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Stability of Power Electronic System

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School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in partial fulfillment of the requirement for the degree of
Master of Engineering

2020

Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original research, is free of plagiarised materials, and has not been submitted for a higher degree to any other University or Institution.

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Supervisor Declaration Statement

I have reviewed the content and presentation style of this thesis and declare it is free of plagiarism and of sufficient grammatical clarity to be examined. To the best of my knowledge, the research and writing are those of the candidate except as acknowledged in the Author Attribution Statement. I confirm that the investigations were conducted in accord with the ethics policies and integrity standards of Nanyang Technological University and that the research data are presented honestly and without prejudice.

29-03-2020

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Authorship Attribution Statement

This thesis contains material from 2 papers from paper accepted at conferences in which I am listed as an author and paper published in the following peer-reviewed journal.

Chapter 3 and chapter 4 is published as L. Miao, J. Z. Xin and B. Zhao, "Revisit Nyquist-Bode Stability Criteria for Power Electronic System with Non-minimum Phase System," 2019 IEEE 4th International Future Energy Electronics Conference (IFEEEC), Singapore, Singapore, 2019, pp. 1-6.

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The contributions of the co-authors are as follows:

- Prof Zhang Xin provided the initial project direction and help me with the structure of the paper.
- I wrote the paper. The manuscript of the paper was revised by Prof Zhang Xin.
- I performed all the laboratory work at Clean Energy Research Lab of the School of Electrical and Electronic Engineering of Nanyang Technological University. I also analyzed the data.
- Group member Zhao assisted in the real time simulation of models using OPAL-RT.

Chapter 5 is published as Z. Zeng, X. Zhang, F. Blaabjerg and L. Miao, "Impedance-Oriented Transient Instability Modeling of SiC MOSFET Intruded by Measurement Probes," in IEEE Transactions on Power Electronics.

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The contributions of the co-authors are as follows:

- Dr Zeng wrote the drafts of the manuscript. The manuscript was revised together with Prof Zhang Xin and me.
- I collected all the probes information from companies and drew the dot diagram for analyzing the relationship of the parasitic of probes.

- Dr. Zeng built the model needed and I assisted in simplify those circuit for later analysis.
- Dr. Zeng and I did the stability analysis according to the poles map in case study.

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Abstract

Power electronic system is widely used and plays great important role in different industries, so the stability of power electronic system becomes a vital problem. To make sure that the system keeps stable, the stability criteria should be used to do the stability analysis of power electronic systems. Usually, the stability analysis of power electronic system can be divided in three parts: switch test, converter and converters. The challenges in the three parts are as follows: In the converter part, the single converter is analysed. The inaccurate parameters of components may affect the stability, and if the converter system is a non-minimum phase system or with integrator in the system, the traditional stability criteria are not useful for it; For converters part, the system which is combined by several converters is discussed, such as the multiple parallel converters and grid-connected inverter with weak grid. Similar to the converter, parameters can cause the instability, and non-minimum phase system cannot use traditional stability criteria to judge the stability; In switch test part, semiconductor material, such as SiC MOSFET, the stability is sensitive to the measurement probes parasitic. The stability evaluation of the probe-device system considering the parasitic of probes is important and is also the challenge. This thesis reviews the Nyquist stability criterion for stability analysis of power electronic systems which contain non-minimum phase system and system with integrator. Stable and unstable cases of boost converter and grid-connected inverter with weak grid are as examples for converter and converters stability analysis, respectively. Impedance models of SiC MOSFET and probes are given to analyse the impact of parasitic and do stability evaluation. All the stability analyses are verified by real-time simulation results.

Key words- Stability analysis, modelling, Nyquist stability criterion, power electronic system

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Chapter 1 Introduction

1.1 Motivation

With power electronic devices are widely used in many different fields, from battery chargers, audio amplifiers, manufacturing machines, to power transmission systems in the interconnected electrical grids, the stability of the power electronic systems becomes a particularly important factor.

However, before a power electronic system put into application, many things need to be checked to make sure the power electronic system is stable. However, there are many factors in the power electronic system which can make the system unstable. These factors can be included in three parts: converter, converters, and switch test part, as shown in Fig. 1.1.

Figure 1.1 Three parts of stability analysis of power electronic system

In the converter part, single converter is the object that should be analysed in this part, for example, the buck converter and boost converter, as shown in the Fig. 1.2, (a) is the circuit model of buck converter, and (b) shows the circuit of boost converter.

To make sure the stability of the converter, all the components of converter should be taken into consideration. If the components mismatch the converter system, the inaccurate parameters of converter components can easily bring the instability problem to the system [1].

(a)

(b)

Figure 1.2 Converter circuit (a) buck converter (b) boost converter

To judge the system is stable or not, stability criteria are commonly used tools. However, the problems are that the converter system may be a non-minimum phase system, which means it may have right-half plane poles or zeros. Also, with controller in the system [2], the transfer function of the converter sometimes has integrator component, so that the Nyquist curve of converter will become an unclosed curve at 0Hz.

With these problems, some traditional stability criteria, like gain margin and phase margin method, which are only applicable in minimum phase system, will be inappropriate. In fact, not all the converters with RHP poles or zeros are unstable. The way to judge the stability of the converter just by RHP poles and gain margin phase margin is conservative and not rigorous. In conclusion, to do a more accurate analysis of the power electronic system, Nyquist stability criterion which considering the RHP poles for non-minimum phase power electronic systems and considering the integrators in the system is needed [3].

In the converters part, the system which is combined by several converters is discussed, such as the multiple parallel converters and grid-connected inverter with weak grid, for the weak grid can be seen as a converter in the small-signal model.

To analyse the stability problem, the impedance-based stability analysis method is commonly used [4]. It can help simplify the derivation procedure of the whole system transfer function and make the analysis easier.

Similar to the converter, inaccurate parameters of the components can also make the converters unstable. During the stability analysis of converters, the potential problem is non-minimum phase system which should be analysed by Nyquist stability criterion rather than traditional stability method. Other stability analysis methods, like forbidden region method, including the Middlebrook criterion [5] – [6], the Gain Margin Phase Margin (GMPM) criterion [7], the opposing argument criterion (OPAC) [8], the energy source analysis consortium criterion (ESAC) [9] – [10] and the root exponential stability criterion (RESC) [11], play great role in the stability analysis of power electronic system, but these stability criteria based on the forbidden region have some limitations so that in some cases, they are not general enough to make sure their judgements on the stability of the converters are accurate.

In the switch test part, semiconductor is widely used in the power electronic system as a switch, for its great advantages, such as high speed in switching [12] – [15]. However, the fast switching speed of some semiconductor is a double-edged sword. On the one hand, it can reduce the switching loss. On the other hand, when using semiconductor in the devices, its stability can be easily affected by the parasitic components of the system [16] – [18], which may affect the transient stability of the semiconductor, and worse, make great damage to the system.

The parasitic-dependent stability of switch can be influenced by parasitic components in device, circuit, and probes. In device part, the parasitic inductances of the package [19] – [20] and snubber capacitor [21] – [22] are analysed to reduce the impact of the parasitic components. In circuit part, multiple degrees of freedom of the gate driver [23] and the ingenious active gate drivers [24] – [26] are used to enhance the transient stability of switch. In the probe part, the parasitic components of probes can cause the propagation delay that has been proved can affect the accurate measurement because of its fast switching speed [27] – [28].

To deal with the problems caused by the parasitic parameters of probes, some innovative probes [29] – [33] and general methodologies [34] – [37] have been proposed to reduce the impact of probes. However, the parasitic components introduced by the measurement probes are usually ignored in the industrial field. The on-off trajectories of semiconductor switches caused by the parasitic components of probes are also ignored. The fact is, besides the parameters of device and circuit, the parameters of measurement probes can also have great impact on the stability of the system, such as amplify the measurement error, or even intrude the transient

instability of the device. The barriers need to be overcome in this part includes how to model the switch and probes considering the parasitic and how to do stability evaluation of the device-probe system.

In summary, the challenges of the stability analysis of power electronic system are showing next:

- Non-minimum phase system of converter and converters cannot be analyzed by traditional stability criteria.
- Nyquist curve of system whose loop gain contains $1/s$ is not a closed curve.
- Model the switches and probes considering parasitic components.
- Stability evaluation of switch-probe system considering the impact of probe parasitic components.

1.2 Objectives

For power electronic systems, the stability problem should be concerned throughout the period from designing to application. This thesis aims to reveal these stability problems which may exist in the whole period.

When analysing the stability problem of converter and converters, the stability criteria are widely used.

In converter part, to analyse the stability of the system with non-minimum phase and integrator, the Nyquist stability criterion is used. Boost converter case is used as a typical example for the converter part.

In converters part, the challenge is also the non-minimum phase system stability analysis. In this part, a grid-connected inverter with weak grid is taken as a typical converters case and the Nyquist stability criterion is applied to the converters to analyse the stability.

In switch test part, taking SiC MOSFET, high impedance passive voltage probe, differential voltage probe and current probe as example to reveal the transient instability of switch caused by the parasitic components in measurement probes. Impedance model of SiC MOSFET and probes and stability evaluation method are given. Cases about the stability of switch with different parasitic parameters of probes and SiC MOSFET are shown.

1.3 Major contribution of the Thesis

This thesis tends to reveal the stability problem of the power electronic systems from the switch to connect to the weak grid. In the thesis, the major contributions can be summary includes:

Reviewing the stability criteria for stability analysis, including forbidden region criteria which is suitable for source-load system, and Nyquist stability criterion which is a general stability criterion for stability analysis of power electronic systems. When reviewing the Nyquist stability criterion, it contains two parts: Firstly, for non-minimum phase system, using the relationship between the number right half plane poles of open-loop transfer function and the Nyquist plot of open-loop transfer function to evaluate the stability. Secondly, the supplementary curve at 0Hz is shown for the loop gain of power electronic system contains integrators.

Then in the converter and converters part of the stability analysis, boost converter and grid connected inverter with weak grid are used as examples, respectively. The stability analysis using some existing stability criteria of power electronic system can be inaccurate, because of the non-minimum phase system and with integrators. Nyquist stability criterion to evaluate the stability of power electronic systems is more accurate and general.

At last, in the switch test part of the stability analysis of power electronic system, the impact of probes is concerned. The parasitic components of probes should be considered in switch test period which can lead to the measurement inaccuracy, or even transient instability of switch rather than ignoring them [18]. SiC MOSFET, differential voltage probe, high impedance passive voltage probe and Rogowski coil are used to verify the disturbance the parasitic components of probes can bring the SiC MOSFET.

1.4 Organization of the Thesis

This thesis will be divided into 6 parts.

Introduction of the thesis is given in Chapter 1. The challenges that exist in the stability analysis of power electronic systems are discussed from three parts: converter, converters and switch test.

Stability criteria, especially the Nyquist stability criterion are revisited in Chapter 2, which is applicable for non-minimum phase system and system with integrator. The

chapter also includes the gain margin phase margin method, the method to count the encirclement of critical point and supplementary method for system contains $1/s^v$, and the forbidden region criteria.

Stable and unstable cases of boost converter are discussed in Chapter 3 as typical cases of converter part. Nyquist stability criterion is used to analyse the stability and deal with the challenges in boost converter like non-minimum phase system and $1/s$ component.

Stable and unstable cases of grid-connected inverter with weak grid are given in Chapter 4 as typical cases of converters part. Also, the Nyquist stability criterion is used to analyse the stability of grid-connected inverter with weak grid system.

The transient instability of SiC MOSFET caused by probes is discussed in Chapter 5. Impedance model of probes and SiC MOSFET considering parasitic components are given to do stability evaluation. Different parasitic of probes and SiC MOSFET have impact on the system stability are shown in the case study.

Finally, the conclusion of the thesis and future work are given in Chapter 6.

Chapter 2 Literature review

2.1 Basic concept of Nyquist diagram

A typical closed-loop power electronic system is given in Fig. 2.1. The closed-loop transfer function of the system is $\frac{G(s)}{1+G(s)H(s)}$. Generally, $G(s)H(s)$ can be written as the format shown in (1):

$$G(s)H(s) = \frac{KN(s)}{s^v D(s)} \quad (1)$$

The roots of the equation $KN(s)=0$ are the zeros of $G(s)H(s)$, and the roots of $s^v D(s) = 0$ are the poles of $G(s)H(s)$. According to the traditional Nyquist criterion, if the real part of all the poles is negative, the system is open-loop stable.

Figure 2.1 Typical closed-loop system

Figure 2.2 shows Nyquist contour on the complex plane, ϵ and R both are the radius of the semicircles. When $v=0$ in (1), which means there is no integrator in the system, as shown in Fig. 2.2 (a), the Nyquist contour along the imaginary axis and then encloses the entire right half plane. For the system with integrators, as shown in Fig. 2.2 (b), poles at the origin should be taken into consideration, so ϵ infinitely close to 0. Usually, $-j\epsilon$ and $j\epsilon$ are equal to $j0^-$ and $j0^+$, respectively. When $j\omega$ rotates along the Nyquist contour on the complex plane clockwise, from $(0, -j\infty)$, and then along the negative imaginary axis to the point $(0, -j\epsilon)$, along the small semicircle to point $(0, j\epsilon)$, along the positive imaginary axis to $(0, j\infty)$, and then rotates clockwise through 180 degrees to coincide with $(0, -j\infty)$, as shown in Fig. 2.2 (b).

When rotates clockwise one turn along the Nyquist contour on the complex plane, plugging every single value along the imaginary axis and all the point along the infinity in the RHP in $G(s)H(s)$, the obtained closed curve on the complex plane is the Nyquist plot of $G(s)H(s)$ [3].

Figure 2.2 Nyquist contour (a) system without integrator (b) system with integrator

2.2 Nyquist plot application: gain margin & phase margin

To get the gain margin and phase margin from the Nyquist plot, as shown in Fig. 2.3 (a), given a $G(j\omega)H(j\omega)$ locus which intersects the real axis and the unit circle at phase crossover (p_c) and gain crossover (g_c). From the phase crossover and gain crossover, gain G_A and phase ϕ_{gc} can be calculated:

$$G_A = |G(j\omega)H(j\omega)| \quad (2)$$

$$\phi_{gc} = \angle G(j\omega)H(j\omega)|_{\omega=\omega_{gc}} \quad (3)$$

So the gain margin K_g and phase margin γ can be expressed as:

$$K_g = 20 \log\left(\frac{1}{G_A}\right) = 20 \log\left(\frac{1}{|G(j\omega)H(j\omega)|_{\omega=\omega_{pc}}}\right) \quad (4)$$

$$\gamma = 180^\circ + \phi_{gc} \quad (5)$$

(a)

(b)

Figure 2.3 Three typical locus (a) in Nyquist diagram and (b) in bode diagram

Fig. 2.3 (a) shows the three lines in the Nyquist diagram. The solid line shows the stable case, according to equation (4) and (5), both the phase margin and gain margin of the solid line are greater than 0. Increasing the gain of the solid line, we can get the two dash lines. The zero line also means an unstable system, whose phase margin and gain margin are both equal to 0. If keeping increasing the gain, it will become the dash-dot line. The phase margin and gain margin are less than 0, so the system is an unstable case. Gain margin and phase margin of the three cases are directly shown in

the bode diagram in Fig. 2.3 (b). The system is stable only when the value of phase margin and gain margin are both greater than 0.

2.3 Nyquist stability criterion

To judge the stability of power electronic systems, Nyquist stability criterion is a commonly used criterion which is applicable for both minimum-phase system and non-minimum phase system. given a closed-loop system, let P be the number of poles on the right-half complex plane of the open-loop transfer function, and Z be the number of poles on the right-half complex plane of the closed-loop transfer function. When rotates clockwise one cycle along the Nyquist contour, set N be the number of times the $G(s)H(s)$ curve encloses $(-1, j0)$ clockwise. The Nyquist stability criterion can be described as: N should be equal to the difference between Z and P , as shown in (6)

$$N = Z - P \quad (6)$$

The closed-loop system is stable if and only if $Z=0$, so $-N=P$.

2.3.1 Characteristic of right-half plane poles and zeros in bode diagram

For simple system, the characteristic of right-half plane poles and zeros can be directly got from the bode plots, as shown in the next table.

In the table, it can be seen that if the system contains one right-half plane real pole, the magnitude curve of bode plot will decrease with 20 dB/dec at high frequency, and at the same time, the phase curve of bode plot will increase 90 degrees from low frequency to high frequency.

On the contrary, if the system contains one right half plane real zero, its magnitude curve is also equal to the 0dB at low frequency but will increases with 20 dB/dec at high frequency. The phase curve of the of the bode plot of the right half plane zero will decreases 90 degrees from low frequency to high frequency.

From the table, if the system contains right half plane conjugate poles, the trend of the magnitude curve and phase curve of the bode plot is same to the right half plane real pole, but they also have some difference. The speed of the decrease of the magnitude curve at high frequency becomes 40 dB/dec, and from low frequency to high frequency, the phase curve will increase 180 degrees instead of 90 degrees. Also, at the vicinity of angular corner frequency f_c , as shown in the figure, its magnitude curve may have a sudden rising oscillation. The shape of the magnitude curve is

related to the quality factor Q , the larger the quality factor is, the larger the oscillation will be, and the value of the quality factor is decided by the parameters of the system. When $Q > 0.5$, the roots of the system become complex.

Similarly, when the system contains right half plane conjugate zeros, the trend of the magnitude curve and phase curve is same to the bode plot of the system with right half plane real zero, and the speed of the increase of the magnitude curve at high frequency is 40 dB/dec, the phase curve from low frequency to high frequency becomes 180 degrees. Like the bode plot of system with right half plane conjugate poles, the magnitude curve also has a falling oscillation at the vicinity of the angular corner frequency f_c , which is associated with the quality factor Q .

Table 1 Characteristic of right half plane poles and zeros in bode plot

Types	Characteristic in bode plots
Right-half plane real poles	<p>The magnitude plot shows a flat line at 0 dB/dec until the corner frequency f_c, then a roll-off at -20 dB/dec. The phase plot shows a step change from 0 to -90 degrees at f_c.</p>
Right-half plane real zeros	<p>The magnitude plot shows a flat line at 0 dB/dec until the corner frequency f_c, then a rise at +20 dB/dec. The phase plot shows a step change from 0 to +90 degrees at f_c.</p>

2.3.2 Times of encirclement

To count the number of times the open-loop function encloses $(-1, j0)$, N , it may be difficult to count in the Nyquist diagram if when the Nyquist plot is complex and hard to read. Under the circumstance, as shown in Fig. 2.4 (a), it can be simplified by counting the number of crossings on the left of the point $(-1, j0)$ of the real negative half-axis in the Nyquist diagram.

In the Nyquist diagram, define the crossing when the phase is decrease as clockwise crossing (CC) and the crossing when the phase is increase as anti-clockwise crossing (ACC), as shown in Fig. 2.4 (a).

Figure 2.4 Type of crossing (a) in Nyquist diagram (b) in bode diagram

Then extending the definition of crossings from the Nyquist diagram to the bode diagram, the crossings counted in the Nyquist diagram have the characteristics that in the bode diagram, the phase of these crossings should equal to -180° and their amplitude should larger than 0 dB. As the result, as Fig. 2.4 (b) shows, that only the crossings where its amplitude is greater than 0 dB will be counted. In the bode diagram, CC is the crossing whose phase decreases and ACC is the crossing whose phase increases.

Figure 2.5 Type of crossing in impedance-based stability method

For the system with two or more subsystems, to count the number of the encirclements, it can also be simplified by counting the number of crossings. As shown in the Fig. 2.5, for example, the system has two subsystems, and their equivalent impedance are Z_1 and Z_2 respectively.

In the bode diagram with equivalent impedance Z_1 and Z_2 , move the phase-frequency characteristic curves of Z_1 up and down by 180° respectively, and then the obtained

upper and lower curves, as the green dot lines in the Fig. 2.5, are the crossing boundaries.

Similar to the analysis in the Fig. 2.4 that the bode diagram only has one curve in magnitude-frequency and phase-frequency plot, the region where the crossings should be counted in is that where the amplitude of the equivalent impedance Z_1 is greater than Z_2 , and it can be called the exterior region. In the exterior region, if the phase-frequency characteristic curve of Z_2 crosses the boundaries upward, as the blue dot in the Fig. 2.5, the crossing is CC, and on the contrary, if the phase-frequency curve crosses the boundaries downward, so the crossing is ACC.

After counting the number of clockwise crossings and anti-clockwise crossings, both the two types of bode diagram, the number of N can be described as:

$$N = N_{CC} - N_{ACC} \quad (7)$$

Where N_{CC} is the number of clockwise crossing, N_{ACC} is the number of anti-clockwise crossing.

2.3.3 Supplementary feature at 0 Hz for system loop gain with $1/s^v$

Nyquist plot is an essential and commonly used tool when analysing the stability of power electronic systems. Generally, the Nyquist diagram we plot should be a smooth closed curve. However, in the system with integrators, after adding the curve at negative frequency, the Nyquist plot is not closed. At 0 Hz, the value of transfer function when $\omega = 0+$ and $0-$ are different.

As shown in Fig. 2.6, when $\omega = 0-$, the Nyquist curve is infinitely close to ∞ and when $\omega = 0+$, the curve becomes infinitely close to $-\infty$.

To solve the problem, there must have a supplementary line at 0 Hz in order to get a smooth, closed Nyquist curve.

$$T(s) = \frac{1}{s^v} \cdot \frac{(\tau_{z,n}s + 1) \dots (\tau_{z,j+1}s + 1)}{(\tau_{p,m}s + 1) \dots (\tau_{p,i+1}s + 1)} \quad (8)$$

$$= \frac{\left(\frac{1}{\omega_{z,j}^2}s^2 + \frac{2\xi_{z,j}}{\omega_{z,j}}s + 1\right) \dots \left(\frac{1}{\omega_{z,1}^2}s^2 + \frac{2\xi_{z,1}}{\omega_{z,1}}s + 1\right)}{\left(\frac{1}{\omega_{p,i}^2}s^2 + \frac{2\xi_{p,i}}{\omega_{p,i}}s + 1\right) \dots \left(\frac{1}{\omega_{p,1}^2}s^2 + \frac{2\xi_{p,1}}{\omega_{p,1}}s + 1\right)}$$

Figure 2.6 Nyquist plot of system with integrator

Given the system transfer function, as shown in (8), if the system has v integrators, when s moves from $(0, j0^-)$ to $(\epsilon, 0)$ to $(0, j0^+)$ in Fig. 2.2, the corresponding phase of the Nyquist plot of $T(s)$ will change from $v\pi/2$ to $-v\pi/2$, which means the integrators give the Nyquist diagram a supplementary clockwise curve of $v\pi$ rad.

Figure 2.7 Supplementary curve of Nyquist diagram when loop gain has $1/s$ (a) Nyquist plot of $\omega=0+$ close to $-\infty$, $\omega=0-$ close to $+\infty$ (b) Nyquist plot of $\omega=0+$ close to $+\infty$, $\omega=0-$ close to $-\infty$

The dashed line in Fig. 2.7 shows the supplementary feature of Nyquist diagrams when $v=1$, so the transfer function has $1/s$, the supplementary curve from $\omega=0-$ to $\omega=0+$ is a semicircle, the radius is infinite. From Fig. 2.7 (a), the Nyquist plot of $\omega=0-$ is close to ∞ and $\omega=0+$ is close to $-\infty$, the supplementary curve from $0-$ to $0+$ crosses the positive real axis once, so that it does not have effect on the number of N . However, when the Nyquist plot of $\omega=0-$ is close to $-\infty$ and $\omega=0+$ is close to $+\infty$, the supplementary curve crosses the negative real axis once, the number of N_{CC} should plus one.

$$v = 2$$

Figure 2.8 Supplementary curve of Nyquist diagram when loop gain has $1/s^2$ (a) Nyquist plot of $\omega=0+$ above real axis, $\omega=0-$ below real axis (b) Nyquist plot of $\omega=0+$ below real axis, $\omega=0-$ above real axis

When $v=2$, the supplementary curve from $\omega=0-$ to $\omega=0+$ will be a circle, as shown in Fig. 2.8. In Fig. 2.8 (a), the Nyquist plot of $\omega=0+$ is above the real axis and $\omega=0-$ below the real axis, the supplementary curve crosses the negative real axis left the point $(-1, j0)$ twice, so the number of N_{CC} should plus 2. In Fig. 2.8 (b), the Nyquist plot of $\omega=0+$ is below the real axis and $\omega=0-$ above the real axis, the supplementary curve from $0-$ to $0+$ just crosses the positive real axis once, which does not change the number of N .

Figure 2.9 Supplementary curve of Nyquist diagram when loop gain has $1/s^v$, v is odd number (a) Nyquist plot of $\omega=0+$ close to $-\infty$, $\omega=0-$ close to $+\infty$ (b) Nyquist plot of $\omega=0+$ close to $+\infty$, $\omega=0-$ close to $-\infty$

And so on, when v is an odd number, as shown in Fig. 2.9, the values of $\omega=0-$ and $\omega=0+$ are close to the imaginary axis. The number of crossings on the left of $(-1, j0)$ is $(n-1)/2$ if the value of $\omega=0-$ is ∞ and the value of $\omega=0+$ is $-\infty$, so the number N_{CC} should plus $(n-1)/2$. If the value of $\omega=0-$ is $-\infty$ and the value of $\omega=0+$ is ∞ , N_{CC} should plus $(n+1)/2$.

When v is an even number, as shown in Fig. 2.10, the values of $\omega=0^-$ and $\omega=0^+$ are close to the real axis, and the number of crossings on the left of $(-1, j0)$ also depends on the values of $\omega=0^-$ and $\omega=0^+$. When Nyquist plot of $\omega=0^-$ and $\omega=0^+$ are on the negative real axis, the supplementary crossings will be $n/2+1$ if the Nyquist plot of $\omega=0^+$ above negative real axis and $\omega=0^-$ below negative real axis, or the crossings will be $n/2-1$ when Nyquist plot of $\omega=0^+$ above negative real axis and $\omega=0^-$ below negative real axis. When the real part of values of $\omega=0^-$ and $\omega=0^+$ both are $+\infty$, the number of crossings is $n/2$, as shown in Fig. 2.10 (c) (d), so N_{CC} need to plus $n/2$.

Figure 2.10 Supplementary curve of Nyquist diagram when loop gain has $1/s^v$, v is even number (a) Nyquist plot of $\omega=0^+$ above negative real axis, $\omega=0^-$ below negative real axis (b) Nyquist plot of $\omega=0^+$ below negative real axis, $\omega=0^-$ above negative real axis (c) Nyquist plot of $\omega=0^+$ above positive real axis, $\omega=0^-$ below positive real axis (d) Nyquist plot of $\omega=0^+$ below positive real axis, $\omega=0^-$ above positive real axis

2.4 Stability criteria based on forbidden region

For stability analysis of converters based on impedance-ratio method, the stability analysis methods include method based on the Nyquist stability criterion and method based on the forbidden region.

Figure 2.11 General source load system

As the Fig. 2.11 shows, to analyse the stability of the system, the impedance ratio $\frac{Z_1}{Z_2}$ can be used for more simple analysis.

As shown in Fig. 2.12, method based on forbidden region including Middlebrook criterion [5] – [6], Gain Margin Phase Margin (GMPM) criterion [7], opposing argument criterion (OPAC) [8], Energy Source Analysis Consortium (ESAC) criterion [9] – [10], root exponential stability criterion (RESC) [11], etc.

Figure 2.12 Forbidden region criteria

As shown in Fig. 2.12, the boundary of the forbidden region of the Middlebrook criterion is a circle. For the impedance ratio $\frac{Z_1}{Z_2}$, it means the forbidden region of the Middlebrook criterion is the region where the magnitude of the impedance ratio $\frac{Z_1}{Z_2}$ is larger than gain shown in the Fig. 2.12, here Gain is equal to the $1/gm$, and gm is the gain margin in the real coordinate, it can be derived as (9)

$$gm = 10^{\frac{K_g}{20}} \quad (9)$$

Where K_g is the gain margin (dB).

If the inverse of gain margin of impedance ratio lies inside the boundary circle of the forbidden region, the system will be considered as a stable system. From the forbidden region of the Middlebrook criterion, it can be seen that the Middlebrook forbidden region considers all the gain margin of the impedance ratio but does not take the phase margin into consideration. The advantage is that the system which satisfy the condition, it will be guaranteed stable. However, the problem is that the criterion is sufficient to analyse the stability, but the restrictions are too harsh for the systems, sometimes the system lies in the forbidden region does not means it is an unstable system.

As shown in the Fig. 2.12, for the impedance ratio $\frac{Z_1}{Z_2}$, the forbidden region of the gain margin and phase margin (GMPM) criterion is that the amplitude of impedance ratio is larger than $1/gm$, and the angle is between $-180^\circ + pm$ and $180^\circ - pm$, where pm is the phase margin of the impedance ratio.

To make sure the system is stable, the impedance ratio should not lie in the forbidden region of the gain margin and phase margin criterion.

Different to the Middlebrook criterion, gain margin and phase margin criterion not only considers the gain margin of the system, but also considers the phase margin, which decreases the conservativeness the forbidden region of Middlebrook criterion brings to the stability analysis.

From the Fig. 2.12, the blue line is the boundary of the forbidden region of the opposing argument criterion, which is a straight line parallel to the imaginary axis and intersects the real axis at the point $(-1/gm, 0)$. The forbidden region is the left of the vertical line. As the result, the relationship of the forbidden region can be got from it are

$$20\lg|Z_1| - 20\lg|Z_2| > -K_g \quad (10)$$

$$\arg(Z_1) - \arg(Z_2) < -90^\circ - \varphi \quad (11)$$

$$\arg(Z_1) - \arg(Z_2) > 90^\circ + \varphi$$

Where

$$\varphi = \arcsin\left(\frac{1}{gm} \frac{Z_2}{Z_1}\right) \quad (12)$$

Different to the criteria above, which are usually used to deal with the stability of the system with single load, the opposing argument criterion can solve the problem that the system with several loads. The impedance ratio $\frac{Z_1}{Z_2}$ for several loads system can be written in the format that

$$\frac{Z_1}{Z_2} = \frac{Z_1}{Z_{2,1}} + \frac{Z_1}{Z_{2,2}} + \dots + \frac{Z_1}{Z_{2,n}} \quad (13)$$

Figure 2.13 Source subsystem with parallel loads

The total impedance ratio can be divided into n individual impedance ratios if the system contains n loads. In practical cases, sometimes it is easier to analyse the individual impedance ratio instead of the total impedance Z_2 , so a criterion is needed

which can make sure the whole system stable by analysing every individual impedance ratio.

The opposing argument criterion is proposed to deal with the problem. It considers the gain margin and phase margin of the system, which at the same time, decreases the conservativeness of the Middlebrook criterion.

The forbidden region of the energy source analysis consortium (ESAC) criterion is the red line shows in the Fig. 2.12. The boundaries of the forbidden region are two-line segments which are parallel to the negative real axis and intersect at the boundary of the unit circle. The two lines are connected by another two lines that start at the intersection point of the unit circle and the parallel boundaries and then converge at the point $(-1/gm, 0)$.

We can see that, comparing with the Middlebrook criterion, gain margin and phase margin criterion, and the opposing argument criterion, the forbidden region of the ESAC is much smaller, so the stability analysis using the energy source analysis consortium criterion may be more accurate, especially for the stable cases.

The root exponential stability criterion (RESC) can be seen as a special case of the energy source analysis consortium criterion. From the Fig. 2.12, the forbidden region of the RESC is similar to the forbidden region of the ESAC, the difference is that the boundary of the RESC is continuously differentiable function, while the ESAC uses linear function. In conclusion, comparing with the ESAC, the RESC is more robust in the definition.

The summaries of the several forbidden region criteria are given in the next table, the difference and the forbidden region can be seen directly.

Table 2 Characteristic of forbidden region criteria

Method		Criterion
Forbidden region	Middlebrook	The amplitude of Z_1/Z_2 within the circle $r=1/gm$
	GMPM (gain margin & phase margin)	Outside the forbidden region which is $ Z_1/Z_2 > 1/gm$ $\angle Z_1/Z_2 < -180^\circ + pm$ $\angle Z_1/Z_2 > 180^\circ - pm$

	OPAC (opposing argument criterion)	The Nyquist plot is required to fall to the right of a line at $s = 1/gm$
	ESAC (the energy source analysis consortium criterion)	The forbidden region is formed by two-line segments which are parallel to the negative real axis and intersect at the boundary of the unit circle. These segments are connected by another two lines which start at the intersection points and converge at the point $(-1/gm, 0)$.
	RESC (root exponential stability criterion)	Represented by a continuous function; more robust than the ESAC

According to the forbidden region method, the system will be stable if the impedance ratio of the system is confined out of the forbidden region.

However, the stability analysis of impedance ratio based on the forbidden region has many limitations.

Firstly, the constraints of the forbidden region criterion are too harsh to make the judgement of the stability. Even the energy source analysis consortium criterion, as we can see from the Fig. 2.12, has the smallest forbidden region among all the given forbidden region criteria, its judgement of the stability is also conservative, and as the result, make the stable system become unstable.

Secondly, the premise of these forbidden region is that the direction of the power flow should be given, under the circumstance, if the converters change the load subsystem and the source subsystem, the forbidden region criteria will not be suitable for the stability analysis of the cases.

Chapter 3 Stability analysis of boost converter

3.1 Modelling of boost converter

Buck and boost converters are commonly used in power electronic system with the rapid development of renewable energy [1]. Taking boost converter as example, the operation principle of boost converter is given next [38].

(a)

(b)

(c)

Figure 3.1 Operation principle of boost converter (a) boost converter circuit (b) switch on (c) switch off

Figure 3.1 shows the circuit of the boost converter. When the switch is on, as shown in Fig. 3.1 (b), the current only through the inductor and switch, the inductor accumulates the power from voltage source V_{in} . At the same time, the capacitor C acts as a source, discharging the power to the load. When the switch is off, as shown in Fig. 3.1 (c), the power from voltage source V_{in} and inductor accumulated when

switch on will pass from the diode to supply to the load and capacitor, so that the output voltage of boost converter is larger than the input voltage.

When talking about the stability problem of the boost converter, many stability criteria will come into mind. Generally, before using these stability criteria, the transfer function of the converter should be derived, as most of these criteria need transfer function to judge the stability.

The simple converter, like buck converter and boost converter, is not difficult to get its transfer function. Taking the boost converter as example.

According to the Fig. 3.1, in one switching period, the inductance current and capacitance voltage can be derived as (14) and (15)

$$v_L = dv_{in} + d'(v_{in} - v_o) = v_{in} - d'v_o \quad (14)$$

$$i_C = d \frac{-v_o}{R} + d' \left(i_L - \frac{v_o}{R} \right) = d' i_L - \frac{v_o}{R} \quad (15)$$

Where d is the duty ratio, $d + d' = 1$, v_{in} is the input voltage, v_o is output voltage, i_C is the current of the capacitance, v_L is the voltage of inductance, R is the load.

Then considering the character of the inductance and capacitance, the current of capacitance and the voltage of inductance can be written as (16) and (17) respectively

$$i_C = C \frac{dv_o}{dt} = d' i_L - \frac{v_o}{R} \quad (16)$$

$$v_L = L \frac{di_L}{dt} = v_{in} - d' v_o \quad (17)$$

To linearize the boost system, small-signal model of boost converter can be used. Assuming the duty ratio d is equal to a DC quiescent value plus a small AC variation, as shown

$$d = D + \Delta d \quad (18)$$

Similarly, the output voltage and the current of inductance can be written as the corresponding DC quiescent value plus small variation as (19) and (20)

$$i_L = I_L + \Delta i_L \quad (19)$$

$$v_o = V_o + \Delta v_o \quad (20)$$

Adding the small-signal equations to the (16) and (17), it can be got as

$$i_c = C \frac{d(V_o + \Delta v_o)}{dt} = (1 - D - \Delta d)(I_L + \Delta i_L) - \frac{V_o + \Delta v_o}{R} \quad (21)$$

$$v_L = L \frac{d(I_L + \Delta i_L)}{dt} = V_{in} - (1 - D - \Delta d)(V_o + \Delta v_o) \quad (22)$$

Because the differential equations of quiescent value V_o , I_L , D are equal to 0, and the value of the small variation is too small to ignore them, so the current of capacitance and the inductance voltage can be simplified as

$$C \frac{d(\Delta v_o)}{dt} = (1 - D)I_L - \frac{V_o}{R} + (1 - D)\Delta i_L - \Delta d I_L - \frac{\Delta v_o}{R} \quad (23)$$

$$L \frac{d(\Delta i_L)}{dt} = V_{in} - (1 - D)V_o + \Delta d V_o - (1 - D)\Delta v_o \quad (24)$$

DC terms of the above equations should be equal to 0, so

$$(1 - D)I_L - \frac{V_o}{R} = 0 \quad (25)$$

$$V_{in} - (1 - D)V_o = 0 \quad (26)$$

So the output voltage and inductance current can be calculated as (27)-(28)

$$V_o = \frac{V_{in}}{1 - D} \quad (27)$$

$$I_L = \frac{V_o}{(1 - D)R} \quad (28)$$

Where D is the duty ratio, V_{in} is input voltage, V_o is the output voltage, R is load, I_L is average inductor current.

The block diagram of the boost converter is shown in Fig. 3.2.

Figure 3.2 Boost converter block diagram

According to the small-signal model of boost converter, after removing the DC terms and small second-order AC terms in the equations (23) and (24), they can be

$$C \frac{d(\Delta v_o)}{dt} = (1 - D)\Delta i_L - \Delta d I_L - \frac{\Delta v_o}{R} \quad (29)$$

$$L \frac{d(\Delta i_L)}{dt} = \Delta d V_o - (1 - D)\Delta v_o \quad (30)$$

Then after the Laplace transformation of the above the equations, they are transformed to

$$sCv_o(s) = (1 - D)i_L(s) - I_L d(s) - \frac{v_o(s)}{R} \quad (31)$$

$$sLi_L(s) = V_o d(s) - (1 - D)v_o(s) \quad (32)$$

So the control-to-output transfer function $G_{vd}(s)$ from $v_o(s)$ to $d(s)$ and gain loop $T(s)$ can be derived as:

$$G_{vd}(s) = \frac{V_o}{1 - D} \cdot \frac{(1 - D)^2 R - sL}{(1 - D)^2 R + sL + s^2 LCR} \quad (33)$$

$$T(s) = G_c(s)G_{vd}(s) \quad (34)$$

$G_c(s)$ in block diagram is the transfer function of the controller. A good controller is intended to maximize the low frequency gain for elimination of steady-state tracking error and rejection of external disturbances, minimize the high frequency gain for noise filtering and maximize the crossover frequency f_c for fast dynamic response. For simple boost converter, $G_c(s)$ is usually an integrator, to eliminate the steady-state tracking error.

$$G_c(s) = G_b \frac{\omega_c}{s} \quad (35)$$

Where

$$\omega_c = 2\pi f_c \quad (36)$$

$$|G_b|_{dB} = -|G_{vd}(j2\pi f_c)|_{dB} \quad (37)$$

If use PI controller in the converter, to choose the suitable K_p and K_i , the transfer function of G_c would become

$$G_c(s) = \frac{G_b}{K} \frac{\omega_c}{s} \frac{(1 + \frac{s}{2\pi f_z})}{(1 + \frac{s}{2\pi f_p})} \quad (38)$$

$$K = \sqrt{\frac{1 + \sin \varphi_b}{1 - \sin \varphi_b}} \quad (39)$$

$$f_p = K f_c \quad (40)$$

$$f_z = \frac{f_c}{K} \quad (41)$$

K depends on the phase boost φ_b needed, which the maximum is 90.

Buck converter circuit and its transfer function derivation is similar to the boost converter.

Before using the Nyquist stability criterion to analyse the stability of boost or buck converter, and according to the bode diagram, Nyquist diagram and pole-zero map to judge the system is non-minimum phase system or minimum phase system, what should be highlighted is that the output voltage is the system output for the boost and buck converter when non-minimum phase theory is applied in. If the output voltage of the non-minimum phase system controls the current of inductor, the system may become minimum phase system [39].

Next part, the stable and unstable cases of simple boost converter will be given, using the Nyquist stability criterion to do the stability analysis.

3.2 Case study

3.2.1 Stable case

For the simple converter systems, a boost converter is used as an example.

Table 3 Parameters of stable boost converter

Parameters	Values	Parameters	Values
V_{in}	30V	K_p, K_i	0.0001, 10
L	0.1mH	R	4Ω
C, R_C	1μF, 10Ω	V_{ref}	60V

A stable case of converter system is given next. Tab. 3 gives parameters of stable boost converter system, including the PI controller in the system. The bode plot, Nyquist plot, and pole-zero map are shown next.

$$T(s) = (k_p s + k_i) \cdot \frac{V_o}{1 - D} \cdot \frac{(1 - D)^2 R - sL}{(1 - D)^2 R + sL + s^2 LCR} \cdot \frac{1}{s} \quad (42)$$

Considering the PI controller of the boost converter, the open-loop-gain $T(s)$ is shown in the (42). As the equation shows, the transfer function has the $1/s$ part, so according to the analysis in chapter 2, at 0 Hz, the Nyquist curve of the system will not be a closed circle. As the result, a semicircle supplementary curve is needed.

As shown in the Nyquist diagram of the boost converter in Fig. 3.3, when ω is equal to 0-, the Nyquist curve is infinitely close to the $+\infty$. On the contrary, when w is equal to 0+, the value of the Nyquist curve becomes $-\infty$. The value of Nyquist curve at 0+

and 0^- are quite the opposite, so the supplementary curve is of the essence in order to get a closed Nyquist plot.

After adding the supplementary curve at 0 Hz, as the blue dot line shows, the Nyquist plot does not encircle the critical point $(-1, j0)$, so the number of encirclements of the point $(-1, j0)$ is still 0.

Figure 3.3 Nyquist diagram of loop gain $T(s)$ of stable boost converter

In Fig. 3.4, the bode diagram of the boost converter is given.

When the magnitude is larger than 0, the phase curve does not cross -180° . The analysis in the bode diagram also gets the conclusion that the number of times the loop gain curve encloses $(-1, j0)$ clockwise is 0, $N=0$, which is the same to the analysis in Nyquist diagram.

Figure 3.4 Bode diagram of loop gain $T(s)$ of stable boost converter

The pole zero map of the boost converter is shown in Fig. 3.5.

From it we can see that one zero can be seen in the right half plane. Because of the right half plane zero, the boost converter is a non-minimum phase system. It does not have a right half plane pole, so $P=0$.

Based on the analysis on Nyquist diagram, bode diagram and pole zero map, $N=-P=0$, as a result, the boost converter is a stable converter.

Figure 3.5 Pole-zero map of loop gain $T(s)$ of stable boost converter

Fig.3.6 is the output waveforms of the boost converter after the real-time simulation using OPAL-RT.

Though some small fluctuation is still existing in the waveforms, the oscillation period is equal to the switching cycle time. The system is stable, the result is same as the above analysis.

Figure 3.6 Experimental waveforms for stable boost converter

3.2.2 Unstable case

Parameters of the unstable boost converter case are given in the Tab. 4.

Table 4 Parameters of unstable boost converter

Parameters	Values	Parameters	Values
V_{in}	30V	K_p, K_i	1, 1
L	0.1mH	R	4 Ω
C	2.8mF	V_{ref}	60V

The Nyquist diagram, bode diagram, and pole-zero map of the boost converter are given. The open-loop transfer function is the same as the stable case shown in (42), still has $1/s$ in the transfer function.

From the Nyquist diagram, when $\omega=0^-$, the Nyquist curve is infinitely close to the $+\infty$ and on the contrary, when $\omega=0^+$, the value of the Nyquist curve becomes $-\infty$, so the Nyquist diagram also needs a semicircle supplementary curve when $\omega=0$.

As shown in Fig. 3.7, the blue dash line is the supplementary curve at 0 Hz. It can be seen from the zoom in figure, that the closed Nyquist curve encloses the point $(-1, j0)$ two times clockwise, so the type of crossing is clockwise crossing and N_{CC} is equal to 2. According to the equation (6), $N = N_{CC} - N_{ACC} = 2$.

Figure 3.7 Nyquist diagram of loop gain $T(s)$ of unstable boost converter

Fig. 3.8 is the bode diagram of the unstable boost converter.

Figure 3.8 Bode diagram of loop gain $T(s)$ of unstable boost converter

From the bode diagram, when magnitude is larger than 0, the phase curve crosses the -180° once with phase decrease, so the type of crossing is also clockwise crossing. If considering the negative frequency, the number of crossing is also 2 clockwise crossings, same to the analysis in the Nyquist diagram.

In Fig. 3.9, the pole-zero map of the boost converter is given. One zero is on the right half plane, the boost converter is a non-minimum phase system. However, there are still not RHP poles exist, so $P=0$.

It can be concluded that $N=2 \neq -P$, the boost converter is unstable.

Figure 3.9 Pole-zero map of loop gain $T(s)$ of unstable boost converter

The real-time simulation result is given in next figure. The output waveforms of the unstable boost converter are shown in Fig. 3.10. The supposed output voltage v_o is a constant value. However, the actual output voltage, as well as the output current, have fluctuations. The cycle time of the fluctuations in Fig. 3.10 is 3.6ms, which is much larger than the switching cycle time 0.05ms set in the model, so the output waveforms also indicate the boost converter is unstable.

Figure 3.10 Experimental waveforms for unstable boost converter

The converter part concentrate on the stability of single converter, like the buck and boost converter in this chapter. However, after analysing the stability of single converter and making sure they are stable, the system contains two or more converters also need to do the stability analysis as a whole system. In the next chapter, the converters system is analysed.

Chapter 4 Stability analysis of Grid-connected inverter with weak grid

4.1 State of arts of stability analysis methods of converters

In converters, according to the impedance-based stability criterion, the impedance ratio of two subsystems can be used as the equivalent loop gain to analyse the stability of converters.

The single grid-connected inverter can be simplified as a current source parallel with a resistor [4]. The equivalent circuit model of a grid-connected inverter with weak grid is shown in Fig. 4.1.

Figure 4.1 Circuit model of current source cascaded system

Before using impedance-based stability criterion in the system, as shown in Fig. 4.2, there are three assumptions:

- (1) The source converter is stable when unloaded.
- (2) The source converter is stable when loaded with an ideal voltage load.
- (3) The load converter is stable when supplied by an ideal current source.

(a)

(b)

(c)

Figure 4.2 Three assumptions for impedance-based stability criterion (a) source converter is stable when unloaded (b) source converter is stable with an ideal voltage load (c) load converter is stable with an ideal current source

According to assumption 1, as shown in Fig. 4.2 (a), we can get that

$$i_o = i_{inv} \quad (43)$$

$$v_{pcc} = 0 \quad (44)$$

So the current source i_{inv} and v_{pcc} are stable.

In Fig. 4.2 (b), according to the assumption 2, it can be derived that

$$v_{pcc} = v_g \quad (45)$$

$$i_o = i_{inv} - \frac{v_g}{Z_{inv}} \quad (46)$$

From the equations above, v_g and $1/Z_{inv}$ is stable.

Then according to the assumption 3, as shown in Fig. 4.2 (c),

$$v_{pcc} = i_o Z_g + v_g \quad (47)$$

$$i_o = i_{inv} \quad (48)$$

So Z_g is stable.

Based on the above assumptions, the output current i_o and voltage v_{pcc} can be derived as the expressions following:

$$i_o = \left(i_{inv} - \frac{v_g}{Z_{inv}} \right) \cdot \frac{1}{1 + \frac{Z_g}{Z_{inv}}} \quad (49)$$

$$v_{pcc} = (v_g + i_{inv}Z_g) \cdot \frac{1}{1 + \frac{Z_g}{Z_{inv}}} \quad (50)$$

Since i_{inv} , $1/Z_{inv}$, v_g and Z_g are all stable, and the right half part of equations (49) and (50) can be seen as the transfer function of closed-loop system, the stable of the output current and voltage can be analysed by the impedance ratio Z_g and Z_{inv} : If the impedance ratio $\frac{Z_g}{Z_{inv}}$ satisfy the Nyquist stability criterion, the system is stable.

4.2 Modelling of grid-connected inverter with weak grid

As shown in Fig. 4.3, (a) is the circuit of a grid-connected inverter with weak grid, and (b) gives the block diagram of the grid-connected inverter. Here closed-loop control system is used for its more advantages, for example, eliminating the deviation, improving the system response and the ability to suppress interference, closed-loop control system is usually used in the single-phase inverter.

(a)

(b)

Figure 4.3 Grid-connected inverter (a) circuit model (b) block diagram

To control the four switches of inverter as shown in Fig. 4.3 (a), a triangle generator and a sine wave generator are needed in the model. The structure of inverter is shown in Fig. 4.3 (a). While the sine wave is higher than the triangular wave, S_1 and S_4 will on, the output voltage is $+V$. Otherwise, S_2 and S_3 will open, and the output voltage become $-V$. Figure 4.4 shows the output waveform of bipolar PWM.

The output waveform is square wave, so in order to get the standard sine wave, LCL type filter is still needed to filter out harmonics. The parameters of inductors and capacitor can be changed to obtain the sine wave we want.

Figure 4.4 Waveforms of bipolar PWM

The stability criterion used to analyse the stability of the grid-connected inverter with weak grid is usually the impedance-based method, to get the impedance ratio of the converters system, the equivalent impedance of the inverter model and grid model are needed respectively.

From Fig. 4.3 (a), the current i_{inv} , i_o and the voltage v_c can be easily depicted as:

$$i_{inv}(s) = \frac{u_f(s) - u_c(s)}{sL_1 + R_1} \quad (51)$$

$$u_c(s) = (i_{inv}(s) - i_o(s)) \cdot \left(\frac{1}{sC_f} + R_c\right) \quad (52)$$

$$i_o(s) = \frac{u_c(s) - u_{pcc}(s)}{sL_2 + R_2} \quad (53)$$

Then from Fig. 4.3 (a) and Fig. 4.1, the equivalent impedance of the grid part Z_g is equal to L_g .

To derive the inverter impedance Z_{inv} , a block diagram of the inverter is given in Fig. 4.3 (b). The admittance Y_o , output current to input voltage transfer function G_{iu} and current loop control transfer function G_c can be obtained in (54)-(56)

$$Y_o(s) = -\frac{i_o}{v_{pcc}} \Big|_{v_f=0} \quad (54)$$

$$= \frac{R_1 + sL_1 + R_c + \frac{1}{sC_f}}{(R_1 + sL_1) \cdot (R_2 + sL_2) + (R_1 + sL_1) \cdot (R_c + \frac{1}{sC_f}) + (R_2 + sL_2) \cdot (R_c + \frac{1}{sC_f})}$$

$$G_{iu}(s) = \frac{i_o}{v_f} \Big|_{v_{pcc}=0} \quad (55)$$

$$= \frac{R_c + \frac{1}{sC_f}}{(R_1 + sL_1) \cdot (R_2 + sL_2) + (R_1 + sL_1) \cdot (R_c + \frac{1}{sC_f}) + (R_2 + sL_2) \cdot (R_c + \frac{1}{sC_f})}$$

$$G_c(s) = PI(s) \cdot \frac{V_{dc}}{V_{tri}} = (K_p + K_i \frac{1}{s}) \cdot \frac{V_{dc}}{V_{tri}} \quad (56)$$

The equivalent impedance of inverter Z_{inv} can be derived as:

$$Z_{inv}(s) = \frac{1 + G_c(s) \cdot G_{iu}(s) \cdot H_s(s)}{Y_o} \quad (57)$$

The stable and unstable case of grid-connected inverter with weak grid system will be given in the following part. Nyquist stability criterion will be used to analyse the stability.

4.3 Case study

4.3.1 Stable case

To analyse the non-minimum phase converters system, a single-phase inverter connected to weak grid model is used as an example. Here the sampling time is 2×10^{-5} , and system switching frequency is 20kHz.

The parameters of stable grid-connected inverter with weak grid case are in Table 5.

Table 5 Stable grid-connected inverter with weak grid parameters

Parameters	Values	Parameters	Values
V_{dc}	405V	K_p, K_i	40, 10000
R_1, L_1	1Ω, 8mH	H_s	0.55
R_2, L_2	0.1Ω, 5mH	L_g	6.5mH
R_C, C_f	15Ω, 30μF	V_{grms}	220V

According to the analysis before, the equivalent impedance of inverter Z_{inv} and equivalent impedance of grid Z_g can be derived as

$$Z_{inv}(s) = \frac{(R_1 + sL_1) \cdot (R_2 + sL_2) + (R_1 + sL_1) \cdot (R_C + \frac{1}{sC_f})}{R_1 + sL_1 + R_C + \frac{1}{sC_f}} + \frac{(R_2 + sL_2) \cdot (R_C + \frac{1}{sC_f})}{R_1 + sL_1 + R_C + \frac{1}{sC_f}} \quad (58)$$

$$+ \frac{(K_p + K_i \frac{1}{s}) \cdot (R_C + \frac{1}{sC_f}) \cdot \frac{V_{dc}}{V_{tri}} \cdot H_s(s)}{R_1 + sL_1 + R_C + \frac{1}{sC_f}}$$

$$Z_g(s) = L_g(s) \quad (59)$$

The Nyquist diagram, bode diagram, and pole-zero map of the converters system are given.

The Nyquist diagram of the impedance ratio Z_g/Z_{inv} is shown in Fig. 4.5. From the zoomed in diagram, it can be seen that the Nyquist curve does not encircle the point (1, j0), so the number of N is zero, N=0.

Figure 4.5 Nyquist diagram of stable grid-connected inverter with weak grid

Bode diagram of the converters system is shown in Fig. 4.6.

The exterior region shown in the figure is the region where the magnitude of Z_g is larger than Z_{inv} , only the crosses in the exterior region are counted as the number of the crossings. In the exterior region, the phase curve of Z_{inv} does not cross the dashed line, which is line that the phase curve of Z_g moves down 180° . It means there is no crossing that exists in the bode diagram, $N=0$, the result is same to the analysis in the Nyquist diagram.

Figure 4.6 Bode diagram of stable grid-connected inverter with weak grid

Fig. 4.7 is the pole-zero map of grid-connected inverter with weak grid system.

According to the pole-zero map, there is no poles or zeros in the right half plane, the number of P is equal to 0.

As a result, the number of the encirclements is equal to the RHP poles, $N=-P=0$, the converters system is a stable system.

Figure 4.7 Pole-zero map of stable grid-connected inverter with weak grid

Fig. 4.8 is the real-time simulation result, shows the voltage and current waveforms of the inverter system. As the figure shows, the output voltage and current are stable sine wave, indicate the grid-connected inverter with weak grid is stable, confirming the analysis of Fig. 4.5.

Figure 4.8 Experimental waveforms of stable grid-connected inverter with weak grid

4.3.2 Unstable case

The parameters of unstable grid-connected inverter and weak grid are given in Table 6. The system switching frequency and sampling frequency is same to the stable case.

Table 6 Unstable grid-connected inverter with weak grid parameters

Parameters	Values	Parameters	Values
V_{dc}	405V	K_p, K_i	28, 60000
R_1, L_1	0.01Ω, 7.5mH	H_s	0.55
R_2, L_2	0.01Ω, 5mH	L_g	7.5mH
R_C, C_f	5Ω, 10μF	V_{grms}	220V

To do the stability analysis, Nyquist diagram, bode diagram, and pole-zero map of Z_g and Z_{inv} or the impedance ratio Z_g/Z_{inv} are given next.

It can be seen from the Nyquist diagram shown in Fig. 4.9 that the Nyquist curve does not enclose the point $(-1, j0)$, so the number $N=0$.

Figure 4.9 Nyquist diagram of unstable grid-connected inverter with weak grid

Fig. 4.10 is the bode diagram of converters, in the exterior region, there still not crossings in the diagram. The analysis result is the same as the analysis in the Nyquist diagram, $N=0$.

From the pole-zero map in Fig. 4.11, a pair of conjugate poles and one zero points can be observed of Z_g/Z_{inv} , the system is non-minimum phase system, and the number of RHP poles is 2.

Figure 4.10 Bode diagram of unstable grid-connected inverter with weak grid

As a result, the number of N is not equal to P , $N=0 \neq -P$.

Figure 4.11 Pole-zero map of unstable grid-connected inverter with weak grid

The real-time simulation experimental result is given in Fig. 4.12. The output waveforms are the voltage and current at PCC, the voltage of capacitance and the inverter current. It can be seen that these waveforms have great fluctuations, the system is unstable, which is the same as the analysis above.

Figure 4.12 Experimental waveforms for inverter connected to grid

This chapter analyse the stability of converters using the Nyquist stability criterion and using the grid-connected inverter with weak grid as the example. However, in the power electronic system, besides the converter components, the stability of switch components using semiconductor material is also important to analyse. In the next chapter, the parasitic components of probe which can affect the stability of switch of converters are analysed.

Chapter 5 Transient instability modeling

of SiC MOSFET intruded by probes

Semiconductor material now is widely used in power electronic systems as the switch material, among them, SiC MOSFET is an advanced power semiconductor material in different industrial fields, for its great advantages, such as high frequency, withstand high temperature, and as well as the high voltage capabilities.

Due to the fast switching speed, SiC MOSFET is sensitive to the parasitic, such as the parasitic components of device, circuit and measurement probes. The parasitic components of device, circuit and probes is vital to the stability of SiC MOSFET, and even to the whole system. If improper parasitic parameters are chosen in the power electronic system, because of the sensitiveness of SiC MOSFET to the parasitic components, it may lead to damage to the transient stability.

Among all the effect by the parasitic components of power electronic devices, circuit and measurement probes, parasitic components of measurement probes are always ignored. However, experiences show that the SiC MOSFET may be lose the transient instability or even damaged if measurement probes and its parasitic parameters are used improperly.

The experimental results of SiC MOSFET transient waveforms are shown in Fig. 5.1, Fig. 5.1 (a) is the stable model, with SiC MOSFET model using the high bandwidth high impedance passive voltage (HIPV) probe and with the short ground lead. On the contrary, the unstable mode is shown in Fig. 5.1 (b), which uses low bandwidth, high impedance passive voltage probe and with the long ground lead. It can be seen that the different probes with different parasitic can affect the test results of SiC MOSFET. If using improper probes, as the Fig. 5.1 (b) shows, the measurement of SiC MOSFET will be inaccurate.

(a)

(b)

Figure 5.1 SiC MOSFET transient waveforms. (a) stable mode with high bandwidth HIPV probe and short ground lead (b) unstable mode with low bandwidth HIPV probe and long ground lead

Besides the high impedance passive voltage probes, the differential voltage (DV) probe and current probe, such as the Rogowski coil, also have parasitic components in the measurement probes, and are widely used in the measurement of power electronic systems, which can affect the transient stability of SiC MOSFET as well. As the result, high impedance passive voltage probe, differential voltage probe and current probe are all taken into consideration when analyse the transient instability caused by the parasitic components of probes of SiC MOSFET.

To find out the impact of the parasitic components of probes on the transient stability of SiC MOSFET, from the impedance model of the probes and SiC MOSFET, there are still some barriers need to be overcome:

Barrier 1: The impedance model of SiC MOSFET and measurement probes considering the parasitic components.

Barrier 2: The model of the interaction between SiC MOSFET and probes and the way to do the stability analysis and reveal the impact of the parasitic components.

5.1 Impedance-oriented modeling

Fig. 5.2 is the circuit model of the SiC MOSFET system measured by passive voltage probe, differential voltage probe, and current probe.

As shown in Fig. 5.2, the gate voltage of S_1 , v_g , the drain-source voltage v_{ds} , and the drain current i_d are measured by the high impedance passive voltage probe, the differential probe and Rogowski coil respectively.

From the Fig. 5.2, it is difficult to do the stability evaluation of the whole system, to simplify the stability analysis, the impedance-oriented modelling of SiC MOSFET and the three types of probes will be analysed respectively, and then combine them together for further stability analysis.

Besides, by analysing some typical probes of some leading companies, such as Tektronix, Keysight and Lecroy, the relationship between their bandwidth, impedance, input capacitance and price will be figured out, which is good for later chosen of probes depending on different situations.

Figure 5.2 SiC MOSFET with different measurement probes

5.1.1 SiC MOSFET

Usually the impedance model of SiC MOSFET is given in the Fig. 5.3. On the left of the impedance model is the picture of the real SiC MOSFET, corresponding to the impedance model, G is the gate terminal, D means the drain terminal and S is the source terminal of SiC MOSFET.

Figure 5.3 Impedance model of SiC MOSFET

From the impedance model, we can see that the SiC MOSFET is modelled as voltage-controlled current source. In the impedance model of SiC MOSFET, i_d is the drain current, v_{gs} is gate-source voltage of SiC MOSFET, the expression of i_d and v_{gs} can be described as

$$i_d = \begin{cases} 0 & 0 \leq v_{gs} < V_{TH} \\ g_m(v_{gs} - V_{TH}) & V_{TH} \leq v_{gs} \leq V_{MI} \end{cases} \quad (60)$$

$$v_{gs} = \begin{cases} G_{gs}(s)v_G & \text{if } v_{gs} < V_{MI} \text{ or } v_{ds} \geq V_{on} \\ V_{MI} & \text{else} \end{cases} \quad (61)$$

In the equations of i_d and v_{gs} , the g_m represents the transconductance, V_{MI} is the voltage of Miller plateau and V_{TH} means the threshold voltage [40], [41]. In the equation (61), $V_{on} = R_{ds,on}I_d$ means the on-state voltage of the drain-source voltage v_{ds} .

From the above two equations, the transfer function $G_{gs}(s)$ can be derived as

$$G_{gs}(s) = \frac{v_{gs}}{v_G} = \frac{1}{(L_g + L_{cs})C_{gs}s^2 + (R_{ge} + R_G + R_{cs})C_{gs}s + 1} \quad (62)$$

5.1.2 High-impedance passive voltage (HIPV) probe

By collecting the information of some typical high impedance passive voltage probes from leading companies, such as Tektronix, Keysight [42] – [45], as shown in Fig. 5.4, the relationship between bandwidth, input capacitance, and price of the high impedance passive voltage probes can be recognized intuitively.

After linear transformation, the general tendencies of the parasitic components of HIPV probes are also given in Fig. 5.4. It can be found that the with bandwidth of HIPV probe increasing, its input capacitance decreases at the same time. On the contrary, as shown in the second figure, the price of high impedance passive voltage probes, different to the input capacitance, in proportion to the bandwidth.

Figure 5.4 Relationship diagram of surveyed HIPV probes (a) input capacitance and bandwidth (b) price and bandwidth

The commonly used impedance model of high impedance passive voltage probe is given in Fig. 5.5.

Corresponding to picture of high impedance passive voltage probe, as shown in the figure, above the impedance model, L_p and L_n are the parasitic components of test point and ground respectively, C_{in} and R_{in} are the capacitance and inductance of probe tip in the figure, C_{cable} is capacitance of coax cable, R_{comp} and C_{comp} are parasitic resistance and capacitance of compensator and the left R_{scope} and C_{scope} are parasitic resistance and capacitance of oscilloscope.

Figure 5.5 Impedance model of high impedance passive voltage (HIPV) probe

By analysing the Fig. 5.5, the impedance Z_{HIPv1} is the equivalent impedance of coax cable, compensator and oscilloscope, which can be derived as

$$Z_{HIPv1} = \frac{1}{sC_{cable} + \frac{sC_{comp}}{sC_{comp}R_{comp} + 1} + \frac{1}{R_{scope}} + sC_{scope}} \quad (63)$$

$$= \frac{As + B}{Cs^2 + Ds + 1}$$

Where

$$\begin{cases} A = R_{scope}R_{comp}C_{comp} \\ B = R_{scope} \\ C = R_{scope}R_{comp}(C_{cable} + C_{scope})C_{comp} \\ D = R_{scope}(C_{cable} + C_{scope} + C_{comp}) + R_{comp}C_{comp} \end{cases} \quad (64)$$

Then taking the test point, ground lead and the probe tip into consideration, the impedance Z_{HIPv2} can be easily got as

$$Z_{HIPv2} = (L_p + L_n)s + \frac{R_{in}}{R_{in}C_{in}s + 1} + Z_{HIPv1} \quad (65)$$

The transfer function G_{HIPv} from the input voltage to the output voltage is equal to the impedance ratio Z_{HIPv1}/Z_{HIPv2} according to the impedance model of the high impedance passive voltage probe, so that the transfer function G_{HIPv} can be easily derived from the above two equations:

$$G_{HIPv}(s) = \frac{v_{out}}{v_{in}} = \frac{Z_{HIPv1}}{Z_{HIPv2}} \quad (66)$$

5.1.3 Differential voltage (DV) probe

Similar to the high impedance passive voltage probes, after collecting the information of the differential voltage probes [46] – [52], the point diagrams of these typical differential voltage probes from some leading companies are drawn in Fig. 5.6.

The relationships between the parasitic components in differential voltage probe can be seen from the Fig. 5.6 intuitively. In the first figure, we can see that when the bandwidth increasing, the input impedance multiplying input capacitance decreases in the meantime. However, the same to high impedance passive voltage probes, the price of differential voltage probe is also dependent on the bandwidth, the larger the bandwidth, the higher the price of DV probe.

Figure 5.6 Relationship diagram of surveyed DV probes (a) input impedance and bandwidth (b) price and bandwidth

The commonly used impedance model and picture of differential voltage probe is shown in Fig. 5.7 [53]- [55].

Usually, the differential voltage probe is symmetrical, so the impedance model is symmetrical as well. As we can see from the figure, generally, the parasitic $L_p = L_n$, $R_1 C_1 = R_3 C_3$, $R_1 = R_2$, $R_3 = R_4$, $R_5 = R_6$, $R_7 = R_8$, $R_9 = R_{10}$.

Figure 5.7 Impedance model of DV probe

By analysing the impedance model, the transfer function from input voltage v_{in} to the operational amplifier voltage v_{in}' is

$$G_{DV1}(s) = \frac{v_{in}'}{v_{in}} = \frac{B_1 s + B_0}{A_3 s^3 + A_2 s^2 + A_1 s + A_0} \quad (67)$$

where

$$\begin{cases} A_3 = R_1 R_3 (L_p + L_n) C_1 C_3 \\ A_2 = (R_1 C_1 + R_3 C_3) (L_p + L_n) \\ A_1 = (L_p + L_n + 2R_3 R_1 C_1 + 2R_1 R_3 C_3) \\ A_0 = 2R_1 + 2R_3 \\ B_1 = 2R_3 R_1 C_1 \\ B_0 = 2R_3 \end{cases} \quad (68)$$

Then to analyse the second part of the impedance model, the transfer function from input voltage of isolated operation amplifier v_{in}' to output voltage v_{out} can be easily derived as

$$G_{DV2}(s) = \frac{v_{out}}{v_{in}'} = \frac{R_9}{R_7} \left(1 + \frac{R_5}{R_{56}} + \frac{R_6}{R_{56}} \right) \quad (69)$$

As the result, the total transfer function of differential voltage probe G_{DV} from input voltage to the output voltage is easy to get

$$G_{DV}(s) = \frac{v_{out}}{v_{in}} = \frac{v_{in}'}{v_{in}} \times \frac{v_{out}}{v_{in}'} = G_{DV1}(s) G_{DV2}(s) \quad (70)$$

Because some parasitic parameters are too small to be ignored, after removing those small parasitic parameters, the input impedance of the DV probe can be simplified as

$$Z_{DV} = (L_p + L_n)s + 2 \frac{R_1}{R_1 C_1 s + 1} \quad (71)$$

5.1.4 Current probe

The point diagram of parameters of some typical current probes in leading companies, like Tektronix, Keysight and Lecroy [56] – [58], are given in Fig. 5.8.

After the linear transformation, the relationship of the parameters of the current probes can be got from the figure directly. As the lines shown in the figure, the product of current and price will decrease with the bandwidth of the current probes increases. From the right figure, the general tendency between impedance and frequency of the current probes is almost direct ratio, which means the higher the frequency, the larger the impedance will be.

The Rogowski coil is usually used to measure the drain current of the SiC MOSFET. Impedance model of the current probe is given in Fig. 5.9 [59]- [61]. In the model, M is the mutual inductance, C_c , C_p , L_c and R_c are parasitic components of the coil, R_d is damping resistance, R_o and C_o are the parasitic components of the passive integrator

and R_i and C_i are the parasitic components of the active integrator. Generally, $R_o C_o = R_i C_i$.

Figure 5.8 Relationship diagram of surveyed Rogowski coil (a) current×price and bandwidth (b) impedance and frequency

From the impedance model of the Rogowski coil, Z_{RC1} and Z_{RC2} can be easily derived as the equations (72) and (73)

$$Z_{RC1} = \frac{R_d(R_o C_o s + 1)}{C_p R_d(R_o C_o s + 1)s + (R_o + R_d)C_o s + 1} \quad (72)$$

$$Z_{RC2} = \frac{L_c s + R_c}{C_c(L_c s + R_c)s + 1} \quad (73)$$

So the input impedance Z_{RC} can be got from Z_{RC1} and Z_{RC2} shown above

$$Z_{RC} = \frac{(Z_{RC1} + Z_{RC2})Ms}{Ms + Z_{RC1} + Z_{RC2}} \quad (74)$$

For some parasitic parameters are very small which can be ignored, the equation of Z_{RC} can be simplified as

$$Z_{RC} = \frac{Ms}{MC_c s^2 + 1} \quad (75)$$

Figure 5.9 Impedance model of current probe

The transfer function of the current probe from the output voltage v_{out} to the input current i can be derived from the above impedance equations

$$G_{RC}(s) = \frac{i}{v_{out}} = Ms \frac{Z_{RC1}}{Z_{RC1} + Z_{RC2}} \frac{1}{R_o C_o s + 1} \quad (76)$$

As a result, to sum up the impedance model of SiC MOSFET, high impedance passive voltage probe, differential voltage probe and current probe above, the mechanism model of the SiC MOSFET with different measurement probes is showing in the following Fig. 5.10

Figure 5.10 Impedance model of SiC MOSFET with probes

5.2 Stability evaluation

As the Fig. 5.10 shows, the impedance model of device-probe system above is too complicated to analyse, according to the transfer function and equivalent impedance derived in the last part can help simplify the circuit as followings:

Figure 5.11 Impedance model of SiC MOSFET with measurement probes (a) impedance model of test bench (b) simplified circuit (c) small-signal model (b) circuit after star-delta transformation

In Fig. 5.11(a), Z_{HIPV} , Z_{DV} and Z_{RC} are the equivalent impedances derived from the single impedance model of high impedance passive voltage probe, differential voltage probe and current probe respectively in the last part.

In the Fig. 5.11 (b), put the impedance model of SiC MOSFET into application, and after the simplification, the simplified circuit can be got.

In order for easy analysis, the small signal model of the impedance model in Fig. 5.11 (b) are given in Fig. 5.11 (c), ignoring some small parasitic parameters which have little impact, the equivalent impedance Z_1 , Z_2 and Z_3 in (c) are derived as

$$\begin{cases} Z_1 = Z_g + \frac{Z_{HIPV}R_{ge}}{(Z_{HIPV} + R_{ge})} \\ Z_2 = Z_{cs} \\ Z_3 = Z_d + \frac{Z'_{DV}(Z_{Loop} + Z_{RC})}{(Z_{Loop} + Z'_{DV} + Z_{RC})} \end{cases} \quad (77)$$

where

$$\begin{cases} Z_g = L_g s + R_g \\ Z_d = L_d s + R_d \\ Z_{cs} = L_{cs} s + R_{cs} \\ Z'_{DV} = \frac{Z_{DV}Z_{sn}}{(Z_{DV} + Z_{sn})} \\ Z_{sn} = R_{sn} + \frac{1}{(C_{sn}s)} \\ Z_{loop} = L_{loop} s + \frac{1}{(C_{loop}s)} \end{cases} \quad (78)$$

For further simplification of the circuit from Fig. 5.11(c) to (d), Z_{GD} , Z_{GS} and Z_{DS} can be expressed by Z_1 , Z_2 and Z_3 in (77)

$$\begin{cases} Z_{GS} = \frac{(Z_1 Z_2 + Z_1 Z_3 + Z_2 Z_3)}{Z_3} \\ Z_{GD} = \frac{(Z_1 Z_2 + Z_1 Z_3 + Z_2 Z_3)}{Z_2} \\ Z_{DS} = \frac{(Z_1 Z_2 + Z_1 Z_3 + Z_2 Z_3)}{Z_1} \end{cases} \quad (79)$$

The transfer function $F(s)$ from the current i_d to the gate-source voltage v_{gs} can be expressed as

$$F(s) = \frac{v_{gs}}{i_d} = -\frac{Z'_{DS}}{Z'_{GD} + Z'_{GS} + Z'_{DS}} Z'_{GS} \quad (80)$$

where

$$\begin{cases} Z'_{GS} = \frac{Z_{GS}}{(C_{GS}sZ_{GS} + 1)} \\ Z'_{GD} = \frac{Z_{GD}}{(C_{GD}sZ_{GD} + 1)} \\ Z'_{DS} = \frac{Z_{DS}}{(C_{DS}sZ_{DS} + 1)} \end{cases} \quad (81)$$

The block diagram of the impedance model is given in the Fig. 5.12.

After analysing the simplified figure above and block diagram, the transfer function of the SiC MOSFET model with measurement probes can be derived as

$$G(s) = \frac{i_d}{v_g} = \frac{A(s)}{1 - A(s)F(s)} = \frac{g_m(Z'_{GD} + Z'_{GS} + Z'_{DS})}{Z'_{GD} + Z'_{GS} + Z'_{DS} - g_m Z'_{GS} Z'_{DS}} \quad (82)$$

Figure 5.12 Block diagram of impedance model

And according to the block diagram and the equation (82), the gain $A(s)$ in the forward path is shown in the next equation

$$A(s) = \frac{\partial i_d}{\partial v_{gs}} = g_m \quad (83)$$

To simplify the stability analysis of the SiC MOSFET with probes, a characteristic equation is needed which can show the stability of the system intuitively and easy to analyse, the characteristic equation is given in (77),

$$\Delta(s) = \prod_i (s - p_i) = 0 \quad (84)$$

Where p_i means poles of $\Delta(s)$.

The poles p_i can also be written as

$$p_i = \sigma_i + j\omega_i = \sigma_i + j2\pi f_i \quad (85)$$

So the damping ratio ξ_i can be expressed as

$$\xi_i = -\frac{\sigma_i}{\sqrt{\sigma_i^2 + \omega_i^2}} = -\frac{\sigma_i}{\omega_{di}} \quad (86)$$

According to the damping ratio in the equation (79), the stability of the SiC MOSFET system can be determined by the real part of the pole, σ_i . If σ_i is positive, which means the damping ratio becomes negative, so the transient stability measurement becomes unstable. On the contrary, if σ_i is negative, or even far away from the imaginary axis, it means the damping ratio is positive, so the system is stable.

Figure 5.13 Flow diagram of transient stability analysis of SiC MOSFET with probes

The flow diagram of the way to figure out the transient stability of the SiC MOSFET device with the measurement probes is given in the Fig. 5.13.

As shown in Fig. 5.13, to build the synthesis impedance model, there are two parallel ways, the first is to build the circuit model and the second part is the impedance model. To get the circuit model, the correct test conditions and test circuit is needed. To set the impedance model, besides the right test conditions, the desired test variables and probes are important factors. After building the synthesis impedance model of the SiC MOSFET device with measurement probes, the stability boundary and limitation can be found according to the stability analysis by using the pole map, which the root of the system can be seen directly to judge the stability.

5.3 Case study

As shown in equation (75), from the closed-loop transfer function $G(s)$, the poles of SiC MOSFET device with measurement probes can be calculated, for example in Fig. 5.14.

The SiC MOSFET with measurement probes system contains more than 40 poles. From the poles map shown in the Fig. 5.14, we can see that among all the poles, there is a pairs of conjugate poles which are very close to the real axis, as the zoom in figure shows, the resonant frequency of the pair of conjugate poles is 24 MHz, and the damping ratio is 0.38.

From the zoom in figure, we can also see that there are two pairs of conjugate poles which are very close to the imaginary axis, their resonant frequencies are 283 MHz and 285 MHz respectively, and the damping ratios are 0.46% and 0.36%, which are very close to 0. These poles are all very close to the unstable area, if there are any disturbance during the measurement, they may move to the right part of the imaginary axis, as the result, the SiC MOSFET device with probes system will lose the transient stability.

Figure 5.14 Poles distribution of the impedance model of SiC MOSFET with measurement probes

Then we consider the effect of parasitic components in HIPV probes, C_{in} and L_n , in the SiC MOSFET system, the poles locus of the system is given in the Fig.5.15.

Figure 5.15 Poles distribution of SiC MOSFET influenced by the parasitic of HIPV probe (a) C_{in} changes from 1pF to 100pF (b) L_n changes from 50nH to 300nH

When the capacitance C_{in} changes from 1 pF to 100 pF, the poles of the system move to the unstable area. The boundary of the C_{in} is 47 pF, if $C_{in} > 47$ pF, it will become unstable.

For the inductance L_n , in the experiments, let L_n changes from 50 nH to 300 nH, from the poles map we can see that with the increase of the inductance L_n , the locus of conjugate poles of the system move close to the unstable area, which also decreases the stability of the system.

The effect of parasitic components of differential voltage probes is also given in the next figure. As shown in Fig. 5.16, with the increase of capacitance C_l and inductance L_n , the poles loci are given in the figure.

Figure 5.16 Poles distribution of SiC MOSFET influenced by the parasitic of DV probe (a) C_1 from 1pF to 100pF (b) L_n from 50nH to 300nH

From the Fig.5.16 (a), with the increase of C_1 , the poles of the system move close to the imaginary axis, which is similar to the HIPV probe, reducing the damping ratio as well as the stability margin.

Similarly, when the inductance L_n increases, the locus of the conjugate poles also moves close to the unstable area, decreases the stability of the system.

Next is the analysis of effect of parasitic components of Rogowski coil on the SiC MOSFET system.

As shown in Fig. 5.17, the mutual inductance M and capacitance C_c change from 1 nH to 100 nH and 0.01pF to 100 pF respectively.

Figure 5.17 Poles distribution of SiC MOSFET influenced by the parasitic of Rogowski coil (a) M changes from 1nH to 100nH (b) C_c changes from 0.01pF to 100pF

From the locus of poles, we can get that the increase of M has little changes on the stability of the system. However, during the period that the capacitance C_c changes

from 0.01 pF to 100 pF, some poles move to the unstable area, which means the changes of C_c can lead to the instability to the system.

At last, we talk about how the parasitic parameters C_{gs} , C_{gd} , C_{ds} , and g_m in SiC MOSFET, which have effect in the stability of the system with probes, as shown in Fig. 5.18.

Figure 5.18 Poles distribution of SiC MOSFET influenced by the parasitic of SiC MOSFET (a) C_{gs} changes from 5pF to 4nF (b) C_{ds} changes from 5pF to 400pF (c) C_{gd} changes from 5pF to 100pF (d) g_m changes from 1S to 25S

According to the Fig. 5.18, it can be seen that, generally, these parasitic components is capacitive and with the increase of the parasitic of the SiC MOSFET, the loci of conjugate poles move to the imaginary axis, making their damping ratio decreases, which reduces the transient stability of the system as well.

Chapter 6 Conclusion and future work

6.1 Conclusion

For stability analysis of power electronic systems, the stability problems in converter part, converters part, and switch test part are all need to be paid attention on. This thesis revisited some stability criteria for stability analysis of power electronic systems, such as the forbidden region criteria, the Nyquist stability criterion, and also the way to calculate the gain margin and phase margin of the system using the Nyquist plot.

It can be seen that among them, the Nyquist criterion, which is applicable for the non-minimum phase power electronic system and considers the case that transfer function contains integrators, is much more widely used than some other traditional stability criteria.

In the converter part, boost converter is a kind of typical and widely used converter. In this part, stable and unstable boost converters are used to show the challenges in the stability analysis. The cases revealing that the parameters of the converter have great impact on system stability. From the Nyquist diagram, bode diagram and pole-zero map of the boost converter case, it can be seen that the boost converter system is non-minimum phase system, and the open-loop gain contains $1/s$, so that Nyquist stability criterion is needed to do the stability analysis, which considering the RHP poles and has supplementary curve at 0 Hz to help analyse the system with $1/s$ in the transfer function.

In the converters part, because in the small signal model of the system, the weak grid can be seen as a converter, the stable and unstable grid-connected inverter with weak grid are used as cases of converters showing the challenges in converters stability analysis. Same to the boost converter, inaccurate parameters of converters also can make the system unstable. The Nyquist diagram, bode diagram, and pole-zero map are drawn to do stability analysis, it shows that the grid-connected inverter with weak grid system sometimes can be a non-minimum phase system, confirming that comparing with traditional stability analysis methods, Nyquist stability criterion is essentially needed.

And in the switch test part, for switches are always use semiconductors as material, like SiC MOSFET, which has great advantages as the switch. However, besides the advantages, it can also bring some problems to the system as a double-edged sword. The advantage of its fast switching speed characteristic is that, the fast switching speed means the high switching frequency, and it can help the system reduce the power loss. However, in contrast, the SiC MOSFET makes the system more complex, and more than this, its fast speed characteristic make it sensitive to the parasitic components, which brings great challenge to the stability of the system. In the thesis, the parasitic of measurement probes which is usually ignored when analysing the factors that have influence to the stability of the SiC MOSFET system. According to the impedance model of SiC MOSFET, high impedance passive voltage probe, differential voltage probe and current probe, the impedance model of SiC MOSFET with probes is derived for stability analysis. By collecting the parasitic components of some typical measurement probes from the leading companies and analysing the poles map of the system, it can be seen that the parasitic components of probes and SiC MOSFET itself has impact on the transient stability of the switch system. Almost, with the parasitic components of probes, like the capacitance and inductance increases, the accuracy of measurement will decrease, or even affect the transient stability.

6.2 Recommendations for further research

The stability analysis on different parts of the power electronic system from design to application has been done. Generally, the grid-connected power electronic systems are more complicated than the single-phase grid-connected inverter discussed in the thesis.

(1) Stability of multiphase parallel-connected inverters with weak grid

Usually, the power electronic system connects to the grid is the multiphase inverter system, or even several inverters parallel connected and then linked to the weak grid system. The stability problems of parallel-connected inverter system and multiphase inverter [62] with weak grid [63], as shown in Fig. 6.1, are more complicated than single-phase grid-connected inverter. The influence of system circuit and controller and interaction between phases, parallel-connected inverters and between inverters and grid systems are all needed to be concerned.

Figure 6.1 Multiple parallel grid-connected inverters [62]

(2) Harmonic analysis in the power electronic systems

The harmonic is also a problem that can affect the stability of the power electronic system. Grid-connected inverters may have unknown instability problems caused by harmonics from the inside and outside of the inverter. The stability of the power electronic system also includes the harmonic factor. Methods like PI controller and harmonic state space [64] have been done for harmonic analysis. One of my future work is to find out how the harmonics and harmonic coupling affect the stability and how to mitigate the useless harmonics need to be discussed.

Author's Publications

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