}^T Q_{k,i} \bar{C} \right. \\
&\quad \left. + [-MC_d \ H]^T R_{k,i} [-MC_d \ H]) \xi(k, i)) \right) \\
&= \sum_{k=0}^{\infty} \left(\sum_{i=0}^{n-1} (\xi^T(k, 0) (\bar{A}^T)^i (\bar{C}^T Q_{k,i} \bar{C} \right. \\
&\quad \left. + [-MC_d \ H]^T R_{k,i} [-MC_d \ H]) (\bar{A})^i \xi(k, i)) \right) \\
&= \sum_{k=0}^{\infty} (\xi^T(k, 0) (\sum_{i=0}^{n-1} ((\bar{A}^T)^i (\bar{C}^T Q_{k,i} \bar{C} \right. \\
&\quad \left. + [-MC_d \ H]^T R_{k,i} [-MC_d \ H]) (\bar{A})^i)) \xi(k, 0)) \\
&= \sum_{k=0}^{\infty} (\xi^T(k, 0) \Theta_k \xi(k, 0)) \tag{3.50}
\end{aligned}$$

where

$$\Theta_k = \sum_{i=0}^{n-1} ((\bar{A}^T)^i (\bar{C}^T Q_{k,i} \bar{C} + [-MC_d \ H]^T R_{k,i} [-MC_d \ H]) (\bar{A})^i). \tag{3.51}$$

In each reset time period, system (3.46) can be written as

$$\begin{pmatrix} \xi_x(k+1, 0) \\ \xi_z(k+1, 0) \end{pmatrix} = (\bar{A})^n \begin{pmatrix} \xi_x(k, 0) \\ \bar{\rho}(k, 0) \end{pmatrix}. \tag{3.52}$$

Partition $(\bar{A})^n$ as,

$$(\bar{A})^n = \begin{pmatrix} \Gamma_A(k) & \Gamma_B(k) \\ * & * \end{pmatrix}. \tag{3.53}$$

Therefore,

$$\xi_x(k+1, 0) = \Gamma_A(k) \xi_x(k, 0) + \Gamma_B(k) \bar{\rho}(k, 0). \tag{3.54}$$

Problem 4. *This problem is to design a control sequence $\bar{\rho}(k)$, $k = 1, 2, \dots$, of considering LQR problems of system (3.54) which ensures asymptotical stability and minimize the cumulative cost function at the same time*

$$J(\infty) = \sum_{k=0}^{\infty} \begin{pmatrix} \xi_x(k, 0) \\ \bar{\rho}(k, 0) \end{pmatrix}^T \Theta_k (\xi_x(k, 0)^T \quad \bar{\rho}(k, 0)^T)^T. \quad (3.55)$$

It is assumed that the reset time period $T_r = nT_s$ and the index weight matrices $Q_{k,i}$, $R_{k,i}$ are constant.

Proposition 3. *Suppose that there exist a positive number $\varepsilon > 0$ such that*

$$\lambda_{\min}(\Gamma_B(k)^T \Gamma_B(k)) \geq \varepsilon > 0, \forall k. \quad (3.56)$$

Then the ORL Problem 3 is equivalent to LQR Problem 4.

Remark 2. *By rearrangement and transformations on system (3.46), $\Gamma_A(k)$ and $\Gamma_B(k)$ are ensured constant and of full rank.*

In order to achieve the above proposition, a conclusion that system (3.54) is asymptotically stable if and only if system (3.46) is asymptotically stable should be proved. System (3.54) is derived from system (3.46), so system (3.54) is asymptotically stable if system (3.46) is asymptotically stable. This is the *if* part. In the following, the proof of the *only if* part is shown,

$$\lim_{k \rightarrow \infty} \xi_x(k, 0) = 0. \quad (3.57)$$

$$\lim_{k \rightarrow \infty} \Gamma_B(k) \bar{\rho}(k, 0) = 0. \quad (3.58)$$

According to (3.58),

$$\lim_{k \rightarrow \infty} \xi_z(k, 0) = \lim_{k \rightarrow \infty} \bar{\rho}(k, 0) = 0. \quad (3.59)$$

That means

$$\lim_{k \rightarrow \infty} \xi(k, 0) = 0. \quad (3.60)$$

Thus the system (3.46) is asymptotically stable. Since Problem 3 and Problem 4 are proved to be equivalent, it is simple to derive the ORL by solving the

corresponding LQR problem. The optimal solutions can directly be obtained by solving some Riccati equations [127][128].

Partition Θ_k ,

$$\Theta_k = \begin{pmatrix} \bar{Q} & \bar{T} \\ \bar{T}^T & \bar{R} \end{pmatrix}. \quad (3.61)$$

Suppose that $(\Gamma_A(k), \Gamma_B(k))$ is controllable, the ORL which stabilizes the previous plant (3.46) and minimizes the cumulative cost function $J(\infty)$ is given by

$$\rho^*(k, 0) - z_r(k, 0) = \xi_z(k, 0) = -K(x(k, 0) - x_r(k, 0)). \quad (3.62)$$

The steady-state points $x_r(k, 0)$ and $z_r(k, 0)$ are solved from (3.35) and (3.40),

$$\begin{aligned} z_r(k, 0) &= (H^T H)^{-1} H^T (B_d^T B_d)^{-1} B_d^T (x_r(k, 0) - x_r(k-1, n-1)) \frac{1}{T_s} \\ &\quad - (H^T H)^{-1} H^T (B_d^T B_d)^{-1} B_d^T A_d x_r(k, 0). \end{aligned} \quad (3.63)$$

The ORL can be expressed as

$$\begin{aligned} \rho^*(k, 0) &= -K(x(k, 0) - x_r) + (H^T H)^{-1} H^T (B_d^T B_d)^{-1} B_d^T (x_r(k, 0) \\ &\quad - x_r(k-1, n-1)) \frac{1}{T_s} \\ &\quad - (H^T H)^{-1} H^T (B_d^T B_d)^{-1} B_d^T A_d x_r(k, 0) \end{aligned} \quad (3.64)$$

K is determined by

$$K = (\Gamma_B(k)^T S \Gamma_B(k) + \bar{R})^{-1} (\Gamma_B(k)^T S \Gamma_A(k) + \bar{T}^T) \quad (3.65)$$

S is the solution of the Riccati equation

$$S = \tilde{Q} + \bar{\Gamma}_A^T S \bar{\Gamma}_A - \bar{\Gamma}_A^T S \bar{\Gamma}_B (\Gamma_B(k)^T S \Gamma_B(k) + \bar{R})^{-1} \Gamma_B(k)^T S \bar{\Gamma}_A \quad (3.66)$$

where $\bar{R}$ and $\tilde{Q}$ are positive definite.

$$\begin{aligned} \bar{\Gamma}_A &= \Gamma_A(k) - \Gamma_B(k) \bar{R}^{-1} \bar{T}^T \\ \tilde{Q} &= \bar{Q} - \bar{T} \bar{R}^{-1} \bar{T}^T. \end{aligned} \quad (3.67)$$

The selection of the system sampling period T_s will impact the performance of ORC to some extent.

3.4 Summary

This chapter investigates optimal reset law design of reset control systems with fixed reset time instants. Firstly, the optimal reset control is introduced of the continuous-time form. By transferring the closed-loop reset control system to a discrete-time linear system with the reset values as control input, the continuous-time optimal reset law design problems can be transformed to LQR problems. Based on this, the continuous-time optimal reset laws are easily obtained by solving algebraic Riccati equations. The design process of the proposed technique is very simple compared to the previous results.

Then, a discrete-time optimal reset control design method is proposed for time-varying reference tracking problem. The original continuous linear plant and the reference are combined to derive the general form of the steady states and control input. All the plant and the the steady-states and inputs are discretized to discrete-time forms. It is proved that the discrete-time ORL problems is also equivalent to LQR problems. By employing in the LQR design method and solving the corresponding algebraic Riccati equations, the discrete-time ORL is given in a simple form. Overall, compared with other similar advanced control schemes, the discrete-time ORC holds a lot of advantages over the traditional control methods as follows: 1) to reduce the overshoot of the step response without degrading other specifications; 2) to suppress controller's saturation effectively; 3) to accelerate the tracking speed with less error.

In the next chapter, the optimal reset control method is applied to PMS machines for designs of PMS machine ORC position controller, current controller and uncertainties estimator. based on ORC principle for PMSM drive systems. As an application, the proposed discrete-time ORC technique is applied to design of speed controller for PMSG system with time-varying reference.

Controls of Permanent Magnet Synchronous Machine

4.1 Introduction

Permanent magnet types have some performance advantages over direct-current excited synchronous types, and have become predominant in fractional horsepower applications. They are smaller, lighter, more efficient and reliable than other singly fed electric machines. Originally all large industrial DC motors used wound field or rotor magnets. Permanent magnets have traditionally only been useful on small motors because it was difficult to find a material capable of retaining a high-strength field. Only recently have advances in materials technology allowed the creation of high-intensity permanent magnets, such as neodymium magnets, allowing the development of compact, high-power motors without the extra real-estate of field coils and excitation means. But as these high performance permanent magnets become more applied in electric motor or generator systems, other problems appear. PMS machines have been widely applied in modern industry [22][23][24].

This thesis takes the wind turbine power generation system installed with PMSG as a representative of the WECS. And the PMSM can be regarded as power inverse systems of PMSG. Therefore, this chapter in-depth gives some

advanced controls for PMS machine systems.

4.2 Position Control of PMSM Drive System

Design robust and precise controller for PMSM positioning system is necessary and significant [129]. In this section, the proposed joint ORC and observer control scheme for PMSM position control design are shown. Based on the basic dynamic model of PMSM, two loops PID plus PI baseline position controllers are designed firstly. To eliminate the uncertainties caused by parametric variations and external load torque disturbances, a linear two pieces cascaded coupled uncertainties observer is designed for the ORC as well. Then the ORC is employed in the baseline controllers to improve the control performance.

The wind turbine pitch system can also be driven by PMS servo motor. Thus the proposed ORC position controller is applied to the pitch positioning servo system of the wind turbines used throughout the whole thesis. All the PMSM and controller parameters are shown in the appendix.

4.2.1 Baseline Two Loops Controllers Design

The PMSM modeling is from section 2.2.1. The overall PMSM position control system structure is shown in Figure 4.1. A two loops baseline control scheme including an inner loop PI current controller and an outer loop PID position controller is given. By conventional method, the two loops control parameters should consider the response speed between the two loop regulators and the interaction bandwidth and coordinations. Therefore, complicated design procedure is unavoidable. Even it requires repeated trial and error verification. This section uses the pole placement method for the global close loop system. It greatly simplifies the design process, while meeting high-performance target. This design method has obvious advantages.

In inner loop current PI controllers design for q and d axis, $L\omega i_d$, $\omega\psi$ and $L\omega i_q$ can be measured or exists as constant. Whereas the feed-forward control method is employed into the PI controllers to decouple the two axis coupling

terms. General inner-loop PI current controllers for PMSM are of the form,

$$\begin{aligned} v_q &= (k_p + k_i \frac{1}{s})(i_q^* - i_q) + L\omega i_d + \omega\psi \\ v_d &= (k_p + k_i \frac{1}{s})(i_d^* - i_d) - L\omega i_q. \end{aligned} \quad (4.1)$$

For simplification, the same controller parameters k_p and k_i are chosen. And d axis current reference $i_d^* = 0$ strategy is applied.

In outer loop position PID controller design for q axis, a feed-forward external load torque compensation term is added to the outer PID position control loop.

$$i_q^* = (k_{pt} + k_{it} \frac{1}{s} + k_{dt}s)(\theta_r^* - \theta_r) + \frac{1}{JK_e} T_L \quad (4.2)$$

where k_{pt} , k_{it} and k_{dt} are the PID controller parameters, and θ_r^* is denoted as the position reference of PMSM drive system.

The global close loop transfer function of PMSM position control system is given as

$$G(s) = \frac{\theta_r}{\theta_r^*} = \frac{\frac{K_e}{J}(k_{dt}s^2 + k_{pt}s + k_{it})(k_ps + k_i)}{D(s)} \quad (4.3)$$

Figure 4.1. Baseline two loops control scheme

where

$$\begin{aligned}
D(s) = & s^5 + \left(\frac{R}{L} + \frac{D}{J} + k_p\right)s^4 + \left(\frac{RD}{LJ} + \frac{Dk_p}{J} + k_i + \frac{DK_ek_pk_{dt}}{J}\right)s^3 \\
& + \left(\frac{K_ek_ik_{dt}}{J} + \frac{Dk_i}{J} + \frac{K_ek_pk_{pt}}{J}\right)s^2 + \left(\frac{K_ek_pk_{it}}{J} + \frac{K_ek_ik_{pt}}{J}\right)s \\
& + \frac{K_ek_ik_{it}}{J}.
\end{aligned} \tag{4.4}$$

Assuming the closed-loop dominant poles of the 5-order double-loop control system are $s_{1,2} = -\xi\omega_n \pm j\omega_n\sqrt{1-\xi^2}$, and the expected non-dominant poles are $s_3 = -m\xi\omega_n$, $s_4 = -n\xi\omega_n$ and $s_5 = -p\xi\omega_n$, respectively. m, n, p are always constant values been chosen as $5 \sim 10$ and ξ and ω_n are chosen as 0.707 and 450 respectively [40]. Therefore, the characteristic equation of the double-loop control system should be

$$D(s) = (s^2 + 2\xi\omega_n s + \omega_n^2)(s + m\xi\omega_n)(s + n\xi\omega_n)(s + p\xi\omega_n). \tag{4.5}$$

Comparing (4.4) and (4.5), it is easy to get 5 groups of control parameters. Deleting the 4 groups with complex numbers, and keeping the group with all real numbers as the our designed control parameters.

4.2.2 Uncertainties Observer Design

In real application of a PMSM position control system, the external load torque T_L is usually not known precisely and may change unpredictably. Assume that the parameters of damping constant D and moment of inertia J may vary around their designed values. The derivatives of parametric variations of D and J are defined as ΔD and ΔJ (including the conditions that $\Delta D = 0$ or $\Delta J = 0$).

The mechanical dynamic equation (2.5) must be rewritten as

$$(J + \Delta J)\dot{\omega}_r = -(D + \Delta D)\omega_r + K_e i_q - T_L. \tag{4.6}$$

Simplifying this equation as

$$J\dot{\omega}_r = -D\omega_r + K_e i_q - \hat{T}_L \quad (4.7)$$

where $\hat{T}_L = T_L + \Delta D\omega_r + \Delta J\dot{\omega}_r$. Then the outer loop position PID controller will be replaced by

$$i_q^* = (k_{pt} + k_{it}\frac{1}{s} + k_{dt}s)(\theta_r^* - \theta_r) + \frac{1}{JK_e}\hat{T}_L. \quad (4.8)$$

Assumption 3. *If the uncertainties observer is fast enough to converge and the moment of inertia, damping constant and external load torque do not change abruptly, the derivatives of J , D and T_L can be considered zero during each sampling period [130].*

Obviously, the electrical torque T_e is available through measurements. The state estimator is designed by propagating the input signals, T_e and $\hat{T}_L$. The state is furthermore updated by a scaling, L , of the error in estimated output as described as

$$\dot{\hat{\omega}}_r = -\frac{D}{J}\hat{\omega}_r + T_e - \hat{T}_L + L(\omega_r - \hat{\omega}_r). \quad (4.9)$$

The observer gain L is designed by the Kalman Filtering (KF) [119] approach in this section.

In PMSM position control drive systems, $\hat{T}_L$ is not measurable which means an extended observer loop should be employed in to estimate $\hat{T}_L$. As a result, a linear two pieces cascaded coupled uncertainties observer is proposed to estimate $\hat{T}_L$. The structure is given in Figure 4.2. The inner loop is a kalman filter designed with respect to (4.9). The outer loop operates as a tracking configuration with ω_r as the tracking objective. Since a PI controller is chosen for the close-loop tracking control, $\hat{T}_L$ can be estimated as the PI output. It can ensure the observer asymptotic tracking as long as that the PI controller is well designed [119]. The pole placement method is introduced to design the PI controller.

The close-loop characteristic equation of the observer system is given as

$$D(s) = s^2 + 2\xi_e\omega_{ne}s + \omega_{ne}^2$$

Figure 4.2. Block diagram of uncertainties observer structure

$$= s^2 + \frac{D - k_{pe}}{J}s + \frac{-k_{ie}}{J} \quad (4.10)$$

where ξ_e and ω_{ne} are chosen as 1 and 450 respectively. So the observer parameters are $k_{pe} = D - 2\xi_e\omega_{ne}J$ and $k_{ie} = -\omega_{ne}^2J$.

Denote that $\hat{e}$ is the derivative between the real value of $\hat{T}_L$ and its estimation value. The observer asymptotic tracking can be ensured by the pole placement designed PI controller. In other words, $\hat{e} \rightarrow 0$ in finite time. With the feed-forward compensation technique for $\hat{T}_L$ term, the global stability can be ensured by the two loops baseline controller.

4.2.3 ORC Position Controller

The ORC is an advanced reset control which consists of a baseline PID controller and an ORL. Obviously, ORC is a kind of nonlinear control scheme. And the ORL can be designed by transferring the problem into LQR problem. Some specifications can be achieved and much better sensor noise suppression without degrading disturbance rejection or losing margins can be realized by ORC. The system stability is guaranteed as the baseline PID controller does.

In this section, the ORC given in section 3.2 is only applied on the outer-loop PID position control. Figure 4.3 shows the block diagram of the ORC with Uncertainties Observer Control Scheme,

Consider the torque dynamic mechanical model of PMSM,

$$\begin{aligned}\dot{\theta}_r &= \omega_r \\ \dot{\omega}_r &= -\frac{D}{J}\omega_r + \frac{1}{J}T_e - \frac{1}{J}\hat{T}_L.\end{aligned}\quad (4.11)$$

Define

$$A = \begin{pmatrix} 0 & 1 \\ 0 & -\frac{D}{J} \end{pmatrix}, \quad B = \begin{pmatrix} 0 \\ \frac{1}{J} \end{pmatrix}, \quad C = 1. \quad (4.12)$$

A general reset controller is of this form

$$\begin{cases} u = k_{pt}e + k_{it}z + k_{dt}\omega_r, \\ \dot{z} = e, & t = t_k, \\ z(t^+) = \rho_k(x, z, \theta_r^*), & t \neq t_k, \end{cases} \quad (4.13)$$

where u expresses the control input of (4.11). $z \in \mathbb{R}^q$ is the state of the controller. $\rho_k(x, z, \theta_r^*)$ is the designed reset law using ORL at time instant $t = t_k$. The reset period is $\Delta t_k = t_{k+1} - t_k$. $e = \theta_r^* - \theta_r$ is the tracking error of the PMSM position.

Figure 4.3. ORC with uncertainties observer control scheme

Combining system (4.11) and the reset controller (4.13) yields

$$\begin{cases} \dot{\bar{x}} = \bar{A}\bar{x} + \bar{B}\theta_r^*, & t \neq t_k \\ z(t^+) = \rho_k(\bar{x}, \theta_r^*), & t = t_k \\ \bar{y} = \bar{C}\bar{x} \end{cases} \quad (4.14)$$

where $\bar{x} = (\theta_r \ \omega_r \ z)^T$, $\bar{C} = (1, 0, 0)$ and

$$\bar{A} = \begin{pmatrix} A - Bk_{pt}C & Bk_{it} \\ -C & 0 \end{pmatrix}, \quad \bar{B} = \begin{pmatrix} Bk_{pt} \\ 1 \end{pmatrix}. \quad (4.15)$$

The baseline PID controller ensure the PMSM position control system (4.14) converge to its steady-state $\bar{x}_r = (\theta_r^* \ \omega_{rr} \ z_r)^T$, such that

$$\begin{cases} \bar{A}\bar{x}_r + \bar{B}\theta_r^* = 0, \\ \bar{C}\bar{x}_r - \theta_r^* = 0. \end{cases} \quad (4.16)$$

Therefore, the steady-state control input u_r is given by

$$u_r = k_{it}z_r. \quad (4.17)$$

For each reset time interval $(t_k, t_{k+1}]$, it define a cost function J_k as

$$J_k = \int_{t_k}^{t_{k+1}} [e^T(s)Q_k e(s) + (u(s) - u_r)^T R_k (u(s) - u_r)] ds \quad (4.18)$$

where Q_k and R_k are positive semi-definite matrices. And the cumulative cost function is described as

$$J(\infty) := \sum_{k=0}^{\infty} J_k. \quad (4.19)$$

Perform a coordinate transformation for (4.14) to its steady state,

$$\begin{cases} \xi_{\theta_r} = \theta_r - \theta_r^* \\ \xi_{\omega_r} = \omega_r - \omega_{rr} \\ \xi_z = z - z_r. \end{cases} \quad (4.20)$$

Thus the system (4.14) can be written as

$$\begin{pmatrix} \xi_{\theta_r}(t_{k+1}) \\ \xi_{\omega_r}(t_{k+1}) \\ \xi_z(t_{k+1}) \end{pmatrix} = e^{\bar{A}\Delta t_k} \begin{pmatrix} \xi_{\theta_r}(t_k) \\ \xi_{\omega_r}(t_k) \\ \bar{\rho}_k \end{pmatrix} \quad (4.21)$$

Partition $e^{\bar{A}\Delta t_k}$ as

$$e^{\bar{A}\Delta t_k} = \begin{pmatrix} \Gamma_A(k) & \Gamma_B(k) \\ * & * \end{pmatrix}, \quad (4.22)$$

then

$$\begin{pmatrix} \xi_{\theta_r}(t_{k+1}) \\ \xi_{\omega_r}(t_{k+1}) \end{pmatrix} = \Gamma_A(k) \begin{pmatrix} \xi_{\theta_r}(t_k) \\ \xi_{\omega_r}(t_k) \end{pmatrix} + \Gamma_B(k) \bar{\rho}_k. \quad (4.23)$$

According to [17], the ORC problem is transferred to such a LQR problem that to design a control sequence $\rho(i)$, $i = 0, 1, \dots$, which both results the original system asymptotically stable and makes the cumulative cost function minimized.

By solving the related Riccati equation from [17], the ORL parameters $K(1)$ and $K(2)$ can be determined and ORL is given as

$$\rho_k^* = -K(1)(\theta_r(t_k) - \theta_r^*) - K(2)(\omega_r(t_k) - \omega_{rr}) + z_r. \quad (4.24)$$

where $\omega_{rr} = 0$ and $z_r = 0$. $K(1) = 25.7491$ and $K(2) = 0.1656$ are determined respectively.

The guidelines for ORC design for PMSM position control are summarized as:

1. Design a baseline PID position controller with dynamic form. It is better to design a baseline controller which makes the close-loop system with small damping ratio. Actually, the PID baseline control should firstly ensure the global system stability.
2. Two parameters should be pre-determined manually to solve the Riccati equation which are R_k/Q_k ratio and reset period Δt_k respectively. Select the system sampling period T_s as your PMSM system required. Note that

larger R_k/Q_k ratio leads to less overshoot and larger Δt_k results longer rising time.

Consider the performance requirements and the computational burden of the controller, $R_k/Q_k = 0.12$, $T_s = 5e^{-5}s$ and $\Delta t_k = 2T_s$ are chosen in this section.

4.2.4 Simulation Results

The proposed method is tested in the simulations using Matlab/Simulink software. The performances comparisons between the proposed method and an advanced Sliding Mode Control (I-SMC) are given in this section with some analysis. All the parameters are given in the appendix.

The I-SMC position control is of the same form in [63]. Denote the tracking error $e = \theta_r^* - \theta_r$. Usually, SMC with boundary layer is used to reduce chattering and boundary layer cause steady-state error when there are parameter variations. To eliminate this error, the sliding surface may be extended using an integrator, so the sliding surface S may be chosen as

$$S = \frac{d^2e}{dt^2} + c_{p1}\frac{de}{dt} + c_{p2} + c_{p3} \int_0^t edt. \quad (4.25)$$

The sliding mode position controller is

$$SMC = \beta \cdot \text{sat}\left(\frac{S}{\Phi}\right) \quad (4.26)$$

where c_{p1} , c_{p2} , c_{p3} and β are the sliding mode parameters, Φ is the boundary layer and the sat function is defined as

$$\text{sat}\left(\frac{S}{\Phi}\right) = \begin{cases} \frac{S}{\Phi}, & |\frac{S}{\Phi}| \leq 1 \\ \text{sgn}(\frac{S}{\Phi}), & |\frac{S}{\Phi}| > 1. \end{cases} \quad (4.27)$$

Normally, the proper sliding surface and controller structure are not easy to be selected, because the sliding surface and controller should be chosen to guarantee the global stability PMSM firstly. Moreover, it is hard to determine an optimal group of the sliding mode parameters, c_{p1} , c_{p2} , c_{p3} and β , because it is necessary

to know the uncertainty boundaries. Thus, the parameters are always determined by trial-and-error method.

Figure 4.4 shows the comparisons of step position responses between I-SMC and ORC with uncertainties observer. The position reference is set as $\theta_r^* = 1\text{rad}$. The external load torque is $T_L = 1.27\text{N} \cdot \text{m}$. Initially, the response speed of the ORC with uncertainties observer control scheme is faster than I-SMC. The two control schemes make the position tracking convergent to the reference almost at the same time. Figure 4.5 shows the zoom of Figure 4.4. As shown, the

Figure 4.4. Step position responses of different control schemes

static errors of both of the control schemes are small enough and close two each other. However, the steady state fluctuations is more severe of the ORC with uncertainties observer control scheme which can be eliminated by considering unreset actions in the ORL design. This will be a topic of our future works.

The simulations with uncertainties caused by parametric variations and external load torque disturbances are given in Figure 4.6. The PMSM parametric variations exists as $\Delta D = +0.1D$ and $\Delta J = +0.1J$. And the external load torque is suddenly changed from $T_L = 1.27\text{N} \cdot \text{m}$ to $T_L = 0.1\text{N} \cdot \text{m}$ at time $t = 0.25\text{s}$. The red ellipses mark the transient fluctuations of the two control schemes in figure (a) and (b) respectively. It is obvious that the anti-uncertainties performance of the proposed control scheme is as good as I-SMC has.

Figure 4.5. Zoom in of position responses steady-state

The responses of position reference changing tracking test with the two control scheme are shown in Figure 4.7. The position reference is suddenly changed from $\theta_r^* = 1rad$ to $\theta_r^* = 2rad$ at time $t = 0.25s$. As seen in Figure 4.7, the response speed of the ORC with uncertainties observer control scheme is much faster than I-SMC at beginning of reference changing. I-SMC make the tracking convergent faster when real position is near the tracking reference.

4.3 Current Control of PMSM Drive System

Since the current control performances directly relate to the machine torque control [42] and the drive performance, it is necessary to design high performance current controllers for PMS machine.

The PMSM parameters R , L and ψ may vary around their designed value [30]. The derivatives of parametric variations of R , L and ψ are defined as ΔR , ΔL and $\Delta\psi$. And denote R_s , L_s and ψ_s as the designed values. So $R = R_s + \Delta R$, $L = L_s + \Delta L$ and $\psi = \psi_s + \Delta\psi$. Therefore, PMSM drive system (2.3) can be rewritten as

$$v_q = (R_s + \Delta R)i_q + (L_s + \Delta L)\frac{di_q}{dt} + (L_s + \Delta L)\omega i_d + \omega(\psi_s + \Delta\psi)$$

Figure 4.6. Position responses of external torque changes with different control schemes

$$v_d = (R_s + \Delta R)i_d + (L_s + \Delta L)\frac{di_d}{dt} - (L_s + \Delta L)\omega i_q. \quad (4.28)$$

Consider that the modeling of PMSM drive system consists of two parts which are the basic dynamic parts described in (2.3) and the uncertainties parts which are shown as σ_d and σ_q ,

$$\frac{di_d}{dt} = -\frac{R_s}{L_s}i_d + \frac{1}{L_s}v_d + \omega i_q - \frac{1}{L_s}\sigma_d$$

Figure 4.7. Position responses of position reference changes with different control schemes

$$\frac{di_q}{dt} = -\frac{R_s}{L_s}i_q + \frac{1}{L_s}v_q - \omega i_d - \frac{1}{L_s}\omega\psi_s - \frac{1}{L_s}\sigma_q \quad (4.29)$$

where $\sigma_q = \varepsilon_q + \Delta R i_q + \Delta L \frac{di_q}{dt} + \Delta L \omega i_d + \omega \Delta \psi$ and $\sigma_d = \varepsilon_d + \Delta R i_d + \Delta L \frac{di_d}{dt} - \Delta L \omega i_q$ express the uncertainties parts of the two axes. ε_d and ε_q are the un-modeled dynamics.

In this section, the ORC principle is shown firstly. Based on the basic electrical model (2.3), the conventional PI current controllers are designed for both the d, q axes. Then the ORC is applied on the current controller designs. To eliminate the uncertainties caused by parametric variations and un-modeled dynamics, ORC two pieces cascaded uncertainty observers are designed as well.

The overall PMSM system structure is shown in Figure 4.8. In the design of PI current controllers for q and d axes, the basic electrical model (2.3) is considered without the parametric variations uncertainties and un-modeled dynamics. The coupling terms $L\omega i_d$, $\omega\psi$ and $L\omega i_q$ can be measured or exists as constant. Whereas the feed-forward control technique is employed into the PI controllers to decouple the two axes coupling terms. Given the general forms of PI current

Figure 4.8. Joint ORC current control scheme of PMSM

controllers as,

$$\begin{aligned} v_q &= (k_p + k_i \frac{1}{s})(i_q^* - i_q) + L_s \omega i_d + \omega \psi_s + \varepsilon_q \\ v_d &= (k_p + k_i \frac{1}{s})(i_d^* - i_d) - L_s \omega i_q + \varepsilon_d. \end{aligned} \quad (4.30)$$

Due to symmetrical structures of the decoupled d, q axes, the same controller parameters k_p and k_i are chosen. This section uses the pole placement method to design the controller parameters.

Then the closed-loop transfer functions of d, q axes are given as

$$G(s) = \frac{i_q}{i_q^*} = \frac{i_d}{i_d^*} = \frac{k_p s + k_i}{L_s s^2 + (R_s + k_p)s + k_i} \quad (4.31)$$

Assuming the closed-loop poles of the 2-order close-loop control system are $s_{1,2} = -\xi \omega_n \pm j \omega_n \sqrt{1 - \xi^2}$ where ω_n and ξ express the desired undamped natural frequency and the damping ratio, respectively. The characteristic equation of the close-loop control system should be

$$D(s) = s^2 + 2\xi \omega_n s + \omega_n^2. \quad (4.32)$$

Comparing (4.31) and (4.32), it is easy to get the corresponding PI controller parameters $k_p = 2\xi\omega_n L - R$ and $k_i = \omega_n^2 L$. Chosen $\xi = 1$ and $\omega_n = 450$ [30].

4.3.1 ORC Application for PMSM Current Control

According to the proposed discrete-time ORC in section 3.3, decoupling the two axes of (4.30), the electrical model of PMSM can be considered as two 1-step individual plants,

$$\begin{aligned}\dot{i}_q &= -ai_q + bv_q - b\Omega_q \\ \dot{i}_d &= -ai_d + bv_d - b\Omega_d\end{aligned}\tag{4.33}$$

where $a = \frac{R}{L}$, $b = \frac{1}{L}$ and $b\Omega_q$ and $b\Omega_d$ express the coupling and uncertainties terms of the two axes, respectively. The ORC is applied on both the current controller designs of the two axes. For the two constant reference tracking problems, the q, d current ORC controllers are given as,

$$\begin{cases} v_q(k, i) = u_q(k, i) \\ u_q(k, i) = k_p(i_q^*(k, i) - i_q(k, i)) + k_i z(k, i) + \Omega_q \\ z(k, i) = E(i_q^*(k, i) - i_q(k, i)) + Dz(k, i - 1) \\ z(k, 0) = \rho_q(k, 0)(i_q(k - 1, n - 1), i_q^*(k - 1, n - 1), i_q^*(k, 0)) \end{cases}\tag{4.34}$$

$$\begin{cases} v_d(k, i) = u_d(k, i) \\ u_d(k, i) = k_p(i_d^*(k, i) - i_d(k, i)) + k_i z(k, i) + \Omega_d \\ z(k, i) = E(i_d^*(k, i) - i_d(k, i)) + Dz(k, i - 1) \\ z(k, 0) = \rho_d(k, 0)(i_d(k - 1, n - 1), i_d^*(k - 1, n - 1), i_d^*(k, 0)) \end{cases}\tag{4.35}$$

$D = 1$ and $E = T_s$ are the constant connection parameters. The ratio $R_c/Q_c = 0.006$ and the reset period $T_{rc} = 2T_s$. k_p and k_i are the parameters of the baseline PI controllers. The ORCs of the two axes controls are

$$\begin{aligned}\rho_q(k, 0) &= \frac{ai_q^*}{bk_i} - K_q(i_q(k - 1, n - 1) - i_q^*(k - 1, n - 1)) \\ \rho_d(k, 0) &= \frac{ai_d^*}{bk_i} - K_d(i_d(k - 1, n - 1) - i_d^*(k - 1, n - 1)).\end{aligned}$$

By solving the Riccati equation, reset parameters $K_q = K_d = 0.0075$ can be determined.

4.3.2 ORC Uncertainties Observer

From the system (4.33), the two axes coupling and uncertainty terms $b\Omega_q$ and $b\Omega_d$ can be described as

$$\begin{aligned} b\Omega_q &= \omega i_d + \frac{1}{L_s} \omega \psi_s + \frac{1}{L_s} \sigma_q \\ b\Omega_d &= -\omega i_q + \frac{1}{L_s} \sigma_d. \end{aligned} \quad (4.36)$$

Obviously, in the system (4.28), the real-time stator end voltages v_q and v_d , stator currents i_q and i_d and the electrical rotational speed ω are all available through measurements.

Assumption 4. *If the uncertainty observers are fast enough to converge to the real uncertainty values while the uncertainties σ_q and σ_d do not change abruptly. It is reasonable to assume that the derivatives of σ_q and σ_d are zero during each sampling period. [36]*

The state observers are designed by propagating the input signals, v_q , Ω_q and v_d , Ω_d . The states are furthermore updated by scalings, L_q and L_d , of the errors in estimated outputs as described as

$$\begin{aligned} \dot{\hat{i}}_q &= -a\hat{i}_q + bv_q - b\Omega_q + L_q(i_q - \hat{i}_q) \\ \dot{\hat{i}}_d &= -a\hat{i}_d + bv_d - b\Omega_d + L_d(i_d - \hat{i}_d). \end{aligned} \quad (4.37)$$

The observer gains L_q and L_d are designed by the Kalman Filtering (KF) [119] approach.

In PMSM, the uncertainties σ_q and σ_d are not measurable which means that extended observer loops should be employed to estimate them. As a result, ORC two pieces cascaded coupled uncertainty observers are proposed to estimate σ_q and σ_d , respectively. The basic structure of it is given in Figure 4.9.

Due to the symmetrical structures of the d, q axes, the ORC uncertainties observer of d axis shares the same structure of the q axis as given. The inner loop is a kalman filter designed with respect to (4.37). The outer loop operates as a tracking configuration with i_q or i_d as the tracking objectives. Theoretically, ORC ensures the perfect performance of outer loop tracking. The the observers' asymptotic tracking can be ensured as long as that the ORC outer-loop controllers are well designed [119]. The uncertainties (σ_q and σ_d) can be estimated as the ORC output plus the kalman filter output, respectively.

The pole placement designed PI baseline controllers for the ORC observers are shown as,

$$\begin{aligned}\frac{1}{L_s}\sigma_q &= (k_{po} + k_{io}\frac{1}{s})(i_q - \hat{i}_q) - L\omega i_d - \omega\psi + v_q \\ \frac{1}{L_s}\sigma_d &= (k_{po} + k_{io}\frac{1}{s})(i_d - \hat{i}_d) + L\omega i_q - v_d.\end{aligned}\quad (4.38)$$

where the observer parameters are $k_{po} = 2\xi_e\omega_{ne} - \frac{R_s}{L_s}$ and $k_{io} = \omega_{ne}^2$. ξ_e and ω_{ne} are chosen as 1 and 450, respectively. The estimated uncertainties $\hat{\sigma}_q$ and $\hat{\sigma}_d$ are expressed as

$$\begin{aligned}\hat{\sigma}_q &= (k_{po} + k_{io}\frac{1}{s})(i_q - \hat{i}_q) - L_q(i_q - \hat{i}_q) \\ \hat{\sigma}_d &= (k_{po} + k_{io}\frac{1}{s})(i_d - \hat{i}_d) - L_d(i_d - \hat{i}_d).\end{aligned}\quad (4.39)$$

Figure 4.9. Block diagram of ORC uncertainties observer structure

So the ORC observer systems can be rewritten as the following systems regarding the real currents i_q and i_d as the time-varying references, respectively,

$$\begin{aligned}\dot{\hat{i}}_q &= -a\hat{i}_q - \hat{\sigma}_q \\ \dot{\hat{i}}_d &= -a\hat{i}_d - \hat{\sigma}_d.\end{aligned}\tag{4.40}$$

Thus the ORC observers are given as

$$\begin{cases} -\hat{\sigma}_q(k, i) = u_{qo}(k, i) \\ u_{qo}(k, i) = k_{po}(i_q(k, i) - \hat{i}_q(k, i)) + k_{io}z(k, i) + \gamma_q \\ z(k, i) = E(i_q(k, i) - \hat{i}_q(k, i)) + Dz(k, i - 1) \\ z(k, 0) = \rho_{qo}(k, 0)(\hat{i}_q(k - 1, n - 1), i_q(k - 1, i - 1), i_q(k, 0)) \end{cases}\tag{4.41}$$

$$\begin{cases} -\hat{\sigma}_d(k, i) = u_{do}(k, i) \\ u_{do}(k, i) = k_{po}(i_d(k, i) - \hat{i}_d(k, i)) + k_{io}z(k, i) + \gamma_d \\ z(k, i) = E(i_d(k, i) - \hat{i}_d(k, i)) + Dz(k, i - 1) \\ z(k, 0) = \rho_{do}(k, 0)(\hat{i}_d(k - 1, n - 1), i_d(k - 1, i - 1), i_d(k, 0)) \end{cases}\tag{4.42}$$

where $\gamma_q = L_q(i_q(k, i) - \hat{i}_q(k, i))$ and $\gamma_d = L_d(i_d(k, i) - \hat{i}_d(k, i))$. The ratio $R_o/Q_o = 0.4$ and the reset period $T_{ro} = T_s$. The corresponding ORLs are

$$\begin{aligned}\rho_{qo}(k, 0) &= \frac{1}{ak_{io}}(1 - e^{-aT_s})(i_q(k, 0) - i_q(k - 1, n - 1))\frac{1}{T_s} \\ &\quad - \frac{1}{ak_{io}}(1 - e^{-aT_s})e^{-aT_s}i_q(k, 0) \\ &\quad - K_{qo}(\hat{i}_q(k - 1, n - 1) - i_q(k - 1, n - 1)) \\ \rho_{do}(k, 0) &= \frac{1}{ak_{io}}(1 - e^{-aT_s})(i_d(k, 0) - i_d(k - 1, n - 1))\frac{1}{T_s} \\ &\quad - \frac{1}{ak_{io}}(1 - e^{-aT_s})e^{-aT_s}i_d(k, 0) \\ &\quad - K_{do}(\hat{i}_d(k - 1, n - 1) - i_d(k - 1, n - 1)).\end{aligned}\tag{4.43}$$

By solving the Riccati equation, ORC observer parameters can be derived as $K_{qo} = K_{do} = -0.0028$.

Denote that $\hat{e}_q = \sigma_q - \hat{\sigma}_q$ and $\hat{e}_d = \sigma_d - \hat{\sigma}_d$ are the derivatives between the real

uncertainty values and their estimation values. According to the ORC principle, the observers' asymptotic tracking can be ensured by the ORC. In other words, $\hat{e}_q \rightarrow 0$ and $\hat{e}_d \rightarrow 0$ can be realized in finite time. With the feed-forward compensation technique of the uncertainty terms, the global current control stability of PMSM can be ensured by the ORC joint control scheme.

The guidelines for design of ORC current controllers and ORC uncertainties observers joint control scheme for PMSM current control are summarized as:

1. Design baseline current controllers with dynamic form. In this section, the pole placement designed PI baseline controllers are considered. It is better to design baseline controllers which makes the close-loop system with small damping ratio.
2. The sampling period of the PMSM current control system is T_s . Then choose the pair of R_c/Q_c and the reset period T_{rc} . Note that larger R_c/Q_c ratio leads to less overshoot and larger T_{rc} results longer rising time.
3. Choose the pair of R_o/Q_o and the reset period T_{ro} of the ORC observers following the law in the second step. Note that $T_{ro} \leq T_{rc}$ always be chosen in order to ensure the assumption 4.

Consider the industrial application and the computational burden of the controllers, the system sampling time $T_s = 2.5 \times 10^{-5}$ is chosen in this problem. Actually, the proposed ORC is an advanced control with open design principles. In other words, according to actual demand of PMSM, H_2 and H_∞ problems can easily be involved in the design of optimal reset law too.

4.3.3 Simulation Results

In this section, the proposed joint ORC current control scheme for PMSM is tested in both the simulations and experiments. The performance comparisons between the conventional decoupling PI control scheme, advanced SMC and the proposed method are given with analysis. All the parameters are in the appendix.

The conventional decoupling PI current controllers are given in (4.30). And the SMC current controls are chosen from [30] for comparisons. Denote the tracking error as $e_q = i_q^* - i_q$ and $e_d = i_d^* - i_d$. The sliding surfaces of the two axes S_q and S_d are

$$\begin{aligned} S_q &= e_q + c_q \int e_q \\ S_d &= e_d + c_d \int e_d. \end{aligned} \quad (4.44)$$

The conventional SMC uses control laws with large control gains yielding the undesired chattering while the control system is in the sliding mode. To eliminate the chattering, a boundary layer technique was usually adopted. However, the boundary layer may cause steady-state error when there are parameter variations. The sliding mode current controllers for the PMSM are

$$\begin{aligned} v_q &= L_s \frac{di_q^*}{dt} + R_s i_q + L_s \omega i_d + \omega \psi_s + c_q e_q + m_q \text{sat}\left(\frac{S_q}{\Phi}\right) \\ v_d &= L_s \frac{di_d^*}{dt} + R_s i_d - L_s \omega i_q + c_d e_d + m_d \text{sat}\left(\frac{S_d}{\Phi}\right). \end{aligned} \quad (4.45)$$

where c_q , c_d , m_q and m_d are the sliding mode parameters, Φ is the boundary layer and the **sat** function is defined as

$$\text{sat}\left(\frac{S}{\Phi}\right) = \begin{cases} \frac{S}{\Phi}, & |\frac{S}{\Phi}| \leq 1 \\ \text{sgn}(\frac{S}{\Phi}), & |\frac{S}{\Phi}| > 1. \end{cases} \quad (4.46)$$

Normally, the proper sliding surfaces and controller structures are not easy to be selected, because they should be chosen to guarantee the global stability of PMSM system at least. The limitation of SMC is that the large control gains are difficult to be designed because the upper bounds of uncertainties are usually unknown. Moreover, it is hard to determine an optimal group of the sliding mode parameters, c_q , c_d , m_q and m_d which are always dependent on complex process.

All the simulations are set up in MatLab/Simulink. The d, q current references are set as $i_d^* = 0A$ and $i_q^* = 4A$. The rotational speed of PMSM is set at $500rpm$.

Figure 4.10 shows the comparisons of current step responses among conven-

tional PI decoupling current controllers, the proposed joint ORC control schemes and the sliding mode current controllers. The transient responses have been emphasized in each sub-figure. In this test, no uncertainty exists in the PMSM system.

(a) current responses of PI

(b) current responses of ORC

(c) current responses of SMC

Figure 4.10. Step current responses of different control schemes without uncertainty

In Figure 4.10 (a), current step responses of conventional decoupling PI controllers are shown. Although good current control performances are achieved, they result in relatively large current fluctuations. The reason which leads to the fluctuations is that conventional decoupling PI controllers are too sensitive to input voltage harmonics and noises. In Figure 4.10 (b) and Figure 4.10 (c), the current step responses of ORC current controls and sliding mode current controls are given, respectively. The response speeds of both ORC and SMC controls are much faster than that of conventional decoupling PI controllers. Moreover, the current fluctuations are restricted in smaller ranges than that of the conventional decoupling PI controllers.

Figure 4.11 shows the current step responses of the PMSM system with uncertainties. The PMSM parametric variations exist as $\Delta R = +0.1R_s$, $\Delta L = +0.2L_s$ and $\Delta\psi = +0.3\psi_s$ while the un-modeled dynamics of d, q axes are assumed as $\varepsilon_d = 0.05V$ and $\varepsilon_q = 0.05V$.

Figure 4.11 (g) gives the current step responses of the aforementioned sliding mode controllers with SMC uncertainty observers. The SMC uncertainties observers are of the following forms,

$$\begin{aligned}\sigma_{sq} &= \frac{1}{\eta_q} \int S_q \\ \sigma_{sd} &= \frac{1}{\eta_d} \int S_d.\end{aligned}\tag{4.47}$$

where η_q and η_d are the SMC observer parameters. The new sliding mode current controllers for the PMSM are given as

$$\begin{aligned}v_q &= L_s \left(\frac{di_q^*}{dt} + \frac{R_s}{L_s} i_q + \omega i_d + \frac{\omega}{L_s} \psi_s + \frac{c_q}{L_s} e_q + \frac{m_q}{L_s} \text{sat}\left(\frac{S_q}{\Phi}\right) + \sigma_{sq} \right) \\ v_d &= L_s \left(\frac{di_d^*}{dt} + \frac{R_s}{L_s} i_d - \omega i_q + \frac{c_d}{L_s} e_d + \frac{m_d}{L_s} \text{sat}\left(\frac{S_d}{\Phi}\right) + \sigma_{sd} \right).\end{aligned}$$

Using the conventional decoupling PI controllers, larger current fluctuations are unavoidable because of the system uncertainties as shown in Figure 4.11 (a). In Figure 4.11 (e), the current step responses of conventional sliding mode controls are given. The conventional SMC will result in big steady-state errors. It

(a) current step responses of PI

(b) voltage step responses of PI

(c) current step responses of ORC

(d) voltage step responses of ORC

(e) current step responses of conventional SMC

(f) voltage step responses of conventional SMC

(g) current step responses of SMC with uncertainty observers

(h) voltage step responses of SMC with uncertainty observers

Figure 4.11. Step current and voltage responses of different control schemes with uncertainties

means that the sliding mode cannot be achieved in finite time which may cause sliding function divergent. If SMC uncertainty observers are employed in the conventional sliding mode controllers. As shown in Figure 4.11 (g), steady-state errors can be eliminated. Although the new sliding mode control schemes ensure fast current control responses with very small steady-state fluctuations, it is hard to determine an optimal group of SMC parameters and SMC observer parameters. In order to make that the SMC observers cooperate well with the SMC current controllers, lots of manual works should be done to adjust the parameters. As a result, it makes the SMC design complicated. Figure 4.11 (c) shows the current step responses of the ORC joint control scheme. It seems that the uncertainties have few effect on the current control performances. The responses are still very fast and the steady-state current fluctuations are small. Figure 4.11 (b)(d)(f)(h) give the d, q voltage responses of the conventional decoupling PI controllers, the ORC joint control scheme, the conventional SMC and the SMC with uncertainty observers, respectively.

In Figure 4.12, the estimation performances of the ORC uncertainty observers and the SMC uncertainty observers are compared in the condition of the former test.

(a) uncertainty estimations of ORC uncertainty observer (b) uncertainty estimations of SMC uncertainty observer

Figure 4.12. Uncertainties estimation performances of different observers

In Figure 4.12 (a), one can see that the ORC uncertainty observers converge very fast which is important for the feed-forward compensations of the uncertainty

terms. According to the linear design principle, the optimal parameters of the ORC uncertainty observers can be determined easily. Figure 4.12 (b) shows the estimation performances of the SMC uncertainty observers. Although q axis SMC uncertainties observer converge very fast, the d axis SMC uncertainties observer converge much slower than the ORC observer. It may cause steady-state error for d axis current control. Furthermore, due to their structures, the estimation values of SMC observers will oscillate in a range, but not the exact values.

4.3.4 Experimental Results and Remarks

The effectiveness of the proposed joint control scheme is investigated via implementation in the PMSM experimental setup. Figure 4.13 shows the entire experimental setup.

Figure 4.13. PMSM drive system experimental setup

The experimental setup was composed of an external BLDC motor controlled by a DSP controller box which works as an external rotor speed drive motor. The rotational speed of the PMSM can be maintained at $500rpm$ by the external BLDC motor. The input terminal of the PMSM are connected to a 3-phase SEMIKRON 6-IGBT gates SKM50Gb 123D converter which works as an inverter for the PMSM power supply. LEM LA-25 current transducers connected to three ADC ports of the dSPACE 1103 controller card are used as PMSM current measurements. The controller board is a 400 MHz Power PC processor dSPACE

1103 which has 6 incremental encoders, analog-to-digital (ADC) and digital-to-analog (DAC) ports. The 32 bit encoder port measures the angular velocity of the PMSM for controller feedback. The power rating of the PMSM is 400 watts. Tests were done at setting the dc power supply to 48V.

The current control references are $i_q^* = 1A$ and $i_d^* = 0A$. The real resistance, impedance and magnet flux of the PMSM stator are R_s , L_s and ψ_s , respectively. In order to test the robust against uncertainty capability of the current control schemes, controllers are designed based on the group of virtual parameters $R' = 0.9R_s$, $L' = 0.8L_s$ and $\psi' = 0.8\psi_s$. It can be seen as an experimental PMSM of parameters $0.9R_s$, $0.8L_s$ and $0.8\psi_s$ which operates with parametric variations of $0.1R_s$, $0.2L_s$ and $0.2\psi_s$.

(a) d, q current responses of PI

(b) d, q current responses of conventional SMC

(c) d, q current responses of ORC

Figure 4.14. Experimental d, q current responses of different controls

Figure 4.14 shows some experimental performance comparisons. The current control responses of the conventional decoupling PI controllers, the conventional

SMC controllers and the proposed ORC joint control schemes are shown in the sub-figures (a), (b) and (c) of Figure 4.14, respectively. From the different experimental results, one can see that the responses of the proposed joint control schemes are smoother with smaller static errors than those by using the conventional decoupling PI controllers. However, the conventional SMC controllers will result in big steady-state errors.

The estimated uncertainties of d, q axes by the proposed ORC uncertainty observers are shown in Figure 4.15. It is obvious that the estimated uncertainties of both of the axes can converge to their desired values well. So the uncertainty terms can be compensated in the current control loops. Thus, the proposed ORC joint control schemes can achieve the best current control performances as shown in Figure 4.14 (c). The stability and robustness of the PMSM current control system can be guaranteed.

Figure 4.16 shows some of the transient responses of the proposed ORC joint control schemes. The d, q currents responses of the ORC joint control scheme of the experimental results are given in Figure 4.16 (a). The transient performances of the corresponding ORC observers in the Figure 4.16 (b). In Figure 4.16 (c), the transient responses of the a, b, c 3-phase currents are shown and the transient rotor speed response is shown in Figure 4.16 (d). It takes about 0.05s for the external BLDC motor to drive the rotational speed of the testing PMSM to the pre-defined value 500rpm in the negative direction. During this transient period, the d, q currents, the outputs of the ORC observers and the a, b, c currents contain small over-shoots or errors to their rated values until that the rotor speed of PMSM reaches 500rpm. However, when the rotor speed of PMSM maintains at the steady-state, the ORC current responses and the ORC observer responses can

Figure 4.15. Experimental d, q uncertainty estimation performances of ORC

be controlled to acceptable performances rapidly. It can come to the conclusion that the proposed ORC joint control schemes realize fast current responses, fast uncertainty estimations and robust against uncertainties in the PMSM current controls.

4.4 Time-varying Speed Control of PMSG

A permanent magnet synchronous generator is a generator where the excitation field is provided by a permanent magnet instead of a coil. In a permanent magnet generator, the magnetic field of the rotor is produced by permanent magnets. Other types of generator use electromagnets to produce a magnetic field in a

Figure 4.16. Transient performances of ORC joint control

rotor winding. The direct current in the rotor field winding is fed through a slip-ring assembly or provided by a brushless exciter on the same shaft. The ORC speed control of PMSG is given in this section.

4.4.1 Baseline Speed Controllers

The overall system structure with control scheme for the rectifier is shown in Figure 4.17.

Figure 4.17. Rectifier control schemes

According to the PMSG model given in Section 2.2.2, take Laplace Transformation for (2.6),

$$(-R_s - sL_s)i_{sd} = U_{sd} - L_s\omega i_{sq} \quad (4.48)$$

$$(-R_s - sL_s)i_{sq} = U_{sq} + L_s\omega i_{sd} - \omega\psi. \quad (4.49)$$

Considering the symmetry of d and q , only the controller design of the q axis current will be shown. Set i_{sq} as the control object and U_{sq} as the output of the controller. According to (4.48), i_{sq} not only corresponds to the q axis reference current i_{sq}^* , but be perturbed by both the d axis current i_{sd} and the magnet flux ψ . So there are two design aims for the PI controller. One of them is fast tracking

to the current loop reference i_{sq}^* with no static error, the other is restraining the disturbances from i_{sd} and ψ to its best. PI is normally hard to achieve the second aim.

$L_s\omega i_{sd}$ and ψ can be measured or exists as a constant. Whereas the feed-forward control method is employed into the PI to decouple the two terms [131]. A general inner-loop PI controller for this system is of the form

$$U_{sq} = -(k_p + k_i \frac{1}{s})(i_{sq}^* - i_{sq}) - L_s\omega i_{sd} + \omega\psi \quad (4.50)$$

Similarly, the direct axis controller should be of the form

$$U_{sd} = -(k_p + k_i \frac{1}{s})(i_{sd}^* - i_{sd}) + L_s\omega i_{sq}. \quad (4.51)$$

For simplification, the same controller parameters value k_p and k_i are chosen. And d axis current reference $i_d^* = 0$ strategy is applied in this section.

4.4.2 Discrete-time ORC Speed Control

Assume that the varying-time shaft rotational speed reference ω_m^* is available. Electrical torque T_{em} should be controlled for the variable ω_m^* tracking. An outer-loop controller of ending i_q^* for the q axis control is built for the turbine MPPT with control parameters k_{pw} and k_{iw} . All the PI parameters k_p , k_i , k_{pw} and k_{iw} used in this section are designed by pole placement method.

Decoupling the relationships between the two axis of (4.48). The PMSG can be considered as three PI controlled 1-step plants. The inner-loop current controllers for PMSG can refer to the current controller design for PMSM. Thus, following the Section 3.3, the discrete-time ORC is applied on the outer-loop speed control which is given with the decoupling term Ω as,

$$\begin{cases} u(k, i) = k_{pw}e(k, i) + k_{iw}z(k, i) + \Omega \\ z(k, i) = Ee(k, i - 1) + Dz(k, i - 1) \\ z(k, 0) = \rho(k, 0)(x(k - 1, n - 1), r(k - 1, n - 1), r(k, 0)) \end{cases} \quad (4.52)$$

$D = 1$ and $E = T_s$ are the connection parameters between system and ORC which are determined by the structure of the system itself. The ORL is given by

$$\begin{aligned}
\rho^*(k, 0) &= -k_1(x(k-1, n-1) - r(k-1, n-1)) \\
&\quad + (H^T H)^{-1} H^T (B_d^T B_d)^{-1} B_d^T (x_r(k-1, n-1) - x_r(k-1, n-2)) \frac{1}{T_s} \\
&\quad - (H^T H)^{-1} H^T (B_d^T B_d)^{-1} B_d^T A_d x_r(k-1, n-1) \\
&= -k_1(x(k-1, n-1) - r(k-1, n-1)) \\
&\quad + \frac{1}{k_{iw}} \frac{-a}{b(1 - e^{-aT_s})} [-e^{-aT_s} r(k-1, n-1) \\
&\quad + \frac{1}{T_s} (r(k-1, n-1) - r(k-1, n-2))]. \tag{4.53}
\end{aligned}$$

By solving the Riccati equation, reset parameter k_1 can be determined. Two parameters should be pre-determined manually to solve the Riccati equation which are R/Q ratio and reset time period $T_r = nT_s$ respectively. Figure 4.18 gives the step responses for different R/Q ratio and different reset time period $T_r = nT_s$, respectively.

Obviously, smaller R/Q ratio leads to faster response. And the step response performance is better when reset time period $T_r = nT_s$ is smaller.

Consider the needs from industrial application and the computational burden of the controller, the pair of optimal values $R/Q = 13$ and $n = 2$ are chosen. In this case, reset parameter $k_1 = -0.0061$.

Remark 3. Figure 4.19 shows the responses for ORC with Different System T_s . On one hand, if T_s is too small, it will cause big computation burden. When choose $T_s = 1 \times 10^{-6}$, it does not improve the performance too much compared to that of $T_s = 2.5 \times 10^{-5}$. However, too small T_s is always unsuitable in industrial applications. On the other hand, if T_s is too big, the ORC may not cooperate well with the baseline system for big cumulative errors. It cannot always improve the transient performance or even make it worse. The curve for $T_s = 1 \times 10^{-4}$ gives a relatively worse dynamic performance than that of the case for $T_s = 2.5 \times 10^{-5}$. $T_s = 2.5 \times 10^{-5}$ is chosen in this problem.

Actually, the ORC speed controller for the PMSG in WTPGS shares the same

Figure 4.18. Step response for different R/Q and different T_r

Figure 4.19. ORC responses for different system sampling time T_s

structure as that of a PMS machine. The PMSG system can be regarded as a inverse system of PMSM. The ORC speed controller will be applied in the PMSG in the next chapter and its effectiveness will also be verified.

The guidelines for ORC design are summarized as:

1. Design a baseline controller with dynamic form. In this section, only the PI baseline controller is considered. It is better to design a baseline controller which makes the close-loop system with small damping ratio. Actually, no matter what your baseline controller's parameters are, ORC will always have a better performance than the applied baseline controller has.
2. Select the minimum system sampling time T_s which can be tolerated by your application. Then choose pairs of R/Q and T_r values and calculate out the corresponding ORL. By comparing, pick up the most suitable parameter pair R/Q , T_r and ORL as your ORC parameters. Note that larger R/Q ratio leads to less overshoot and larger T_r results longer rising time.
3. Prolong the system sampling time T_s step by step. Choose the one which makes the control system under the minimum computational burden and guarantees the system an acceptable dynamic performance as well. Keep this T_s as the final system sampling time.

4.5 Concluding Remarks

In this chapter, the ORC technique has been applied on design of PMS machine position controller. The ORC position controller is designed based on PID and PI two loops linear baseline position controller. However, the transient performance can be improved far beyond the ability of linear baseline position controller. Overall, compared with other advanced control schemes, the ORC position controller holds its advantages such as relatively simpler linear design principle and easily application to other PMS machine position control drive systems. Due to that the uncertainties caused by parametric variations and external load torque disturbances are unavoidable, a linear two pieces cascaded coupled uncertainties

observer is proposed for the uncertainties feed-forward compensation. Since the observer can easily be ensured convergent by the design, the global PMS machine position control system stability is guaranteed easily. The simulation results show the performance comparisons between the ORC and uncertainties observer joint control scheme and an advanced SMC. It is obvious that the position control performance of ORC with uncertainties observer joint control scheme is as well as that of the advanced SMC. However, the design process is relatively easier.

Subsequently, the ORC technique is used to design current controllers and uncertainties observers of the PMS machine system. The ORC current controllers are designed based on linear PI baseline controllers. And the ORC uncertainty observers are designed based on linear two pieces cascaded coupled uncertainty observers for the uncertainty feed-forward compensations. The ORC current controllers are applied to improve the transient current control performances and reduce the steady-state current ripples. Meanwhile, the ORC uncertainty observers are proposed to accelerate the convergent speed of observers. Since the observers can easily be ensured fast convergent to the real uncertainties, the system stability won't be influenced by the uncertainties in finite time and the global stability can be guaranteed easily. Overall, compared with other advanced control schemes, the ORC joint method holds its advantages such as relatively simple linear design principle and easily application in other PMS machine systems. Both the simulation and experimental results show the performance comparisons between the ORC joint current control scheme and some other advanced control schemes. It is obvious that the ORC joint control scheme can achieve good current control performances.

Furthermore, the discrete-time optimal reset control design method is employed to design the speed controller for PMS machine systems. In order to reduce the overshoot of the step response without degrading other specifications, suppress controller's saturation effectively and accelerate the tracking speed with less error, the discrete-time ORC technique is proposed for speed reference tracking control systems, especially for time-varying speed reference. The simulation results guide us the methods for determination of the ORC parameters. Actually, the ORC speed controller for the PMSG in WTPGS shares the same structure as

that of a PMS machine. The ORC speed controller will be applied in the PMSG and its effectiveness will also be verified in the next chapter.

With the advanced controllers proposed in this chapter, the generator control performances of the WTPGS installed with PMSG will be greatly improved. And then, there are some advanced control methods for WTPGS given in the next chapter including MPPT control, wind speed estimation, WTPGS sensorless control and pitch control.

Controls of Wind Energy Conversion System

5.1 Introduction

In recent years there has been an increasing awareness about the climate change and the harmful effects that the emissions of carbon have. This created a higher demand for clean and sustainable energy sources. The wind energy has experienced the biggest growth in the past decade. However, the main drawback of the wind energy is that it is irregular in occurrence [132]. The problem becomes how to maximize the energy capture from the wind.

Due to the disadvantages of the fixed speed wind generators such as the low efficiency, poor power quality, high mechanical stress, the variable-speed wind turbine [133] became the most attractive option in the past years because of the development of power electronic devices. In the ideal case of control, the wind turbine generator in variable speed can extract the maximum power [134]. The variable-speed wind turbine is proved to yield more energy than the fixed-speed wind turbine. Actually, for a WTPGS, it is always desirable to capture as much power as possible from the wind. This depends on both that the WTPGS is able to obtain accurate and real-time wind information and efficient and accurate controls are employed in the WTPGS control systems. In the case that the wind

speed is definitely obtained, the problem goes to design advanced rotational speed controller for WTPGS which is effective for time-varying reference, responds fast and accurate, can reasonably protect the system and has anti-windup capacity. In the case that the wind speed cannot be accurately measured, the WTPGS's main concern is to real-time estimate wind speed and then realize the sensorless maximum power extracted control.

The topology with PMSG and full-scale converter has an increasing market share today which is considered as the development trend of WTPGS to some extent. Compared with the induction generator, the permanent magnet synchronous generator is more efficient, smaller in size and easier to control. The efficiency of the PMSG wind turbine, was assessed to be higher than other variable speed wind turbine concepts [135]. In this chapter, take PMSG WTPSG as a representative, the WTPGS control problems are studied in the both conditions that wind speed is available and wind speed cannot be accurately measured, respectively.

5.2 Maximum Power Point Tracking (MPPT) Control of WTPGS

First of all, the case that that wind speed is available is considered in this section. According to Section 2.3.1,

$$\begin{aligned} P_w &= \frac{1}{2}\pi R^2 \rho C_p(\lambda, \beta) V^3 \\ \lambda &= \frac{\omega_m R}{V} \\ P_w &= T_w \omega_m \\ T_w &= \frac{1}{2}\pi R^3 \rho V^2 \frac{C_p(\lambda)}{\lambda} \end{aligned} \tag{5.1}$$

The pitch system isn't considered here, thus, $\beta = 0$. Then the power coeffi-

cient C_p is a function versus tip speed ratio λ from (2.11),

$$C_p = \sum_{i=0}^6 a_i \lambda^{i+1} \quad (5.2)$$

Read from the $C_p - \lambda$ characteristic curve, the maximum power coefficient is C_{pmax} which appears at the optimal Tip Speed Ratio (TSR) λ_{opt} . In normal operation, in order to capture the maximum power, mechanical angular velocity of the turbine shaft should be controlled to make the TSR always at the optimal value λ_{opt} . The torque dynamic equation on the shaft can be described as an one mass lumped system,

$$J_g \frac{d\omega_r}{dt} = -K_t \omega_r + T_w - T_g. \quad (5.3)$$

The shaft rotational speed reference ω_m^* is only proportional relevant to the wind speed V in the considered weak wind velocity region. Electrical torque T_{em} should be controlled for the variable ω_m^* tracking [117].

$$\omega_m^* = \omega_{ropt} = \frac{\lambda_{opt} V}{R}. \quad (5.4)$$

Figure 5.1 shows typical wind speed data during a short period which means that the WTPGS MPPT control is equivalent to PMSG sharply time-varying speed reference tracking problem [136].

Figure 5.1. A typical string of wind speed data during a short period

An outer-loop controller of enduing i_q^* for the q axis control is built for the turbine MPPT with control parameters k_{pw} and k_{iw} . All the PI parameters k_p , k_i , k_{pw} and k_{iw} used in this section are designed by pole placement method. It becomes to design speed controller for PMSG system which can refer to Section 4.4.

5.2.1 ORC MPPT Control

In order to realize MPPT, the PMSG ORC speed controller is designed according to the theory given in section 4.4.

The ORC MPPT controller is tested on a small WTPGS installed with a PMSG. All the simulations are set up in MatLab/Simulink. The performances of conventional PI and ORC are compared by the simulation and experimental results. All the parameters are given in the appendix.

5.2.2 Simulations and Remarks

a. Reducing Mechanical and Electrical Stresses : During PMSG transient process, PI always cause temporarily electrical and mechanical surges. So the generator will always be under huge stresses of stator windings start-up currents and start-up torque overshoots. Of course, if such impacts exist on generator's shaft and in generator's electric circuit frequently, it will cause the generator to be at the high critical condition or even cause the damage of the generator.

In parameters design of conventional PI, smaller damping factor will make the response faster but leads to larger overshoot as well. And larger bandwidth can achieve faster response and shorter settling time but results in worse steady-state stability. ORC is proposed to mitigate the conflict. The ORC's function is very similar to the soft start of electric motor and it can also be used in electric motor control. A testing $2.5m/s$ wind speed is suddenly given to the wind turbine. The simulation results will show the step responses comparison between the conventional PI and ORC.

In Figure 5.2, dotted curve expresses the reference speed, while, the solid curve represents the rotational speed transient response for ORC and the dashed curve

stands for the transient response for PI. The latter figure zooms in the former figure on both the axis. Mainly due to the overshoot, the convergence speed of the PI controller is slower than that of the ORC. However the overshoot has been virtually eliminated. In addition, ORC makes the system convergent faster than PI does with less static error as well.

In Figure 5.3, red curve is the wind torque response on the generator shaft. And the blue curve expresses the electrical torque response. Figure 5.3(a) indicates that the sudden change of wind speed will produce a relatively big electrical torque surge on the rotational shaft of PI controlled WTPGS. The frequent torque impacts will possibly damage generator's rotating part which will shorten lifespan of the PMSG. Figure 5.3(b) shows the ORC can protect the PMSG against

Figure 5.2. Shaft rotational speed to estimated reference tracking with zoom

the torque impacts. The peak value of the electrical torque has been reduced by about 75% compared to PI's. Especially in a big wind turbine generation system, the reduced stress is enormous which should be avoided as far as possible.

In Figure 5.4, the currents responses of phase a, phase b and phase c are presented as red curve, blue curve and green curve, respectively. Figure 5.3(a) gives that a sudden change of wind speed will make the generator winding circuits under the big stresses of current impulses. Obviously, it is a huge hidden safety danger for the generator's operation. Unfortunately, if a current limiter

Figure 5.3. Wind torque and electrical torque responses

Figure 5.4. Generator 3-phase currents responses

is employed in the current control loop against the impulse stresses, the speed converges to the tracking reference very slowly. Figure 5.3(b) shows the transient responses of ORC controlled WTPGS. The peak values of the 3-phase currents have been reduced to about 25% of the PI's. In a big WTGS, the winding circuits and electronic devices are both very sensitive to the current. As the same as the torque, the stresses of currents are enormous which should be avoided as far as possible.

The simulation results show that the ORC can restrain both the impulsive

impacts of the mechanical and electrical stresses far beyond the ability of conventional PI.

b. Tracking the Estimated Reference : Rapid and accurate reference tracking is an important prerequisite for the maximum power capturing of the wind turbine. In order to realize the MPPT, the shaft rotational speed should always track the estimated optimal rotor speed to make the TSR maintaining at its optimal λ_{opt} . Following the wind speed variation, the shaft rotational speed should also be in sharp variation. So another purpose for the application of ORC on WTPGS is to enhance the dynamic tracking performance.

In Figure 5.5, a sharp variable wind speed is employed in this simulation. Red curve expresses the reference shaft rotational speed estimated from (5.4), meanwhile, the blue curve presents the real shaft rotational speed. The dynamic response of PI controlled WTPGS is described in Figure 5.3(a). The dynamic response cannot avoid chattering. Figure 5.3(b) shows that the deviation of dynamic performance of ORC installed WTPGS can be reduced to only 16% of that of the PI controlled WTPGS. Obviously, the ORC controlled WTPGS can achieve a perfect dynamic tracking performance.

c. Anti – windup Characteristic : When considering the saturation characteristics of PMSG, the performance of the PI based on linear theory is relatively poor. And the windup phenomenon is always hard to be avoided in conventional PI due to the non-zero initial conditions caused by the integration. Some sudden wind speed increasing may cause the controller's integration windup inevitably. If the control output exceed its saturation, windup phenomenon will result in the system a large state overshoot, a long settling time and even cause system instability. A testing $3.03m/s$ wind speed is employed in the simulations.

In Figure 5.6, (a) indicates that there is a big deviation between the estimated reference and real rotational speed. But Figure 5.3(b) expresses that the dynamic deviation is eliminated by the ORC method.

In Figure 5.7, the simulations explain the reason why the deviation exists in Figure 5.6. (a) shows the q axis voltage is staying at the saturated value during a long time. The saturation of the controller makes the wind turbine MPPT out of control. However, ORC demonstrates its anti-windup capability in (b). So the

Figure 5.5. Shaft rotational speed to estimated reference tracking

dynamic performance of ORC is perfect far beyond the ability of PI.

In Figure 5.8, red curve is the wind torque response on the generator shaft. And the blue curve expresses the electrical torque response. Exploration shows that ORC can inhibit the controller windup.

5.2.3 Application on Hardware Setup

The optimal reset control design is verified via implementation in the Air-X WTPGS system setup. The dynamic performance of ORC controlled WTPGS is

Figure 5.6. Shaft rotational speed to estimated reference tracking

investigated. Figure 5.9 shows the whole experimental setup used for testing in details.

The experimental set-up was composed of an Alpha series 600 watt BLDC motor controlled by a DSP controller box for an emulated wind turbine. The output terminal of the generator are connected to a 3-phase SEMIKRON 6-IGBT gates SKM50Gb 123D converter which works as a rectifier. Then, the dc end of the rectifier is connected to the dc-link and a load resistance in parallel. The rated voltage value of the dc-link is 24V. And the load resistance is used for the

Figure 5.7. Responses of generator d, q voltages compare to saturation

generator output power consumption. LEM LA-25 current transducers connected to three ADC ports of the dSPACE 1103 controller card are used as PMSG current measurement. The controller board is a 400 MHz Power PC processor dSPACE 1103 which has 6 incremental encoders, analog-to-digital (ADC) and digital-to-analog (DAC) ports. The 32 bit encoder port measures the position and angular velocity of the PMSG for controller feedback.

In Figure 5.10, a sharply variable wind speed is employed in the MPPT speed tracking experiment. Red curve expresses the estimated rotational reference, and

Figure 5.8. Wind torque and electrical torque responses

the blue curve gives the real rotational speed. (a) shows the time-varying reference tracking by PI controlled WTPGS while (b) shows that of the ORC controlled WTPGS. 10 times of this testing have been tried and the average tracking deviation of ORC controlled WTPGS is only 16% of the PI controlled WTPGS. It can come to the conclusion that the dynamic rotational speed chattering can be eliminated by the ORC very well and the ORC leads to more smooth tracking performance with less deviation.

Figure 5.9. WTPGS experimental setup

Figure 5.10. Experimental estimated rotational reference tracking

5.3 Sensorless Control of WTPGS

The sensorless control in this section is employed to achieve sensorless MPPT and pitch controls of WTPGS. However, this theory can not only be used for maximum power tracking control, but also for specific defined power point tracking or other

objectives controls in the WTPGS. For example, the transient controls of WTPGS in chapter 7 is also based on this sensorless mapping characteristic.

In the case that wind speed cannot be accurately measured, in order to implement maximum wind power extraction, most controller designs of the variable-speed WTPGS employ anemometers to measure wind speed in order to derive the desired optimal shaft speed for sensorless maximum power control. In most cases, a number of anemometers are placed surrounding the wind turbine at some distance to provide adequate wind speed information. These mechanical sensors increase the cost (e.g., equipment and maintenance costs) and reduce the reliability of the overall WTPGS system [79][75]. The anemometer installed on the top of nacelle may be a source of inaccurate measurement of the wind speed and the use of anemometers raises a problem of calibration and measurement accuracy, as well as increasing the initial cost of the wind generation systems. For these reasons, it is desirable to replace the mechanical anemometers by the digital wind-speed estimator based on the turbine characteristics.

In this section, a two-loop linear observer is firstly proposed for estimation of the turbine mechanical characteristics, the wind mechanical torque on the shaft and the rotational speed. Then the ELM principle is introduced and applied on the design of the wind speed estimator and the pitch angle controller for WTPGS sensorless maximum power controls.

5.3.1 Two-loop Linear Observer

The system (2.12) can be written as the general state space form with $x = [\omega_m \ \omega_g \ \theta_\Delta]^T$ as the state vector,

$$\begin{aligned} \dot{x} &= Ax + D_m T_w + D_g T_g \\ \omega_m &= C_m x \\ \omega_g &= C_g x \end{aligned} \tag{5.5}$$

where

$$A = \begin{bmatrix} -\frac{B_m+\mu}{J_m} & \frac{\mu}{J_m} & -\frac{K}{J_m} \\ \frac{\mu}{J_g} & -\frac{B_g+\mu}{J_g} & \frac{K}{J_g} \\ 1 & -1 & 0 \end{bmatrix}, \quad D_m = \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix},$$

$$D_g = \begin{bmatrix} 0 \\ -1 \\ 0 \end{bmatrix}, \quad C_m = \begin{bmatrix} 1 & 0 & 0 \end{bmatrix}, \quad C_g = \begin{bmatrix} 0 & 1 & 0 \end{bmatrix}.$$

The generator electrical torque T_g is proportional to the q -axis current and the generator speed ω_g can be measured. The state estimator is designed by propagating the input signals, T_w and T_g , through (5.5). The states are furthermore updated by a scaling, L , of the error in estimated output as described as

$$\dot{\hat{x}} = A\hat{x} + B_m T_w + B_g T_g + L(\omega_g - \hat{\omega}_g). \quad (5.6)$$

where the observer gain L is designed by the Kalman Filtering (KF) approach. An ORC two pieces cascaded coupled linear observer is proposed to estimate T_w and ω_m . The basic structure is given in Figure 5.11 [137]. The inner-loop is a

Figure 5.11. Block diagram of ORC uncertainties observer structure

Kalman filter designed with respect to (5.6). The outer-loop ORC operates as a dynamic tracking configuration with the real-time ω_g as the tracking objective. Normally, conventional PI or PID controller is used for the outer-loop dynamic tracking. However, the performance of ORC loop was proved perfect far beyond

Figure 5.12. Performance comparisons of different observers

the capability of PI or PID loop. Figure 5.12 gives the comparisons of step responses of observing performance with ORC and PID tracking loops. For a given objective $\omega_g = 8.785$, the observer with ORC loop responds and converges much faster than that with conventional PID loop. Theoretically, ORC supports the outer-loop tracking with perfect performance. The asymptotic tracking of the observer can be ensured as long as that the ORC observer is well designed. The turbine mechanical torque T_w can be estimated as the ORC output. The previous chapters give the design process of ORC.

5.3.2 Wind Speed Estimation

Section 1.4 reviews wind speed estimation method in the previous literatures. It has been proved that ANN methods can overcome the drawbacks of the previous traditional methods and are effective for wind speed real-time estimation. In this section, ELM ANN based input-output mapping of a specific design characteristic of the wind turbine is proposed for the WTPGS sensorless maximum power control with wind speed on-line estimation.

5.3.2.1 Extreme Learning Machine (ELM)

The basic principle of ELM is shown as follows.

Single-hidden layer feed-forward neural networks (SLFN) are widely used to approximate complex nonlinear mappings directly from the input samples. The input weights and hidden layer biases of SLFN can be randomly assigned if the activation functions in the hidden layer are infinitely differentiable. After the input weights and the hidden layer biases are randomly chosen, SLFNs can be simply considered as a linear system and the output weights of SLFNs can be analytically determined through simple generalized inverse operation of the hidden layer output matrices. ELM [82][83][84][85], which can obtain better generalization performance, is emerged based on this concept. Different from traditional learning algorithms the ELM not only tends to reach the smallest training error but also the smallest norm of weights which makes the regression performance better.

According to [80], for N arbitrary distinct samples $(\mathbf{x}_i, \mathbf{t}_i)$, where $\mathbf{x}_i = [x_{i1}, x_{i2}, \dots, x_{in}]^T \in \mathbf{R}^n$ and $\mathbf{t}_i = [t_{i1}, t_{i2}, \dots, t_{im}]^T \in \mathbf{R}^m$, standard SLFNs with $\tilde{N}$ hidden nodes and activation function $g(x)$ are mathematically modeled as

$$\sum_{i=1}^{\tilde{N}} \beta_i g_i(\mathbf{x}_j) = \sum_{i=1}^{\tilde{N}} \beta_i g_i(\mathbf{w}_i \mathbf{x}_j + b_i) = \mathbf{o}_j, j = 1, \dots, N. \quad (5.7)$$

where $\mathbf{w}_i = [w_{i1}, w_{i2}, \dots, w_{in}]^T$ is the weight vector connecting the i th hidden node and the input nodes, $\beta_i = [\beta_{i1}, \beta_{i2}, \dots, \beta_{im}]^T$ is the weight vector connecting the i th hidden node and the output nodes, and the b_i is the threshold of the i th hidden node. $\mathbf{w}_i \mathbf{x}_j$ denotes the inner product of $\mathbf{w}_i$ and $\mathbf{x}_j$. That standard SLFNs with $\tilde{N}$ hidden nodes with activation function $g(x)$ can approximate these N samples with zero error means that $\sum_{j=1}^N \|\mathbf{o}_j - \mathbf{t}_j\| = 0$, i.e., there exist β_i , $\mathbf{w}_i$ and b_i such that

$$\sum_{i=1}^{\tilde{N}} \beta_i g_i(\mathbf{w}_i \mathbf{x}_j + b_i) = \mathbf{t}_j, j = 1, \dots, N. \quad (5.8)$$

The above N equations can be written compactly as $\mathbf{H}\beta = \mathbf{T}$, where

$$\mathbf{H}(\mathbf{w}_1, \dots, \mathbf{w}_{\tilde{N}}, b_1, \dots, b_{\tilde{N}}, \mathbf{x}_1, \dots, \mathbf{x}_N)$$

$$\begin{aligned}
&= \begin{bmatrix} g(\mathbf{w}_1 \mathbf{x}_1 + b_1) & \cdots & g(\mathbf{w}_{\tilde{N}} \mathbf{x}_1 + b_{\tilde{N}}) \\ \vdots & \cdots & \vdots \\ g(\mathbf{w}_1 \mathbf{x}_N + b_1) & \cdots & g(\mathbf{w}_{\tilde{N}} \mathbf{x}_N + b_{\tilde{N}}) \end{bmatrix}_{N \times \tilde{N}}, \\
\beta &= \begin{bmatrix} \beta_1^T \\ \vdots \\ \beta_{\tilde{N}}^T \end{bmatrix}_{\tilde{N} \times m}, \mathbf{T} = \begin{bmatrix} \mathbf{t}_1^T \\ \vdots \\ \mathbf{t}_N^T \end{bmatrix}_{N \times m}. \tag{5.9}
\end{aligned}$$

Training an ELM is simply equivalent to find a least-squares solution $\hat{\beta}$ of above the linear system, $\hat{\beta} = \mathbf{H}^\dagger \mathbf{T}$, where $\mathbf{H}^\dagger$ is the Moore-Penrose generalized inverse of matrix $\mathbf{H}$, $\mathbf{w}_i$ and b_i are randomly chosen and $g(x)$ is the sigmoid function in this section.

5.3.2.2 ELM Based Wind Speed Estimation

Commonly, no matter what the wind speed estimation methods are, the ideas are based on the power characteristic function. Given the turbine power P_w , the rotational speed ω_m and the blade pitch angle β , the wind speed can be calculated from the nonlinear inverse function of (2.10).

The ANN technique has been successfully used to solve input-output mapping problem of nonlinear inverse function. The $[P_w, \omega_m, \beta]$ three input vector is always set as the input of the ANN and the $\hat{V}$ is the output of the ANN which represents the estimated wind speed. The traditional ANNs [75] such as BP-SLFN and conventional RBF [81] are trained off-line using a training data set that covers the entire operating range of the wind turbine. The samples of the rotor speed ω_m , wind speed V and the pitch angle β are generated evenly in the operating range with the increments of $\Delta\omega$, ΔV and $\Delta\beta$, respectively. At each data sample of them $\omega(i)$, $V(i)$ and $\beta(i)$, the turbine power sample $P_w(i)$ is calculated from (2.10). However, these methods have some disadvantages as the following:

1. Although the training data set don't need to be stored in the WTPGS hardware, the large amount of data for training are hard to be obtained by nominal computer. They may make the training speed very slow even

reduce the approximation performance. For example, the training data set of the three training inputs may include the number of elements of $1000_{(V)} \times 900_{(\beta)} \times 200_{(\omega_m)}$ in a 4-byte float format. The total 4 input-and-outputs requires 2880 Megabytes memory for storing the training data set. The larger the operating range of the wind turbine or the smaller the sampling interval is, the larger amount of training data and longer the training time will be. It always makes the normal computer out of memory which means that the training process cannot easily be done by normal computers. That is inconvenient for the estimator design. In order to avoid the problem, the sampling intervals always should be increased to decrease the number of elements of the training data set which inevitably will decrease the estimation performance and accuracy of the wind speed.

2. With a same number of hidden neurons, the estimation accuracy of a three inputs ANN is worse than that of a two inputs ANN when operating as function approximation tools trained by the same algorithm. In order to achieve a similar approximation accuracy, there must be more hidden neurons for the three inputs ANN theoretically which will make the wind speed estimation slow and need more memory space of WTPGS hardware.
3. The traditional ANNs are designed on a rated constant air density without considering the variations. However, the air density decreases with increasing altitude, as does air pressure. It also changes with variances in temperature or humidity. For example, generally, the air temperature changing of $\pm 10^\circ\text{C}$ results the air density variation of $\pm 4\%$. The diurnal temperature varies around $\pm 15^\circ\text{C}$ or even more (seasonal/annual more) in some places with WTPGS installed. It will make the real-time data P_w varying so as the output estimated wind speed. From (2.10), its inverse function of wind speed V is not proportional to P_w which will vary with air density ρ . So a new set of training data should be regenerated and the ANN should be retrained on-line with respected to the new air density manually. That's unrealistic.
4. From Figure 5.13, the inverse function of wind speed V from (2.10) is not

always a function in the operating range. It means that given a set of P_w , ω and β , the corresponding wind speed V sometimes is not unique (V_1 and V_2). In other words, not all the data in the operating range are suitable for training the ANN. However, no wind speed selection or exclusion law is given in the conventional ANN methods for picking the training data.

In addition, no wind speed estimation is tested under the condition when pitch control system is activating in the previous works.

Figure 5.13. Characteristic of turbine captured power

In this section, a wind speed estimation method based on ELM is proposed. The basic structure is shown in Figure 5.14. ELM has been proven as a high performance function approximation tool due to its perfect regression capability beyond other kinds of ANNs.

To overcome the disadvantages of previous ANN methods, $\frac{C_p(\lambda, \beta)}{\lambda^3}$ which is a function with respect to only two states λ and β is used as the objective function for ANN training instead of P_w . $\frac{C_p(\lambda, \beta)}{\lambda^3}$ can be derived from the aerodynamic modeling of wind turbine in section 2.3.1. Since $\frac{C_p(\lambda, \beta)}{\lambda^3} = \frac{2\hat{T}_w}{\rho\pi R^5 \hat{\omega}_m^2}$, $\frac{C_p(\lambda, \beta)}{\lambda^3}$ can be online calculated by the real-time estimated $\hat{T}_w$ and $\hat{\omega}_m$. Due to that ρ is only used for calculation of $\frac{C_p(\lambda, \beta)}{\lambda^3}$, it is independent to the proposed ELM wind speed estimator.

Figure 5.14. ELM based wind speed estimation

The characteristic of $\frac{C_p(\lambda, \beta)}{\lambda^3}$ versus λ of different β is given in Figure 5.15. For a determined β , there exists an unique λ denoted as λ_m which corresponds

Figure 5.15. Characteristic curves of $\frac{C_p(\lambda, \beta)}{\lambda^3}$ Versus λ of different β

the maximum value of $\frac{C_p(\lambda, \beta)}{\lambda^3}$. In the condition of any determined β , the corresponding λ_m and the maximum value of $\frac{C_p(\lambda, \beta)}{\lambda^3}$ will be given. The λ_m separates the whole function of $\frac{C_p(\lambda, \beta)}{\lambda^3}$ into two parts. According to this two parts, there are two solutions of inverse function of λ with respect to $\frac{C_p(\lambda, \beta)}{\lambda^3}$ and β which denote as a low value λ_1 and a high value λ_2 , respectively. Since that the system

sampling period is very short (i.e. 10^{-4} sec) compared to the wind speed variation, the λ of each two adjacent sampling time instants can be regarded as equivalent. Denote the $\hat{V}_p$ as estimated wind speed at the previous time instant, the TSR of the previous time instant λ_p can be calculated as $\lambda_p = \frac{\omega_m R}{\hat{V}_p}$. So the real-time TSR can be estimated as

$$\lambda = \begin{cases} \lambda_2, & \lambda_p \geq \lambda_m \\ \lambda_1, & \lambda_p < \lambda_m. \end{cases} \quad (5.10)$$

Then the online wind speed can be estimated as $\hat{V} = \frac{\omega_m R}{\lambda}$. In this two inputs ELM based estimation strategy, the training data may include the number of elements of $1000_{(\lambda)} \times 900_{(\beta)}$ in a 4-byte float format for the 3 input-and-outputs. It requires only 10.8 Megabytes memory for the training data set. That can easily be taken by normal computers and the training precess can be very fast and precise.

5.3.3 Sensorless MPPT and Pitch Controls of WTPGS

For the commonly used variable-speed variable-pitch (VSVP) WTPGS, the pitch control system for β is activated when the wind speed exceeds the rated value. In other words, the pitch angle should be controlled to make the power captured by the turbine maintaining at its rated value when the wind speed exceeds its rated value. According to (2.10), the pitch control can be regarded as a controller for limiting the power coefficient.

Several pitch angle control methods have been reported so far, such as the classical PID controller. However, most of the designs are based on a linearized turbine model at a specific operating point [138]. It is obviously that the controllers may not provide identical performances when the turbine parameters or operating point deviates greatly. Some other pitch control strategies employ in nonlinear control schemes. However, the design principles are always complicated which make the redesigns and extension applications hard for different kinds of WTPGS.

Assume that the rated turbine extraction power is P_{wr} , the corresponding

objective power coefficient $C_p(\lambda, \beta)^*$ is given as $C_p(\lambda, \beta)^* = \frac{2P_{wr}}{\rho\pi R^2 \hat{V}^3}$, while $\lambda = \frac{\omega_m R}{\hat{V}}$ is obtained from Section 5.3.2.2. The pitch angle reference is given as $\beta^* = f'(C_p^*(\lambda, \beta), \lambda)$ from its inverse β function of $C_p(\lambda, \beta)$. Using the same ELM mapping technique as shown in Section 5.3.2.2, a two-inputs ELM pitch controller can be established.

The structure is given in Figure 5.16 and Figure 5.17. Similarly, a simple se-

Figure 5.16. ELM based pitch control

Figure 5.17. Block diagrams of pitch control systems

lection law should be set to provide the pitch control command which corresponds

to the turbine captured power at the previous sampling time instant P_{wp} .

$$\beta^* = \begin{cases} f'(C_p^*(\lambda, \beta), \lambda), & P_{wp} > P_{wr} \\ 0, & P_{wp} \leq P_{wr}. \end{cases} \quad (5.11)$$

For the sensorless wind turbine, the proposed ELM pitch angle controller operates with considering the best cooperation of the turbine modeling, the real-time estimated rotational speed and the real-time estimated wind speed. The main advantages of the proposed controller are that it is more robust for different operating points with fast response capability than traditional linear controller. Moreover, it is much easier to be designed and applied in different kinds of WTPGS than some of the nonlinear controllers.

The proposed wind speed estimation and pitch control methods are obviously suitable for all classes of WTPGS. For simplification, a direct-drive variable-pitch PMSG WTPGS is employed to test the proposed methods. Due to that the observer given in Section 5.3.1 is also effective for one-order direct-drive wind turbine, it is reasonable to verify the effectiveness of the proposed methods on a direct-drive variable-pitch PMSG WTPGS as a representative of all kinds of WTPGS. The overall PMSG WTPGS configuration and sensorless control structures are shown in Figure 5.18.

Figure 5.18. Entire sensorless control scheme for variable-pitch PMSG WTPGS

Define the TSR optimal value as λ_{opt} and the rotational speed rated value as ω_{mr} . The rotational speed should be real-time adjusted to the optimal value for MPPT. So the turbine rotational speed command is given as

$$\omega_m^* = \begin{cases} \frac{\lambda_{opt}\hat{V}}{R}, & \omega_m^* > \omega_{mr} \\ \omega_{mr}, & \omega_m^* \leq \omega_{mr}. \end{cases} \quad (5.12)$$

When the turbine extracted power goes beyond the rated power value, the pitch angle control is activated to decrease the extracted power back to the rated value. The speed PI controller and PMSG current controllers are all designed by pole-placement method.

5.3.4 Simulation Results

The proposed wind speed estimation and pitch control methods is tested on a small distributed direct-drive variable-pitch PMSG WTPGS. All the simulations are set up in MatLab/Simulink. Some performance comparisons of conventional methods and the proposed methods will be given. Some experimental testings on an emulated WTPGS are also implemented to verify the proposed methods. The cut-in wind speed and cut-out wind speed for both the simulation and experimental systems are given as $3m/s$ and $16m/s$, respectively. All other WTPGS parameters are shown in the appendix.

The torque dynamic model for the simulated small distributed direct-drive variable-pitch PMSG WTPGS uses the one-mass lumped modeling from (2.13). A testing pitch control system as shown in Section 5.3.3 is added to this simulated WTPGS for testing the performance of the proposed ELM pitch controller. The simulation performances of the proposed method for both low wind speed and high wind speed conditions are given. This section also gives the comparisons and analysis of the performances between the proposed method and conventional RBF for wind speed estimation as well as the proposed method and conventional PID pitch controller for pitch control. The rated rotational speed is set to be $\omega_{mr} = 33.75rad/s$ and the turbine rated extracted power is $P_{wr} = 4000w$.

In order to protect the wind turbine, a rotor speed upper limit is manually

set to prevent the wind turbine from excessively fast rotation. It is defined as a speed slightly higher than the rated rotor speed ω_{mr} . And it is realized by inserting a limiter for the rotor speed reference in the control loops. It should be declared that the speed upper limit does not have any impact on our simulations, because there isn't any information about the speed upper limit contained in our proposed sensorless control.

Under normal air density condition, the wind speed estimation performances and errors of ELM method and conventional RBF ANN method are given in Figure 5.19, respectively. Although both the two methods can achieve good estimation performances, the proposed ELM based method is relatively more superior with less estimation error.

Define an abnormal air density condition which is that the real air density increases by 6% to the design nominal value in this test. Under the abnormal condition, the wind speed estimation performances and errors of ELM method and conventional RBF ANN method are given in Figure 5.20, respectively. It is obvious that the wind speed estimation accuracy of the conventional RBF ANN will be influenced by the air density variations. Once the air density changes by the variation of air temperature, the wind speed estimation error will increase evidently. Certainly, under this condition, the rotor speed command cannot be given properly which will make the MPPT control failed. However, the proposed ELM based method won't be affected by the factor of air density.

In Figure 5.21, the comparison of real rotational speed between the ELM method and the conventional RBF ANN method under the abnormal air density condition is shown. The comparison of TSR between the two methods is given as well. The sensorless MPPT rotational speed control is perfect by the ELM method. Using the conventional RBF ANN method instead, the rotational speed is always under relatively big deviation to the optimal rotor speed reference. So does the tip speed ratio. The TSR cannot be controlled to the optimal value (around 6.9) by the conventional RBF ANN based sensorless control which will make the turbine maximum power extraction control failed.

Figure 5.22 shows the wind speed estimation performance by the ELM method when the wind speed is mostly above the rated value. Due to the dynamics of the

activated pitch angle system, the estimation error increases slightly, but the overall performance is still acceptable. This result also proves that the more dynamic inputs the ANN estimator has, the larger estimation error will be caused. In other words, the proposed mapping characteristic used in ELM method for decreasing the number of estimator's inputs is reasonable and effective for improving the wind speed estimation accuracy.

In Figure 5.23, the pitch control performances of ELM pitch controller and conventional PI controller are compared. The wind speed is in the same high wind speed condition of that shown in Figure 5.22. The first sub-figure gives the comparison of pitch angle responses and the turbine extracted power with the two pitch control methods are shown in the second sub-figure. It shows that the response of the ELM pitch controller is much faster than the conventional PI controller. The proposed pitch control can eliminate the big power extraction oscillations when the wind speed changes fast.

However, the extracted power of ELM pitch controlled system sometimes has small fluctuations around the rated power which means that the ELM pitch controller results in more static error than the conventional PI controller. It may cause by the inevitable small wind speed estimation error. A correction law including the information of the real-time extracted power may be used for improving the pitch control performance and eliminating the static error. That is an interesting topic to be explored in the future work.

5.3.5 Experimental Tests

The proposed sensorless control is verified via implementation in an emulated WTPGS setup which is shown in Figure 5.9. The ELM wind speed estimation and ELM pitch control performances are investigated. The experimental setup was composed of a 600 watts BLDC motor controlled by a DSP controller box which emulates the variable-pitch wind turbine, a 400 watts PMSG and a dSPACE controller. The entire experimental system operates in low power and low wind speed conditions. The wind turbine is emulated by the motor. 60 hidden neurons are chosen to build the ELM wind speed estimator which can refer to

the method proposed in [139]. Actually, the number of hidden neurons can be determined by manual adjustment to make the wind speed estimation error in the given tolerance.

In Figure 5.24, the experiment runs when the wind speed is under its rated value $3m/s$. Then the pitch control system is deactivated. (a) gives the real wind speed in red and the estimated wind speed in blue. In (b), the red curve expresses the optimal rotational speed reference and the blue one expresses the real rotor speed. In (c), the red curve and blue curve describe the turbine extracted power P_w and the stator generated power P_g , respectively. The wind speed estimation error (blue) and real-time pitch angle (red) are shown in (d). According to the Figure 5.24, the proposed ELM wind speed estimation and sensorless control are proven to be effective. In most of the time, the wind speed estimation error is less than $0.05m/s$.

In Figure 5.25, the experiment runs for when the wind speed is above its rated value $3m/s$. So the pitch control system is activated. The results are shown in the same sequence as in Figure 5.24.

It is easy to get the conclusion that the proposed ELM wind speed estimation and sensorless control are also effective when the pitch control system is activated. And the ELM based pitch control can achieve its power control objective. Due to the fast response of ELM pitch controller, the extracted power can be well maintained at its rated value (around $23w$).

5.4 Concluding Remarks

In this chapter, take PMSG WTPSG as a representative, the WTPGS control problems are studied in the both conditions that wind speed is available and wind speed cannot be accurately measured, respectively. In order to reduce the overshoot of the step response without degrading other specifications, suppress controller's saturation effectively and accelerate the tracking speed with less error, the discrete-time ORC is proposed for the MPPT control design of WTPGS. The simulation results show that the discrete-time ORC can improve the dynamic

responses to excellent performances beyond the ability of conventional PI. The experiments validate the effectiveness of the proposed approach.

In the case that the wind speed cannot be accurately measured, an ELM based wind speed estimation and sensorless maximum power control scheme is proposed for WTPGS. Due to that the estimation accuracy of the real-time wind speed is the key point for sensorless controlling of the WTPGS, a highly precise wind speed estimator based on ELM is designed to achieve the real-time wind speed precise estimation. The real-time turbine information are estimated by a fast two-loop ORC observer from the measurable generator information. Overall, The proposed method is much faster, more accurate and less time-consuming over the conventional lookup table, root calculation and SVR methods. Compared to traditional ANN methods, the proposed method is more precise and suitable for all classes of WTPGS which holds the advantages such as high accuracy training algorithm and special designed mapping characteristic.

In addition, ELM is used for design the pitch controller. The ELM pitch controller is designed based on the mapping characteristic between the power coefficient and the turbine real-time information. This pitch control scheme not only avoids bad performance of conventional linear controllers caused by the fixed linearized operating point, but also removes the complicated design process of some conventional nonlinear controllers. The effectiveness of the proposed wind speed estimation and pitch control methods are verified on a direct-drive variable-pitch PMSG WTPGS as a representative by both the simulation and experiments.

In next chapters, the attention will shift to electric power systems control and the control of grid-connected WTPGS. The power system backstepping control scheme for transient stabilization with voltage regulation is introduced in Chapter 6. And the controls for both synchronous generator excitations and decentralized controller of WTPGS are proposed for SMIB power system with PCC installed WTPGS and multi-machine power system with integration of WTPGS in Chapter 7.

(a) wind speed estimation performance of ELM

(b) wind speed estimation performance of RBF

(c) wind speed estimation error of ELM

(d) wind speed estimation error of RBF

Figure 5.19. Wind speed estimation performances and errors under normal air density condition

(a) wind speed estimation performance of ELM

(b) wind speed estimation performance of RBF

(c) wind speed estimation error of ELM

(d) wind speed estimation error of RBF

Figure 5.20. Wind speed estimation performances and errors under abnormal air density condition

Figure 5.21. Sensorless control performances by ELM and RBF methods

Figure 5.22. ELM wind speed estimation in high wind speed region

Figure 5.23. Pitch control performances by ELM and PI controllers

Figure 5.24. Experimental results for low wind speed test

Figure 5.25. Experimental results for high wind speed test

Backstepping Control of Power System

6.1 Introduction

In modern power system, improving the transient stability becomes increasingly important for reliability and economical operation. Transient stability is the ability of power system to maintain synchronism when subjected to a severe transient disturbance [1][140]. It evaluates the power system's ability to withstand the large disturbance, usually short-circuits, and survive transition to the pre-fault or a new operating condition. Transient stabilization is one of the most challenging problems in the control of power systems [141]. Since the problems arise of how to prevent generators from losing synchronism and dampening the subsequent oscillation quickly as the system is subjected to the severe disturbances, by improving the generator excitation control the transient stability can be greatly enhanced. As a result, the performance of control system plays an important role in the power system transient stability.

However, power system controls are also with other objective including regulation of voltage. The power system controllers should coordinate the two different, sometimes inherently conflicting, control objectives with varying operating regions. The conventional approaches were always focused on design coordination

laws to combine each controller designed individually for each control objective such as switching laws of coordination controllers and soft-switching membership function of global control [142]. The problems of the previous controllers are: 1) theoretically, the global stability cannot be guaranteed by each of the two controllers unless the appropriate switching membership function can be found. It is because that the two individual robust controllers, transient stabilization controller and voltage controller, are designed respectively for different subsystems of the power system; 2) when other power regulation devices are simultaneously installed in the system for the same purposes of transient stability enhancement and voltage regulation, the previous controllers cannot provide a coordination control method for different power components.

In this section, a backstepping control scheme is proposed to deal with the mentioned two problems. Nowadays, the backstepping technique is widely used in the control of PMS machine system [143]. The backstepping technique is employed to integrate the transient stabilization control and voltage regulation of synchronous generator in one control structure. From the power point of view, the power angle and frequency controls are decoupled from the power and terminal voltage controls. Thus, the single voltage control structure with a power objective index can be used for both the transient stabilizing control and voltage regulation. Moreover, the proposed power objective index can offer a power reference for the power regulation device. And the device may work with its decentralized controller to track the power reference which provides good coordination and controls between the original power system and the power regulation device. To the best of my knowledge, only a few papers have been published on trying to apply the backstepping control principle on power system controls [144][145][146].

The basic idea of backstepping control is to treat the system partitively and design the controller separately based on the cascade characteristics of the non-linear system. In the design process, some of the system states can be considered

as virtual controls for other states. Consider the following single-input system,

$$\begin{cases} \dot{x}_1 = x_2 + f_1(x_1) \\ \dot{x}_2 = x_3 + f_2(x_1, x_2) \\ \vdots \\ \dot{x}_n = f_n(x_1, x_2, \dots, x_n) + u \end{cases} \quad (6.1)$$

where $x \in R^n$ and $u \in R$ are the states and input variable of the system, respectively. The conventional backstepping control means considering the x_{i+1} as the virtual control input in each subsystem $\dot{x}_i = x_{i+1} + f_i(x_1, \dots, x_i)$ which can make the previous states asymptotically stable through the proper virtual feedback $x_{i+1} = \alpha_i (i = 1, 2, \dots, n-1)$. However, the solution of the system $x_{i+1} = \alpha_i$ cannot be ensured under normal conditions. Thus, it is necessary to define some error variables to achieve an asymptotical feature between x_{i+1} and the virtual feedback α_i which is shown as

$$\begin{cases} z_1 = x_1 \\ z_2 = x_2 - \alpha_1(x_1) \\ \vdots \\ z_n = x_n - \alpha_{n-1}(x_1, \dots, x_{n-1}) \end{cases} \quad (6.2)$$

where $\alpha_i (i = 1, 2, \dots, n-1)$ are pending to be determined. And define a Lyapunov function V_i for each step to ensure proper asymptotical stability for each state. Then, the asymptotical stability of the entire system can be realized by calming the error z between the state x_{i+1} and the virtual feedback α_i .

The backstepping control is employed in the controller designs for the conventional SMIB power system, the SMIB power system with PCC installed SMES unit and multi-machine power system, respectively, for both transient stabilization and voltage regulation. All the system parameters are given in Appendix.

6.2 Backstepping Control of SMIB Power System

For the SMIB power system, the classical three-order model given in Section 2.4 with DFL technique applied from Section 2.6.1 is employed in this section for backstepping control scheme design. The overall structure for the SMIB power system is shown in Fig.6.1.

Figure 6.1. Classical SMIB power systems

Employing the DFL compensating law and substituting (2.39) in (2.43),

$$\dot{V}_t(t) = f_1(t)\omega(t) - \frac{f_2(t)}{T'_{d0}}\Delta P_e(t) + \frac{f_2(t)}{T'_{d0}}v_f(t). \quad (6.3)$$

Selecting $[\Delta V_t(t), \Delta\omega(t), \Delta P_e(t)]^T$ as the system state vector, a new linearized system can be developed. Since $f_1(t)$ and $f_2(t)$ are dependent on the operating conditions, their bounds can be found within a certain operating region. The conventional robust linear voltage controller can be obtained as

$$v_f(t) = -k_V\Delta V_t - k_\omega\omega - k_P\Delta P_e. \quad (6.4)$$

where k_V , k_ω and k_P are the gains of the voltage controller obtained by solving the Algebraic Riccati Equation.

Since the terminal voltage is introduced as a feedback variable in the controller's design, the post-fault terminal voltage can be recovered to its pre-fault value effectively. If the power system structure has been changed during the fault,

the generator power angle regulation won't be realized definitely. However, when the power system is subjected to a severe disturbance which cause the system operates in a wider range outside the tolerance of the voltage controller, the voltage controller may not ensure the transient stability of the system. Thus, the effect of the power angle should be significantly considered in the design of the feedback controller to keep the synchronism of the system.

6.2.1 Backstepping Control of SMIB Power System for Transient Stability Enhancement and Voltage Regulation

For the power systems, the proposed backstepping control is applied on the DFL-compensated model. The mechanical part and the electrical part of the model are treated separately. The state $\Delta P_e(t)$ is designed as a virtual control input for the mechanical dynamics of (2.39). Denote the $\Delta\delta(t)$ and $\Delta\omega(t)$ as the deviations of the power angle and the frequency, respectively.

Combine the system (2.39) and (6.3) yields the entire model of the power system,

$$\dot{\mathbf{X}} = \mathbf{A}\mathbf{X} + \mathbf{B}u. \quad (6.5)$$

where

$$\begin{aligned} \mathbf{X} &= [\delta(t), \Delta\omega(t), \Delta P_e(t), \Delta V_i(t)]^T, \quad u = v_f(t) \\ \mathbf{A} &= \begin{bmatrix} 0 & 1 & 0 & 0 \\ 0 & -\frac{D}{2H} & -\frac{\omega_0}{2H} & 0 \\ 0 & 0 & -\frac{1}{T'} & 0 \\ 0 & f_1(t) & \frac{1}{T'_{d0}}f_2(t) & 0 \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 & 0 & \frac{1}{T'} & \frac{1}{T'_{d0}}f_2(t) \end{bmatrix}^T. \end{aligned}$$

Consider the first two order of this system as one subsystem,

$$\dot{\delta}(t) = \omega(t)$$

$$\dot{\omega}(t) = -\frac{D}{2H}\omega(t) - \frac{\omega_0}{2H}\Delta P_e(t). \quad (6.6)$$

Then, define an error variable of this subsystem,

$$r = \frac{-2H}{\omega_0}e_2 + \lambda e_1 \quad (6.7)$$

where λ is a positive parameter, $e_1 = \Delta\delta(t)$ and $e_2 = \omega(t)$.

Differentiate (6.7) with substituting (2.39) into, it gives

$$\begin{aligned} \dot{r} &= \frac{-2H}{\omega_0}\dot{e}_2 + \lambda\dot{e}_1 \\ &= \frac{-2H}{\omega_0}\dot{\omega}(t) + \lambda\dot{\delta}(t) \\ &= \lambda\omega(t) + \frac{-2H}{\omega_0}\left(-\frac{D}{2H}\omega(t) - \frac{\omega_0}{2H}\Delta P_e(t)\right) \\ &= -r + \frac{-2H}{\omega_0}\omega(t) + \lambda\Delta\delta(t) + \lambda\omega(t) + \frac{D}{\omega_0}\omega(t) + \Delta P_e(t). \end{aligned} \quad (6.8)$$

Define an error variable of $\Delta P_e(t)$ which is described as

$$z_P = \Delta P_e(t) - \alpha_P. \quad (6.9)$$

If $V_m = \frac{1}{2}r^2$ is chosen as the Lyapunov function and $\alpha_P = -(\frac{-2H}{\omega_0}\omega(t) + \lambda\Delta\delta(t) + \lambda\omega(t) + \frac{D}{\omega_0}\omega(t))$, thus there exists

$$\begin{aligned} \dot{r} &= -r + \Delta P_e(t) - \alpha_P \\ &= -r + z_P \\ \dot{V}_m &= r(-r + z_P) = -r^2 + rz_P. \end{aligned} \quad (6.10)$$

Obviously, when the error variable z_p possesses an expected asymptotical behavior($z_p \rightarrow 0$) and ($\alpha_P = -(\frac{-2H}{\omega_0}\omega(t) + \lambda\Delta\delta(t) + \lambda\omega(t) + \frac{D}{\omega_0}\omega(t))$), according to (6.10), r is asymptotically stable ($\dot{V}_m < 0$). So the problem becomes that design proper control of the system to make the error variable z_P possessing an expected asymptotical behavior.

To achieve acceptable design objective of z_P , the subsystem including the

terminal voltage dynamic is considered for the controller design

$$\dot{\mathbf{X}} = \mathbf{A}\mathbf{X} + \mathbf{B}u. \quad (6.11)$$

where

$$\begin{aligned} \mathbf{X} &= [\Delta\omega(t), z_P, \Delta V_t(t)]^T, \quad u = v_f(t) \\ \mathbf{A} &= \begin{bmatrix} -\frac{D}{2H} & -\frac{\omega_0}{2H} & 0 \\ 0 & -\frac{1}{T'} & 0 \\ f_1(t) & \frac{1}{T'_{d0}}f_2(t) & 0 \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 & \frac{1}{T'} & \frac{1}{T'_{d0}}f_2(t) \end{bmatrix}^T. \end{aligned}$$

It goes to a conventional voltage controller design problem. Based on (6.4), the proposed backstepping controller of the global power system can be described as

$$\begin{aligned} v_f(t) &= -k_V \Delta V_t - k_\omega \omega - k_P z_P \\ &= -k_V \Delta V_t - k_\omega \omega - k_P (\Delta P_e + \frac{-2H}{\omega_0} \omega(t) + \lambda \Delta \delta(t) + \lambda \omega(t) + \frac{D}{\omega_0} \omega(t)) \\ &= -k_V \Delta V_t - (k_\omega + k_P \frac{2H}{\omega_0} - k_P \lambda - k_P \frac{D}{\omega_0}) \omega - k_P \Delta P_e - k_P \lambda \delta. \end{aligned} \quad (6.12)$$

Since the system stability enhancement and voltage regulation are considered simultaneously in the controller design, the backstepping control scheme should ensure the successful achievement of the both goals.

According to [95], the subsystem (6.11) model can be described with the input-output relations with uncertainties as

$$\dot{x}(t) = (A + \Delta A(t))x(t) + (B + \Delta B(t))u(t). \quad (6.13)$$

where $x(t) \in \mathfrak{R}^n$ is state variables; $u(t) \in \mathfrak{R}^n$ is the input variables of the subsystem. And $\Delta A(t)$ and $\Delta B(t)$ represent uncertainties and can be decomposed as the following assumption.

Assumption 5. *The parametric uncertainties in $\Delta A(t)$ and $\Delta B(t)$ can be de-*

composed as follows

$$[\Delta A(t) \quad \Delta B(t)] = LF(t)[E_1 \quad E_2] \quad (6.14)$$

where L , E_1 and E_2 are known constant matrices which characterize the structure of the uncertainty. The uncertainty $F(t)$ satisfies $F^T(t)F(t) \leq I$; I is an unitary matrix.

The control objective of the backstepping control design is to determine the robust feedback voltage controller for the model (6.13) such that the closed-loop subsystem is stable and the stability of the subsystem with uncertainties is guaranteed. The stability of the backstepping controller for the uncertain subsystem is discussed.

The following robust voltage controller can be designed to guarantee the stability of close loop subsystem

$$u(t) = Kx(t). \quad (6.15)$$

The global closed loop uncertain subsystem of (6.13) is then described by

$$\dot{x}(t) = \{(A + \Delta A(t)) + (B + \Delta B(t))K\}x(t). \quad (6.16)$$

The stability of this global closed loop subsystem is then studied by using quadratic Lyapunov stability theory.

Theorem 1. *The subsystem model described in equation (6.13) is quadratically stable for all admissible uncertainties, satisfying Assumption 5, via the proposed feedback control law if there exists positive definite symmetric matrices P for the following algebraic Riccati equation*

$$\begin{aligned} (A - BR^{-1}E_2^T E_1)^T P + P(A - BR^{-1}E_2^T E_1) + P(LL^T - BR^{-1}B^T)P \\ + E_1^T(I - E_2 R^{-1} E_2^T)E_1 + Q = 0 \end{aligned} \quad (6.17)$$

where Q is a symmetric positive defined matrix, which can be chosen by the designer, $R = E_2^T E_2 > 0$ [147][148][149]. In this case, the stabilizing state-feedback

control law is given by

$$u(t) = -R^{-1}(B^T P + E_2^T E_1)x(t). \quad (6.18)$$

Lyapunov Stability Proofs: Define an energy function for the subsystem (6.11):

$$V_v = x^T P x. \quad (6.19)$$

where P is a positive-definite symmetric matrix, which make the closed loop subsystem asymptotically stable. It is obvious that $V_v > 0, \forall x \neq 0$. The derivative of the energy function of the uncertain subsystem (6.11) can be represented as the following form:

$$\begin{aligned} \dot{V}_v &= \dot{x}^T P x + x^T P \dot{x} \\ &= x^T [(A + \Delta A)^T P + P(A + \Delta A)]x - x^T K^T (B + \Delta B)^T P x - x^T P (B + \Delta B) K x \\ &= x^T [(A + \Delta A)^T P + P(A + \Delta A)]x - 2x^T P (B + \Delta B)^T K x. \end{aligned} \quad (6.20)$$

Substituting the feedback control law (6.18) and the Assumption 5 yields

$$\begin{aligned} \dot{V}_v &= x^T [(A + LF(t)E_1)^T P + P(A + LF(t)E_1)]x \\ &\quad - 2x^T P (B + LF(t)E_2)^T R^{-1} (B^T P + E_2^T E_1)x \\ &= x^T (AP + PA - 2PBR^{-1}B^T P - 2PBR^{-1}E_2^T E_1)x \\ &\quad + x^T [(LF(t)E_1)^T P + PLF(t)E_1 - 2PLF(t)E_2 R^{-1}(B^T P + E_2^T E_1)]x \\ &= x^T (AP + PA - 2PBR^{-1}B^T P - PBR^{-1}E_2^T E_1 - E_1^T E_2 R^{-1}B^T P)x \\ &\quad + 2x^T PLF(t)[E_1 - E_2 R^{-1}(B^T P + E_2^T E_1)]x \\ &\leq x^T [(A - BR^{-1}E_2^T E_1)^T P + P(A - BR^{-1}E_2^T E_1) - 2PBR^{-1}B^T P]x \\ &\quad + x^T PLF(t)F^T(t)L^T P x \\ &\quad + x^T [E_1 - E_2 R^{-1}(B^T P + E_2^T E_1)]^T [E_1 - E_2 R^{-1}(B^T P + E_2^T E_1)]x \\ &\leq x^T [(A - BR^{-1}E_2^T E_1)^T P + P(A - BR^{-1}E_2^T E_1) - PBR^{-1}B^T P \\ &\quad + PLL^T P + E_1^T E_1 - E_1^T E_2 R^T E_2^T E_1]x \\ &= -x^T Q x \end{aligned}$$

$$\leq 0 \quad (6.21)$$

where $Q = -[(A - BR^{-1}E_2^T E_1)^T P + P(A - BR^{-1}E_2^T E_1) - PBR^{-1}B^T P + PLL^T P + E_1^T E_1 - E_1^T E_2 R^T E_2^T E_1]$ which is consistent with that given in Theorem 1. So the proposed robust voltage controller (6.12) for the subsystem (6.11) can make the error variable z_P possessing an expected asymptotical behavior ($z_P \rightarrow 0$).

The global Lyapunov function candidate for the proposed backstepping controlled global power system (6.5) can be given by

$$V = V_m + V_v = \frac{1}{2}r^2 + V_v. \quad (6.22)$$

Thus

$$\dot{V} = \dot{V}_m + \dot{V}_v = -r^2 + rz_P + \dot{V}_v. \quad (6.23)$$

Since $z_P \rightarrow 0$ is ensured by (6.12), it follows immediately that $\dot{V} < 0, \forall x \neq 0$. The global the power system (6.5) is globally uniformly asymptotically stable for all admissible uncertainties in the sense of Lyapunov.

Obviously, the power angle δ is introduced in the feedback control for the improvement of the transient stability. In practice, both the transient stability enhancement and good post-fault performance of the system should be achieved. In other words, good post-fault performance means the excitation should be controlled to regulate the generator terminal voltage V_t after the transient period. When the system is operating around its nominal operating point, the operation of the synchronous generator is near the power system synchronous speed. In other words, the risk that the synchronous generator loses the synchronization is relatively small in this case. Thus, the weight of power angle in feedback control can be reduced accordingly. Thus a soft-switch function is employed in the proposed backstepping control to dominate the weight of the power angle (λ) in the defined error variable (6.7).

Define a positive constant λ_0 ,

$$\lambda = \mu_\lambda \lambda_0, \quad 0 \leq \lambda \leq \lambda_0 \quad (6.24)$$

where

$$\begin{aligned}\mu_\lambda &= 1 - \left(\frac{1}{1 + \exp(-120(z_b + 0.4))} - \frac{1}{1 + \exp(-120(z_b - 0.4))} \right) \\ z_b &= \sqrt{\omega^2 + (\Delta P_e)^2}.\end{aligned}$$

In this section, the proposed control scheme for power systems is tested in MatLab/Simulink. The performance comparisons between the global control and the proposed backstepping control are given with some analysis. Then a PCC bus installed SMES unit is introduced in the simulation. All the parameters are given in the appendix.

The fault in this section is a symmetrical three phase short circuit fault with its sequences described as:

Case 1. Temporary Fault:

Stage 1: The system is in a pre-fault steady state;

Stage 2: A fault occurs at $t = t_0$;

Stage 3: The fault is removed by opening the breakers of the transmission line at $t = t_1$;

Stage 4: The transmission line are restored at $t = t_3$.

Case 2. Permanent Fault:

Stage 1: The system is in a pre-fault steady state;

Stage 2: A fault occurs at $t = t_0$;

Stage 3: The fault is removed by opening the breakers of the transmission line at $t = t_1$;

Stage 4: The system is in a post-fault state.

Case 3. Permanent Fault + Increase of the Mechanical Input Power:

Stage 1: The system is in a pre-fault steady state;

Stage 2: A fault occurs at $t = t_0$;

Stage 3: The fault is removed by opening the breakers of the transmission line at $t = t_1$;

Stage 4: The mechanical input power of the generator has a 30% step increase at $t = t_2$;

Stage 5: The system is in a post-fault state.

$t_0 = 0.1s$, $t_1 = 0.25s$, $t_2 = 1s$, $t_3 = 1.4s$ are chosen in the simulations. The fault location is indexed by a positive constant λ_l which is the fraction of the line to the left of the fault. In other words, the fault location is indexed by a positive constant for convenient representation which means the proportion of the distance between PCC bus and the fault location to the distance between PCC bus and infinity bus. $\lambda_l = 0.1$ and $\lambda_l = 0.01$ are employed in the simulations for testing, respectively.

The Existing Global Controller

The global control shown for comparisons can be found in [100][150]. The global control law is the average of the individual control laws, weighted by the operating region membership function and the input v_f takes the form as,

$$v_f = \mu_\delta v_{f1} + \mu_V v_{f2} \quad (6.25)$$

where v_{f1} expresses the DFL nonlinear transient controller and v_{f2} is the voltage controller which are given respectively as,

$$\begin{aligned} v_{f1} &= 22.36\delta + 12.81\omega - 82.45\Delta P_e \\ v_{f2} &= -40.14\Delta V_t + 10.11\omega - 30.81\Delta P_e. \end{aligned}$$

And the membership functions are given as,

$$\begin{aligned} \mu_V &= \left(1 - \frac{1}{1 + \exp(-120(z - 0.08))}\right) \cdot \left(\frac{1}{1 + \exp(-120(z + 0.08))}\right) \\ \mu_\delta &= 1 - \mu_V \end{aligned}$$

where $z = \sqrt{\alpha_\omega \omega^2 + \alpha_{\Delta V_t} (\Delta V_t)^2}$. α_ω and $\alpha_{\Delta V_t}$ are constant parameters which should be well tuned manually. In this section, $\alpha_\omega = 0.05$ and $\alpha_{\Delta V_t} = 0.05$ are chosen.

6.2.2 Backstepping Control of SMIB Power System Installed with Super-conducting Magnetic Energy Storage

The concepts of first-swing stability enhancement using the SMES have been presented in [103]. During the period of severe disturbance such as three-phase fault imposed on the power system, $\Delta P_e(t)$ of the generator changes drastically. To prevent the generator from losing its synchronism too rapidly, the high speed thyristor-controlled SMES system is set into its dynamic operation to minimize $\Delta P_e(t)$ as soon as the disturbance is detected. This is accomplished by varying the input to the SMES system so as to vary its power $P_{sm}(t)$. In this subsection, the simultaneous and coordinated use of generator excitation with the proposed backstepping excitation control and an externally PCC bus connected SMES unit with power reference tracking control for transient stability enhancement are discussed to show the effectiveness. The comparison to generator excitation with global control and the SMES with conventional PI control are given as well.

The schematic diagram of Figure 6.2 shows the configuration of a thyristor controlled SMES unit. The main components of the SMES system are the DC super-conducting coil and two sets of six pulses bridge power converters. The converter units are fed from a set of $Y - \Delta / Y - Y$ connected transformers. The converter impresses $\pm V_{sm}$ volts on the super-conducting coil. The converter unit is forced commutated and α is the firing angle of the thyristor. Charge and discharge are controlled through simple change of the commutation angle α . If $\alpha < 90^\circ$, the converter acts as a converter mode (charging mode). If $\alpha > 90^\circ$, the converter then works as an inverter mode (discharging mode). Since the bridge current I_{sm} is not reversible, the bridge output power P_{sm} is uniquely a function of α , which can be positive or negative. Neglecting the transformer and

the converter losses, the DC voltage is given by

$$V_{sm} = 2V_{do} \cos(\alpha) - 2I_{sm}R_c \quad (6.26)$$

The SMES system dynamical equations are expressed as follows

$$\begin{aligned} \Delta \dot{I}_{sm}(t) &= \frac{1}{L_{sm}} \Delta V_{sm}(t) \\ \Delta \dot{V}_{sm}(t) &= -\frac{1}{T_v} \Delta V_{sm}(t) + \frac{k_v}{T_v} u_v(t) \end{aligned} \quad (6.27)$$

where

$$\Delta V_{sm}(t) = V_{sm}(t) - V_{sm0}, \quad \Delta I_{sm}(t) = I_{sm}(t) - I_{sm0}.$$

The physical limits of the current and voltage of the super-conducting inductor are

$$-0.235p.u. \leq V_{sm}(t) \leq 0.235p.u., \quad 0.31I_{sm0} \leq I_{sm}(t) \leq 1.38I_{sm0}.$$

Under normal operating conditions in the steady-state, the values of $I_{sm}(t)$ and $V_{sm}(t)$ are I_{sm0} and V_{sm0} , respectively. The output power of the SMES can

Figure 6.2. The configuration of SMES unit

Figure 6.3. The SMES unit control block diagrams

be described as,

$$P_{sm}(t) = V_{sm}(t)I_{sm}(t). \quad (6.28)$$

Thus, system (2.39) becomes,

$$\begin{aligned} \dot{\delta}(t) &= \omega(t) \\ \dot{\omega}(t) &= -\frac{D}{2H}\omega(t) - \frac{\omega_0}{2H} [\Delta P_e(t) + P_{sm}(t)] \\ \Delta \dot{P}_e(t) &= -\frac{1}{T'}\Delta P_e(t) + \frac{1}{T'}v_f(t) \end{aligned} \quad (6.29)$$

SMES, in this thesis, acts as a component (real power storage and discharge) in the power system which makes a preliminary validation of backstepping co-ordination control theory. Then the proposed method will be extended to the non-conventional power system in Chapters 7.

Since the error variable $z_P = \Delta P_e(t) - \alpha_P$ is expected to possess an asymptotical behavior, the real-time error $\Delta P_e(t) - \alpha_P$ should be eliminated as possible as it can. Thus, ideally, this real-time power error can be compensated by the SMES from the power point of view. It is reasonable to employ the z_P in the SMES output power control as the power reference. Figure 6.3 shows the SMES

unit control block diagrams of conventional system frequency based control from [102][110][151] and the proposed power reference based control, respectively.

6.2.3 Simulation Results

According to [95], the subsystem robust voltage controller for the proposed backstepping feedback control law can be determined with parameters $k_V = 40.14$, $k_\omega = -10.11$ and $k_P = 30.81$. In the backstepping feedback control law (6.12) with (6.24), $\lambda_0 = 0.6$ is chosen. The system operating point is given in the following table,

Table 6.1. Operating point of the SMIB power system

Synchronous generator	
Power Angle	$\delta_0 = 72^\circ$
Electrical Frequency	$\omega_0 = 314.159rad/s$
Electrical Power	$P_{m0} = 0.9p.u.$
Terminal Voltage	$V_{t0} = 1p.u.$

The power system responses with the different controllers subjected to different faults are shown in Figures 6.4-6.7. In this section, comparisons are mainly based on the performances between the existing global control and the proposed backstepping control scheme. All the dash-dot lines express the responses of global controlled SMIB power system, and the performances of the proposed backstepping controlled SMIB power system are indicated by solid lines.

Figure 6.4 exhibits the closed-loop power system responses for the two kinds of controllers in fault sequence *case 1* with fault location $\lambda_l = 0.1$. Although the global control and the proposed backstepping control both can achieve good transient control performance and restoration performance of post-transient voltage, it is obvious that the control performances of the power angle, the system frequency and the terminal voltage of the proposed backstepping control have been improved significantly compared to those of the global control.

Figure 6.5 shows the responses for the global controller and the proposed backstepping controller in *case 2* with fault location $\lambda_l = 0.01$. In this case, there

Figure 6.4. Power system responses for *case 1* with fault location $\lambda_l = 0.1$, “solid lines”: backstepping control and “dashed lines”: global control

is no significant improvement of control performances by the proposed backstepping controller. The recovery speeds of the system frequency and the SG output electrical power are accelerated. However, that of the terminal voltage is decelerated.

In fault sequence *case 3*, the responses for the two controllers with fault location $\lambda_l = 0.01$ and $\lambda_l = 0.1$ are shown in Figures 6.6 and 6.7, respectively. The simulation results of Figure 6.6 are consistent with those of Figure 6.4. Significant improvements of the proposed backstepping control performances of the power angle, the system frequency, the SG output electrical power and the terminal voltage are all observed compared to those of the global control. In the case of Figure 6.7, the disturbed power system loses its synchronism by the global controller. It can be identified that the soft switching membership function of the global control cannot ride through some of the existing blind spots. Theoretically, no principle can be employed into design of an optimal soft switching

membership function which not only ensures good switching performance of the two sub-controllers but also guarantees the global stability. On the contrary, the principle of the proposed backstepping control ensures both good control performances and global stability as shown in Figure 6.7.

For the global controller, it seems that the global control performance is too dependent on the performance of the soft switching membership function since an inappropriate one may cause the loss of synchronism. It is because that every sub-controller is designed individually based on its corresponding sub-system. Each of them cannot guarantee the stability of the global system independently. Thus, unsatisfactory performances will appear unless appropriate soft switching membership function can be determined. Exactly, the shortcomings of the global control can be overcome by the proposed backstepping control. In order to coordinate the controls of the original power system and the power regulation devices, in the following, a cooperation control based on the proposed backstepping law

Figure 6.5. Power system responses for *case 2* with fault location $\lambda_l = 0.01$

will be given from the power point of view. And its reliability will be verified by the power system installed with a SMES unit. The overall structure for a particular SMIB power system with an PCC installed SMES unit is shown in Figure 6.8.

In the following figures of power system responses, the dash-dot lines express the responses of global controlled SMIB power system with conventional system frequency base controlled SMES, and the performances of the proposed back-stepping controlled SMIB power system with the proposed power reference base controlled SMES are indicated by solid lines. The parameters of the SMES unit is given in the following table,

Table 6.2. Parameters of the SMES unit

SMES unit	
$I_{sm0} = 0.649p.u.$	$V_{sm0} = 0p.u.$
$L_{sm} = 0.5p.u.$	$k_v = 1.83$
$T_v = 0.026s$

Figure 6.6. Power system responses for *case 3* with fault location $\lambda_l = 0.01$

Figure 6.9 shows the system performance when subjected to fault in the sequence of *case 2*. Although both of the mentioned controls can ensure synchronization of the power system after severe fault, but the post-fault equilibrium points of the power system by the two controls are different. Due to the lack of coordination in the global controller and the conventional system frequency based SMES controller, SMES unit cannot identify the structure change of power system the post-fault equilibrium point so that the states of the SMES unit won't be restored to the pre-fault situation. As shown in Figure 6.9, the SMES output power isn't recovered back to the pre-fault situation 0. Since the energy storage capacity is limited, it is unrealistic for the SMES to supply power continually. However, the proposed backstepping controlled SMIB power system with the proposed power reference base controlled SMES successfully overcomes this disadvantage of the traditional control. The states of the SMES unit can be recovered as well.

Figure 6.7. Power system responses for *case 3* with fault location $\lambda_l = 0.1$

Figure 6.8. A single machine infinite bus power system with PCC installed SMES

Figure 6.10 shows the system performance which contrasts with that in Figure 6.9. In addition, it exhibits better transient performance and restoration of normal post-transient voltage of the proposed control than those of the conventional global control with system frequency based SMES control. In this Figure, it shows faster recovery responses of the power angle, system frequency, SG electrical power output and terminal voltage by the proposed control with respect to the fault sequence *case 3*. From the simulations, one can see that the proposed backstepping controller with the proposed power reference base SMES control achieves the proposed control task and is robust with respect to different faults.

6.3 Backstepping Control of Multi-Machine Power System

In this section, the proposed backstepping control scheme is extended to multi-machine power systems. For the multi-machine power system, the classical three-order model given in Section 2.5 with DFL technique applied from Section 2.6.2 is employed for backstepping control scheme design.

Employing the DFL compensating law and substituting (2.47) in (2.49),

$$\dot{V}_{ti}(t) = f_{1i}(t)\Delta\omega_i(t) - \frac{f_{2i}(t)}{T'_{d0i}}\Delta P_{ei}(t) + \frac{f_{2i}(t)}{T'_{d0i}}v_{fi}(t). \quad (6.30)$$

Selecting $[\Delta V_{ti}(t), \Delta\omega_i(t), \Delta P_{ei}(t)]^T$ as the system state vector of the i th generator, the DFL-compensated model can be developed. Since $f_{1i}(t)$ and $f_{2i}(t)$ are

Figure 6.9. Responses of power system installed with SMES for *case 2* with fault location $\lambda_l = 0.01$

dependent on the operating conditions, their bounds can be found within certain operating regions. The robust linear voltage controllers can be obtained as

$$v_{fi}(t) = -k_{Vi}\Delta V_{ti} - k_{\omega i}\omega_i - k_{Pi}\Delta P_{ei}. \quad (6.31)$$

where k_{Vi} , $k_{\omega i}$ and k_{Pi} are the gains of the linear voltage controllers obtained by solving the Algebraic Riccati Equation (ARE).

Figure 6.10. Responses of power system installed with SMES for *case 3* with fault location $\lambda_l = 0.01$

Since the terminal voltages are introduced as feedback variables in the controllers' design, the post-fault terminal voltages can be recovered to their pre-fault values effectively. When the power system is subjected to a severe disturbance which cause the system operates in a wider range outside the tolerance of the voltage controllers, the voltage controllers may not ensure the transient stability of the system. Thus, the power angles should be significantly introduced in the

design of the feedback controllers to keep the synchronism of the system and ensure the transient stability.

6.3.1 Backstepping Control of Multi-Machine Power System for Transient Stability Enhancement and Voltage Regulation

For the multi-machine power systems, the proposed backstepping control are applied on each SG DFL-compensated model. The mechanical part and the electrical part of each generator are treated separately. The state $\Delta P_{ei}(t)$ are designed as virtual control inputs for the mechanical dynamics of (2.47). Denote the $\Delta\delta_i(t)$ and $\Delta\omega_i(t)$ as the deviations of the power angle and the frequency of the i th synchronous generator, respectively.

Combine the system (2.47) and (6.30) yields,

$$\dot{\mathbf{X}} = \mathbf{A}\mathbf{X} + \mathbf{B}u + \mathbf{G}_i \quad (6.32)$$

where

$$\begin{aligned} \mathbf{X} &= [\delta_i(t), \Delta\omega_i(t), \Delta P_{ei}(t), \Delta V_{ti}(t)]^T, \quad u = v_{fi}(t) \\ \mathbf{A} &= \begin{bmatrix} 0 & 1 & 0 & 0 \\ 0 & -\frac{D_i}{2H_i} & -\frac{\omega_0}{2H_i} & 0 \\ 0 & 0 & -\frac{1}{T'_{d0i}} & 0 \\ 0 & f_{1i}(t) & \frac{1}{T'_{d0i}}f_{2i}(t) & 0 \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 & 0 & \frac{1}{T'_{d0i}} & \frac{1}{T'_{d0i}}f_{2i}(t) \end{bmatrix}^T \\ \mathbf{G}_i &= \begin{bmatrix} 0 \\ 0 \\ \sum_{j=1}^n E'_{qi}(t)[\dot{E}'_{qj}(t)B_{ij} + E'_{qj}(t)\dot{B}_{ij}]\sin\delta_{ij}(t) \\ 0 \end{bmatrix} \\ &\quad - \begin{bmatrix} 0 \\ 0 \\ \sum_{j=1}^n E'_{qi}(t)E'_{qj}(t)B_{ij}\cos\delta_{ij}(t)\Delta\omega_j(t) \\ 0 \end{bmatrix}. \end{aligned}$$

Consider the large-scale nonlinear system composed of N interconnected subsystems, where i th subsystem is described as:

$$\dot{x}_i(t) = f_i(x_i, u_i, t) + \Delta f_i(x_i, u_i, t) + \sum_{j=1}^n g_{ij}(x_i, x_j) \quad (6.33)$$

where for the i th ($i = 1, 2, \dots, n$) subsystem; $x_i(t) \in \mathfrak{R}^n$ and $u_i(t) \in \mathfrak{R}^n$ are the vectors of state variables and the input variables of the i th subsystem, respectively; f_i is a nonlinear function with respect to (x_i, u_i, t) representing the nominal model of the i th subsystem which is continuously differentiable with respect to (x_i, u_i, t) . Δf_i represents the model uncertainties of the i th subsystem, $g_{ij}(x_i, x_j)$ are unknown nonlinear vector functions that represent the nonlinearities of the i th subsystem and the interactions between the i th subsystem and j th subsystem.

The n interconnected subsystems in the large-scale nonlinear system can be described by the following state space form:

$$\begin{aligned} S_i : \dot{x}_i(t) = & [A_i + \Delta A_i(t)]x_i(t) + [B_i + \Delta B_i(t)]u_i(t) \\ & + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} \{[G_{kij} + \Delta G_{kij}(t)]g_{kij}(x_i, x_j)\} \end{aligned} \quad (6.34)$$

where k is the number of the nonlinearities and interconnection in the i th subsystem ($k = 1, 2, \dots, T_{jk}$); the matrices A_i , B_i and G_{kij} are known real constant matrices of appropriate dimension that describe the nominal model; $\Delta A_i(t)$, $\Delta B_i(t)$ and $\Delta G_{kij}(t)$ are real time varying parameter uncertainties; $g_{kij}(x_i, x_j)$ are unknown nonlinear vector functions that represent nonlinearities in the subsystem and the interactions with other subsystems.

Consider the first two order of the i th generator model as one subsystem,

$$\begin{aligned} \dot{\delta}_i(t) &= \Delta \omega_i(t) \\ \Delta \dot{\omega}_i(t) &= -\frac{D_i}{2H_i} \Delta \omega_i(t) - \frac{\omega_0}{2H_i} \Delta P_{ei}(t). \end{aligned} \quad (6.35)$$

Then, define an error variable of this subsystem of the i th generator,

$$r_i = \frac{-2H_i}{\omega_0} e_{2i} + \sigma_i e_{1i} \quad (6.36)$$

where σ_i are positive parameters, $e_{1i} = \Delta\delta_i(t)$ and $e_{2i} = \Delta\omega_i(t)$.

Differentiate (6.36) with substituting (2.47) into, it gives

$$\begin{aligned} \dot{r}_i &= \frac{-2H_i}{\omega_0} \dot{e}_{2i} + \sigma_i \dot{e}_{1i} \\ &= \frac{-2H_i}{\omega_0} \dot{\omega}_i(t) + \sigma_i \dot{\delta}_i(t) \\ &= \sigma_i \Delta\omega_i(t) + \frac{-2H_i}{\omega_0} \left(-\frac{D_i}{2H_i} \omega_i(t) - \frac{\omega_0}{2H_i} \Delta P_{ei}(t) \right) \\ &= -r_i + \left(\frac{-2H_i}{\omega_0} + \sigma_i + \frac{D_i}{\omega_0} \right) \Delta\omega_i(t) + \sigma_i \Delta\delta_i(t) + \Delta P_{ei}(t). \end{aligned} \quad (6.37)$$

Define error variables of $\Delta P_{ei}(t)$ which are described as

$$z_{Pi} = \Delta P_{ei}(t) - \alpha_{Pi}. \quad (6.38)$$

If $V_{mi} = \frac{1}{2} r_i^2$ are chosen as the Lyapunov function and $\alpha_{Pi} = -\left(\frac{-2H_i}{\omega_0} \Delta\omega_i(t) + \sigma_i \Delta\delta_i(t) + \sigma_i \Delta\omega_i(t) + \frac{D_i}{\omega_0} \Delta\omega_i(t)\right)$, thus there exist

$$\begin{aligned} \dot{r}_i &= -r_i + \Delta P_{ei}(t) - \alpha_{Pi} \\ &= -r_i + z_{Pi} \\ \dot{V}_{mi} &= r_i(-r_i + z_{Pi}) = -r_i^2 + r_i z_{Pi}. \end{aligned} \quad (6.39)$$

Obviously, when the error variables z_{Pi} possess expected asymptotical behaviors ($z_{Pi} \rightarrow 0$) and ($\alpha_{Pi} = -\left(\frac{-2H_i}{\omega_0} \Delta\omega_i(t) + \sigma_i \Delta\delta_i(t) + \sigma_i \Delta\omega_i(t) + \frac{D_i}{\omega_0} \Delta\omega_i(t)\right)$), according to (6.39), r_i are asymptotically stable ($\dot{V}_{mi} < 0$). So the problem becomes that design proper controls of the i th system to make the error variable z_{Pi} possessing expected asymptotical behaviors.

To achieve acceptable design objectives of z_{Pi} , the subsystem of the i th generator including the terminal voltage dynamic is considered for the controller

design

$$\dot{\mathbf{X}} = \mathbf{A}\mathbf{X} + \mathbf{B}u + \mathbf{G}_i \quad (6.40)$$

where

$$\begin{aligned} \mathbf{X} &= [\Delta\omega_i(t), \Delta P_{ei}(t), \Delta V_{ti}(t)]^T, \quad u = v_{fi}(t) \\ \mathbf{A} &= \begin{bmatrix} 1 & 0 & 0 \\ -\frac{D_i}{2H_i} & -\frac{\omega_0}{2H_i} & 0 \\ 0 & -\frac{1}{T'_{d0i}} & 0 \\ f_{1i}(t) & \frac{1}{T'_{d0i}}f_{2i}(t) & 0 \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 & \frac{1}{T'_{d0i}} & \frac{1}{T'_{d0i}}f_{2i}(t) \end{bmatrix}^T \\ \mathbf{G}_i &= \begin{bmatrix} 0 \\ \sum_{j=1}^n E'_{qi}(t)[\dot{E}'_{qj}(t)B_{ij} + E'_{qj}(t)\dot{B}_{ij}] \sin \delta_{ij}(t) \\ 0 \end{bmatrix} \\ &\quad - \begin{bmatrix} 0 \\ \sum_{j=1}^n E'_{qi}(t)E'_{qj}(t)B_{ij} \cos \delta_{ij}(t)\Delta\omega_j(t) \\ 0 \end{bmatrix}. \end{aligned}$$

It goes to conventional voltage controllers design problem. According to [96] and based on (6.31), the proposed backstepping controllers of the i th generator of the multi-machine power system can be described as

$$\begin{aligned} v_{fi}(t) &= -k_{Vi}\Delta V_{ti} - k_{\omega i}\Delta\omega_i - k_{Pi}z_{Pi} \\ &= -k_{Vi}\Delta V_{ti} - k_{\omega i}\Delta\omega_i \\ &\quad -k_{Pi}(\Delta P_{ei} + \frac{-2H_i}{\omega_0}\Delta\omega_i(t) + \sigma_i\Delta\delta_i(t) + \sigma_i\Delta\omega_i(t) + \frac{D_i}{\omega_0}\Delta\omega_i(t)) \\ &= -k_{Vi}\Delta V_{ti} - (k_{\omega i} + k_{Pi}\frac{2H_i}{\omega_0} - k_{Pi}\sigma_i - k_{Pi}\frac{D_i}{\omega_0})\Delta\omega \\ &\quad -k_{Pi}\Delta P_{ei} - k_{Pi}\sigma_i\Delta\delta_i. \end{aligned} \quad (6.41)$$

Obviously, the power angles δ_i are introduced in the feedback controls for the improvement of the transient stability. In practice, both the transient stability enhancement and good voltage regulation performance of each SG of the multi-

machine power system should be achieved. When the multi-machine system is operating around its nominal operating point, the operations of the synchronous generators are near the power system synchronous speed. In other words, the risk that the synchronous generator loses the synchronization is relatively small in this case. Thus, the weight of power angle in feedback control can be reduced accordingly. Soft-switch functions are employed in the proposed backstepping excitation controls of SGs, respectively, to dominate the weights of the power angles (σ_i) in the defined error variables (6.36).

Define positive constants σ_{0i} ,

$$\sigma_i = \mu_{\sigma_i} \sigma_{0i}, \quad 0 \leq \sigma_i \leq \sigma_{0i} \quad (6.42)$$

where

$$\begin{aligned} \mu_{\sigma_i} &= 1 - \left(\frac{1}{1 + \exp(-120(z_{bi} + 0.4))} - \frac{1}{1 + \exp(-120(z_{bi} - 0.4))} \right) \\ z_{bi} &= \sqrt{\Delta\omega_i^2 + (\Delta P_{ei})^2}. \end{aligned}$$

Since the system stability enhancement and voltage regulations are considered simultaneously in the controllers' designs, the backstepping control schemes can ensure the successful achievement of the both goals.

Subsequently, the design processes of the voltage controllers for the second subsystem of the i th generator will be briefly introduced. The subsystem (6.40) model can be described with the input-output relations with uncertainties as

$$\begin{aligned} S_i : \dot{x}_i(t) &= [A_i + \Delta A_i(t)]x_i(t) + [B_i + \Delta B_i(t)]u_i(t) \\ &+ \sum_{j=1}^n [G_{1ij} + \Delta G_{1ij}(t)]g_{1ij}(x_i, x_j) \\ &+ \sum_{j=1}^n [G_{2ij} + \Delta G_{2ij}(t)]g_{2ij}(x_i, x_j) \end{aligned} \quad (6.43)$$

where for the i th ($i = 1, 2, \dots, n$) subsystem, the matrices A_i , B_i , G_{1ij} and G_{2ij} are known real constant matrices of the appropriate dimensions that describe the

nominal model, ΔA_i , ΔB_i , ΔG_{1ij} and ΔG_{2ij} are real time-varying parameter uncertainties and $g_{1ij}(x_i, x_j)$ and $g_{2ij}(x_i, x_j)$ are unknown nonlinear vector functions that represent nonlinearities in the i th subsystem and in the interactions between the i th subsystem and other subsystems.

$$\begin{aligned} G_{1ij} &= \begin{bmatrix} 0 \\ 0 \\ 0 \end{bmatrix}, \quad G_{2ij} = \begin{bmatrix} 0 \\ 0 \\ 0 \end{bmatrix}, \quad g_{1ij} = \sin \delta_{ij}(t), \quad g_{2ij} = \Delta \omega_j(t), \\ \Delta G_{1ij} &= \begin{bmatrix} 0 \\ \sum_{j=1}^n E'_{qi}(t) [\dot{E}'_{qj}(t) B_{ij} + E'_{qj}(t) \dot{B}_{ij}] \\ 0 \end{bmatrix}, \\ \Delta G_{2ij} &= \begin{bmatrix} 0 \\ -\sum_{j=1}^n E'_{qi}(t) E'_{qj}(t) B_{ij} \cos \delta_{ij}(t) \\ 0 \end{bmatrix} \end{aligned}$$

The uncertain matrices $\Delta A_i(t)$, $\Delta B_i(t)$, ΔG_{1ij} and ΔG_{2ij} are assumed to be of the following structures in the assumption:

Assumption 6. *The uncertain matrices $\Delta A_i(t)$, $\Delta B_i(t)$, ΔG_{1ij} and ΔG_{2ij} are assumed to be decomposed of the following structure:*

$$\begin{aligned} [\Delta A_i(t) \quad \Delta B_i(t)] &= L_i F_i(t) [E_{1i} \quad E_{2i}] \\ \Delta G_{1ij,l} &= L_{1ij} F_{1ij}(t) E_{1ij} \\ \Delta G_{2ij,l} &= L_{2ij} F_{2ij}(t) E_{2ij} \end{aligned}$$

where L_i , E_{1i} , E_{2i} , L_{1ij} , L_{2ij} , E_{1ij} and E_{2ij} are real constant matrices, $F_i(t)$, $F_{1ij}(t)$ and $F_{2ij}(t)$ are unknown matrix function with Lebesgue measurable elements and satisfying $F_i^T(t) F_i(t) \leq I_i$, $F_{1ij}(t) F_{1ij}^T(t) \leq I_{1ij}$ and $F_{2ij}(t) F_{2ij}^T(t) \leq I_{2ij}$; I_i , I_{1ij} and I_{2ij} are unitary matrices.

Assumption 7. *There exist known matrices W_{1i} , W_{2i} , W_{1ij} and W_{2ij} such that the nonlinear terms $g_{1ij}(x_i, x_j)$ and $g_{2ij}(x_i, x_j)$ satisfy:*

$$\|g_{1ij}(x_i, x_j)\| \leq \|W_{1i} x_i(t)\| + \|W_{1ij} x_j(t)\|$$

$$\|g_{2ij}(x_i, x_j)\| \leq \|W_{2i}x_i(t)\| + \|W_{2ij}x_j(t)\|.$$

The control objective of the backstepping control designs are to determine the robust decentralized feedback controllers for each of the i th generator (6.33), such that the closed-loop global multi-machine systems are stable and the stability of the interconnected nonlinear multi-machine systems with uncertainties is guaranteed. From [152][96], the following robust controllers can be designed to guarantee the stability of close loop subsystem of the i th generator in the multi-machine power system

$$u_i(t) = -K_i x_i(t). \quad (6.44)$$

It is noted that the backstepping stabilizing feedback laws are in general nonlinear functions of the system states $x_i(t)$. The stability of this global closed loop system is then studied by using quadratic Lyapunov stability theory. The closed loop system for the i th subsystem is then described by the following equation:

$$\begin{aligned} \dot{x}_i(t) = & [A_i + \Delta A_i(t) + (B_i + \Delta B_i(t))K_i]x_i(t) \\ & + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} [(G_{kij} + \Delta G_{kij}(t))g_{kij}(x_i, x_j)] \end{aligned} \quad (6.45)$$

where k stands the number of the uncertainty and nonlinearity terms which includes 3 terms in this proposed multi-machine power system. The stability of the n interconnected systems (6.45) is then studied by using quadratic Lyapunov stability theory.

Theorem 2. *The interconnected uncertain systems in equation (6.33) is quadratically stable for all admissible uncertainties and nonlinearities, satisfying Assumptions 6 and 7, via the decentralized control law and if there exists positive definite symmetric matrices P_i for the following algebraic Riccati equations*

$$(A_i - B_i R_i^{-1} E_{2i}^T E_{1i})^T P_i + P_i (A_i - B_i R_i^{-1} E_{2i}^T E_{1i}) + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} (W_{kij}^T W_{kij} + W_{ki}^T W_{ki})$$

$$\begin{aligned}
& +P_i\{L_iL_i^T - B_iR_i^{-1}B_i^T + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} [G_{kij}(I - \varepsilon_{kij}^2 E_{kij}^T E_{kij})G_{kij}^T + \frac{1}{\varepsilon_{kij}^2} L_{kij}L_{kij}^T]\}P_i \\
& +Q_i = 0
\end{aligned} \tag{6.46}$$

where Q_i are symmetric positive defined matrices, which can be chosen by the designer, $R_i = E_{2i}^T E_{2i} > 0$, ε_{kij} are scaling parameters to be chosen which should satisfy $\varepsilon_{kij}^2 E_{kij}^T E_{kij} < I, \forall i, j = 1, 2, \dots, n$. In this case, the stabilizing state-feedback control laws are given by

$$u_i(t) = -R_i^{-1}(B_i^T P_i + E_{2i}^T E_{1i})x_i(t). \tag{6.47}$$

If the linearized i th system described in (6.33) are controllable, using the corresponding state feedback control laws described in (6.47), then the global interconnected nonlinear system is asymptotically stable.

Lyapunov Stability Proofs: The following preliminary results are introduced [152].

Lemma 1. For any matrices X and Y with appropriate dimensions, there is

$$X^T Y + Y^T X \leq \beta X^T X + \frac{1}{\beta} Y^T Y, \quad \forall \beta > 0. \tag{6.48}$$

Lemma 2. Let G, L, E and $F(t)$ be real matrices of appropriate dimensions with $F(t)$ being a matrix function. Then, for any $\varepsilon > 0$ such that $\varepsilon^2 E^T E \leq I$ and $F^T(t)F(t) \leq I$, there exists

$$[G + LF(t)E][G + LF(t)E]^T \leq G(I - \varepsilon^2 E^T E)^{-1}G^T + \frac{1}{\varepsilon^2} LL^T. \tag{6.49}$$

Define an energy function for the second subsystem of the i th linearized generator model of the multi-machine system (6.33):

$$\begin{aligned}
V_{vi}(x_i, t) &= x_i^T(t)P_i x_i(t) + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} \int_0^t \{[W_{kij}x_j(s)]^T [W_{kij}x_j(s)] \\
&\quad + [W_{kii}x_i(s)]^T [W_{kii}x_i(s)]\} ds
\end{aligned}$$

$$-g_{kij}^T(x_i(s), x_j(s))g_{kij}(x_i(s), x_j(s))\}ds. \quad (6.50)$$

where $P_i (i = 1, 2, \dots, n)$ are positive-definite symmetric matrices, which make the i th closed loop generator subsystem asymptotically stable. In the view of Assumptions 6 and 7, it is obvious that $V_{vi} > 0, \forall x_i \neq 0$. The Lyapunov function candidate for the interconnected nonlinear uncertain multi-machine system is as follows:

$$\begin{aligned} V_v &= \sum_{i=1}^n V_{vi}(x_i, t) \\ &= \sum_{i=1}^n \{x_i^T(t)P_i x_i(t) + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} \int_0^t [(W_{kij}x_j(s))^T(W_{kij}x_j(s)) \\ &\quad + [W_{kii}x_i(s)]^T[W_{kii}x_i(s)] - g_{kij}^T(x_i(s), x_j(s))g_{kij}(x_i(s), x_j(s))]ds\} \end{aligned}$$

In the view of Assumptions 6 and 7, it is obvious that $V_v > 0, \forall x \neq 0$.

The i th subsystem (6.47) with the proposed feedback control law can be represented as

$$\begin{aligned} \dot{x}_i(t) &= (A_i + \Delta A_i(t))x_i(t) + (B_i + \Delta B_i(t))u_i(t) \\ &\quad + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} [(G_{kij} + \Delta G_{kij}(t))g_{kij}(x_i, x_j)] \\ &= (\bar{A}_i + L_i F_i(t) \bar{E}_i)x_i(t) \\ &\quad + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} [(G_{kij} + L_{kij} F_{kij}(t) E_{kij})g_{kij}(x_i, x_j)] \end{aligned} \quad (6.51)$$

where $\bar{A}_i = A_i - B_i K_i$ and $\bar{E}_i = E_{1i} - E_{2i} K_i$.

The derivative of the energy function of the second subsystem of the i th generator model for interconnected multi-machine systems can be represented as

$$\begin{aligned} L(x, t) &= \dot{V}_v \\ &= \sum_{i=1}^n \{ \dot{x}_i^T(t)P_i x_i(t) + x_i^T(t)P_i \dot{x}_i(t) + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} [(W_{kij}x_j)^T(W_{kij}x_j) \end{aligned}$$

$$\begin{aligned}
& + (W_{ki}x_i)^T (W_{ki}x_i) - g_{kij}^T(x_i, x_j)g_{kij}(x_i, x_j)] \} \\
= & \sum_{i=1}^n \{ x_i^T(t) [(\bar{A}_i + L_i F_i(t) \bar{E}_i)^T P_i + P_i (\bar{A}_i + L_i F_i(t) \bar{E}_i)] x_i(t) \\
& + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} \{ [(G_{kij} + L_{kij} F_{kij}(t) E_{kij}) g_{kij}(x_i, x_j)]^T P_i x_i(t) \\
& + x_i^T(t) P_i (G_{kij} + L_{kij} F_{kij}(t) E_{kij}) g_{kij}(x_i, x_j) \} \\
& + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} [(W_{kij} x_j)^T (W_{kij} x_j) + (W_{ki} x_i)^T (W_{ki} x_i) \\
& - g_{kij}^T(x_i, x_j) g_{kij}(x_i, x_j)] \} \\
= & \sum_{i=1}^n \{ x_i^T(t) [\bar{A}_i^T P_i + (L_i F_i(t) \bar{E}_i)^T P_i + P_i \bar{A}_i + P_i L_i F_i(t) \bar{E}_i] x_i(t) \\
& + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} [g_{kij}^T(x_i, x_j) (G_{kij} + L_{kij} F_{kij}(t) E_{kij})^T P_i x_i(t) \\
& + x_i^T(t) P_i (G_{kij} + L_{kij} F_{kij}(t) E_{kij}) g_{kij}(x_i, x_j)] \\
& + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} [(W_{kij} x_j)^T (W_{kij} x_j) + (W_{ki} x_i)^T (W_{ki} x_i) \\
& - g_{kij}^T(x_i, x_j) g_{kij}(x_i, x_j)] \}. \tag{6.52}
\end{aligned}$$

According to the Lemma 1 with $\beta = 1$, thus

$$\begin{aligned}
L(x, t) & \leq \sum_{i=1}^n \{ x_i^T(t) [\bar{A}_i^T P_i + P_i \bar{A}_i + \bar{E}_i^T F_i^T(t) F_i(t) \bar{E}_i + P_i L_i L_i^T P_i] x_i(t) \\
& + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} [g_{kij}^T(x_i, x_j) g_{kij}(x_i, x_j) \\
& + x_i^T(t) P_i (G_{kij} + L_{kij} F_{kij}(t) E_{kij}) (G_{kij} + L_{kij} F_{kij}(t) E_{kij})^T P_i x_i(t)] \\
& + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} [(W_{kij} x_j)^T (W_{kij} x_j) + (W_{ki} x_i)^T (W_{ki} x_i) - g_{kij}^T(x_i, x_j) g_{kij}(x_i, x_j)] \} \\
= & \sum_{i=1}^n \{ x_i^T(t) [\bar{A}_i^T P_i + P_i \bar{A}_i + \bar{E}_i^T F_i^T(t) F_i(t) \bar{E}_i + P_i L_i L_i^T P_i] x_i(t)
\end{aligned}$$

$$\begin{aligned}
& + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} x_i^T(t) P_i (G_{kij} + L_{kij} F_{kij}(t) E_{kij}) (G_{kij} + L_{kij} F_{kij}(t) E_{kij})^T P_i x_i(t) \\
& + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} x_i^T (W_{kij}^T W_{kij} + W_{ki}^T W_{ki}) x_i(t) \}.
\end{aligned} \tag{6.53}$$

Then by employing Lemma 2, due to that $F_i^T(t) F_i(t) \leq I$, it follows that

$$\begin{aligned}
L(x, t) & \leq \sum_{i=1}^n \{ x_i^T(t) (\bar{A}_i^T P_i + P_i \bar{A}_i + \bar{E}_i^T \bar{E}_i + P_i L_i L_i^T P_i) x_i(t) \\
& + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} x_i^T(t) P_i [G_{kij} (I - \varepsilon_{kij}^2 E_{kij}^T E_{kij}) G_{kij}^T + \frac{1}{\varepsilon_{kij}^2} L_{kij} L_{kij}^T] P_i x_i(t) \\
& + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} x_i^T (W_{kij}^T W_{kij} + W_{ki}^T W_{ki}) x_i(t) \}.
\end{aligned} \tag{6.54}$$

Noting that $R_i^{-1} = E_{2i}^{-1} (E_{2i}^T)^{-1} = (R_i^{-1})^T$ and substituting the $\bar{A}_i$, $\bar{E}_i$ and the proposed feedback control laws into

$$\begin{aligned}
L(x, t) & \leq \sum_{i=1}^n x_i^T(t) \{ [A_i - B_i R_i^{-1} (B_i^T P_i + E_{2i}^T E_{1i})]^T P_i \\
& + P_i [A_i - B_i R_i^{-1} (B_i^T P_i + E_{2i}^T E_{1i})] \\
& + [E_{1i} - E_{2i} R_i^T (B_i^T P_i + E_{2i}^T E_{1i})]^T [E_{1i} - E_{2i} R_i^T (B_i^T P_i + E_{2i}^T E_{1i})] \\
& + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} (W_{kij}^T W_{kij} + W_{ki}^T W_{ki}) \\
& + P_i \{ \sum_{j=1}^n \sum_{k=1}^{T_{jk}} [G_{kij} (I - \varepsilon_{kij}^2 E_{kij}^T E_{kij}) G_{kij}^T + \frac{1}{\varepsilon_{kij}^2} L_{kij} L_{kij}^T] + L_i L_i^T \} P_i \} x_i(t) \\
& = \sum_{i=1}^n x_i^T(t) \{ (A_i - B_i R_i^{-1} E_{2i}^T E_{1i})^T P_i + P_i (A_i - B_i R_i^{-1} E_{2i}^T E_{1i}) \\
& - P_i B_i (R_i^{-1})^T B_i^T P_i \\
& - P_i B_i R_i^{-1} B_i^T P_i + E_{1i}^T E_{1i} - E_{1i}^T E_{2i} R_i^{-1} B_i^T P_i - E_{1i}^T E_{2i} R_i^{-1} E_{2i}^T E_{1i} \\
& - P_i B_i (R_i^{-1})^T E_{2i}^T E_{1i} - E_{1i}^T E_{2i} (R_i^{-1})^T E_{2i}^T E_{1i} + E_{1i}^T E_{2i} R_i^{-1} B_i^T P_i \\
& + E_{1i}^T E_{2i} (R_i^{-1})^T E_{2i}^T E_{1i} + P_i B_i (R_i^{-1})^T B_i^T P_i + P_i B_i (R_i^{-1})^T E_{2i}^T E_{1i}
\end{aligned}$$

$$\begin{aligned}
& + P_i \left\{ \sum_{j=1}^n \sum_{k=1}^{T_{jk}} [G_{kij} (I - \varepsilon_{kij}^2 E_{kij}^T E_{kij}) G_{kij}^T + \frac{1}{\varepsilon_{kij}^2} L_{kij} L_{kij}]^T + L_i L_i^T \right\} P_i \\
& + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} (W_{kij}^T W_{kij} + W_{ki}^T W_{ki}) \} x_i(t) \\
= & \sum_{i=1}^n x_i^T(t) \{ (A_i - B_i R_i^{-1} E_{2i}^T E_{1i})^T P_i + P_i (A_i - B_i R_i^{-1} E_{2i}^T E_{1i}) - P_i B_i R_i^{-1} B_i^T P_i \\
& + P_i \left\{ \sum_{j=1}^n \sum_{k=1}^{T_{jk}} [G_{kij} (I - \varepsilon_{kij}^2 E_{kij}^T E_{kij}) G_{kij}^T + \frac{1}{\varepsilon_{kij}^2} L_{kij} L_{kij}]^T + L_i L_i^T \right\} P_i \\
& + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} (W_{kij}^T W_{kij} + W_{ki}^T W_{ki}) \} x_i(t) \\
= & -x_i^T(t) Q_i x_i(t) \\
\leq & 0
\end{aligned} \tag{6.55}$$

where

$$\begin{aligned}
Q_i = & -\{ (A_i - B_i R_i^{-1} E_{2i}^T E_{1i})^T P_i + P_i (A_i - B_i R_i^{-1} E_{2i}^T E_{1i}) \\
& + P_i \left\{ \sum_{j=1}^n \sum_{k=1}^{T_{jk}} [G_{kij} (I - \varepsilon_{kij}^2 E_{kij}^T E_{kij}) G_{kij}^T + \frac{1}{\varepsilon_{kij}^2} L_{kij} L_{kij}]^T + L_i L_i^T \right\} P_i \\
& - P_i B_i R_i^{-1} B_i^T P_i + \sum_{j=1}^n \sum_{k=1}^{T_{jk}} (W_{kij}^T W_{kij} + W_{ki}^T W_{ki}) \}.
\end{aligned} \tag{6.56}$$

which is consistent to that given in Theorem 2. So the proposed robust voltage controller (6.41) for the subsystem (6.40) can make the error variable z_{Pi} possessing an expected asymptotical behavior ($z_{Pi} \rightarrow 0$).

The global Lyapunov function candidate for the proposed backstepping controlled global power system (6.32) can be given by

$$V = \sum_{i=1}^n V_{mi} + V_v = \sum_{i=1}^n \frac{1}{2} r_i^2 + V_v. \tag{6.57}$$

Thus

$$\dot{V} = \sum_{i=1}^n \dot{V}_{mi} + \dot{V}_v = \sum_{i=1}^n (-r_i^2 + r_i z_{Pi}) + \dot{V}_v. \quad (6.58)$$

Since $z_{P_i} \rightarrow 0$ are ensured by (6.41), it follows immediately that $\dot{V} < 0, \forall x \neq 0$. The global the power system (6.32) is globally uniformly asymptotically stable for all admissible uncertainties in the sense of Lyapunov.

6.3.2 Simulation Results

In this section, the proposed control scheme for power systems is tested in Mat-Lab/Simulink. The performance comparisons between the global control and the proposed backstepping control are given with some analysis. The overall structure of a typical three-machine nine-bus power system is shown in Fig.6.11.

Figure 6.11. Classical three-machine nine-bus power systems

The fault in this section is a symmetrical three phase short circuit fault with its sequences described as:

Case 1. Temporary Fault:

Stage 1: The system is in a pre-fault steady state;

Stage 2: A fault occurs at $t = t_0$;

Stage 3: The fault is removed by opening the breakers of the transmission line at $t = t_1$;

Stage 4: The transmission line are restored at $t = t_3$.

Case 2. Permanent Fault:

Stage 1: The system is in a pre-fault steady state;

Stage 2: A fault occurs at $t = t_0$;

Stage 3: The fault is removed by opening the breakers of the transmission line at $t = t_1$;

Stage 4: The system is in a post-fault state.

$t_0 = 0.1s$, $t_1 = 0.25s$, $t_2 = 1s$, $t_3 = 1.4s$ are chosen in the simulations. The fault is supposed occurring on one of the transmission lines between bus 7 and 8. The fault location is indexed by a positive constant λ_l which is the fraction of the line to the left of the fault. $\lambda_l = 0.1$ and $\lambda_l = 0.01$ are employed in the simulations for testing, respectively.

The global control which is the average of the individual control laws is employed for comparisons. The power system responses with the different controllers subjected to different faults are shown in Figures 6.12-6.13. In this section, comparisons are mainly based on the performances between the existing global control and the proposed backstepping control scheme. All the dash-dot lines express the responses of global controlled power system, and the performances of the proposed backstepping controlled power system are indicated by solid lines.

The overall structure for the three-machine nine-bus power system is shown in Figure 6.11. Define the Generator 3 as the reference so that its power angle is 0. According to the theory given in [96], it is easy to estimate the state uncertainties of the power system. Thus, the baseline voltage controller for the

proposed backstepping feedback excitation control law of Generator 1 can be determined with parameters $k_{V1} = 28.23$, $k_{\omega1} = -20.01$ and $k_{P1} = 180.31$. The parameter $\sigma_{01} = 0.05$ of Generator 1 is chosen. The corresponding backstepping control parameters for Generator 2 are chosen as $k_{V2} = 27.74$, $k_{\omega2} = -21.12$, $k_{P2} = 175.81$ and $\sigma_{02} = 0.05$. The system operating point is given in the following table,

Table 6.3. Operating point of the three-machine nine-bus power system

Generator 1	
Power Angle	$\delta_{01} = 21.4^\circ$
Electrical Frequency	$\omega_{01} = 376.9911rad/s$
Electrical Power	$P_{m01} = 1.63p.u.$
Terminal Voltage	$V_{t01} = 1.025p.u.$
Generator 2	
Power Angle	$\delta_{02} = 14.62^\circ$
Electrical Frequency	$\omega_{02} = 376.9911rad/s$
Electrical Power	$P_{m02} = 0.85p.u.$
Terminal Voltage	$V_{t02} = 1.025p.u.$
Generator 3	
Electrical Power	$P_{m03} = 0.7164p.u.$
Terminal Voltage	$V_{t03} = 1.04p.u.$

Figure 6.12 exhibits the closed-loop multi-machine power system responses for the two global control and the proposed backstepping control scheme in fault sequence *case 1* with fault location $\lambda_l = 0.01$. As shown in Figure 6.12, although both of the control schemes can achieve good transient control performance and restoration performance of post-transient voltage, the recovery performances of the power angle and terminal voltage of the proposed backstepping control have been improved compared to those of the global control, especially for Generator 2. And it also shows that the SGs frequencies control performances have been improved some but not much by the proposed backstepping control scheme. It is worth pointing out that the recoveries of Generator 1 and Generator 2 terminal voltages have been accelerated by the proposed backstepping control scheme which is of great significance for the power system post-transient recovery. The Generator 1 terminal voltage recovery has been accelerated for about 2sec, mean-

while, about 4sec for the Generator 2 terminal voltage.

The simulation results of *case 2* with fault location $\lambda_f = 0.01$ are shown in Figure 6.13. The simulation results are consistent with those of the previous one situation. Significant improvements of the proposed backstepping control performances of the power angles, the SGs frequencies and the SGs terminal voltages of both Generator 1 and Generator 2 are all observed compared to those of the traditional controls. From the simulations, one can see that the proposed backstepping control scheme achieves the proposed control task and is robust with respect to different faults.

6.4 Concluding Remarks

In this chapter, a backstepping control scheme of power system is presented. The backstepping control objectives for power system are defined as not only achieving satisfactory control performances of transient control and voltage regulation for the power system when subjected to a severe disturbance, but ensuring the global stability during the disturbance. For the existing global controller, every sub-controller is designed individually based on its corresponding subsystem. Each sub-controller can make the corresponding subsystem stable, but cannot ensure the entire system stability. Therefore, the soft switching membership function is not only used to combine the sub-controllers but also try to ensure the entire system stability. So the global performance of the global method is dependent on the performance of the soft switching membership function. It is a high risk for ensuring the global stability of the entire power system. Some simulation results have been given to support this claim. The proposed backstepping control scheme successfully resolved the inadequacies of the global control. Based on backstepping principle, the proposed backstepping control for power system can achieve satisfactory control performances of transient control and voltage regulation and ensure the global stability as well. Moreover, the proposed backstepping control provides a method of coordination and control between the synchronous generator and other power components in the power system from the power point of view. The proposed method is verified on both a single machine infinite bus

power system and a three-machine nine-bus multi-machine power system. The proposed backstepping control technique is proved to be effective for the highly nonlinear and highly interconnected multi-machine power systems and robust with respect to different faults.

In the next chapter, the applications of backstepping control theory are extended to complex power system including the traditional synchronous generator power sources and grid-integrated WTPGS. The backstepping controller will be responsible for both of the two different power sources and to provide effective coordination control of the SG and WTPGS during power system transient period.

Figure 6.12. Multi-machine power system responses for *case 1* with fault location $\lambda_l = 0.01$

Figure 6.13. Multi-machine power system responses for *case 2* with fault location $\lambda_l = 0.01$

Backstepping Control of Power Systems with Grid-connected Wind Turbine Power Generation System

Large-scale use of wind power raises many questions in integration into the existing utility power grid. The connection of large wind turbines to the grid has a large impact on grid stability [153]. Basically, the grid codes, which cover a wide range of voltage levels from medium voltage to very high voltage, require wind turbines to have an operational behavior more similar to that of conventional generation capacity and more responsibility in network operation [7][154]. The status of wind turbines is thus changing from being simple energy sources to having power plant status. In PMSG WTPGS, these requirements have focus on power controllability. The power control capability means mainly that the wind turbines have to share some of the duties carried out traditionally by conventional power plants, such as regulating active and reactive power and performing frequency and voltage control on the grid. Wind farms connected at the transmission level shall act as a conventional power plant providing a wide range of controlling the output power.

The overall Detail Control SG WTPGS is shown in Figure 7.1.

The backstepping control scheme illustrated in the previous chapter is enriched

Figure 7.1. Detail Control SG WTPGS

to deal with the transient stabilization control and voltage regulation problems in SMIB power system with PCC bus connected WTPGS and dominate the controls of the WTPGS, simultaneously. Above all, the backstepping technique can guarantee the system stability theoretically. Subsequently, the controls of WTPGS including the rotational speed and pitch controls are considered within the backstepping control design of the SG transient stabilization and voltage regulation controls. So in power grid transient process, it makes the rotational speed and pitch controls of the WTPGS more active by its decentralized rotor speed controller and pitch controller under the condition of ensuring the self stability of wind turbine. It provides an integrated and coordinated control structure between SG and WTPGS from the real power point of view.

The WTPGS mentioned in this chapter expresses a wind farm which is composed of a large number of wind turbine power generation system sharing the same capacity and parameters. All the individual wind turbines are regarded as

power lossless systems. Compared to the entire capacity of the non-conventional power system, the power losses on the back-to-back converters are negligible. Thus, in the real-power point of view, the back-to-back converters of WTPGS can be considered ideal. Since the back-to-back converters are ideal, the wind turbine and PMSG part can be considered to be decoupled to the grid. And all the back-to-back converters are not taken into account in the overall modeling.

7.1 Backstepping Control of SMIB Power Systems with Grid-connected WTPGS

In this section, a SMIB power system with a WTPGS installed at PCC bus is considered. The overall structure for the particular power system is shown in Figure 7.2.

Figure 7.2. A single machine infinite bus power system with PCC installed WTPGS

7.1.1 Backstepping Control of SMIB Power System with Grid-connected WTPGS

The modeling of the SMIB power system with a PCC installed WTPGS are obtained from Section 2.3 and Section 2.4. Considering only the electrical behavior of the WTPGS, a simplified modeling of the wind turbine blade and shaft is normally used [3]. This section mainly focuses on the effects of the wind turbine

activation and control on the power systems, thus the converters of the WTPGS are considered operating ideally.

Firstly, make an assumption that the mentioned WTPGS in the power system behaves as a wind farm which is composed of lots of identical PMSG wind turbine components. According to Section 2.3, the each wind turbine extracted power P_w from the wind is a cubic function with respect to the wind speed,

$$P_w = \frac{1}{2}\pi R^2 \rho C_p(\lambda, \beta) V^3 = T_w \omega_m \quad (7.1)$$

where ρ is the air density (kg/m^3), R is the radius of the blade swept area (m), V is the wind speed (m/s), ω_m is the mechanical rotational speed of the turbine shaft, T_w is wind mechanical torque on the turbine shaft, the power coefficient $C_p(\lambda, \beta)$ is given in (2.14) with respect to the tip speed ratio (TSR) $\lambda = \frac{\omega_m R}{V}$ and the pitch angle β . The torque dynamic equation on the generator rotor can be described as one-mass lumped systems,

$$J_t \frac{d\omega_g}{dt} = T_{wg} - T_g. \quad (7.2)$$

where J_t is the moment of inertia of the lumped generator system, ω_g is the generator rotor speed.

The wind turbine installed with a direct-drive Permanent Magnet Synchronous Generator (PMSG) [51][115] is employed in the investigation as a representative of the WTPGS. ψ is the magnet flux and ω is the generator electric speed. Note that T_{wg} is the equivalent wind mechanical torque on the generator rotor which is proportional to T_w . Given the turbine gear ratio n , thus, $nT_{wg} = T_w$.

If the pole pair number is p , then $p\omega = \omega_g$. The electromagnetic torque is given as

$$T_g = \frac{3}{2}p(\psi i_{sq} - (L_d - L_q)i_{sd}) = \frac{3}{2}p\psi i_{sq}. \quad (7.3)$$

According to (7.3), the electromagnetic torque is only relevant to the quadrature current component i_{sq} . A general PI controller for the rotor speed is of the form

$$i_{sq}^* = -(k_p + k_i \frac{1}{s})(\omega_g^* - \omega_g). \quad (7.4)$$

The stator generated power is $P_g = T_g \omega_g$. The WTPGS output active and reactive power are expressed as P_{wg} and Q_{wg} , respectively.

State that the WTPGS is regarded as a power lossless system. Thus, $P_{wg} = P_g$. The output active and reactive power controls can be achieved by controlling quadrature and direct current components, respectively. Since no reactive power control is considered in this work, the output terminal power factor is maintained at 1. In other words, $Q_{wg} = 0$ is kept by the ideal power inverter during transient process.

Figure 7.3. WTPGS pitch control system

The conventional pitch control system of WTPGS is shown in Figure 7.3..

Figure 7.4. Torque characteristic and torque coefficient curve of wind turbine

The wind turbine torque coefficient can be described as $C_T(\lambda, \beta) = \frac{C_p(\lambda, \beta)}{\lambda}$ with respect to tip speed ratio (TSR) λ and pitch angle β . From Figure 7.4(a), the stable operation region of wind turbine can easily be determined. It shows that the operating point P_1 on curve OP_1A is unstable. When the wind speed

is maintaining at a constant and turbine speed has a small positive (negative) disturbance, the mechanical torque will increase (decrease) accordingly, which leads to a increase (decrease) of the turbine speed. However, a disturbance of turbine speed around point P_2 on curve AP_2B can be retrieved to P_2 in finite time. As a result, the wind turbine has a stable operating point which is with respect to unique λ and unique β [106][155]. Figure 7.4(b) gives the entire torque coefficient stable operating region of the wind turbine.

Assume that the output power of the WTPGS is P_{w0} under normal operation condition. And denote $\Delta P_{wg} = P_{wg} - P_{w0}$ as the output power deviation of the WTPGS. With considering the power flow effects of the WTPGS on the SMIB power system, the system can be modeled with considering the total input power deviation of the entire power system,

$$\begin{aligned}\dot{\delta}(t) &= \omega(t) \\ \dot{\omega}(t) &= -\frac{D}{2H}\omega(t) - \frac{\omega_0}{2H}(\Delta P_e(t) - \Delta P_{wg}) \\ \Delta \dot{P}_e(t) &= -\frac{1}{T'_{d0}}\Delta P_e(t) + \frac{1}{T'_{d0}}v_f(t) + E'_q(t)\dot{E}'_{qp}(t)B_p \sin \delta_p(t) \\ &\quad - E'_q(t)E'_{qp}(t)B_p \cos \delta_p(t)\omega_p(t)\end{aligned}\quad (7.5)$$

where $E'_{qp}(t)$ is the transient voltage of the PCC bus, B_p expresses the susceptance between the generator and the PCC bus and $\delta_p(t)$ dedicates the phase angle difference between $E'_q(t)$ and the PCC bus.

In order to ensure the entire system stability, the controller designs for the WTPGS in transient process should be considered within the error variable z_p asymptotical behavior to satisfy the entire non-conventional power system stability. Since the error variable $z_p = \Delta P_e(t) - \alpha_P$ is expected to possess an asymptotical behavior, the real-time error $\Delta P_e(t) - \alpha_P$ should be eliminated as possible as it can. Thus, ideally, this real-time power error can be compensated by the WTPGS deviation from the power point of view which makes the whole system on an optimal state of asymptotically stable. It is reasonable to employ the z_p in the control of WTPGS output power P_{wg} as the power reference $P_{wg}^* = z_p$.

According to the torque dynamic equation of the generator rotor (7.2) and all

the power losses of the WTPGS are negligible, the real-time power reference of the wind turbine captured power P_w^* can be derived from P_{wg}^* .

$$\begin{aligned} J_t \omega_g \frac{d\omega_g}{dt} &= T_{wg} \omega_g - T_g \omega_g = P_w^* - P_{wg}^* \\ \Rightarrow P_w^* &= J_t \omega_g \frac{d\omega_g}{dt} + P_{wg}^* \end{aligned} \quad (7.6)$$

where ω_g and $\frac{d\omega_g}{dt}$ are available.

With turbine extraction power reference P_w^* , the corresponding objective torque coefficient $C_T^*(\lambda, \beta)$ is given as $C_T^*(\lambda, \beta) = \frac{2P_w^*}{\rho \pi R^2 V^3 \lambda}$. It is reasonable to set an assumption that the wind speed is maintaining at a constant during the transient process of several seconds. Undoubtedly, the turbine and generator rotational speeds are measurable.

According to Figure 7.4, the entire stable operation region versus the torque coefficient $C_T(\lambda, \beta)$ of WTPGS can be determined. With the given analysis, the wind turbine has a stable operating point which is with respect to unique λ and unique β when torque coefficient reference $C_T^*(\lambda, \beta)$ is determined. Thus, the inverse λ and β functions of $C_T(\lambda, \beta)$ can be derived uniquely, respectively.

$$\begin{aligned} \beta &= f'_\beta(C_T(\lambda, \beta), \lambda) \\ \lambda &= f'_\lambda(C_T(\lambda, \beta), \beta). \end{aligned} \quad (7.7)$$

Since λ and β are both measurable, the real-time control references of λ and β can be determined as

$$\begin{aligned} \beta^* &= f'_\beta(C_T^*(\lambda, \beta), \lambda) \\ \lambda^* &= f'_\lambda(C_T^*(\lambda, \beta), \beta) \longrightarrow \omega_m^* \longrightarrow \omega_g^*. \end{aligned} \quad (7.8)$$

In this thesis, referring to the application of ELM mapping characteristic based sensorless control in Section 5.3 in details, the ELM based ANNs input-output mapping technique is implemented to build dynamic searching laws for searching the rotor speed reference and pitch angle reference within the stable operation region, respectively. It is essentially using the trained ELM ANNs for

the inverse functions (7.8) regression [137]. Evidently, according to Figure 7.4, the stability of the WTPGS is guaranteed by the proposed WTPGS control.

The remaining problem only lies in the design of appropriate backstepping controller for the SG which can refer to Chapter 6.

7.1.2 Simulation Results

In this section, the proposed backstepping control scheme for the SMIB power system with PCC installed WTPGS is tested in MatLab/Simulink. The performance comparisons between the SG global control with system frequency based WTPGS pitch control and the proposed controls are given with some analysis.

The fault in this section is a symmetrical three phase short circuit fault with its sequences described as:

Case 1. Temporary Fault:

Stage 1: The system is in a pre-fault steady state;

Stage 2: A fault occurs at $t = t_0$;

Stage 3: The fault is removed by opening the breakers of the transmission line at $t = t_1$;

Stage 4: The transmission line are restored at $t = t_3$.

Case 2. Permanent Fault:

Stage 1: The system is in a pre-fault steady state;

Stage 2: A fault occurs at $t = t_0$;

Stage 3: The fault is removed by opening the breakers of the transmission line at $t = t_1$;

Stage 4: The system is in a post-fault state.

$t_0 = 0.1s$, $t_1 = 0.25s$, $t_2 = 1s$, $t_3 = 1.4s$ are chosen in the simulations. The fault location is indexed by a positive constant λ_l which is the fraction of the line to

the left of the fault. $\lambda_l = 0.5$ and $\lambda_l = 0.01$ are employed in the simulations, respectively.

The global control shown for comparisons can be found in [100]. The global control law is the average of the individual control laws, weighted by the soft switching membership function and the input v_f takes the form as,

$$v_f = \mu_\delta v_{f1} + \mu_V v_{f2} \quad (7.9)$$

where v_{f1} expresses the DFL nonlinear transient controller and v_{f2} is the voltage controller which are given respectively as,

$$\begin{aligned} v_{f1} &= 22.36\delta + 12.81\omega - 82.45\Delta P_e \\ v_{f2} &= -22.03\Delta V_t + 1.01\omega - 180.81\Delta P_e. \end{aligned}$$

And the membership functions are given as,

$$\begin{aligned} \mu_V &= \left(1 - \frac{1}{1 + \exp(-120(z - 0.08))}\right) \cdot \left(\frac{1}{1 + \exp(-120(z + 0.08))}\right) \\ \mu_\delta &= 1 - \mu_V \end{aligned}$$

where $z = \sqrt{\omega^2 + (\Delta V_t)^2}$.

The system frequency based WTPGS pitch control shown for comparisons can be found in [110]. As shown in Figure 7.3, a general PI control loop for system frequency based WTPGS pitch system with respect to the PCC bus frequency $\omega_p(t)$ is of the form,

$$\beta^* = -(k_{pp} + k_{ip}\frac{1}{s})(\omega_p^*(t) - \omega_p(t)). \quad (7.10)$$

Using the method given in [95] to estimate the state uncertainties of the power system, the baseline voltage controller for the proposed backstepping feedback control law for SG can be determined with parameters $k_V = 22.03$, $k_\omega = -1.01$ and $k_P = 180.81$. In the backstepping feedback control law (6.41) with (6.42), $\sigma_0 = 0.1$ is chosen. The system operating point is given in the following table,

The power system SG responses and WTPGS responses with the different

Table 7.1. Operating point of the SMIB power system and PCC bus installed WTPGS

Synchronous generator	
Power Angle	$\delta_0 = 34.3^\circ$
Electrical Frequency	$\omega_0 = 314.159\text{rad/s}$
Electrical Power	$P_{m0} = 0.8\text{p.u.}$
Terminal Voltage	$V_{t0} = 1\text{p.u.}$
WTPGS	
Wind Power Supply	$P_{w0} = 0.4\text{p.u.}$
PCC Bus 8 Voltage	$V_{PCC0} = 0.981\text{p.u.}$

controllers subjected to different faults are shown in Figures 7.5-7.10. All the dash-dot lines express the responses of global SMIB SG control with the system frequency based WTPGS pitch control, and the performances of the proposed backstepping control scheme are indicated by solid lines.

Figure 7.5. Power system responses for *case 1* with fault location $\lambda_l = 0.5$

Figures 7.5 and 7.6 exhibit the closed-loop power system responses for the two kinds of control schemes in fault sequence *case 1* with fault location $\lambda_l =$

0.5. As shown in Figure 7.5, although both of the control schemes can achieve good transient control performance and restoration performance of post-transient voltage, it is obvious that the control performances of the power angle, the SG frequency and the electrical power of the proposed backstepping control scheme have been improved compared to those of the global control. It is worth pointing out that the recovery of SG terminal voltage has been accelerated about 2sec by the proposed backstepping control scheme.

Figure 7.6. WTPGS responses for *case 1* with fault location $\lambda_l = 0.5$

Figure 7.6 shows the corresponding WTPGS responses. Due to the limitation of pitch system response speed, the response of turbine captured power for the conventional pitch controller lags far behind that of the proposed backstepping control scheme. Since the proposed control provides a kind of active control method for the rotor speed, this more flexible control overcomes disadvantage of slow response when only the conventional system frequency based pitch control exists. Not only a greater range of turbine captured power can be controlled, but the recovery speed of captured power back to the pre-fault rated value is much faster by the proposed control. More importantly, the PCC bus voltage by the proposed backstepping control scheme is recovered about 2sec ahead of that of the conventional system frequency based pitch control.

Figure 7.7. Power system responses for *case 2* with fault location $\lambda_l = 0.01$

Figures 7.7 and 7.8 shows the responses for for the two kinds of control schemes in *case 2* with fault location $\lambda_l = 0.01$. In this case, there is no significant improvement of the power angle, the SG frequency and the electrical power control performances by the proposed backstepping control. However, the recovery

speeds of the SG terminal and PCC bus voltages are still significantly accelerated.

Figure 7.8. WTPGS responses for *case 2* with fault location $\lambda_l = 0.01$

It is noted that although both of the mentioned control schemes can ensure synchronization of the SG after severe fault, but the post-fault equilibrium points of the SG by the two controls are different. Due to the lack of coordination in the global control and the conventional system frequency based WTPGS pitch controller, WTPGS cannot identify the structure change of power system in permanent fault so that the states of the WTPGS won't be restored to the pre-fault

Figure 7.9. Power system responses for *case 2* with fault location $\lambda_l = 0.5$

situation. As shown in Figure 7.8, the WTPGS captured power isn't recovered back to the pre-fault rated value. This is not allowed. So new pitch controller should be designed instead of the conventional system frequency based WTPGS pitch controller. However, the proposed backstepping controlled system successfully overcomes this disadvantage. The states of the WTPGS can be recovered as well.

The simulation results of *case 2* with fault location $\lambda_l = 0.5$ are shown in Figures 7.9 and 7.10. The simulation results are consistent with those of the previous two situations. Significant improvements of the proposed backstepping control performances of the power angle, the SG frequency, the SG output electrical power, the terminal voltage and the PCC bus voltage are all observed compared to those of the traditional controls. Meanwhile, the states of the WTPGS successfully restores to the pre-fault situation. From the simulations, one can see that the proposed backstepping control scheme achieves the proposed control task and is robust with respect to different faults.

Figure 7.10. WTPGS responses for *case 2* with fault location $\lambda_l = 0.5$

For the global SG control with the conventional system frequency based WTPGS pitch control, on one hand, it seems that the traditional controls cannot guarantee the global stability of the entire system in theory. On the other hand, they cannot provide good integration and coordination control performances for the whole power system, because of lacking of overall consideration in design. Exactly, the shortcomings of the the traditional controls can be overcome by the proposed backstepping control scheme.

7.2 Backstepping Control of Multi-machine Power System with Grid-connected WTPGS

The proposed backstepping control scheme for power system with grid-connected WTPGS proposed in the previous section is extended to a large scale power system including multi-machine synchronous generators and a grid-integrated WTPGS. A three-machine nine-bus power system with the WTPGS installed at one of the SG PCC bus is employed in simulation. Power system uncertainties and interactions are taken into account in the design of decentralized controllers for every SG excitation. Overall, the controller designs for the WTPGS is based on the best pursuit of system input-output power balance, in considering the entire multi-machine system as a whole which should consider the on-line information of all the SGs during the transient period.

A typical three-machine nine-bus power system with a WTPGS installed at the PCC bus 8 is considered. The overall structure for the particular power system is shown in Figure 7.11. The WTPGS shares the same model as shown in Section 7.1.

7.2.1 Backstepping Control of Multi-Machine Power System with Grid-connected WTPGS

Assume that the output power of the WTPGS is P_{w0} under normal operation condition. And denote $\Delta P_{wg} = P_{wg} - P_{w0}$ as the output power deviation of the WTPGS. With considering the power flow effects of the WTPGS on the global multi-machine power system, the entire multi-machine system input-output power should be balanced. Denote ΔP_{wgn} as the real power deviation sent to the n th synchronous generator from the WTPGS. Thus, $\sum_{j=1}^n \Delta P_{wgn} = \Delta P_{wg}$. the multi-machine system modeling can be considered as,

$$\begin{aligned}\dot{\delta}_i(t) &= \Delta\omega_i(t) \\ \Delta\dot{\omega}_i(t) &= -\frac{D_i}{2H_i}\Delta\omega(t) - \frac{\omega_0}{2H_i}(\Delta P_{ei}(t) - \Delta P_{wgn})\end{aligned}$$

Figure 7.11. A three-machine nine-bus power system with grid-connected WTPGS

$$\begin{aligned}
\Delta \dot{P}_{ei}(t) = & -\frac{1}{T'_{d0i}} \Delta P_{ei}(t) + \frac{1}{T'_{d0i}} v_{fi}(t) + \sum_{j=1}^n E'_{qi}(t) [\dot{E}'_{qj}(t) B_{ij} + E'_{qj}(t) \dot{B}_{ij}] \sin \delta_{ij}(t) \\
& - \sum_{j=1}^n E'_{qi}(t) E'_{qj}(t) B_{ij} \cos \delta_{ij}(t) \Delta \omega_j(t).
\end{aligned} \tag{7.11}$$

Since the error variables $z_{Pi} = \Delta P_{ei}(t) - \alpha_{Pi}$ are expected to possess asymptotical behaviors, the real-time errors $\Delta P_{ei}(t) - \alpha_{Pi}$ should be eliminated as possible as they can. Thus, ideally, the total real-time power errors can be compensated by the WTPGS deviation from the power point of view. It is reasonable to employ the total of z_{Pi} in the control of WTPGS output power P_{wg} as the power reference $P_{wg}^* = \sum_{j=1}^n z_{Pi}$.

According to the torque dynamic equation of the generator rotor (7.2) and all the power losses of the WTPGS are negligible, the real-time power reference of the wind turbine captured power P_w^* can be derived from P_{wg}^* .

$$J_t \omega_g \frac{d\omega_g}{dt} = T_{wg} \omega_g - T_g \omega_g = P_w^* - P_{wg}^*$$

$$\implies P_w^* = J_t \omega_g \frac{d\omega_g}{dt} + P_{wg}^* \quad (7.12)$$

where ω_g and $\frac{d\omega_g}{dt}$ are available.

With turbine extraction power reference P_w^* , the corresponding objective torque coefficient $C_T^*(\lambda, \beta)$ is given as $C_T^*(\lambda, \beta) = \frac{2P_w^*}{\rho\pi R^2 V^3 \lambda}$. It is reasonable to set an assumption that the wind speed is maintaining at a constant during the transient process of several seconds. Undoubtedly, the turbine and generator rotational speeds are measurable.

According to Figure 7.4, the entire stable operation region versus the torque coefficient $C_T(\lambda, \beta)$ of WTPGS can be determined. With the given analysis, the wind turbine has a stable operating point which is with respect to unique λ and unique β when torque coefficient reference $C_T^*(\lambda, \beta)$ is determined. Thus, the inverse λ and β functions of $C_T(\lambda, \beta)$ can be derived uniquely, respectively.

$$\begin{aligned} \beta &= f'_\beta(C_T(\lambda, \beta), \lambda) \\ \lambda &= f'_\lambda(C_T(\lambda, \beta), \beta). \end{aligned} \quad (7.13)$$

Since λ and β are both measurable, the real-time control references of λ and β can be determined as

$$\begin{aligned} \beta^* &= f'_\beta(C_T^*(\lambda, \beta), \lambda) \\ \lambda^* &= f'_\lambda(C_T^*(\lambda, \beta), \beta) \longrightarrow \omega_m^* \longrightarrow \omega_g^*. \end{aligned} \quad (7.14)$$

As done in Section 7.1, the ELM based ANNs input-output mapping technique is also implemented to build dynamic searching laws for searching the rotor speed reference and pitch angle reference within the stable operation region, respectively. It is essentially using the trained ELM ANNs for the inverse functions (7.14) regression. Evidently, according to Figure 7.4, the stability of the WTPGS is guaranteed by the proposed WTPGS control. The remaining problem only lies in the design of appropriate backstepping controller for the SG which can refer to Chapter 6.

In this section, the proposed backstepping control scheme for the multi-

machine power system with PCC bus 8 installed WTPGS is tested in Mat-Lab/Simulink. The performance comparisons between the SG global control [100] with system frequency based WTPGS pitch control [110] and the proposed controls are given with some analysis. All the parameters are given in the appendix.

The fault in this section is a symmetrical three phase short circuit fault with its sequences described as:

Case 1. Temporary Fault:

Stage 1: The system is in a pre-fault steady state;

Stage 2: A fault occurs at $t = t_0$;

Stage 3: The fault is removed by opening the breakers of the transmission line at $t = t_1$;

Stage 4: The transmission line are restored at $t = t_3$.

Case 2. Permanent Fault:

Stage 1: The system is in a pre-fault steady state;

Stage 2: A fault occurs at $t = t_0$;

Stage 3: The fault is removed by opening the breakers of the transmission line at $t = t_1$;

Stage 4: The system is in a post-fault state.

$t_0 = 0.1s$, $t_1 = 0.25s$, $t_2 = 1s$, $t_3 = 1.4s$ are chosen in the simulations. The fault is supposed occurring on one of the transmission lines between bus 7 and 8. The fault location is indexed by a positive constant λ_l which is the fraction of the line to the left of the fault. $\lambda_l = 0.95$ and $\lambda_l = 0.01$ are employed in the simulations, respectively.

The global control shown for comparisons can be found in. The global control law for the i th SG is the average of the individual control laws, weighted by the operating region membership function and the input v_{fi} takes the form as,

$$v_{fi} = \mu_{\delta i} v_{fi1} + \mu_{V i} v_{fi2} \quad (7.15)$$

where v_{fi1} expresses the DFL nonlinear transient controller and v_{fi2} is the voltage controller which are given respectively as,

$$\begin{aligned} v_{fi1} &= K_{\delta i1}\delta + K_{\omega i1}\omega + K_{Pi1}\Delta P_e \\ v_{fi2} &= K_{Vi2}\Delta V_t + K_{\omega i2}\omega + K_{Pi2}\Delta P_e. \end{aligned}$$

And the membership functions are given as,

$$\begin{aligned} \mu_{Vi} &= \left(1 - \frac{1}{1 + \exp(-120(z_i - 0.08))}\right) \cdot \left(\frac{1}{1 + \exp(-120(z_i + 0.08))}\right) \\ \mu_{\delta i} &= 1 - \mu_{Vi} \end{aligned}$$

where $z_i = \sqrt{\omega_i^2 + (\Delta V_{ti})^2}$.

The system frequency based WTPGS pitch control shown for comparisons can be found in [110]. As shown in Figure 7.3, a general PI control loop for system frequency based WTPGS pitch system with respect to the frequency of PCC bus 8 ($\omega_8(t)$) is of the form,

$$\beta^* = -(k_{pp} + k_{ip}\frac{1}{s})(\omega_8^*(t) - \omega_8(t)). \quad (7.16)$$

7.2.2 Simulation Results and Analysis

Define the Generator 3 as the reference so that its power angle is 0. According to [95][96] to estimate the state uncertainties of the power system, the baseline voltage controller for the proposed backstepping feedback excitation control law of Generator 1 can be determined with parameters $k_{V1} = 28.23$, $k_{\omega 1} = -20.01$ and $k_{P1} = 180.31$. The parameter $\sigma_{01} = 0.5$ of Generator 1 is chosen. The corresponding backstepping control parameters for Generator 2 are chosen as $k_{V2} = 27.74$, $k_{\omega 2} = -21.12$, $k_{P2} = 175.81$ and $\sigma_{02} = 0.5$. The system operating point is given in the following table,

The multi-machine power system SGs responses and WTPGS responses with the different controllers subjected to different faults are shown in Figures 7.12-7.20. All the dash-dot lines express the responses of SGs global control with

Table 7.2. Operating point of the three-machine nine-bus power system and bus 8 installed WTPGS

Generator 1	
Power Angle	$\delta_{01} = 29.8^\circ$
Electrical Frequency	$\omega_{01} = 376.9911rad/s$
Electrical Power	$P_{m01} = 1.63p.u.$
Terminal Voltage	$V_{t01} = 1.025p.u.$
Generator 2	
Power Angle	$\delta_{02} = 19.7^\circ$
Electrical Frequency	$\omega_{02} = 376.9911rad/s$
Electrical Power	$P_{m02} = 0.85p.u.$
Terminal Voltage	$V_{t02} = 1.025p.u.$
Generator 3	
Electrical Power	$P_{m03} = 0.192p.u.$
Terminal Voltage	$V_{t03} = 1.04p.u.$
WTPGS	
Wind Power Supply	$P_{w0} = 0.8p.u.$
PCC Bus 8 Voltage	$V_{PCC0} = 1.02p.u.$

the system frequency based WTPGS pitch control, and the performances of the proposed backstepping control scheme are indicated by solid lines.

Figures 7.12, 7.13 and 7.14 exhibit the closed-loop power system responses for the two kinds of control schemes in fault sequence *case 1* with fault location $\lambda_l = 0.01$. As shown in Figure 7.12, although both of the control schemes can achieve good transient control performance and restoration performance of post-transient voltage, it is obvious that the control performances of the power angles, the SGs frequencies and the electrical powers of the proposed backstepping control scheme have been improved compared to those of the global control, especially for Generator 1. It is worth pointing out that the recoveries of Generator 1 and Generator 2 terminal voltages have been accelerated by the proposed backstepping control scheme. The Generator 1 terminal voltage recovery has been accelerated for about 4sec, meanwhile, about 1sec for the Generator 2 terminal voltage.

Figure 7.14 shows the corresponding WTPGS responses. Due to the limitation of pitch system response speed, the response of turbine captured power for the

conventional pitch controller lags far behind that of the proposed backstepping control scheme. Since the proposed control provides a kind of active control method for the rotor speed, this more flexible control overcomes disadvantage of slow response when only the conventional system frequency based pitch control exists. Not only a greater range of turbine captured power can be controlled, but the recovery speed of captured power back to the pre-fault rated value is much faster by the proposed control. More importantly, the voltage of WTPGS PCC bus 8 by the proposed backstepping control scheme is recovered about $3sec$ ahead of that of the conventional system frequency based pitch control.

Figures 7.15, 7.16 and 7.17 shows the responses for for the two kinds of control schemes in *case 2* with fault location $\lambda_l = 0.01$. In this case, the simulation results are consistent with the previous test. Significant improvements of the proposed backstepping control performances of the power angles, the SGs frequencies and the electrical powers of both Generator 1 and Generator 2 are all observed compared to those of the traditional controls. However, improvement of terminal voltage recovery of Generator 1 is still very significant, but the improvement for Generator 2 is not obvious.

It is noted that although both of the mentioned control schemes can ensure synchronization of the SGs after severe system fault, but the post-fault equilibrium points of the multi-machine power system by the two controls are different. Due to the lack of coordination in the global controls and the conventional system frequency based WTPGS pitch controller, WTPGS cannot identify the structure change of power system in permanent fault so that the states of the WTPGS won't be restored to the pre-fault situation. As shown in Figure 7.17, the WTPGS captured power and the pitch angle cannot be recovered back to the pre-fault rated value. These are not allowed in the system. So new pitch controller should be designed instead of the conventional system frequency based WTPGS pitch controller to ride through the permanent fault. However, the proposed backstepping controlled system successfully overcomes this disadvantage. The states of the WTPGS can be recovered as well. Significant improvement of PCC bus 8 voltage recovery is still observed which is helpful for the stabilization control of WTPGS.

The simulation results of *case 2* with fault location $\lambda_l = 0.95$ are shown in

Figures 7.18, 7.19 and 7.20. The simulation results are consistent with those of the previous two situations. From the simulations, one can see that the proposed backstepping control scheme achieves the proposed control task and is robust with respect to different faults.

For the SGs global control with the conventional system frequency based WTPGS pitch control, on one hand, it seems that the traditional controls cannot guarantee the global stability of the entire system in theory. On the other hand, they cannot provide good integration and coordination control performances for the entire multi-machine power system, because of lacking of overall consideration in design. Exactly, the shortcomings of the the traditional controls can be overcome by the proposed backstepping control scheme.

7.3 Concluding Remarks

In this chapter, backstepping control scheme composed of backstepping controller for SG and coordination control law for WTPGS is presented. And the proposed control scheme is verified in both a SMIB power system with PCC installed WTPGS and a three-machine nine-bus multi-machine power system with a WTPGS installed at one of the PCC bus. The backstepping control objectives for SGs system are defined as not only achieving satisfactory control performances of transient stabilization and voltage regulation when subjected to a severe disturbance, but ensuring the global stability during the disturbance. The proposed backstepping control scheme successfully resolved the inadequacies of the global control. Some simulation results have been given to support this claim. Moreover, the proposed backstepping control provides a coordination control law for WTPGS of coordination and controls between the SGs and WTPGS from the real power point of view. This control law is designed within the stable operation region of the WTPGS which guarantees the basic stability requirement of WTPGS. With the help of the ELM ANNs, the proposed control is more flexible and active for both of the turbine rotor speed control and pitch control. Therefore, the transient control performance of WTPGS has been significantly improved. Simulation results demonstrate the achievements of the proposed control task. The

proposed backstepping control technique is proven to be effective for the highly nonlinear and highly interconnected complex non-conventional power systems.

Figure 7.12. Power system responses for *case 1* with fault location $\lambda_l = 0.01$

Figure 7.13. Excitation control responses for *case 1* with fault location $\lambda_l = 0.01$

Figure 7.14. WTPGS responses for *case 1* with fault location $\lambda_l = 0.01$

Figure 7.15. Power system responses for *case 2* with fault location $\lambda_l = 0.01$

Figure 7.16. Power system responses for *case 2* with fault location $\lambda_l = 0.01$

Figure 7.17. WTPGS responses for *case 2* with fault location $\lambda_l = 0.01$

Figure 7.18. Power system responses for *case 2* with fault location $\lambda_l = 0.95$

Figure 7.19. Power system responses for *case 2* with fault location $\lambda_l = 0.95$

Figure 7.20. WTPGS responses for *case 2* with fault location $\lambda_l = 0.95$

Conclusions and Recommendations

8.1 Conclusions

Recently, due to the environmental problem and shortage of fossil fuel, as a representative of renewable energy, wind power has received much interest and considerable attention all over the world. However, as a new grid-connected power generation system, the reliability, stability and dynamic behavior of controls of the wind power generation systems are really great challenges for modern industrial engineers.

Related to the interconnection of WTPGS and the utility grid, the control objectives of the WTPGS should be considered from the both sides, the WTPGS themselves and the power grid. From a WTPGS or a wind farm's point of view, it always want to capture as much power as allowed from the wind and send the most power to the utility grid. However, on the perspective of power market and planning, theoretically, the power grid wishes that a wind farm can support a stable power supply and doesn't negatively affect the global stability and reliability of the power systems. Even during a system fault, the control of the wind power plants cannot only ensure enough capabilities of the low voltage ride through of themselves, but can contribute to the fast recovery of the entire power systems.

This thesis deals with the controls of the both sides mentioned above. On the WTPGS side, on one hand, it is necessary to get accurate wind speed information

which will be transferred to optimal rotational speed and pitch angle control commands for the WTPGS. On the other hand, high effective controllers for the generator should be proposed to make that the responses of the generator as fast as possible. On the grid side, robust and efficient global controllers should be employed in the power systems in particular consideration of high degree of nonlinearities, uncertainties and interconnections of SG and WTPGS.

In variable-speed WTPGS controls, the turbine shaft rotational speed should be controlled optimally with respect to the variable wind speed in order to achieve MPPT. Such turbine speed command should base on the real-time information of wind speed. Normally, wind speed anemometers are employed for the wind speed estimation. However, high cost of precise anemometer limits the extensive usage of this equipment. A precise real-time wind speed estimation method and sensorless control for variable-speed variable-pitch WTPGS are proposed in this work. The wind speed estimation is realized by an Extreme Learning Machine based nonlinear input-output mapping neural network. A specific design characteristic of the wind turbine is used for improving the mapping accuracy with considering the variable pitch angle. The proposed wind speed estimation is established by using the information of the wind mechanical torque and the turbine rotational speed which are estimated by a two-loop linear observer. The estimated wind speed is then used to determine the optimal rotational speed and pitch angle commands.

In the wind generator control case, in order to obtain high control performance and system robustness, the advanced controller must achieve fast dynamic response during the transient state and robustness against parametric variations and unstructured dynamics. An Optimal Reset Control ORC scheme based on linear principle is proposed for the generator's control. Theoretically, it has been proved that the ORC design problems are equivalent to LQR problems and the ORL can be obtained by solving algebraic Riccati equations. The major advantages of the ORC method over the traditional control methods are as follows: 1) to reduce the overshoot of the step response without degrading other specifications; 2) to suppress controller's saturation effectively; 3) to accelerate the tracking speed with less error. With these advantages, the ORC has also been

used in the design of uncertainty observer. Due to the fast convergence of the ORC uncertainty observers, feed-forward compensations of uncertainty terms can be employed online in the ORC control loops to eliminate the uncertainty effects. The ORC control scheme can achieve fast dynamic response during the transient process and robustness against uncertainties in the generators' controls.

The proposed wind speed estimation, sensorless control and generator control techniques have been tested in a WTPGS installed with a PMSG. From the power point of view, the PMSG can be regarded as an inverse PMSM in the negative power direction. In the control of generators, the proposed ORC has been applied in WTPGS MPPT control to enhance the power capture capability of the PMSG WTPGS, in the PMSM current controls with ORC uncertainty observers to improve the current control performance and system robustness and in the PMSM position controls to realize robust and high-performance position responses of PMSM, respectively.

For the controls of the grid side, eyes are mainly focused on the control designs for power system transient stability enhancement with voltage regulations. Modern power systems are large scale and complex. They consist of a collection of several independent and spatially distributed dynamic subsystems with grid-connected wind power plants interconnected together. Due to that the transient stability and voltage regulation are both important properties of power system controls, the controllers were always designed separately to deal with the two problems in the previous literatures. However, the conventional controllers may not dominate the global controls of a power system with grid-connected WTPGS and cannot prompt the cooperation of the SG and the WTPGS which may negatively affect the global stability.

In this research, a robust back-stepping control scheme is proposed for the SG which can both achieve transient control and voltage regulation within one control structure. Based on the nominal DFL power systems, the conventional robust voltage regulators can easily be designed. Although the post-fault voltage regulation problems are realized by such voltage controllers, they are only effective around the normal operating point. When a large disturbance occurs, the power systems may not be survived by the voltage controllers. Thus, back-stepping laws

are employed to define variable electrical power output objectives of the SG which can correct the actions of the conventional voltage controllers from the power point of view. Since the power angle and frequency controls of SG are decoupled from the power and terminal voltage controls with the back-stepping controls, one control structure for both the transient stabilizing control and voltage regulation is realized.

Moreover, the proposed robust back-stepping control scheme provides a method of coordination and control of the original power system and the WTPGS. The power deviation between the electrical power output objective of the SG and its real output power can be offered as the reference for the wind turbine captured power. When the power system is subjected to a severe disturbance, the rotational speed and pitch angle commands can be given by the power deviation through the mapping characteristic of the wind turbine in the transient process which can dominate the WTPGS cooperating with the global system and helping to enhance the transient stability and voltage regulation capacity of the global system. The proposed back-stepping control scheme is thus successfully extended to uncertain power systems to coordinate different control objectives of different power sources within wide operating regions. The proposed controller design technique has been tested on a SMIB power system with grid-connected PMSG WPTGS in order to gain a better understanding into the dynamics and control of the compensated system. The applications of the proposed controllers to a multi-machine power system with grid-connected PMSG WPTGS to enhance the transient stability with voltage regulation have also been considered.

8.2 Recommendations for Further Research

In the proposed discrete-time ORC, there are several interesting topics for future research. One is the analysis and design of reset control systems with input saturation. Another one is the baseline controller and reset law co-optimization.

In ELM based wind speed estimation and sensorless control of WTPGS, it shows that the response of the ELM pitch controller is much faster than the conventional PI controller. The proposed pitch control can eliminate the big

power extraction oscillations when the wind speed changes fast. However, the extracted power of ELM pitch controlled system sometimes has small fluctuations around the rated power which means that the ELM pitch controller results in more static error than the conventional PI controller. It may cause by the inevitable small wind speed estimation error. Alternatively, a correction law for the pitch control action against the inevitable wind speed estimation error can be explored in the future work to eliminate the static error and improve the pitch control performance further.

In the proposed power system backstepping control scheme designs, the weight function of the power angle in the defined error variable can be optimized. Although this parameter within a certain range will not affect the stability of the control power system, but the optimization of this function can be expected to further improve the backstepping control performance. In future work, we can make a more optimal form of the weight function and summarize a theoretical design approach of the function.

In application of backstepping control scheme on the multi-machine power system with PCC installed WTPGS, according to the simulation results, it comes to the conclusion that there are still some areas need to be improved in the future work for the proposed backstepping control scheme,

1. Actually, the WTPGS mentioned in this thesis expresses a wind farm which is composed of a large number of wind turbine power generation system sharing the same capacity and parameters. Considering that all the wind turbines in the wind farm are identical is still a simplification. The discrepancies of each wind turbine power generation system in a big wind farm and their respective controls will be taken into account in the future work.
2. The coordination control for a non-conventional power system which includes several wind farms distributed in different regions is another topic. There is a challenge of how to allocate the duty for each wind farm.
3. The performance improvements are always more obvious for Generator 1. It is because that the WTPGS is installed at PCC bus 8 which shares the same

PCC bus with Generator 1. In other words, the proposed backstepping based control for WTPGS will more positively affect the power system control performances locally. In the future work, the proposed backstepping based control law for WTPGS will be optimized with considering the non-conventional power system structure, power generation source distributions, distances between every two generation sources and transmission systems. It aims to improve the backstepping control performances globally.

4. The reactive power output of WTPGS isn't considered. The reactive power flow control is another issue apart from rotor angle stability issue. However, as you know, the PMSG and DFIG wind energy conversion systems have the abilities of adjustments of their output reactive powers. Since they can be expected to play an important role in the grid voltage control, the reactive power flow control for WTPGS will have to be considered in the future work in order to accomplish the defined entire advanced control for large scale non-conventional power systems.

Parameters of PMS machines

A.1 PMS machines Parameters

In Chapter 4, the parameters of the PMS machine systems are given as,

The total stator resistance and inductance with filter: $R = 1.5\Omega$, $L = 4.5mH$.

The magnet flux: $\psi = 0.08333T$.

The moment of inertia of the lumped system: $J = 1.24 \times 10^{-4}kg \cdot m^2$.

The damping constant of turbine: $D = 1.08 \times 10^{-4}N \cdot m \cdot s/rad$.

The pole pair number: $p = 4$.

A.2 PMS Machine Position Control Parameters

In Section 4.2, the proposed ORC position control scheme for PMSM is tested in both the simulations and experiments. The performances comparisons between the proposed method and an advanced sliding mode control (I-SMC) are given with some analysis.

The parameters of the proposed ORC position control scheme are given as,

The inner-loop PI parameters: $k_p = 3265.796$ and $k_i = 457534.8042$.

The outer-loop PID parameters: $k_{pt} = 13.64$, $k_{it} = 763.56$ and $k_{dt} = 0.1051$.

The observer parameters: $k_{pe} = -0.1115$ and $k_{ie} = -25.11$.

The parameters of the advanced I-SMC employed for comparison are given as,

The I-SMC parameters: $c_{p1} = 900$, $c_{p2} = 250000$, $c_{p3} = 5000$ and $\beta = 3$.

The boundary layer: $\Phi = 2$.

A.3 PMS Machine Current Control Parameters

In Section 4.3, the proposed joint ORC current control scheme for PMSM is tested in both the simulations and experiments. The performance comparisons between the conventional decoupling PI control scheme, advanced SMC and the proposed method are given with analysis.

The parameters of the proposed joint ORC current control scheme are given as,

The ORC current controllers parameters: $k_p = 0.1115$ and $k_i = 25.11$.

The ORC observers parameters: $k_{po} = 0.1115$ and $k_{io} = 25.11$.

The ORC observers kalman filter gains: $L_q = 67.283$ and $L_d = 67.283$.

The parameters of the advanced sliding mode control scheme employed for comparison are given as,

The SMC parameters: $c_q = 4.3$, $c_d = 4.3$, $m_q = 1.2$ and $m_d = 1.2$.

The SMC observers parameters: $\eta_q = 0.001$ and $\eta_d = 0.001$.

The boundary layer: $\Phi = 3$.

Parameters of Wind Turbine Power Generation System

B.1 Parameters of PMSG WTPGS for MPPT Control

In Section 5.2, the parameters of the wind turbine power generation system in design of MPPT control are given as,

The turbine blade radius: $R = 0.5m$.

The air density: $\rho = 1.25kg/m^3$.

The gearbox Ratio: $n_g = 1$.

The PMSG pole pair number: $p = 4$.

The damping constant of turbine: $D_g = 1.08 \times 10^{-4}$.

The moment of inertia of the lumped system: $J_g = 1.24 \times 10^{-4}kg \cdot m^2$.

The total stator resistance and inductance with filter: $R_s = 1.5\Omega$, $L_s = 4.5mH$.

The magnet flux: $\psi = 0.0833T$.

The wind turbine power coefficient parameters: $a_0 = 0.0052$, $a_1 = -1.109 \times 10^{-3}$, $a_2 = 6.909 \times 10^{-3}$, $a_3 = -8.315 \times 10^{-4}$, $a_4 = -5.58 \times 10^{-5}$, $a_5 = 1.111 \times 10^{-5}$, $a_6 = -3.873 \times 10^{-7}$.

The maximum power coefficient: $C_{pmax} = 0.4063$.

The optimal tip speed ratio: $\lambda_{opt} = 6.9$.

The rectifier inner-loop PI parameters: $k_p = 20.3$ and $k_i = 8.6$.

The rectifier outer-loop MPPT PI parameters: $k_{pw} = 0.0788$ and $k_{iw} = 25.11$.

B.2 Parameters of WTPGS for Wind Speed Estimation and Sensorless Maximum Power Control

In Section 5.3, the parameters of the wind turbine power generation system in design of wind speed estimator and sensorless maximum power control are given as,

The WTPGS for simulations: turbine blade radius $R = 2m$; damping constant of turbine $D_g = 2 \times 10^{-2}$; moment of inertia of the lumped system $J_g = 1.08 \times 10^{-2} kg \cdot m^2$.

The WTPGS for experiments: turbine blade radius $R = 1m$; damping constant of turbine $D_g = 1 \times 10^{-4}$; moment of inertia of the lumped system $J_g = 1.2 \times 10^{-4} kg \cdot m^2$.

The maximum power coefficient: $C_{pmax} = 0.4412$.

The optimal tip speed ratio: $\lambda_{opt} = 6.9077$.

The normal air density for estimator design: $\rho = 1.25 kg/m^3$.

The pitch rate limit: $6^\circ/s$.

The pitch system time constant: $T_{servo} = 0.05s$.

B.3 Parameters of Grid-connected WTPGS

In Chapter 7, the parameters of the grid-connected wind turbine power generation system are given as,

The wind turbine and generator parameters: $C_{pmax} = 0.4412$, $\lambda_{opt} = 6.9077$,
 $R = 30m$, $J_t = 108kg.m^2$.

The air density: $\rho = 1.25kg/m^3$.

The pitch system time constant: $T_{servo} = 0.05s$.

The pitch rate limit: $10^\circ/s$.

The wind speed during transient period: $V = 10m/s$ (maintains a constant).

The WTPGS different operating points in each system are shown in tables in the simulation sections of each section, respectively.

Parameters of Power Systems

C.1 Parameters of SMIB Power System

In Chapter 6 and Section 7.1, the SMIB power system basic parameters are given as,

The synchronous generator: $x_d = 1.863p.u.$, $x'_d = 0.257p.u.$, $x_T = 0.127p.u.$,
 $T'_{d0} = 6.9s$, $x_L = 0.4853p.u.$, $H = 4s$, $D = 5p.u.$, $k_c = 1p.u.$, $x_{ad} = 1.712p.u.$.

The SMIB power system different operating points in each chapter are shown as tables in the simulation sections of each chapter, respectively.

C.2 Parameters of Three-machine Nine-bus Power System

In Chapter 6 and Section 7.2, the three-machine nine-bus power system basic parameters are given as,

The synchronous generator 1: $x_{d1} = 0.8958p.u.$, $x'_{d1} = 0.1198p.u.$, $T'_{d01} = 6.0s$, $H_1 = 6.40s$, $D_1 = 1p.u.$, $k_{c1} = 1p.u.$, $x_{ad1} = 1.712p.u.$.

The synchronous generator 2: $x_{d2} = 1.1325p.u.$, $x'_{d2} = 0.1813p.u.$, $T'_{d02} = 5.89s$, $H_2 = 3.01s$, $D_2 = 1p.u.$, $k_{c2} = 1p.u.$, $x_{ad2} = 1.712p.u.$.

The synchronous generator 3 (reference generator): $x_{d3} = 0.1460p.u.$, $x'_{d3} = 0.6080p.u.$, $T'_{d03} = \infty$, $H_3 = \infty$.

The transformer impedances: $x_{T1} = 0.0625p.u.$ between bus 2 – 8; $x_{T2} = 0.0586p.u.$ between bus 3 – 6; $x_{T3} = 0.0576p.u.$ between bus 1 – 4.

The bus impedances: 4 – 9: $0.010 + j0.085p.u.$; 4 – 5: $0.017 + j0.092p.u.$; 8 – 9: $0.032 + j0.161p.u.$; 5 – 6: $0.039 + j0.170p.u.$; 7 – 8: $0.0085 + j0.072p.u.$; 6 – 7: $0.0119 + j0.1008p.u.$.

The shunt admittances: 4–0: $j0.1670p.u.$; 8–0: $j0.2275p.u.$; 6–0: $j0.2835p.u.$.

The load admittances: A : $0.9106 + j0.3187p.u.$; B : $0.7812 + j0.1953p.u.$; C : $0.6637 + j0.2655p.u.$.

The power system different operating points in each chapter are shown as tables in the simulation sections of each chapter, respectively.

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Author's Publications

1. S. Wu, Y. Y. Wang and S. J. Cheng, "Optimal reset control design for current control and uncertainties estimation in permanent magnet synchronous motor drive systems," *Electric Power Applications, IET*, volume: 6, issue: 2, Feb. 2012, page(s): 122-132.
2. S. Wu, Y. Y. Wang and S. J. Cheng, "Extreme learning machine based wind speed estimation and sensorless control for wind turbine power generation system," *Neurocomputing*. (to be published)
3. S. Wu, Y. Y. Wang, Y. Q. Guo and S. J. Cheng, "Discrete-time optimal reset control for time-varying reference system and its application on PMSG wind generation system," *Control Systems Technology, IEEE Transactions on*. (Revised and resubmitted)
4. S. Wu, Y. Y. Wang and S. J. Cheng, "Backstepping control for power systems with grid-connected wind turbine power generation system," *Sustainable Energy, IEEE Transactions on*. (Revised and resubmitted)
5. S. Wu, Y. Y. Wang, X. L. Zhu and S. J. Cheng, "Power systems backstepping control for transient stability enhancement and voltage regulation," *Control Systems Technology, IEEE Transactions on*. (Revised and resubmitted)
6. S. Wu, Y. Y. Wang and S. J. Cheng, "Advanced position control design based on linear theory for permanent magnet synchronous motor drive sys-

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8. D. Q. Dang, S. Wu, Y. Y. Wang and W. J. Cai, “Model predictive control for maximum power capture of variable speed wind turbines,” *IPEC, 2010, Conference Proceedings*, pp.274-279.