# Modeling and Control of the Modular Multilevel Converter (MMC) based Solid State Transformer (SST) with Magnetic Integration

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Abstract-Solid state transformer (SST) can provide more functionalities compared advanced with conventional transformer, and has great potential in smart grid application. Recently, the SST with medium frequency (MF) isolation link and magnetic integration feature has been proposed, which can reduce the system volume and thus increase the power density. However, the magnetic integration also introduces strong coupling between the line frequency (LF) and MF variables, which poses a great challenge on modeling and control issues. This paper proposes a modeling and control method for an SST with magnetic integration and mixed-frequency modulation. A mathematical model based on dual d-q references is deduced, and then a cascaded control system is designed according to the model. Parameters of the controller for the variables at one frequency are properly designed to avoid disturbance from the variables at the other frequency. The simulation and experimental results show good decoupling effect and satisfactory dynamics performance of the proposed control system.

*Index Terms*—solid state transformer (SST), modular multilevel converter (MMC), control system, mathematical model

# I. INTRODUCTION

Compared with line frequency (LF) power transformers, solid state transformers (SSTs) have more advanced functions. It can not only provide galvanic isolation and voltage matching, but also realize port power control, reactive power compensation and fault protection. Therefore, they are recommended to act as the "energy routers" in the future smart grid [1], [2].

Since SSTs convert the port voltage and current (usually in LF-AC or DC form) into medium frequency (MF) AC form and applied it on the isolation transformer, the volume of the transformer can be greatly decreased thanks to the higher operation frequency [3]. However, power electronics converters with heatsinks and extra passive components are required in both the primary and secondary side to realize the

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conversion from LF-AC, or DC, to MF-AC, which takes a lot of volume. Taking the SST tapping from the medium voltage (MV) AC grid as an example, modular converters are usually utilized to convert the LF-AC grid voltage to DC, then to MF-AC voltage [4]. The objectives of the control system for such a SST are to realize active and reactive powers control of the LF and MF port, as well as realize inherent energy balancing. Such a configuration requires a two-stage energy conversion with lots of active and passive components, which degrades the power density.

One of the ways to improve the power density of SST is to increase the integration level of the SST topology, so as to realize LF-AC to MF-AC energy conversion in one conversion stage. Although the matrix converter can realize such a conversion within one stage, the two frequencies cannot be fully decoupled and the reactive power of the port cannot be freely controlled [5-7]. Ref. [8] proposed that the power generated by different frequency components are orthogonal to each other, hence the MMC arm could output active power at one frequency while absorb the same amount of power at a different frequency by using mixed-frequency modulation. With this concept, ref. [8] focused on DC-AC converter by regarding the DC value as zero-frequency value. However, it could not be applied to AC-AC conversion directly; Ref. [9] proposed an AC-AC MMC to realize directly LF-AC to MF-AC conversion. However, resonant capacitor was needed to extract the MF component, which was unsuitable for high voltage and high power applications. In Ref. [10], a MMC based SST with magnetic integration was proposed. The arm inductors were integrated with the MF transformer winding to create the MF loop on the basis of conventional MMC structure. Such magnetic integration design could reduce the system volume and increase the power density. However, with the increase of topology integration, coupling between variables at different frequencies in such an SST was also strengthened, which posed a challenge on modeling and high-performance controlling. Although the control frameworks were mentioned in [10], neither the detail modeling method nor the high-performance controller have been studied. To bridge this gap, this paper thus presents the detailed dynamical mathematical model deduction, and then designs its decoupled control system. The main contributions of this paper are as follows:

1. Since the state variables of the integrated SST have two different frequencies, a mathematical model of the SST



Fig. 1. Topology of the SST with magnetic integration

based on dual d-q reference frames is proposed, which facilitates the design of the control system;

- 2. A cascaded control system with energy balancing outer loop and current tracking inner loops is presented, which can provide effective power flow control of the LF and MF ports, respectively;
- 3. A control parameter design method is presented. Benefited from the dual *d-q* references, simple PI controllers can be applied and they can naturally immune to the influence generated by the other frequency component. Therefore, decoupled control between the two frequency components can be achieved.

The rest of this paper is organized as follows. In Section II, the topology of the SST with integration feature is reviewed. In Section III, the mathematical model of the SST based on dual d-q reference is deduced and an overview of the control system is demonstrated. The design principle of the current and energy-balancing controllers of the control system is given in section IV and V, respectively. The simulation and experiment results are given in Section VI. Besides, some related importance issues about this paper are discussed in Section VII. The conclusion is drawn in Section VIII.

# II. OVERVIEW OF THE SST TOPOLOGY

# A. Topology Introduction

The topology of the SST proposed in [10] is shown in Fig. 1, which is utilized to interconnect MVAC distribution network and LVDC microgrid. A three-phase MMC is used in the MV side to withstand the high voltage. Each MMC arm contains *N* half-bridge submodules (SMs) in series. Different from traditional MMC, the arm inductors of the MMC are replaced by the primary windings of three MF transformers  $T_1 - T_3$ . The MF value should be properly designed. As the switching frequency increases, the transformer volume could be decreased but the submodule heatsink volume would be increased. In this paper, the MF is selected as 500Hz to compromise the overall system volume. For interested readers,

more details can be found in [10], where the compromising between transformer volume and submodule heatsink volume was well explained. The central taps of primary windings are connected to the MVAC grid through the filter inductors  $L_{g}$  and equivalent resistances  $R_{g}$ . The secondary windings are in star-connection, and are connected to the LVDC microgrid  $V_{\rm LVDC}$  through a three-phase full-bridge converter. Such a topology can reduce the volume of the magnetic components as it integrates the arm inductors with the MF transformer. With detailed design procedure given in [10], an optimal design results which compromise voltage and current stress, submodule volume, transformer volume, transformer losses and magnetic saturation can be found. It should be noted that usually several secondary windings with rectifiers are utilized to divide the current stress in the LVDC side, and they are regarded as an equivalent one in this paper for convenience. K is the turning ratio of the transformer.  $L_{\sigma}$  is the leakage inductance of the windings, which has all been converted into the primary side.  $R_{\rm arm}$  is the arm equivalent resistances.

 $v_{Ux}$  and  $v_{Lx}$  (x=A, B, C) are the arm output voltages.  $v_{gx}$  are the phase voltages of the MVAC grid.  $v_{sy'}$  (y=a, b or c) are the secondary winding voltages of the transformers, and are controlled as symmetrical voltages by the three-phase full-bridge converter.  $i_{Ux}$ ,  $i_{Lx}$ ,  $i_{gx}$ ,  $i_{cx}$  and  $i_{sy}$  are the upper arm current, lower arm currents, grid currents, arm circulating currents and secondary currents, respectively. Their relationships are:

$$\begin{cases} i_{gx} = i_{Lx} - i_{Ux} \\ i_{cx} = \frac{i_{Ux} + i_{Lx}}{2} \\ i_{sy} = K i_{cx} \end{cases}$$
(1)

# B. Modulation

Mixed-frequency modulation is applied on the MMC arms to handle the two frequency components [8-10]. The modulation references are defined as:

$$\begin{cases} m_{Ux} = 0.5 - m_{LFx} + m_{MFx} \\ m_{Lx} = 0.5 + m_{LFx} + m_{MFx} \end{cases}$$
(2)

where  $m_{LFx}$  and  $m_{MFx}$  are the LF and MF modulation references, respectively. Ignoring the switching frequency components and assuming that the SM capacitor voltages are well balanced, the arm output voltages can be expressed as:

$$\begin{cases} v_{Ux} = m_{Ux} v_{\Sigma Ux} \\ v_{Ux} = m_{Ux} v_{\Sigma Ux} \end{cases}$$
(3)

where  $v_{\Sigma Ux}$  and  $v_{\Sigma Lx}$  are the SM capacitor voltage summation of the corresponding arms. In the following section, it can be seen that currents can be controlled by adjusting the LF and MF components references  $m_{LFx}$  and  $m_{MFx}$ . Besides, defining the voltage summation and difference of the upper and lower arms as:

$$\begin{cases} v_{\Sigma x} = v_{\Sigma U x} + v_{\Sigma L x} \\ v_{\Delta x} = v_{\Sigma U x} - v_{\Sigma L x} \end{cases}$$

$$\tag{4}$$

where  $v_{\Sigma x}$  denotes the phase energy, and  $v_{\Delta x}$  denotes the energy difference between the upper and lower arms.



Fig. 2. Diagram of the control system.

#### III. MODELING AND CONTROL OF THE SST

In [10], a three-phase mathematical model of the SST is presented to explain the operation principle. However, it is not suitable for the control system design. In this section, a model based on dual d-q reference is proposed. And based on that, the control system for the SST is demonstrated.

# A. Dual d-q reference model

# 1) LF Current Relationship

The Kirchhoff voltage equation of the loops P-A-O-B-P and P-A-O-C-P can be written as:

$$v_{\rm UA} + L_{\sigma} \frac{di_{\rm UA}}{dt} + R_{\rm arm} i_{\rm UA} + \frac{1}{2} K v_{\rm sa} - L_{\rm g} \frac{di_{\rm gA}}{dt} - R_{\rm g} i_{\rm gA} + v_{\rm gA}$$
(5a)

$$= v_{\rm UB} + L_{\sigma} \frac{d_{\rm UB}}{dt} + R_{\rm arm} i_{\rm UB} + \frac{1}{2} K v_{\rm sb} - L_{\rm g} \frac{g_{\rm g}}{dt} - R_{\rm g} i_{\rm gB} + v_{\rm gB}$$

$$v_{\rm s} + L \frac{di_{\rm UA}}{dt} + R_{\rm g} i_{\rm gB} + \frac{1}{2} K v_{\rm sb} - L \frac{di_{\rm gA}}{dt} - R_{\rm g} i_{\rm gB} + v_{\rm gB}$$

$$= v_{\rm UC} + L_{\sigma} \frac{di_{\rm UC}}{dt} + R_{\rm arm} i_{\rm UC} + \frac{1}{2} K v_{\rm sc} - L_{\rm g} \frac{di_{\rm gC}}{dt} - R_{\rm g} i_{\rm gB} + v_{\rm gC}$$
(5b)

Adding (5a) and (5b) together gives:

$$3L_{\sigma} \frac{di_{UA}}{dt} + 3R_{arm}i_{UA} + \frac{3}{2}Kv_{sa} - 3L_{g}\frac{di_{gA}}{dt} - 3R_{g}i_{gA} + 3v_{gA}$$
(6a)  
=  $v_{UB} + v_{UC} - 2v_{UA}$ 

$$3L_{\sigma}\frac{di_{LA}}{dt} + 3R_{am}i_{LA} + \frac{3}{2}Kv_{sa} + 3L_{g}\frac{di_{gA}}{dt} + 3R_{g}i_{gA} - 3v_{gA}$$
(6b)

 $= v_{\rm LB} + v_{\rm LC} - 2v_{\rm LA}$ 

Subtracting (6a) from (6b) and substituting (1) into it, the dynamics of the grid current  $i_{gA}$  can be expressed as:

 $L_{\rm LF} \frac{di_{\rm gA}}{dt} = \frac{1}{6} (v_{\rm LB} - v_{\rm UB}) + \frac{1}{6} (v_{\rm LC} - v_{\rm UC}) - \frac{1}{3} (v_{\rm LA} - v_{\rm UA}) + v_{\rm gA} - R_{\rm LF} i_{\rm gA} (7)$ where  $L_{\rm LF} = L_{\rm g} + 0.5L_{\sigma}$  and  $R_{\rm LF} = R_{\rm g} + 0.5R_{\rm arm}$ . Similarly, the grid current  $i_{\rm gB}$  and  $i_{\rm gC}$  can be calculated, which forms the three-phase state variables equations as:

$$\begin{cases} L_{\rm LF} \frac{di_{gA}}{dt} = \frac{1}{6} (v_{\rm LB} - v_{\rm UB}) + \frac{1}{6} (v_{\rm LC} - v_{\rm UC}) - \frac{1}{3} (v_{\rm LA} - v_{\rm UA}) + v_{gA} - R_{\rm LF} i_{gA} \\ L_{\rm LF} \frac{di_{gB}}{dt} = \frac{1}{6} (v_{\rm LC} - v_{\rm UC}) + \frac{1}{6} (v_{\rm LA} - v_{\rm UA}) - \frac{1}{3} (v_{\rm LB} - v_{\rm UB}) + v_{gB} - R_{\rm LF} i_{gB} \\ L_{\rm LF} \frac{di_{gC}}{dt} = \frac{1}{6} (v_{\rm LA} - v_{\rm UA}) + \frac{1}{6} (v_{\rm LB} - v_{\rm UB}) - \frac{1}{3} (v_{\rm LC} - v_{\rm UC}) + v_{gC} - R_{\rm LF} i_{gC} \end{cases}$$
(8)

Decomposing  $m_{\text{LFx}}$ ,  $m_{\text{MFx}}$ ,  $i_{\text{gx}}$ ,  $v_{\text{gx}}$   $v_{\Sigma x}$  and  $v_{\Delta x}$  into the *d-q* reference synchronizing to the grid voltage  $v_{\text{gx}}$  (we denote the variables in this reference with a superscript "L"), (8) can be transformed into the LF *d-q* reference as:

$$\begin{cases} L_{\rm LF} \frac{di_{\rm gd}^{\rm L}}{dt} = v_{\rm gd}^{\rm L} + \omega_{\rm LF} L_{\rm LF} i_{\rm gq}^{\rm L} - R_{\rm LF} i_{\rm gd}^{\rm L} - v_{\rm LFd}^{\rm L} \\ L_{\rm LF} \frac{di_{\rm gq}^{\rm L}}{dt} = v_{\rm gq}^{\rm L} - \omega_{\rm LF} L_{\rm LF} i_{\rm gd}^{\rm L} - R_{\rm LF} i_{\rm gq}^{\rm L} - v_{\rm LFq}^{\rm L} \end{cases}$$
(9a)

where

$$\begin{cases} v_{\rm LFd}^{\rm L} = \frac{1}{2} \left( 0.5 v_{\rm Ad}^{\rm L} + m_{\rm MFd}^{\rm L} v_{\rm A0}^{\rm L} + m_{\rm LFd}^{\rm L} v_{\Sigma 0}^{\rm L} \right) \\ v_{\rm LFq}^{\rm L} = \frac{1}{2} \left( 0.5 v_{\rm Aq}^{\rm L} + m_{\rm MFq}^{\rm L} v_{\rm A0}^{\rm L} + m_{\rm LFq}^{\rm L} v_{\Sigma 0}^{\rm L} \right) \end{cases}$$
(9b)

which can be used as the design basis of the LF current controller. Assuming that the *d*-axis is in phase with the synthesized vector of  $v_{gx}$  so that  $v_{gd}=V_g$  and  $v_{gq}=0$  can be obtained, the relationship between LF power, voltage and current can be obtained as:

$$\begin{cases} P_{\rm LF} = \frac{3}{2} V_{\rm g} i_{\rm gd}^{\rm L} \\ Q_{\rm LF} = \frac{3}{2} V_{\rm g} i_{\rm gq}^{\rm L} \end{cases}$$
(10)

### 2) MF Current Relationship

Similarly, adding (6a) and (6b) together gives the dynamics of the three-phase circulating current  $i_{cx}$  as:

$$\begin{cases} L_{\rm MF} \frac{di_{\rm cA}}{dt} = \frac{1}{3} \left( v_{\rm UB} + v_{\rm LB} \right) + \frac{1}{3} \left( v_{\rm UC} + v_{\rm LC} \right) - \frac{2}{3} \left( v_{\rm UA} + v_{\rm LA} \right) - R_{\rm MF} i_{\rm cA} - K v_{\rm sa} \\ L_{\rm MF} \frac{di_{\rm cB}}{dt} = \frac{1}{3} \left( v_{\rm UC} + v_{\rm LC} \right) + \frac{1}{3} \left( v_{\rm UA} + v_{\rm LA} \right) - \frac{2}{3} \left( v_{\rm UB} + v_{\rm LB} \right) - R_{\rm MF} i_{\rm cB} - K v_{\rm sb} \ (11) \\ L_{\rm MF} \frac{di_{\rm cC}}{dt} = \frac{1}{3} \left( v_{\rm UA} + v_{\rm LA} \right) + \frac{1}{3} \left( v_{\rm UB} + v_{\rm LB} \right) - \frac{2}{3} \left( v_{\rm UC} + v_{\rm LC} \right) - R_{\rm MF} i_{\rm cC} - K v_{\rm sc} \end{cases}$$

where  $L_{MF}=2L_{\sigma}$  and  $R_{MF}=2R_{arm}$ . Decomposing  $m_{LFx}$ ,  $m_{MFx}$ ,  $i_{cx}$ ,  $v_{sy}$   $v_{\Sigma x}$  and  $v_{\Delta x}$  into the *d-q* reference synchronizing to  $v_{sy}$  (and the variables in this reference are denoted with a superscript "M"), (11) can be transformed as:

$$\begin{cases} L_{\rm MF} \frac{di_{\rm cd}^{\rm M}}{dt} = -Kv_{\rm sd}^{\rm M} + \omega_{\rm MF}L_{\rm MF}i_{\rm cq}^{\rm M} - R_{\rm MF}i_{\rm cd}^{\rm M} - v_{\rm MFd}^{\rm M} \\ L_{\rm MF} \frac{di_{\rm cq}^{\rm M}}{dt} = -Kv_{\rm sq}^{\rm M} - \omega_{\rm MF}L_{\rm MF}i_{\rm cd}^{\rm M} - R_{\rm MF}i_{\rm cq}^{\rm M} - v_{\rm MFq}^{\rm M} \end{cases}$$
(12a)

where

$$\begin{cases} v_{MFd}^{M} = \left(0.5 v_{\Sigma d}^{M} + m_{MFd}^{M} v_{\Sigma 0}^{M} + m_{LFd}^{M} v_{\Delta 0}^{M}\right) \\ v_{MFq}^{M} = \left(0.5 v_{\Sigma q}^{M} + m_{MFq}^{M} v_{\Sigma 0}^{M} + m_{LFq}^{M} v_{\Delta 0}^{M}\right) \end{cases}$$
(12b)

which can be used as the design basis of the MF current controller. And the active and reactive powers are:

$$\begin{cases} P_{\rm MF} = \frac{3}{2} K V_s i_{\rm ed}^{\rm M} \\ Q_{\rm MF} = \frac{3}{2} K V_s i_{\rm eq}^{\rm M} \end{cases}$$
(13)

# 3) Power Balancing

To facilitate the deduction, all the SM capacitor voltage are considered as stable and well balanced. Taking the phase A as an example, the arm voltages  $v_{UA}$  and  $v_{LA}$  are:

$$\begin{cases} v_{\rm UA} = m_{\rm UA} V_{\rm \Sigma arm} = (0.5 - m_{\rm LFA} + m_{\rm MFA}) V_{\rm \Sigma arm} \\ v_{\rm LA} = m_{\rm LA} V_{\rm \Sigma arm} = (0.5 + m_{\rm LFA} + m_{\rm MFA}) V_{\rm \Sigma arm} \end{cases}$$
(14)

where  $V_{\Sigma arm}$  is the summation of the SM capacitor voltages of the MMC arm. According to (1), the arm currents are:

$$\begin{cases} i_{UA} = -\frac{1}{2}i_{gA} + i_{cA} \\ i_{LA} = \frac{1}{2}i_{gA} + i_{cA} \end{cases}$$
(15)

The arm active power is defined as the mean value of the product of the instantaneous values of voltage and current. As discussed in [8], different frequency components are orthogonal to each other and their products will not contribute to the active power (i.e., the periodical integrals of all the cross product of terms with different frequencies are zero). Since  $m_{\text{LFA}}$  and  $i_{\text{gA}}$  are LF components while  $m_{\text{MFA}}$  and  $i_{\text{cA}}$  are MF components, the active power of the upper and lower arms in phase A can be simplified as:

$$p_{\rm A} = \frac{1}{\underline{T}} \underbrace{\int_{0}^{T} \left( m_{\rm LFA} i_{\rm gA} V_{\Sigma \rm arm} \right) dt}_{P_{\rm LFA}} + \frac{1}{\underline{T}} \underbrace{\int_{0}^{T} \left( 2m_{\rm MFA} i_{\rm cA} V_{\Sigma \rm arm} \right) dt}_{P_{\rm MFA}}$$
(16)

where *T* is the period corresponding to the line frequency, and  $P_{\text{LFA}}$  and  $P_{\text{MFA}}$  are the active power inputted to the phase A during *T*. Since all the capacitors are controlled as stable,  $p_{\text{A}}$  should be zero, (16) can be further simplified as:

$$P_{\rm LFA} = -P_{\rm MFA} \tag{17}$$

Phase B and C can also be analyzed in similar manners as:

$$P_{\rm LFB} = -P_{\rm MFB} \text{ and } P_{\rm LFC} = -P_{\rm MFC}$$
(18)

From (10) and (13),  $P_{\rm LF}$  and  $P_{\rm MF}$  are the power inputted from the LF port and the power outputted to the MF port, respectively, which gives:

$$\begin{cases} P_{\rm LF} = P_{\rm LFA} + P_{\rm LFB} + P_{\rm LFC} \\ P_{\rm MF} = -\left(P_{\rm MFA} + P_{\rm MFB} + P_{\rm MFC}\right) \end{cases}$$
(19)

Therefore, the total active power of LF and MF satisfies:

$$P_{\rm LF} = P_{\rm MF} \tag{20}$$

which also denotes that the input and output power of the converter are balanced. Since the voltages of capacitors are controlled as stable, the capacitors operate as the medium to deliver the power from LF to MF. It should also be noted that the harmonics power and conversion losses, which are ignored in the deduction, will cause small unbalance between the input and output power. Therefore, a controller for energy balancing between the input and output power should be implemented, which will be discussed later.

# 4) SM Capacitor Voltage Relationship

According to [11], the dynamic relationships between the SM capacitor voltage summation and the modulation reference can be expressed as:

$$\begin{cases} \frac{C_{\rm SM}}{N} \frac{\mathrm{d}v_{\Sigma \mathrm{U}x}}{\mathrm{d}t} = m_{\mathrm{U}x} i_{\mathrm{U}x} \\ \frac{C_{\rm SM}}{N} \frac{\mathrm{d}v_{\Sigma \mathrm{L}x}}{\mathrm{d}t} = m_{\mathrm{L}x} i_{\mathrm{L}x} \end{cases}$$
(21)

Defining  $v_{\Delta x}$  as the difference between  $v_{\Sigma Ux}$  and  $v_{\Sigma Lx}$  as:

$$v_{\Delta x} = v_{\Sigma L x} - v_{\Sigma U x} \tag{22}$$

Substituting (1), (2) and (22) into (21) gives:

$$\frac{C_{\rm SM}}{N} \frac{\mathrm{d}v_{\Delta x}}{\mathrm{d}t} = \left(0.5 + m_{\rm MFx}\right)i_{\rm gx} + 2m_{\rm LFx}i_{\rm cx} \tag{23}$$

which can be used as the design principle of the energy balancing controller to balancing the voltage between the upper and lower arms.

# B. Control System Overview

Based on the above modeling results, a cascaded control system is adaptively designed. The overall control system is shown in Fig. 2. The basic value of the current references  $i_{gd,q}^{L^*}$  and  $i_{ed,q}^{M^*}$  are generated by the power order according to (10) and (13). Nevertheless, they are also fine-tuned by the energy balancing controller. The modulation references  $m_{LFd}^{L}$ ,  $m_{LFd}^{L}$ ,  $m_{MFd}^{M}$  and  $m_{MFq}^{M}$  are generated by the current controller based on the dynamic equations in (9) and (12). They are converted to three-phase reference  $m_{LFx}$  and  $m_{MFx}$  and then combined as the arm voltage modulation indexes  $m_{Ux}$  and  $m_{Lx}$  according to (2).

Both pulse width modulation (PWM) and nearest level modulation (NLM) can be used to realize the mixed-frequency modulation. Whether NLM or PWM should be used depends on the SM number of the MMC arm. In applications when the SM number is large, NLM is usually selected to reduce the total switching times, and thus reduce the switching loss. When the SM number is small, PWM is used to improve the waveforms quality. No matter which modulation method is used, the arm voltage can still be simplified as (3) by ignoring the switching frequency component and the above deduction is held.

# IV. CURRENT LOOP CONTROLLERS DESIGN

#### A. LF Current loop controller

According to (9), it is proposed to adjust  $m_{\text{LFd},q}$  to control the MMC LF output voltage  $v_{\text{LFd},q}$ , and hence, to control the grid current  $i_{\text{gd},q}$ . Relationships among them can be obtained according to (9) as:

$$\begin{cases} m_{\rm LFd}^{\rm L} = \frac{2}{v_{\Sigma0}^{\rm L}} \left[ -\left(sL_{\rm LF} + R_{\rm LF}\right) i_{\rm gd}^{\rm L} + v_{\rm gd}^{\rm L} + \omega_{\rm LF} L_{\rm LF} i_{\rm gq}^{\rm L} - \frac{1}{4} v_{\Delta d}^{\rm L} - \frac{1}{2} m_{\rm MFd}^{\rm L} v_{\Delta 0}^{\rm L} \right] \\ m_{\rm LFq}^{\rm L} = \frac{2}{v_{\Sigma0}^{\rm L}} \left[ -\left(sL_{\rm LF} + R_{\rm LF}\right) i_{\rm gq}^{\rm L} + v_{\rm gq}^{\rm L} - \omega_{\rm LF} L_{\rm LF} i_{\rm gd}^{\rm L} - \frac{1}{4} v_{\Delta q}^{\rm L} - \frac{1}{2} m_{\rm MFq}^{\rm L} v_{\Delta 0}^{\rm L} \right] \end{cases}$$
(24)



Fig. 3. Details of the LF Controller. (a) control diagram; (b) Bode diagrams.



Fig. 4. Details of the MF Controller. (a) control diagram; (b) Bode diagrams.

Fig. 5. Energy balancing controller diagrams (a) voltage balancing between upper and lower arms in the same leg; (b) total energy balancing controller

which is the design basis of the LF current controller. Therefore, the LF current controller is designed as (25) and the control diagram has also been illustrated in Fig. 3(a).

$$\begin{cases} m_{\text{LFd}}^{\text{L}} = \frac{2}{v_{\Sigma0}^{\text{L}}} \left[ \underbrace{-\left(\frac{K_{\text{LFi}}}{s} + K_{\text{LFp}}\right) \left(i_{\text{gd}}^{\text{L}*} - i_{\text{gd}}^{\text{L}}\right)}_{\text{PI controller}} + \underbrace{v_{\text{gd}}^{\text{L}} + \omega_{\text{LF}}L_{\text{LF}}i_{\text{gq}}^{\text{L}} - \frac{1}{4}v_{\Delta \text{d}}^{\text{L}}}_{\text{feedfoward compensation}} \right] \\ m_{\text{LFq}}^{\text{L}} = \frac{2}{v_{\Sigma0}^{\text{L}}} \left[ \underbrace{-\left(\frac{K_{\text{LFi}}}{s} + K_{\text{LFp}}\right) \left(i_{\text{gq}}^{\text{L}*} - i_{\text{gq}}^{\text{L}}\right)}_{\text{PI controller}} + \underbrace{v_{\text{gq}}^{\text{L}} - \omega_{\text{LF}}L_{\text{LF}}i_{\text{gd}}^{\text{L}} - \frac{1}{4}v_{\Delta \text{q}}^{\text{L}}}_{\text{feedfoward compensation}} \right] (25)$$

where the PI controller with proportional and integral parameters  $K_{\rm LFi}$  and  $K_{\rm LFp}$  is utilized to eliminate the tracking error between  $i_{\rm gd,q}^{L*}$  and  $i_{\rm gd,q}^{L}$ . The feedforward compensation item is used to eliminate the coupling and improve the dynamics performance. It is worth mentioning that the terms  $m_{\rm MFd}^{\rm L} v_{\Delta 0}^{\rm L}$  and  $m_{\rm MFq}^{\rm L} v_{\Delta 0}^{\rm L}$ , which are near 500Hz under the LF *d-q* reference frame, are not considered in (25). This is because our control bandwidth will be properly set later to immune from 500Hz component.

Substituting (25) into (9), relationship between  $i_{gd,q}^{L*}$  and  $i_{gd,q}^{L}$  can be calculated as:

$$i_{gd,q}^{L} = \underbrace{\frac{sK_{LFp} + K_{LFi}}{s^{2}L_{LF} + s\left(R_{LF} + K_{LFp}\right) + K_{LFi}}_{G_{LF}(s)}}_{-\frac{0.5s}{\frac{s^{2}L_{LF} + s\left(R_{LF} + K_{LFp}\right) + K_{LFi}}}} i_{gd,q}^{L*} m_{MFd,q}^{L} v_{\Delta 0}^{L}}$$
(26)

It can be found that  $i_{gd,q}^{L}$  tracks  $i_{gd,q}^{L*}$  and the frequency response is determined by the transfer function  $G_{LF}(s)$ . And a

disturbance is introduced by the terms  $m_{MFd}^L v_{\Delta 0}^L$  and  $m_{MFq}^L v_{\Delta 0}^L$ (as mentioned, this is because the feedforward terms for  $m_{MFq}^L v_{\Delta 0}^L$  and  $m_{MFq}^L v_{\Delta 0}^L$  are not considered) and the frequency response is determined by the transfer function  $G_{LD}(s)$ . And the impact of this disturbance item can be minimized by properly setting the bandwidth of the controller (i.e., the parameter  $K_{LFp}$  and  $K_{LFi}$ ). In our case, the PI controller parameters  $K_{LFp}$  and  $K_{LFi}$  are set as:

$$K_{\rm LFp} = \frac{L_{\rm LF}}{\tau_{\rm LF}}$$
, and  $K_{\rm LFi} = \frac{R_{\rm LF}}{\tau_{\rm LF}}$  (27)

where  $\tau_{LF}$  is the time constant of the controller. Hence the transfer function  $G_{LF}(s)$  simplified as a first-order element as:

$$G_{\rm LF}(s) = \frac{1}{\tau_{\rm LF}s + 1}$$
 (28)

Here,  $\tau_{LF}$ =2.5ms is selected so that the current can tracks the reference within 10ms - 13ms (usually the step response of a first-order system reaches 98% of the step value in  $4\tau_{\rm LF}$ ). Considering the discretization process with 200µs sampling time and using bilinear transform method,  $G_{LF}(s)$  and  $G_{LD}(s)$ can be converted into discretization form as  $G_{LF}(z)$  and  $G_{LD}(z)$ , respectively. Their bode diagrams are drawn in Fig. 3(b). It can be found that the cut-off frequency  $f_{c LF}$  of  $G_{LF}(z)$  is about 65Hz. It is noted that  $m_{\rm MF}$  is the MF reference (500Hz in our case), and since  $m_{\rm MFd}^{\rm L} v_{\Delta 0}^{\rm L}$  and  $m_{\rm MFq}^{\rm L} v_{\Delta 0}^{\rm L}$  are decomposing into the LF reference frame (50Hz), their frequencies in the LF reference frame is near 500Hz. And from the Bode diagrams, it can be found that  $G_{LD}(z)$  provides a -40dB decay (i.e. 0.0001 magnitude) for the 500Hz component and as the frequency goes higher, the decay will become larger. Supposing  $m_{\rm MF} \approx 0.5$  and  $v_{\Delta 0} \approx 1000$ , it will be attenuated to 0.05, whose impact on the LF current can be negligible.

#### B. MF Current Loop Controller

Similar to the LF current loop controller, the MF current controller can be designed as (29) with control diagram shown in Fig. 4(a).

$$\begin{bmatrix} m_{MFd}^{M} = \frac{1}{v_{\Sigma0}^{M}} \begin{bmatrix} -\left(\frac{K_{MFi}}{s} + K_{MFp}\right) \left(i_{cd}^{M*} - i_{cd}^{M}\right) - Kv_{sd}^{M} + \omega_{MF}L_{MF}i_{cq}^{M} - \frac{1}{2}v_{\Sigmad}^{M} \end{bmatrix} \\ m_{MFq}^{M} = \frac{1}{v_{\Sigma0}^{M}} \begin{bmatrix} -\left(\frac{K_{MFi}}{s} + K_{MFp}\right) \left(i_{cq}^{M*} - i_{cq}^{M}\right) - Kv_{sq}^{M} - \omega_{MF}L_{MF}i_{cd}^{M} - \frac{1}{2}v_{\Sigmaq}^{M} \end{bmatrix} \\ p_{I \text{ controller}} \begin{bmatrix} -\left(\frac{K_{MFi}}{s} + K_{MFp}\right) \left(i_{cq}^{M*} - i_{cq}^{M}\right) - Kv_{sq}^{M} - \omega_{MF}L_{MF}i_{cd}^{M} - \frac{1}{2}v_{\Sigmaq}^{M} \end{bmatrix} \end{bmatrix}$$
(29)

And the relationship between  $i_{cd,q}^{M^*}$  and  $i_{cd,q}^{M}$  can be obtained as:

$$\underbrace{\frac{M}{_{cd,q}}}_{cd,q} = \underbrace{\frac{sK_{MFp} + K_{MFi}}{s^{2}L_{MF} + s\left(R_{MF} + K_{MFp}\right) + K_{MFi}}_{G_{MF}(s)} i_{cd,q}^{M^{*}} - \underbrace{\frac{s}{s^{2}L_{MF} + s\left(R_{MF} + K_{MFp}\right) + K_{MFi}}_{G_{MD}(s)} m_{LFd,q}^{M} v_{\Delta 0}^{M}}_{M}$$
(30)

It can be found that  $i_{cd,q}^{M}$  tracks  $i_{cd,q}^{M^*}$  and the frequency response is determined by the transfer function  $G_{MF}(s)$ . Similar to that in (26), a disturbance is introduced by terms  $m_{LFd}^{M} v_{\Delta 0}^{M}$ 



Fig. 6. Simulation waveforms of the port voltage and current of the SST during load switching

and  $m_{\text{LFq}}^{\text{M}} v_{\Delta 0}^{\text{M}}$ , and the frequency response is determined by the transfer function  $G_{\text{MD}}(s)$ .

The PI parameter design method is similar to that in the LF current loop controller design and the time constant is also set as 2.5ms. When  $m_{\text{LFd}}^{\text{M}}v_{\Delta 0}^{\text{M}}$  and  $m_{\text{LFq}}^{\text{M}}v_{\Delta 0}^{\text{M}}$  are decomposing into the MF reference frame (500Hz), their frequencies in the MF reference frame are near 500Hz. Similar to the LF controller, the Bode diagrams of the transfer functions after discretizing is shown in Fig. 4(b), it can be found that  $G_{\text{MD}}(z)$  also provides a -42dB decay (i.e., about 0.0001 magnitude) for the 500Hz component, which makes the controller immune to the disturbance terms  $m_{\text{LFq}}^{\text{M}}v_{\Delta 0}^{\text{M}}$  and  $m_{\text{LFq}}^{\text{M}}v_{\Delta 0}^{\text{M}}$ .

### V. ENERGY BALANCING CONTROLLER DESIGN

Energy balancing between the MMC arms is required for the normal operation of the MMC. The energy balancing controller in this paper is designed as the outer loop controller to adjust the current references, which has two aspects: the energy balancing between the upper and lower arms in the same leg, and the total energy balancing.

# A. Energy Balancing Between the Upper and Lower Arms

The energy balancing controller can be applied to either MF or LF channel. However, if it is applied to the MF channel, a MF component (corresponding to the MF d-q reference) needs to be injected, which challenges the PI controller bandwidth design. Instead, applying the energy balancing controller on the LF channel is much more convenient, and the design is given as follow.

The arm energy is represented as the SM capacitor voltage summation. From (23), the DC value of the  $v_{\Delta x}$  can be controlled by injecting a DC bias of  $i_{gx}$ :

$$\frac{C_{\rm SM}}{N} \frac{\mathrm{d}v_{\Delta x\_dc}}{\mathrm{d}t} = 0.5 i_{\rm gx\_dc} \tag{31}$$

Therefore, the controller is designed as:

$$i_{gx\_dc}^* = \left(\frac{K_{\Delta i}}{s} + K_{\Delta p}\right) \left(v_{\Delta x\_dc}^* - v_{\Delta x\_dc}\right)$$
(32)

The injected DC component  $i_{gx\_dc}$  will be converted to LF component in the LF *d-q* reference, which is within the LF controller's bandwidth. This is the reason that we applied this controller to the LF channel. Design of the parameters  $K_{\Delta i}$  and  $K_{\Delta p}$  are similar with those in the previous section, and thus, are not repeated here. The outputs of the PI controllers ( $i_{gx\_dc}^*$ , x=A, B, C) are decomposing into the LF *d-q* reference as shown in Fig. 5(a), and compensated to the current reference  $i_{gd,q}^{L*}$  as shown in Fig. 2.

# B. Total Energy Balancing

In the control system, although the active power order  $P_{LF}^*$ and  $P_{MF}^*$  for the LF and MF port are designed as the same, the port power may still be different in practical due to two reasons: *1*) the non-ideal factors, such as harmonics and conversion losses, lead to difference between the input and output power; *2*) the terms  $i_{gy\_de}^*$  for energy balancing also impact the reactive power of the LF loop. And the active power difference between LF and MF loop will charge or discharge the SM capacitors. Hence the total SM capacitor energy  $W_{sum}$  can be treated as the indicator for total energy balancing control. The dynamic of  $W_{sum}$  can be expressed as:

$$\frac{dW_{\rm sum}}{dt} = P_{\rm LF} - P_{\rm MF} = \frac{3}{2} V_{\rm g} i_{\rm gd}^{\rm L} - \frac{3}{2} K V_{\rm s} i_{\rm cd}^{\rm M}$$
(33)

And the dynamic of  $W_{sum}$  can also be expressed by the SM

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Fig. 7 Frequency spectrum of the arm current.



Fig. 8. Performance of the current loop controllers

capacitor energy summation:

$$\frac{\mathrm{d}W_{\mathrm{sum}}}{\mathrm{d}t} = \frac{\mathrm{d}}{\mathrm{d}t} \left(\frac{1}{2}C_{\mathrm{SM}}\sum v_{\mathrm{SM}i}^2\right) \approx C_{\mathrm{SM}}V_{\mathrm{SM}}\frac{\mathrm{d}}{\mathrm{d}t}v_{\mathrm{SM\Sigma}}$$
(34)

where  $v_{\text{SM}\Sigma}$  is the summation of all the SM capacitor voltages. A current compensation to  $i_{\text{cd}}^{\text{M}*}$  is introduced to realize total energy balancing of the SST as (35) and the control diagram is shown in Fig. 5(b):

$$i_{\rm cd}^{\rm M*} = -\left(\frac{K_{\rm W1i}}{s} + K_{\rm W1p}\right) \left(v_{\rm SM\Sigma}^* - v_{\rm SM\Sigma}\right) + \frac{P^*}{1.5KV_{\rm s}}$$
(35)

Assuming that  $i_{cd}^{M*} = i_{cd}^{M}$  and substituting (34) and (35) into (33), the transfer function of  $v_{SM\Sigma}$  can be expressed as:



Fig. 9. Simulation waveforms with the dual loop control system

$$v_{\rm SM\Sigma} = \frac{1.5KV_{\rm s}K_{\rm Wli}s + 1.5KV_{\rm s}K_{\rm Wlp}}{C_{\rm SM}V_{\rm SM}s^{2} + 1.5KV_{\rm s}K_{\rm Wli}s + 1.5KV_{\rm s}K_{\rm Wlp}}v_{\rm SM\Sigma}^{*} + \frac{\left(1.5V_{\rm g}i_{\rm gd}^{\rm L} - P^{*}\right)s}{C_{\rm SM}V_{\rm SM}s^{2} + 1.5KV_{\rm s}K_{\rm Wli}s + 1.5KV_{\rm s}K_{\rm Wlp}}$$
(36)

Since the energy balancing controller is serve as the outer loop of the control system, the cut-off frequency is set as 5Hz. The detailed design process is not repeated here since it is similar to those in Section IV.

### VI. VERIFICATION

# A. Simulation

The model of the SST with the proposed control system shown in Fig. 2 is built in PLECS. The circuit parameters of the SST is shown in TABLE I.

# 1) Inner Current Loop Control

The transient performance of the inner LF and MF current loop controller is shown in Fig. 6, 7 and 8. Fig. 6 shows the three-phase voltage and current waveforms. The reference  $i_{gd}^{L*}$ is stepped down from 81.63A to 40.82A and  $i_{cd}^{M*}$  are stepped down from 83.33A to 41.67A, which is equivalent to the situation that the active power of the SST switches from rated power to half rated power condition. And the reference  $i_{gq}^{L*}$  and  $i_{cq}^{M*}$  are all set as zero.  $v_{UA}$  and  $i_{UA}$  is demonstrated as an example of the arm voltages and currents. Since mixed-frequency modulations are used,  $v_{UA}$  and  $i_{UA}$  contains both LF and MF components as described in (1) – (3). The frequency spectrum of the arm current in rated load condition is also illustrated in Fig. 7. It can be found that the arm currents are dominated by the LF and MF components. The amplitude of the LF component (41A) is about half of the amplitude of  $i_{gA}$  (82A, see Fig. 6), while the amplitude of the MF component (83A) equals to the amplitude of  $i_{cA}$ . These all match with the arm current expression in (15). The voltage and current of the LF and MF ports are also demonstrated. The port currents and voltages are in phase, indicating that both ports were controlled to operate in unit power factor condition. When the current references changed, the current switched smoothly without overshoot.

The *d*-axis and *q*-axis current components when the power is switching from rated load to half load are illustrated in Fig. 8(a). It can be found that the proposed control method can provide good tracking performance of  $i_{gd}^{L}$  and  $i_{ed}^{M}$ . At the same time, the impacts on the *q*-axis components, i.e.,  $i_{gq}^{L}$  and  $i_{eq}^{M}$ , is negligible, which implies good decoupling effect between the *d*- and *q*- channels. The tracking time of the current is 10ms – 15ms, which matches with the controller design in Section IV with 2.5ms time constants. Also, small ripples at medium frequency can be found in the *d*-axis and *q*-axis current, which is introduced by the disturbance terms as discussed in (26) and (30). And their amplitudes are small enough thanks to the decay for MF components provided by the controller as shown in the bode diagrams in Fig. 3(b) and Fig. 4(b).

Fig. 8(b) shows the situation that the active power is stepped up from half rated power to rated power condition. Similar conclusion can be made that the controllers provides good tracking performance and effective decoupling.

#### 2) Dual Loop control

The outer energy balancing control is enabled to balance the voltage of the upper and lower arms, as well as balance the input and output voltage. And the simulation results are shown in Fig. 9 when the active power reference is stepped up from half rated power to rated power condition. It can be seen that the active and reactive power tracks the reference well. And the SM capacitor summation is stabilize as the given value with only a little oscillation when the load switches. This is because the tracking process of the active power in the two ports are not exactly the same during transient process. This power difference, as well as the differences that introduced by other non-ideal factors, will be compensated by the total energy controller. And it is seen that the waveforms of  $v_{\Delta x}$  have no DC bias, indicating that the energy of the upper and lower arm are well balanced.

#### B. Experiment

To further verify the proposed control strategy, a scale-down prototype is built in the laboratory. The parameters are shown in TABLE II.

Fig. 10 shows the key waveforms of the prototype. The grid voltage  $v_{gA}$ ,  $v_{gB}$  and current  $i_{gA}$ ,  $i_{gB}$  is shown in Fig. 10(a). It can be found that  $v_{gA}$  and  $v_{gB}$  are in phase with  $i_{gA}$  and  $i_{gB}$ , respectively, which denotes that the grid port is operating in unit power factor condition. And the peak value of the grid current is controlled as 5A. Fig. 10(b) shows the transformer secondary voltage  $v_{sa}$  and the circulating current  $i_{cA}$ . It can be found that  $i_{cA}$  is in phase with the fundamental component of  $v_{sa}$ ,



Fig. 10. Key waveforms of the prototype.



Fig. 11. Dynamic waveforms of the prototype

indicating unit power factor operation of the MF port. Fig. 10(c) shows the arm current  $i_{LA}$ . It can be found that  $i_{LA}$  contains the LF and MF components, matching with the expression in (15).

Fig. 11 shows the step response waveforms when the required active power steps up from 100W to 500W. Good tracking performance can still be found in  $i_{gx}$  and  $i_{cx}$ . The tracking time is within 20ms, which is very close to the controller design target. And the energy of the upper and lower arms, which are represented as the summation of SM capacitor voltages  $v_{\Sigma UA}$  and  $v_{\Sigma LA}$ , are balanced well.

#### VII. FURTHER DISCUSSION

Some important issues are briefly discussed in this section.

## A. Mixed-frequency Feature Comparison

The MMC-based SST in this paper and the MMC-based APF both use the submodule as the buffer to exchange the active power at different frequencies. However, they have distinct differences as follows:

- The MMC-based APF absorbs active power at LF and outputs it in harmonics forms near LF. Such process is carried out within the same and only port of the converter, and the APF is used for improving the electric quality, not for energy conversion;
- As for the MMC-based SST in this paper, it is designed to realize isolated energy conversion. It absorbs active power at LF and outputs it in MF form to the isolation transformer, *vice versa*. And the topology of such a SST is different from the traditional MMC, hence the modeling and controller design need to be reconsidered.

#### B. Controller Comparison

Actually, apart from the PI controller based on d-q transformation used in this paper, resonant controller can also be utilized as the current loop controller. Both of them have their pros and cons. Resonant controller has high gain to specific frequency component and thus, has good tracking performance at the selected frequency. However, resonant controller has high requirement on the discretization process and requires higher sampling frequency [12]. Using PI controller is much simple, but it should be associated with d-q transformation when applying on three-phase system.

For three-phase application, d-q transformation is inevitable to realize instantaneous powers monitoring and feedback. Therefore, using PI controller is much more simple and straightforward. In many literatures, d-q transformation plus PI controller is preferred for instantaneous power control while resonant controller can be applied for circulating current suppression [13, 14]. Since the goals of our application is to control the power of LF and MF ports, applying dual d-qtransformation plus PI controller for different control channels is more convenient.

#### VIII. CONCLUSION

A mathematical model for the SST with high integration level is proposed in this paper to describe the dynamics of the state variables at different frequencies. Based on the model, a control system with inner current loop controllers and outer energy balancing controllers is designed. The current loop controllers are designed based on the dual *d-q* reference model to provide good tracking performance of the LF and MF currents. The controller parameters are properly set to makes the controller immune to the disturbance at the other frequency. And the energy balancing is achieved by the outer controllers to change the current references.

Simulations and experiments are performed to verify the proposed control system. The results show good decoupling between d and q channels with the proposed control method.

And the dynamics response of each channel is fast without overshoot.

The controllers designed in this paper do not consider the fault conditions yet. To enhance fault ride-through ability, extra implementations should be added. For example, when the LVDC short-circuit occurs, the secondary voltages  $v_{sy}$  (y=a, b or c) would be clamped to zero. In this case, the MF voltages  $v_{MFx}$  should be reduced to match with  $v_{sy}$ , so as to limit the fault current; and when the DC bus of MMC is short-circuit, the SM should be protected by extra protection circuit such as thyristors [15]-[17]. The detail analysis and design will be done as the future work.

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