

Model Predictive Current Control for High-frequency Link Matrix DC/AC Converter with Loosely Coupled Transformer

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Abstract—Among various DC/AC matrix converters, the isolated single-phase DC to three-phase AC matrix converter has attracted significant research attention due to its advantages of low loss, simple topology and modulation. PI control is widely used in industrial manufacturing, but the increase in control target error and environmental factors will lead to decline of control system accuracy. We introduce a finite-control-set model predictive control (FCS-MPC) to predict the d- and q-axis currents in a high-frequency link matrix DC/AC converter (HFLMC) with a loosely coupled high-frequency transformer (LCHFT). The modulation and commutation modes of the HFLMC are introduced, transmission power of the HFLMC with air gap is given, and the output side compensated predictive model is also established. Design of LCHFT is given. The coupling coefficient of LCHFT under different air gap affects transmission efficiency. Finally, experiment results on a 300 W prototype verify correctness of the theoretical analysis and simulation.

Index Terms—High-frequency link matrix DC/AC converter (HFLMC), loosely coupled high-frequency transformer (LCHFT), model predictive control (MPC), PI control.

I. INTRODUCTION

RECENTLY, wireless power transmission (WPT) effectively solves the problem that power supply is difficult to access conveniently and safely [1]–[3]. The loosely coupled high-frequency transformer (LCHFT) is used to replace the brush and slip ring in rotor power supply of doubly fed machines, which effectively solves a series of problems during operation of the motor [4], [5]. DC/AC converter is used as the driving circuit of motor to realize different forms of energy conversion. High-frequency transformers (HFT) are used to enhance reliability and dynamic performance of the

system [6], [7]. In a high-frequency link matrix DC/AC converter (HFLMC), full bridges in the front stage convert DC voltage to a high-frequency AC voltage at the primary side. A cascaded HFAC-AC with HFAC-DC and DC-AC is employed on the secondary side [8]. However, large filtering capacitors are necessary in the DC bus in the intermediate rectification process, which makes the whole system have large power loss and low power density [2].

Recently, a single-stage HFAC-AC matrix converter has been researched to reduce efficiency loss and increase power density. Mahlein [9] first introduced a 3-phase HFAC to 3-way AC matrix converter (3-3 matrix converter, 3-3MC). Compared with the 3-3MC, the single-phase HFAC to three-phase AC converter (1-3 matrix converter, 1-3MC) [10] can simplify topology and facilitate modulation, and is especially suitable for application in the transformer second side [11]. That is, HFLMC with an H bridge on the HFT's primary side and a 1-3MC on the HFT secondary side is attractive.

Up to now, research of the HFLMC has mostly focused on modulation of switching signal and optimal design of the circuit. Hu [12] presented a four-segment space vector PWM (4S-SVPWM) scheme by coordinating the front and back stage PWM signals to minimize switching loss and eliminate DC bias. Yan [13] proposed a SPWM modulation method to realize the decoupling operation. Sayed [14] proposed a space vector modulation with variable switching period to improve transmission efficiency. Fang [15] changed the zero-vector action time in bipolar modulation to mitigate low order harmonics. Regarding topology optimization [16], [17], parameters of components were analyzed and selected, and topology loss performance was analyzed to achieve higher operation efficiency.

In the field of motor drive, LCHFT with small gaps has been used in synchronous motors [18]. A brushless doubly-fed motor with a CLLC converter and a rotating LCHFT was presented in [19]. The secondary DC bus is on the rotor, and the large filter capacitor impacts stability and security of the motor rotor system. In this paper, a topology of HFLMC with an LCHFT is proposed.

Traditional control modes of DC/AC converters mainly include PI control [20], PI control with disturbance compensation [21], model predictive control (MPC) [22]–[24], and so on. In HFLMC, PI control was adopted in [25]–[27]. However,

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the MPC has not been specifically applied to the HFLMC, as yet. This paper is to introduce an MPC with steady-state compensation to the proposed HFLMC with an LCHFT.

The rest of this paper is organized as follows: Section II presents topology of the proposed HFLMC with a LCHFT, and introduces the SVM modulation method, working principle, commutation method and power transmission with coupling coefficient. Section III establishes the mathematical model of the converter, gives the prediction model with compensation. Section IV compares simulation results with PI control and FCS-MPC. To validate analysis and simulations, Section V presents a prototype and experimental results, followed by the conclusion.

II. MAIN TOPOLOGY AND MODES OF OPERATION

A. Main Topology

Figure 1 shows the topology of the HFLMC with an LCHFT. HFLMC consists of a full bridge, an LCHFT, a 1-3MC, and an LC filter. The full bridge contains four switch tubes (S_1 – S_4). L_k represents auxiliary inductance and leakage of the primary side of the transformer, the number of turns on both sides is the same. There are six bidirectional switches S_{xy} ($x = a, b, c, y = p, n$), which constitute 1-3MC. Each group of bidirectional switches comprises a pair of reverse series switches, upper switch S_{xy+} , and lower switch S_{xy-} .

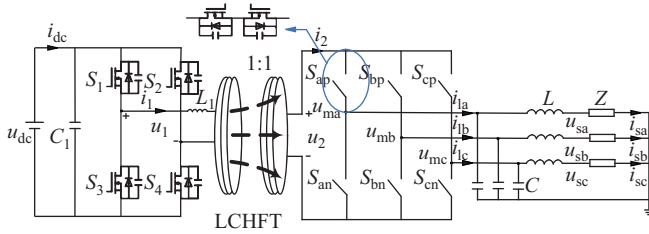


Fig. 1. The topology of the HFLMC.

C_1 represents the DC filter capacitor, i_{dc} and u_{dc} represent DC current and voltage, respectively. i_{1a} , i_{1b} , i_{1c} and u_{ma} , u_{mb} , u_{mc} represent phase currents and voltages of 1-3MC side, Z represents the output impedance, L and C represent the AC filter inductance and capacitor, i_{sa} , i_{sb} , i_{sc} and u_{sa} , u_{sb} , u_{sc} represent output currents and voltages, respectively. u_1 and i_1 represent transformer primary voltage and current, u_2 and i_2 represent transformer secondary voltage and current, respectively.

DC input voltage is transformed to bipolar HFT primary voltage through the full bridge converter. Electric energy is transmitted through the primary and secondary sides of the LCHFT. Bipolar HFT secondary voltage is transformed into low-frequency voltage by the 1-3MC. Ripples in the input DC voltage and high-frequency harmonics are eliminated by the input capacitor and output LC filter, respectively.

B. Space Vector Modulation

To avoid short circuits in the load side and open circuit on the transformer, switching functions and constraints of each

switch group are given as follows:

$$\begin{cases} S_{ap} + S_{bp} + S_{cp} = 1 \\ S_{an} + S_{bn} + S_{cn} = 1 \end{cases} \quad (1)$$

where $S_{xy} = 1, 0$ (turn on, turn off).

Figure 2 shows the current space vector synthesis of the 1-3MC. Fig. 2(a) and (b) represent the space vector of the upper-arms and lower-arms, respectively. Directions of these vector states in Fig. 2(a) and (b) are opposite, and their amplitudes are the same. When current on the transformer secondary side is positive, Fig. 2(a) is employed to generate switching vectors; when current on the transformer secondary side is negative, Fig. 2(b) is employed to generate switching vectors. For example, i_1 contains $i_{1+}(i_{ab+})$ and $i_{1-}(i_{ab-})$, where i_{1+} generates u_{ab} voltage while i_{1-} generates u_{ba} voltage. Table I shows each vector's corresponding switching states and output voltage.

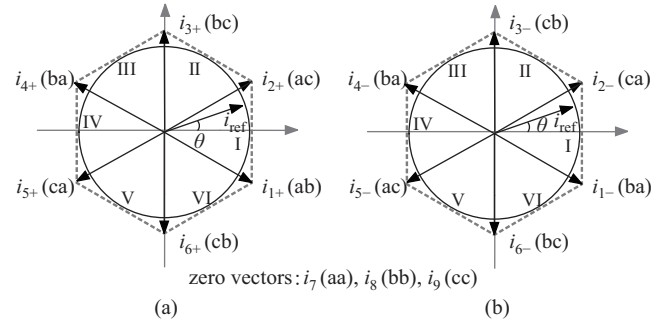


Fig. 2. Current space vector synthesis. (a) Positive sector. (b) Negative sector.

TABLE I
SPACE VECTOR CONTROL OF HFLMC

Space vectors	Switching gates						Output voltage
	S_{ap}	S_{an}	S_{bp}	S_{bn}	S_{cp}	S_{cn}	
i_1+/i_4-	1	0	0	1	0	0	u_{ab}
i_2+/i_5-	1	0	0	0	0	1	u_{ac}
i_3+/i_6-	0	0	1	0	0	1	u_{bc}
i_4+/i_1-	0	1	1	0	0	0	u_{ba}
i_5+/i_2-	0	1	0	0	1	0	u_{ca}
i_6+/i_3-	0	0	0	1	1	0	u_{cb}
i_7	1	1	0	0	0	0	0
i_8	0	0	0	0	1	1	0
i_9	0	0	1	1	0	0	0

Load output voltage u and current i are given as follows:

$$u = \begin{bmatrix} u_{sa} \\ u_{sb} \\ u_{sc} \end{bmatrix} = U_m \begin{bmatrix} \cos(\omega t) \\ \cos(\omega t - 2\pi/3) \\ \cos(\omega t + 2\pi/3) \end{bmatrix} \quad (2)$$

$$i = \begin{bmatrix} i_{sa} \\ i_{sb} \\ i_{sc} \end{bmatrix} = I_m \begin{bmatrix} \cos(\omega t) \\ \cos(\omega t - 2\pi/3) \\ \cos(\omega t + 2\pi/3) \end{bmatrix} \quad (3)$$

where ω represents angular frequency, U_m and I_m represent amplitude of output phase voltage and phase current, respectively.

For sector I, output side line voltage can be expressed as:

$$\begin{bmatrix} u_{ab} \\ u_{ac} \end{bmatrix} = \sqrt{3}U_m \begin{bmatrix} \cos(\omega t + \pi/6) \\ \cos(\omega t - \pi/6) \end{bmatrix} \quad (4)$$

When the ON-OFF state changes, the closed switch group is turned on, and another opened bidirectional switch group is turned off. According to Fig. 2, there are six effective switching vectors and three zero vectors. i_{ref} can be composed of two effective switching vectors and one zero vector.

Figure 3 shows a cycle of the transformer primary side switch signal, voltage and secondary side switch signal, voltage and inductive voltage, current of one cycle in sector I. In the first half of the cycle, reference current vector is synthesized by vectors i_{ab} , i_{ac} , i_0 . In the second half of the cycle, vectors i_{ba} , i_{ca} , and i_0 are synthesized, and corresponding duty cycles of bidirectional switches from S_{ap} to S_{cn} are $0.5T_{02}$, T_{ab} , T_{ac} , T_{01} , T_{ba} , T_{ca} , $0.5T_{02}$, respectively.

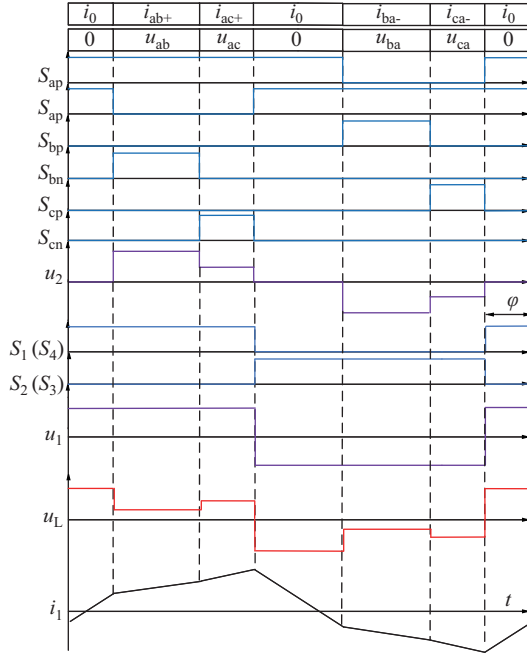


Fig. 3. Primary switch signal and voltage, secondary switch signal and voltage, inductive voltage and current in sector I.

Duty calculation is given in (5). Since amplitudes of effective vectors are the same in the first half cycle and the second one, the duty cycle of each switch vector in the first half cycle can be calculated as:

$$\begin{cases} T_{ab} = T_{ba} = \frac{1}{2}mT \sin\left(\frac{\pi}{6} - \theta\right) \\ T_{ac} = T_{ca} = \frac{1}{2}mT \sin\left(\frac{\pi}{6} + \theta\right) \\ T_{01} = T_{02} = \frac{T}{2} - T_{ab} - T_{ac} \end{cases} \quad (5)$$

where current reference angle θ represents the angular displacement between the angle of the α -axis and the current reference, m represents the modulation coefficient of HFLMC, and T represents switching time of one cycle.

For the full-bridge converter, duty cycle of the conduction signal is set to 48%, and dead time is 2%, correspondingly. Conduction signals on Switches S_1 and S_3 are the same, and so are Switches S_2 and S_4 .

C. Four-step Commutation

Safe commutation is one of the important valuation criteria of matrix converters. If the current type commutation

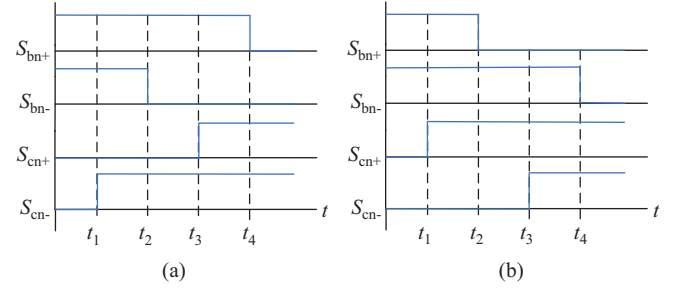


Fig. 4. Four-step commutation process. (a) $u_{ac} > u_{ab}$. (b) $u_{ac} < u_{ab}$.

strategy is adopted, the output zero vector of the bipolar vector modulation would not judge current polarity, resulting in failure of commutation. Therefore, voltage type four-step commutation strategy is adopted in this paper. Fig. 4 shows a set of commutation processes.

When $u_{mb} > u_{mc}$, as shown in Fig. 4(a), before time t_1 , both S_{bn+} and S_{bn-} are on, while both S_{cn+} and S_{cn-} are off; at t_1 , S_{cn-} is turned on; at t_2 , S_{bn-} is turned off; at t_3 , S_{cn+} is turned on; at t_4 , S_{bn+} is turned off and the commutation process is completed. During the process, each switch alternatively performs a turn-on operation and a turn-off operation, and there are not inter-phase short circuits and open circuits in the converter.

Alternatively, four-step commutation can be simplified to two-step commutation proposed by Ziegler and Hofmann [16]. In four-step commutation, the first two steps are simplified to one step, and the last two steps are simplified to one step. Two-step commutation greatly increases the proportion of free commutation and reduces switching stress during vector switching.

D. Transmission Efficiency with Coupling Coefficient

Application of CLLC converter and a LCHFT in doubly-fed motor is introduced in [19]. Capacity of the intermediate DC filter capacitor is large, and rotation of the motor will produce greater inertia. The converter with high power density is adopted, and the LCHFT is used on the rotor of doubly fed machine instead of ordinary HFT.

The u_1 on the primary side of the transformer in Fig. 3 is analyzed by Fourier transform

$$u_1 = \frac{4}{\pi} u_{dc} \sum_{n=1,3,5}^{\infty} \frac{1}{n} \sin(2n\omega_s t) \quad (6)$$

The RMS of u_1 is

$$U_1 = \sum_{n=1,3,5}^{\infty} \frac{2\sqrt{2}}{n\pi} u_{dc} \quad (7)$$

where n represents the order of harmonics, and ω_s represents switching angular frequency.

When using loosely coupled transformers, different air gaps may lead to different coupling coefficients k between the two sides of the transformer. In the first sector, the two effective vectors are u_{ab} and u_{ac} , and the secondary side voltage u_2

can be calculated by

$$u_2 = \sum_{n=1}^{\infty} \{U_n \sin(n\omega_s t - n\varphi)\} \quad (8)$$

where φ represents the lead angle of the primary voltage of the transformer relative to the secondary voltage.

The corresponding secondary voltage amplitude can be expressed as:

$$U_n = \frac{2k \cos(n\pi)}{n\pi} \left\{ u_{ab} \begin{pmatrix} \cos(n\omega_s(T_{02} + T_{ab} + T_{ac})/2) \\ -\cos(n\omega_s(T_{02} + T_{ac})/2) \end{pmatrix} \right. \\ \left. + u_{ac} \begin{pmatrix} \cos(n\omega_s(T_{02} + T_{ac})/2) \\ -\cos(n\omega_s T_{02}/2) \end{pmatrix} \right\} \quad (9)$$

By bringing (4), (5) and (9) to (8), u_2 can be rewritten as:

$$u_2 = \sum_{n=1}^{\infty} \left\{ \frac{2\sqrt{3}kU_m \cos(n\pi)}{n\pi} \left(A \sin\left(\frac{\pi}{3} - \theta\right) + B \sin\theta - C \sin\left(\frac{\pi}{3} + \theta\right) \right) \sin(n\omega_s t - n\varphi) \right\} \quad (10)$$

where

$$A = \sin\left(\frac{\pi}{3} - \theta\right) \cos\left(\frac{n\pi}{2}\right) \\ B = \cos\left(\frac{n\pi}{2} \left(1 - m \sin\left(\theta - \frac{\pi}{6}\right)\right)\right) \\ C = \cos\left(\frac{n\pi}{2} (1 - m \cos\theta)\right)$$

The RMS of u_2 can be given as:

$$U_2 = \sum_{n=1,3,5}^{\infty} \frac{U_n [\cos(n\varphi) - j \sin(n\varphi)]}{\sqrt{2}} \quad (11)$$

Apparent power S can be calculated from the RMS of the primary voltage and primary current flowing through L_k by

$$S = U_1 I_1 = U_1 \frac{U_1 - NU_2}{\omega_s L_k} = P_t + jQ_t \quad (12)$$

Then, powers of the HFLMC are given by

$$P_t = \sum_{n=1,3,5}^{\infty} \frac{8u_{dc}^2 - 2n\pi U_n u_{dc} \cos(n\varphi)}{(n\pi)^2 \omega_s L_k} \quad (13)$$

$$Q_t = \sum_{n=1,3,5}^{\infty} \frac{2U_n u_{dc} \sin(n\varphi)}{(n\pi)^2 \omega_s L_k} \quad (14)$$

By neglecting the high-order harmonics, transmission power P_1 in one cycle can be expressed as:

$$P_1 = \frac{8u_{dc}^2 - 2\pi U_n u_{dc} \cos\varphi}{\pi^2 \omega_s L_k} \quad (15)$$

III. PROPOSED APPROACH

A. Mathematical Model of Converter

As the AC side is connected to the load through a filter inductor and filter capacitor in this paper, corresponding AC side voltage and current equations can be expressed as

$$\begin{cases} u_{mx} = L \frac{di_{sx}}{dt} + u_{sx} + r i_{sx} \\ C \frac{du_{mx}}{dt} = i_{lx} - i_{sx} \end{cases} \quad (16)$$

where r is the equivalent resistance of each phase on the AC side.

In order to simplify the mathematical model of HFLMC, the rotating coordinate system is employed, and output voltage and current equations are given as:

$$\begin{cases} \frac{d}{dt} i_{sd} = \omega i_{sq} + \frac{1}{L} u_{md} - \frac{1}{L} u_{sd} - \frac{r}{L} i_{sd} \\ \frac{d}{dt} i_{sq} = -\omega i_{sd} + \frac{1}{L} u_{mq} - \frac{1}{L} u_{sq} - \frac{r}{L} i_{sq} \end{cases} \quad (17)$$

$$\begin{cases} \frac{du_{md}}{dt} = \frac{1}{C} i_{ld} + \omega u_{mq} - \frac{1}{C} i_{sd} \\ \frac{du_{mq}}{dt} = \frac{1}{C} i_{lq} - \omega u_{md} - \frac{1}{C} i_{sq} \end{cases} \quad (18)$$

where u_{sd} , u_{sq} and i_{sd} , i_{sq} represent phase voltage and phase current components of load side, respectively, u_{md} , u_{mq} and i_{ld} , i_{lq} represent input phase voltage and phase current components, respectively.

MPC can predict the current of the matrix converter quickly and robustly. In this paper, the i_{ld} and i_{lq} of the matrix converter are controlled by MPC. Combined with formula (17), (18), the discrete equation under predictive control is obtained as:

$$\begin{cases} i_{sd}(k+1) = -\frac{1}{\omega L} \Delta u_{mq}(k+1) - \frac{T}{L} u_{sd}(k) + \frac{T}{\omega C L} i_{lq}(k) \\ \quad + \left(1 - \frac{rT}{L}\right) i_{sd}(k) + \left(\omega - \frac{1}{\omega C L}\right) T i_{sq}(k) \\ i_{sq}(k+1) = \frac{1}{\omega L} \Delta u_{md}(k+1) - \frac{T}{L} u_{sq}(k) - \frac{T}{\omega C L} i_{ld}(k) \\ \quad + \left(1 - \frac{rT}{L}\right) i_{sq}(k) + \left(-\omega + \frac{1}{\omega C L}\right) T i_{sd}(k) \end{cases} \quad (19)$$

where $u_{sd(k+1)}$ and $u_{sq(k+1)}$ represent predicted voltage components of output side at $k+1$ time, $u_{sd(k)}$ and $u_{sq(k)}$ represent measured voltage components of output side at k time, $i_{sd(k)}$ and $i_{sq(k)}$ represent measured current components of output side at k time, $i_{ld(k)}$ and $i_{lq(k)}$ represent calculated current components of the 1-3MC side at k time. $\Delta u_{md(k+1)}$ and $\Delta u_{mq(k+1)}$ represent calculated difference voltage components of the 1-3MC side at $k+1$ time.

Since the value of $\Delta u_{md(k+1)}$, $\Delta u_{mq(k+1)}$ cannot be obtained at $k+1$ time, the difference between k time and $k-1$ time is used to express as:

$$\begin{cases} \Delta u_{md}(k+1) \approx u_{md}(k) - u_{md}(k-1) \\ \Delta u_{mq}(k+1) \approx u_{mq}(k) - u_{mq}(k-1) \end{cases} \quad (20)$$

where $\Delta u_{md(k)}$, $\Delta u_{mq(k)}$ and $\Delta u_{md(k-1)}$, $\Delta u_{mq(k-1)}$ represent the calculated difference voltage components of the 1-3MC at k time and $k-1$ time.

Steady-state compensation quantities at k time of d axis and q axis are respectively, expressed as:

$$\begin{cases} f_d(k) = e_1 (i_{sd}(k) - i_{sdm}(k)) \\ f_q(k) = e_2 (i_{sq}(k) - i_{sqm}(k)) \end{cases} \quad (21)$$

where e_1 , e_2 are the steady-state compensation coefficients in the d axis and q axis, respectively.

The mathematical model of the load side can be expressed as:

$$\begin{cases} i_{sdy}(k+1) = i_{sdm}(k) + f_d(k) \\ i_{sqy}(k+1) = i_{sqm}(k) + f_q(k) \end{cases} \quad (22)$$

where $i_{sd(k+1)}$, $i_{sq(k+1)}$ represent the predicted current components of output side at $k+1$ time after introducing steady-state compensation.

B. Prediction Model of Converter

In order to make the output result of the prediction module of HFLMC closer to the given reference current and make the current stress of the secondary side smaller, the cost function is

$$J = \min \left\{ m_1 [i_{sd}(k+1) - i_{sd}^*(k+1)]^2 + n_1 [i_{sq}(k+1) - i_{sq}^*(k+1)]^2 + m_2 [i_{ld}(k)]^2 + n_2 [i_{lq}(k)]^2 \right\} \quad (23)$$

where $i_{sd}^*(k+1)$ and $i_{sq}^*(k+1)$ are the d-axis and q-axis reference currents on the load side at $k+1$ time, respectively; m_1 and m_2 are output tracking control factors, respectively; n_1 and n_2 represent constraint factors of the control quantity, respectively.

To minimize differences of reference and prediction current on the load side and obtain the stress of the converter side current of the HFLMC, the minimum value of the cost function can be given by

$$\frac{\partial J}{\partial i_{ld}} = 0, \quad \frac{\partial J}{\partial i_{lq}} = 0 \quad (24)$$

Then, the minimum currents of the HFLMC can be expressed as:

$$i_{ld} = \frac{-cn_2 i_{sq}^*(k+1) + cn_2 \{-a[u_{md}(k) - u_{md}(k-1)] - bu_{sq}(k) + di_{sq}(k) - gi_{sd}(k) + f_q(k)\}}{n_1 c^2 + m_2} \quad (25)$$

$$i_{lq} = \frac{-cn_1 i_{sd}^*(k+1) - cn_1 \{-a[u_{mq}(k) - u_{mq}(k-1)] \cdot i_{sd}(k) + bu_{sd}(k) - di_{sd}(k) + gi_{sq}(k) + f_d(k)\}}{n_2 c^2 + m_1} \quad (26)$$

where $a = \frac{1}{\omega L}$, $b = \frac{T}{L}$, $c = \frac{T}{\omega CL}$, $d = 1 - \frac{rT}{L}$, $g = (\omega - \frac{1}{\omega CL})T$.

C. Control Diagram of the Converter

Control strategy of MPC with steady-state compensation is given in Fig. 5. Output phase u_{sx} and i_{sx} are sampled. Three-phase u_{sx} are converted to u_{sd} , u_{sq} and three-phase i_{sx} to i_{sd} , i_{sq} through the Park transformation. Basic MPC is established, and steady-state compensation of output side prediction results is carried. The minimum cost function with control quantity constraint is given. Optimal m of HFLMC is obtained by minimizing the current of 1-3MC. The appropriate φ between the front and rear stages of HFLMC is set as in [14] to guarantee control requirements of system energy transmission.

Control strategy of PI is given in Fig. 6. The whole control system is designed according to the closed-loop transfer function to ensure high control accuracy of the whole HFLMC.

IV. SIMULATION RESULTS

Conventionally, PI controller is used in the matrix converter to control d - and q -axis current. However, with PI controller, the transient process of load changes is too long, which decreases stability of the whole system. In this section, two

kinds of control of HFLMC with LCHFT are simulated using MATLAB/Simulink model. Comparison results of the steady-state and dynamic response under the two kinds of control are given. Corresponding parameters are listed in Table II.

Figure 7(a) and (b) show transient processes of the converter with PI controller and MPC controller when output reference current changes from 1 to 2 A at 0.1 s. Both controllers can realize good agreement between reference current and measured one. Transient processes cost 8.7 ms in the former and 2.5 ms in the latter, which shows better dynamic response of the latter.

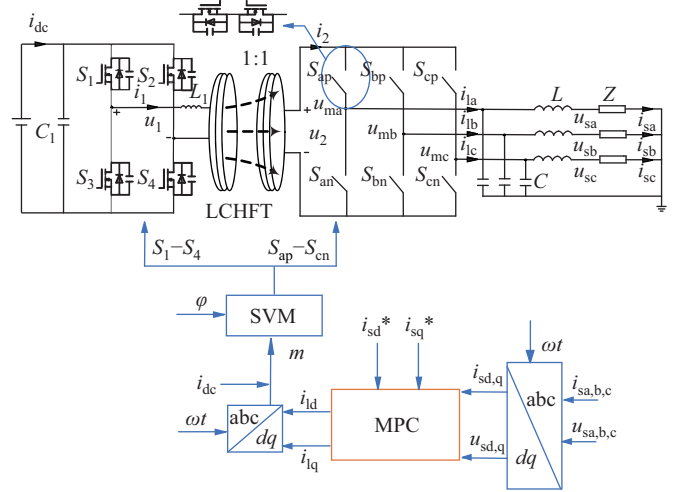


Fig. 5. The control strategy of MPC with steady-state compensation.

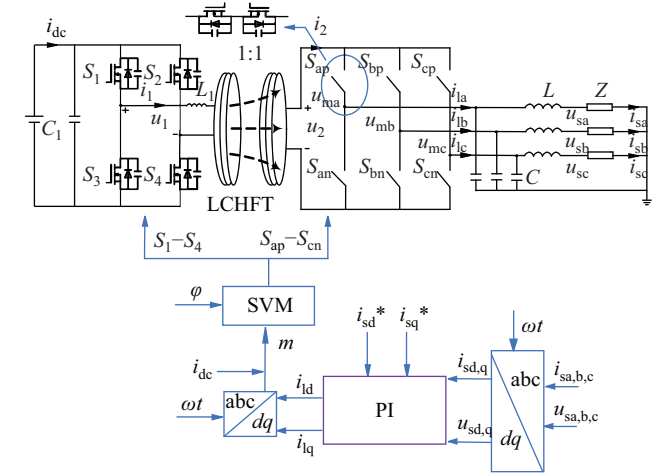


Fig. 6. The control strategy of PI.

TABLE II
SIMULATION PARAMETERS

Parameters	Value
DC input voltage (U_{dc})	70 V
Leakage inductance (L_k)	0.23 mH
Transformer ratio (N)	1:1
DC filter capacitor (C_1)	3600 uF
AC filter inductor (L_f)	2.5 mH
AC filter capacitor (C_f)	12 uF
Load (R)	10 Ω , 20 Ω
Switching frequency (f_s)	10 kHz

Figure 8(a) and (b) show the transient processes of the converter with PI controller and MPC controller when output reference current is constant and output three-phase resistance of the load is changed from 10 to 20 Ω at 0.1 s. It can be seen that output current stabilizes at 1 A, and the transient process of the voltage changing are 7.6 ms with PI controller and 1.3 ms with MPC controller.

Figure 9 shows output current i_d under different controls when reference current is 1 A. Fig. 9(a) shows PI control has good stability. Fig. 9(b) and Fig. 9(c) show MPC without steady-state compensation and with steady-state compensation. MPC without compensation will cause deviation of output current i_d relative to the reference current during load switch-

ing. During vector switching, the harmonic content will be increased. Adding compensated MPC can ensure sufficient system control accuracy and reduce harmonic content.

V. EXPERIMENTAL RESULTS

A 300 W prototype is manufactured to verify correctness of the above theoretical analysis and simulation results. Fig. 10 shows the experiment platform and prototype. MOSFET (4227) is used as the switching device of the whole HFLMC. The DSP28379 and CPLD EPM570 platforms are employed simultaneously to generate driving signals and complete the SVM modulation strategy. Experimental parameters are the same as those in the simulation.

Size design of magnetic core material is calculated by the AP method.

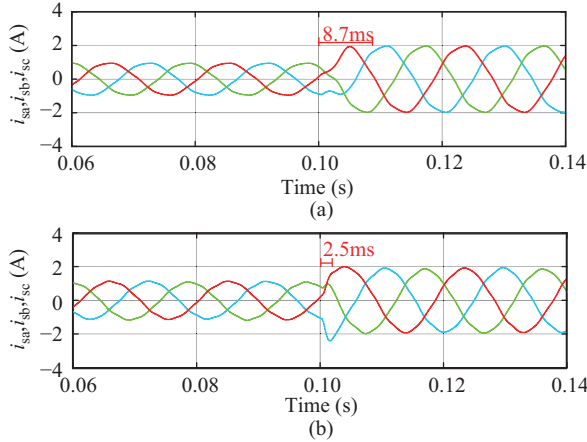


Fig. 7. Output current reference value changed from 1 A to 2 A. (a) PI. (b) MPC.

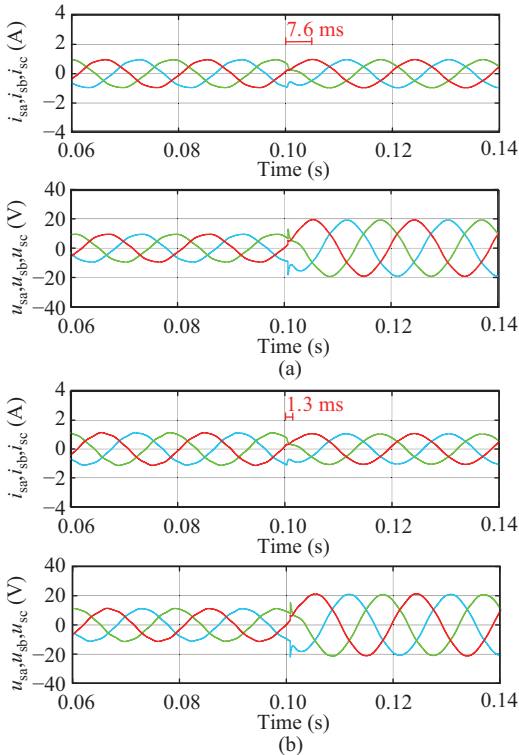


Fig. 8. Output three-phase current and voltage of load switched from 10 Ω to 20 Ω . (a) PI. (b) MPC.

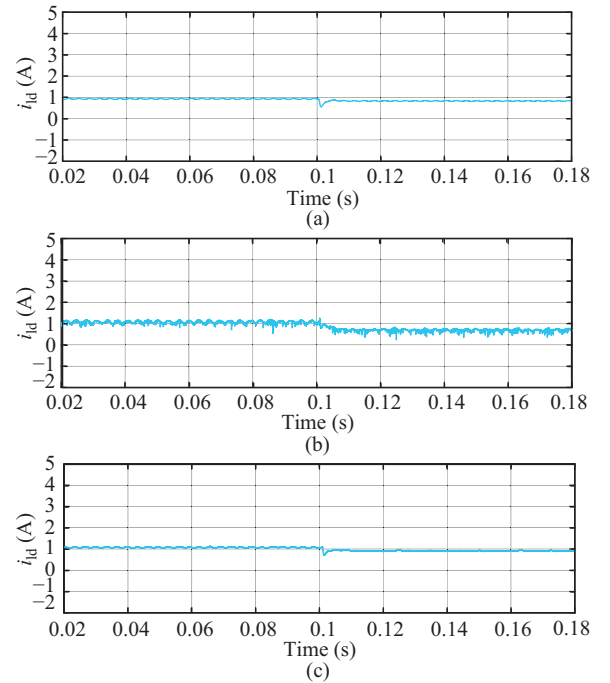


Fig. 9. Output current i_d during load switching. (a) PI. (b) Uncompensated MPC. (c) Compensated MPC.

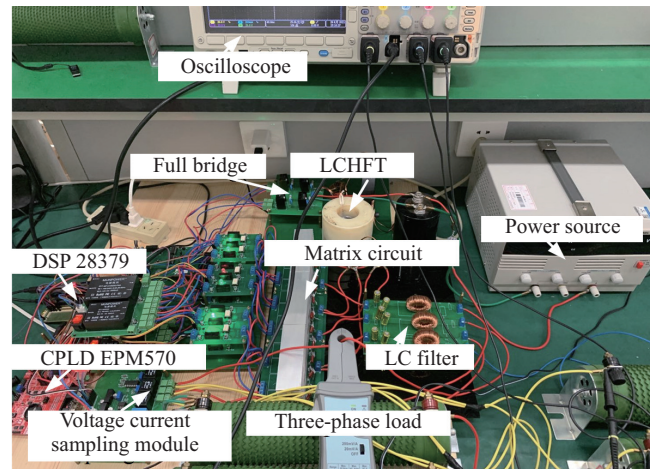


Fig. 10. The prototype of HFLMC with LCHFT.

$$A_p = A_c W_a = \frac{P_t \times 10^4}{4.44 f_s B_m k_f j k_u} \quad (27)$$

where A_c is the geometric cross-sectional area of the magnetic core, W_a is the window area of the magnetic core, P_t is the apparent power, f_s is the working frequency, B_m is the maximum magnetic flux density of the magnetic core, k_f is the filling coefficient of the magnetic core material, j is current density and k_u is the utilization coefficient of the magnetic core window.

Existence of an air gap will produce a large amount of leakage inductance and affect the coupling coefficient. Therefore, selection of air gap size is particularly important for transmission efficiency of the high-frequency transformer. Results show that air gap size is negatively correlated with the coupling coefficient.

In order to obtain the existing window utilization coefficient of LCHFT, we introduce the correction coefficient k_g to correct window utilization coefficient k_u

$$k_{u(g)} = k_g \times k_u = \frac{L_1 \times 2}{L_1 \times 2 + g} k_u \quad (28)$$

where k_g represents the coefficient of reduction of window utilization coefficient caused by existence of air gap, L_1 is self inductance at the primary side of transformer, and g is air gap distance.

Physical diagram of LCHFT is shown in Fig. 11. The LCHFT has two pot-shaped PC40 material ferrite cores and two winding with 0.2*35 Litz wire. The air gap between the cores changes from 0 to 3 mm in our experiment. Detailed parameters of the LCHFT are listed in Table III.

In the experiment, DC input voltage is 70 V and load is 10 Ω . Waveforms of the load side output voltage u_a , transformer primary voltage u_1 , current and secondary voltage

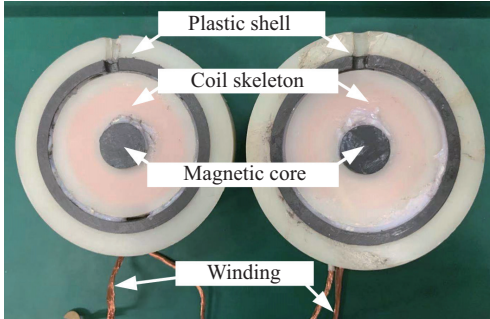


Fig. 11. Physical diagram of the loosely coupled transformer.

TABLE III
PARAMETERS OF THE LCHFT

Parameters	Value
Area A_p	258.23 cm ²
Core cross-sectional area A_e	23.39 cm ²
The window area A_w	11.04 cm ²
Outer diameter of the core	72
Height of the core	22
Outer diameter of the winding	1.53 mm
Cross-sectional area of the winding	1.1 mm ²
Turns of winding	3×6
Turns of primary winding N_p	18
Turns of secondary winding N_s	18

u_2 corresponding to the reference currents under PI control and MPC respectively are shown in Figs. 12 and 13.

When the current reference value is 1 A (or 2 A), the main characteristics of the two control methods are similar, including achievement of accurate output voltages and currents, constant of the φ , and increase of m and voltage wave duty ratio. It can be seen m of secondary side voltage changes from 0.28 to 0.45 with PI controller and 0.29 to 0.45 with MPC controller. Their power quality is different. Fig. 14 shows harmonic analysis of the secondary voltage of a transformer with two controls.

When the three-phase load is switched from 10 Ω to 20 Ω , three-phase output voltage and the b-phase output current

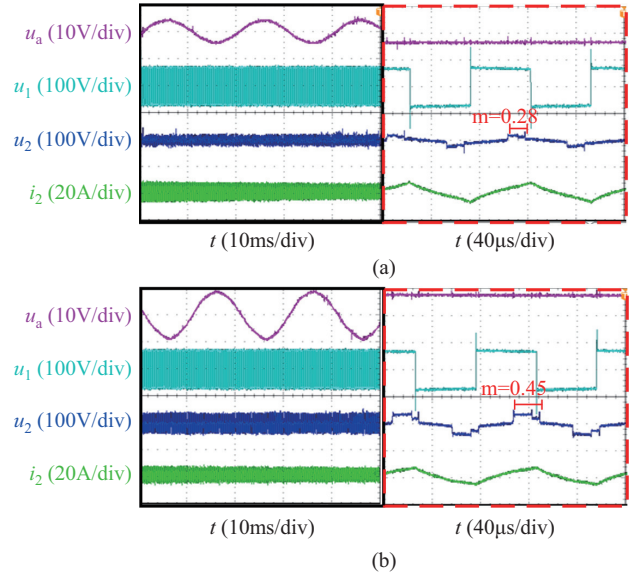


Fig. 12. Output voltage in phase A output voltage, transformer primary voltage, current, secondary voltage under PI control. (a) $u_{dc} = 70$ V, $R = 10$ Ω , $i_{ref} = 1$ A. (b) $u_{dc} = 70$ V, $R = 10$ Ω , $i_{ref} = 2$ A.

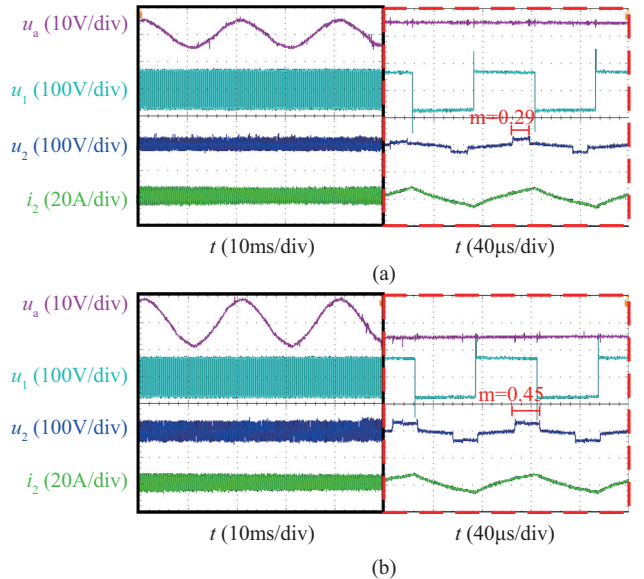


Fig. 13. Output voltage in phase A output voltage, transformer primary voltage, current, secondary voltage under MPC. (a) $u_{dc} = 70$ V, $R = 10$ Ω , $i_{ref} = 1$ A. (b) $u_{dc} = 70$ V, $R = 10$ Ω , $i_{ref} = 2$ A.

under PI control and MPC are shown in Fig. 15, respectively. Under PI control, the three-phase output voltage is changed from 10 to 20 V, and output current of b-phase reaches the stable state of 1 A within 12.1 ms, while under control of MPC, within 2 ms. Obviously, experiment results are consistent with simulations and MPC performs better than PI control in terms of dynamic response.

For these two schemes, harmonic order of the phase voltage is shown in Fig. 16. Harmonic contents under both controls are relatively low, with THD of 3.1% and 2.3% at loads of 10 Ω , respectively. Harmonic order of phase current is shown in Fig. 17. Output current has large current distortion, with

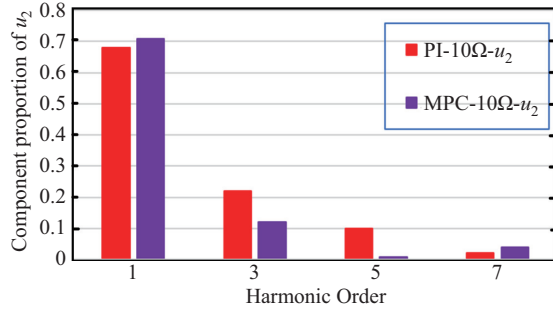


Fig. 14. Harmonic analysis of transformer secondary voltage. (a) PI. (b) MPC.

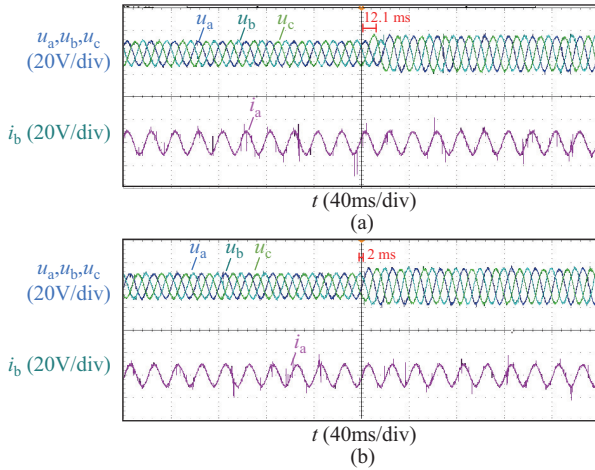


Fig. 15. The corresponding output voltage and b-phase output current under load switching. (a) PI. (b) MPC.

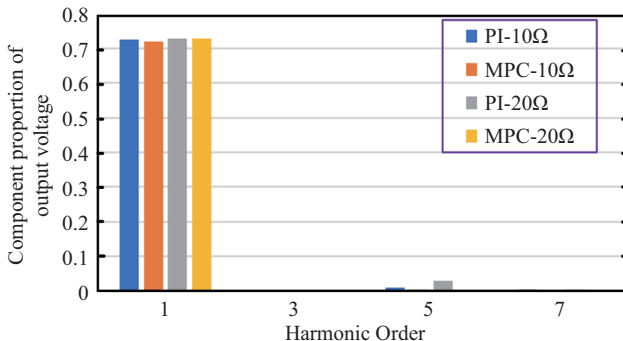


Fig. 16. The corresponding harmonic order of the phase voltage at $i_{ld}^* = 1$ A.

THD of 4.2% and 4.6% at loads of 10 Ω , respectively. Fifth and seventh harmonics in MPC controlled system are larger than those in PI. Fig. 18 shows the dynamic change of output power during load switching. MPC has shorter switching time and better dynamic performance during load switching.

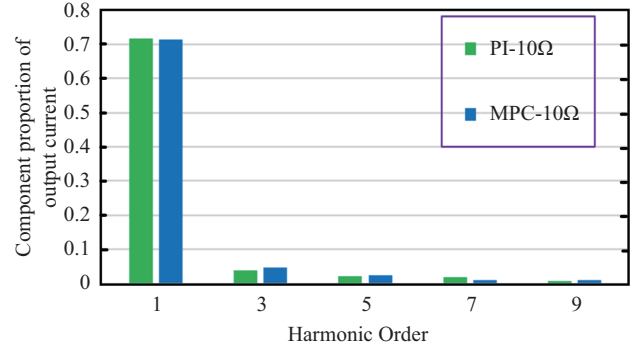


Fig. 17. The corresponding harmonic order of the phase current at $i_{ld}^* = 1$ A.

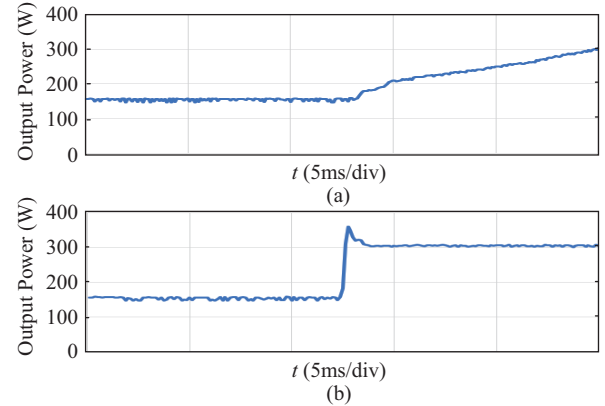


Fig. 18. Dynamic change of output power during load switching. (a) PI. (b) MPC.

The air gap of LCHFT affects its coupling coefficient, thus affecting bidirectional transmission efficiency of the HFLMC. Transmission efficiency of PI control and MPC under different air gaps is given in Fig. 19. When the air gap distance of the LCHFT is 0 mm, transmission efficiencies of PI control and MPC are 79.12% and 79.54%, respectively. When the air gap distance is 1 mm, transmission efficiency increases from goes up to 83.64% and 83.51% respectively. As the air gap continues to increase, the coupling coefficient of the transformer decreases and transmission efficiency is decreased. When the air gap is 3 mm, transmission efficiency of PI control and MPC are 75.34% and 76.02%, respectively.

Figure 19(b) illustrates the efficiency of different transformers at different input powers. Compared with ferrite transformers, a nanocrystalline transformer has higher transmission efficiency. When input power is 300 W, transmission efficiency is 91.12% and 87.93%, respectively.

VI. CONCLUSION

This paper presents a compensated MPC for HFLMC with LCHFT. An MPC output current tracking control method

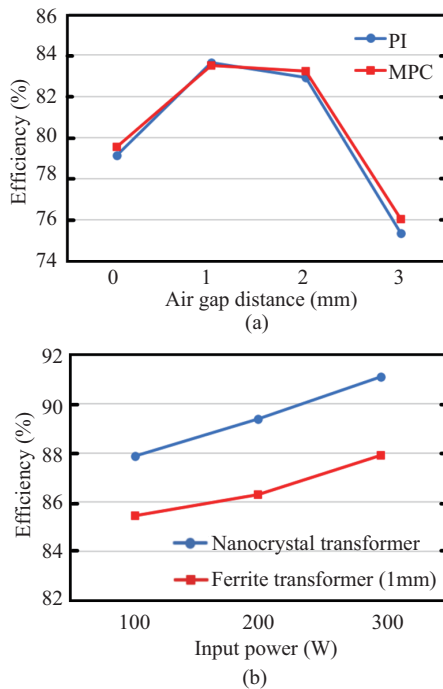


Fig. 19. Efficiency diagram. (a) The efficiency under PI and MPC control at different air gaps. (b) The efficiency of different transformers at different input power.

with steady-state compensation is proposed, and its steady-state and dynamic response are compared with PI control. The corresponding modulation method, commutation mode and power transmission with coupling coefficient were given. The corresponding mathematical model on the load side was established, and the prediction equation was obtained by discretization, and the optimal cost function was selected. The models predictive control with compensation has better stability and less harmonic content. When reference current and load change, MPC has better dynamic response and less overshoot, but increases harmonic content of output current. LCHFT with air gap has the ability to compensate leakage inductance and improve transmission efficiency. Experimental results show that HFLMC is feasible by using the LCHFT.

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