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Finite-Control-Set Model Predictive Control for Inductive Power Transfer Charging EV systems with Constant Voltage Load

Zeinab Karami* zeinab.k.h.abadi@ntnu.no
Jiayu Zhou* jiayu.zhou@ntnu.no
Giuseppe Guidi† Giuseppe.Guidi@sintef.no
Jon Are Suul*† jon.are.suul@ntnu.no
*Department of Engineering Cybernetics, *Norwegian University of Science and Technology,* Trondheim, Norway
†*SINTEF Energy Research,*

***Abstract*—This paper presents a finite-control-set model predictive control (FCS-MPC) strategy for achieving power control in inductive power transfer (IPT) -based electric vehicle (EV) charging systems. The proposed control method is applied to a series-series-compensated IPT system when charging a battery appearing as a constant voltage load (CVL). It has been previously shown that IPT systems with CVL can exhibit large current/power oscillations due to the low damping of a critical mode that can be excited by pulse skipping modulation techniques such as pulse density modulation (PDM). By predicting the next optimal switching states and the input power, the proposed control method will suppress the oscillations caused by pulse skipping operation, while ensuring a fast dynamic response and zero voltage switching (ZVS) operation. The effectiveness and performance of the proposed control method in comparison with a PI-based PDM method are verified by the simulation results.**

***Index Terms*—Constant voltage load, finite-control-set model predictive control, inductive power transfer, pulse skipping modulation**

I. Introduction

Inductive power transfer (IPT) technologies have become increasingly popular for a wide range of applications such as mobile devices, marine transport, and electric vehicles (EVs) [1]. It is critical for IPT systems to be able to accomplish soft switching and guarantee the control of power converters to achieve the maximum power efficiency. Furthermore, dynamic charging systems usually require a fast response during the power transfer process. Therefore, in order to design a control strategy with robust and fast dynamic response, an accurate model of the IPT system is required. Series-series (SS)-compensated IPT systems are often most suitable for high-power battery charging applications [2]. In addition, the use of a diode-rectifier on the receiving side is an effective option for minimizing the number of components, costs and losses of high-power systems. In [3] a nonlinear model of IPT system with a constant resistance load (CRL) is investigated for constant output voltage regulation.

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However, the configuration of IPT system with a diode rectifier interfaced directly to a battery results in a system with constant voltage load (CVL) characteristics. CVL or a slowly varying voltage load is the most common receiving side interface for battery charging in practical applications [4]. In this regard, a comparison of the state-space model of an IPT system with CRL and CVL is presented in [4], where the results show that the dynamic response of CVL is significantly different from an equivalent CRL model. Thus, the battery charging systems should be modeled in such a way that CVL characteristics can be accurately represented.

The PI-based phase-shift modulation (PSM) method is the most common technique for power control and voltage regulation among control methods in IPT charging systems [5]. However, this technique leads to increased losses and reduced system efficiency at low output voltage due to the hard switching. To deal with this challenge, the PI control method combined with pulse density modulation (PDM) is used in such systems to accomplish soft switching in the full operating range [6], [7]. These methods can always achieve zero voltage switching (ZVS) under different power conditions by skipping pulses for adjusting the average sending voltage. Nevertheless, a poorly damped oscillation mode of IPT systems with CVL can be excited by pulse-skipping modulations such as the PDM pattern, which can cause high current/power ripple [8]. To solve this problem, an active damping method to attenuate the oscillations caused by the skipped pulses of the PDM is presented in [8]. However, the active damping method transiently introduces a PSM, which implies hard switching and increased losses.

Recently, model predictive control (MPC) methods have been proposed for wireless power transfer (WPT) charging systems in order to improve the dynamic response of power control and voltage regulation [9]–[11]. MPC is an optimal control method that aims to solve an optimization problem with a fast dynamic response over a prediction horizon at each sampling step [12].

MPC has become a well-known technology in the research and development stages of power converter control due to the significant increase in the computing power of microprocessors, which can perform the large amount of calculations with high speed and lower cost [13]. There are several advantages to using this technology, including fast dynamic response, ease of implementation in multivariable systems, and the ability to use nested control loops in one loop in some conditions. Furthermore, this method takes into account non-linearity dynamics, uncertainty, and constraints of state variables and input variables [12].

Generally, MPC is divided into two categories: continuous control set MPC (CCS-MPC) and finite control set MPC (FCS-MPC). The CCS-MPC calculates a continuous control signal and then uses a modulator to generate the output voltage of converter. In contrast to the CCS-MPC, the FCS-MPC considers the switching behavior of the power converter to formulate the MPC algorithm and does not require an external modulator to evaluate each possible switch position [14].

This paper presents FCS-MPC in the IPT charging EV system by considering a nonlinear dynamic model of CVL. The main purpose of this paper is to regulate the power flow and determine the optimal ON or OFF state of the switches by minimizing the cost function at each sampling time. Furthermore, the proposed control method should guarantee the soft switching performance, while maintaining fast dynamic response with low current/power ripples.

In contrast to the PI-based control methods, the proposed FCS-MPC strategy provides a faster dynamic response and can effectively overcome the effects of disturbances, and nonlinearities, while reducing the current/power ripples and ensuring the ZVS operation [8]. Compared with the MPC-based control methods for IPT charging system, the proposed method can guarantee efficiency and power regulation by considering the nonlinear dynamic model of the system in the presence of CVL, which has been ignored in previous methods. Furthermore, the modulators such as PSM, PDM, etc., have been used in these papers to achieve the switching sequences for power converters [10], [11]. As opposed to the proposed FCS-MPC control method avoids the delay between the control and modulator stage by considering the switching characteristic in the control algorithm.

The FCS-MPC can provide accurate power with the low current/power ripple, however in the OFF state when zero voltage is applied, the input power fluctuates significantly. Therefore, a low-pass filter (LPF) is used in the cost function to ensure a minimum error between the predicted power and the power reference. By optimization the cost function over the control horizon, it is provided the low damping of the critical oscillation frequency caused by the CVL characteristics and ensures that the power transfer can be accurately regulated while overcoming the mentioned challenge.

The main advantages of proposed control method are as follows:

- Easy to implement zero voltage switching (ZVS),
- Low ripples compared to traditional PI+PDM,
- Fast dynamic response compared to traditional methods,
- Low computation cost due to avoiding need to modulator.

The validity and performance of the proposed FCS-MPC control method for IPT system with CVL compared to PI-PDM based on delta-sigma modulator (DSM) is confirmed by the simulation results.

II. Nonlinear Model of IPT System with CVL

Fig. 1(a) shows the general electric circuit model of the studied IPT charging system for EV, where the primary side (i.e., charging station) induces the secondary current for the receiving side. Then, the energy is transferred to the load through the high-frequency rectifier. In the proposed IPT system, a SS compensation network is considered due to the simple structure and constant resonant frequency. In addition, due to the fact that the FCS-MPC control method is a model-based controller, to determine the design of the proposed control method and achieve good performance, the IPT system should be represented by an accurate dynamic model. The proposed load in Fig. 1(a) can be considered as an equivalent CRL with respect to steady-state performance. However, in practical applications, the most common receiving side interface for battery charging will be as a CVL or a slowly varying voltage load, which is considered in this paper. To reduce system cost and complexity, a diode rectifier is used on the pickup side, in which the dc terminal is connected directly to a battery, which is modeled as a CVL (Vdc, out).

The equivalent circuit of the SS-compensated IPT charging system with CRL/CVL is shown in Fig. 1(b), which consist of a high-frequency full bridge inverter with a constant DC voltage source as input voltage $V_{dc,in}$ on the primary side. R_1, R_2, C_1, C_2, L_1, and L_2 are series equivalent resistance, capacitance and inductance of primary and secondary coils, respectively, and M is the mutual inductance between two coils. Furthermore, the sending and receiving currents and voltages are expressed by I_1, I_2, V_1 and V_2, respectively.

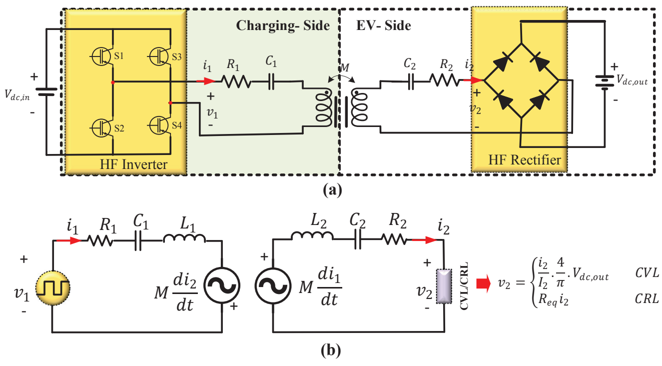

Fig. 1. (a) Schematic diagram of SS-compensated IPT charging system with CVL, (b) Equivalent circuits using fundamental harmonic approximation with CVL or CRL.

According to [3] it can be stated that the nonlinear model of the IPT system is extracted by considering CVL. In this regard, a non-linear state-space model of the SS-compensated

IPT system including the CVL characteristics is represented in dq-axis reference frame as follows:

$$\dot{x} = f(x, u) \tag{1}$$

$$y = g(x, u) \tag{2}$$

where $f(x, u)$ is a nonlinear IPT system with CVL assuming that the state variables in a first harmonic approximation in a synchronously rotating dq reference frame which is given by:

$$\begin{aligned}
\dot{x}_1 &= -\frac{R_1}{L_{\alpha1}} \cdot x_1 + \omega \cdot x_2 - \frac{MR_2}{L_{\alpha1}L_2} \cdot x_3 - \frac{1}{L_{\alpha1}} \cdot x_5 \\
&\quad - \frac{M}{L_{\alpha1}L_2} \cdot x_7 + \frac{1}{L_{\alpha1}} \cdot v_{1,d} - \frac{M}{L_{\alpha1}L_2} \cdot v_{2,d} \\
\dot{x}_2 &= -\omega \cdot x_1 - \frac{R_1}{L_{\alpha1}} \cdot x_2 - \frac{MR_2}{L_{\alpha1}L_2} \cdot x_4 - \frac{1}{L_{\alpha1}} \cdot x_6 \\
&\quad - \frac{M}{L_{\alpha1}L_2} \cdot x_8 + \frac{1}{L_{\alpha1}} \cdot v_{1,q} - \frac{M}{L_{\alpha1}L_2} \cdot v_{2,q} \\
\dot{x}_3 &= -\frac{MR_1}{L_{\alpha2}L_1} \cdot x_1 + \frac{R_2}{L_{\alpha2}} \cdot x_3 + \omega \cdot x_4 + \frac{M}{L_{\alpha2}L_1} \cdot x_5 \\
&\quad + \frac{1}{L_{\alpha2}} \cdot x_7 - \frac{M}{L_{\alpha2}L_1} \cdot v_{1,d} + \frac{1}{L_{\alpha2}} \cdot v_{2,d} \\
\dot{x}_4 &= -\frac{MR_1}{L_{\alpha2}L_1} \cdot x_2 - \omega \cdot x_3 + \frac{R_2}{L_{\alpha2}} \cdot x_4 + \frac{M}{L_{\alpha2}L_1} \cdot x_6 \\
&\quad + \frac{1}{L_{\alpha2}} \cdot x_8 - \frac{M}{L_{\alpha2}L_1} \cdot v_{1,q} + \frac{1}{L_{\alpha2}} \cdot v_{2,q} \\
\dot{x}_5 &= \frac{1}{C_1} \cdot x_1 + \omega \cdot x_6, \qquad \dot{x}_6 = \frac{1}{C_1} \cdot x_2 - \omega \cdot x_5 \\
\dot{x}_7 &= \frac{1}{C_2} \cdot x_3 + \omega \cdot x_8, \qquad \dot{x}_8 = \frac{1}{C_2} \cdot x_4 - \omega \cdot x_7
\end{aligned} \tag{3}$$

where $x = [i_{1,d}\ i_{1,q}\ i_{2,d}\ i_{2,q}\ v_{c1,d}\ v_{c1,q}\ v_{c2,d}\ v_{c2,q}]^T$ is state variable, $u = [v_{1,d} \quad v_{1,q}]^T$ is input control signal which is output voltage of the full-bridge (FB) inverter on the primary-side. By aligning the voltage vector to the d axis, the output voltage of the q channel $v_{1,q} = 0$, therefore; the input control signal is simplified as $u = v_{1,d}$. $L_{\alpha1} = L_1 - M^2/L_2$ and $L_{\alpha2} = L_2 - M^2/L_1$ are leakage inductance, $\omega = 2\pi f_0$ (f_0 is resonance frequency). The resonance frequency of the secondary side can be set to slightly higher than that of the primary side in order to minimize losses in all operations while satisfying the ZVS conditions. Thereby, the compensation capacitances can be designed accordingly as (4), with a detuning factor x_C, which is slightly larger than 1 [15].

$$C_1 = x_c \cdot C_2 \cdot \frac{L_2}{L_1} \tag{4}$$

The receiving side voltage of IPT system which is different for the CRL and CVL models [8], in dq-axis reference frame is defined as dq-components (i.e. $v_{2,d}$ and $v_{2,q}$). Indeed, this represents the nonlinear term of IPT system with CVL which is defined as:

$$v_{2,dq} = \frac{4}{\pi} \cdot V_{dc,out} \cdot \frac{i_{2,dq}}{\sqrt{i_{2,d}^2 + i_{2,q}^2}}$$

Furthermore, in (2) $y = P_{in}$, the input power which is represented as:

$$P_{in} = v_{1,d} \cdot i_{1,d} + v_{1,q} \cdot i_{1,q} \tag{5}$$

III. Proposed Control System Design

In this section, the implementation of the proposed control method based on FCS-MPC on the primary side of the IPT system with CVL is evaluated. In general, MPC utilizes a discrete mathematical model of a system to predict the future behavior of the system over a prediction horizon. By evaluating the predicted values and minimizing a cost function, future optimal actions are determined repetitively. At each sampling time, the optimal value is obtained by solving the optimization problem, and the first element is applied to the system as the output of the MPC controller [12]. Currently, MPC control method has become increasingly popular in the field of power electronics due to the several advantages. First, the availability of accurate mathematical models that can be used to predict the behavior of the variables in electrical and mechanical systems. Second, the computing power of modern microprocessors is able to carry out the intensive computations of MPC quickly and economically [16]. Finally, the use of MPC control strategies with the discrete nature of power converters can be beneficial. This allows for the optimization problem to be solved by assessing the cost function for the possible switching states of the power converters. As this involves evaluating a limited number of control actions, the approach is referred to as FCS-MPC. This technique has been extensively used in power converter applications because of the limited number of switching states that are possible [17].

The IPT charging systems with CVL have very low damping at the resonant frequency, which leads to large, poorly damped, and fluctuation in response to sudden changes. Moreover, there is a large current/power ripple that can be excited by pulse-skipping modulation techniques. This paper presents an FCS-MPC approach for an SS-compensated IPT system, which not only regulates the system power properly but also reduces the current/power fluctuations and ensures ZVS by applying the optimal ON/OFF signal to the switching pattern. Based on the MPC aim, the proposed FCS-MPC approach uses the predictive strategy to predict the power of the IPT charging system, and allows the system to reach the desired values by minimizing the error between the predicted power value and reference power. This leads to the current/power fluctuations being inherently suppressed.

Fig. 2 presents a schematic diagram of the proposed FCS-MPC control method and PI-based PDM applied to the IPT system in order to control the power flow and the switching states. To calculate the MPC control commands, first, the future behaviors of the IPT system (inductor currents, capacitor voltages of both sides and input power in the current sample k) are predicted by utilizing the measured or estimated values. Then using these predicted values, as well as the reference values, the cost function is calculated to achieve the control

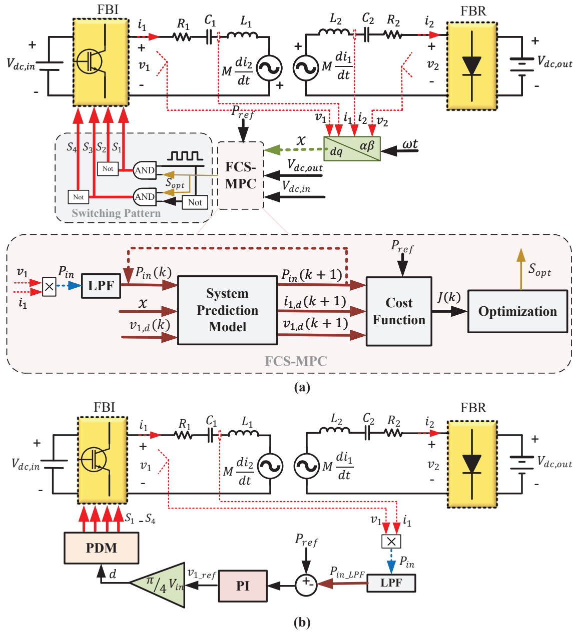

Fig. 2. Distributed implementation of control methods: (a) FCS-MPC , and (b) PI-PDM.

objective (e.g., power regulation). Finally, the optimization problem in the FCS-MPC controller is determined by minimizing the cost function, and the optimal ON/OFF states (S_{opt}) is applied to the switching pattern.

To this end, there are four switching states $S = [0\ 0; 0\ 1; 1\ 0; 1\ 1]^T$ based on switching operation of S_1 and S_3 which effect the operation of the control signal. The output voltage of the FB inverter in time domain (v_1) and d-axis $v_{1,d}$ corresponding to these switching states are given in Table I. It can be seen that the output voltage of the FB inverter has three different pole voltages depending on the switching operation of S_1 and S_3. Apparently, when S_1 and S_3 are closed or opened simultaneously, the resulting voltage v_1 will be zero. On the other hand, when S_2 is closed and S_3 is opened, the inverter produces a positive voltage $v_1 = +V_{dc,in}$. Finally, the inverter generates a negative voltage $v_1 = -V_{dc,in}$, when S_2 is opened and S_3 is closed.

TABLE I
POSSIBLE SWITCHING STATES.

S_1	S_2	S_3	S_4	Operating State	v_1	$v_{1,d}$	State Signal
0	1	0	1	0	0	0	OFF
0	1	1	0	-1	$-V_{dc,in}$	$4/\pi \cdot V_{dc,in}$	ON
1	0	0	1	1	$V_{dc,in}$	$4/\pi \cdot V_{dc,in}$	ON
1	0	1	0	0	0	0	OFF

Briefly, it can be stated that, the proposed control method selects one state from the relevant states according to the current circuit variables to apply as optimal switching operation in the next control cycle. The selected switching state obtains the smaller error between the predicted input power and reference power at each sampling time. The proposed FCS-MPC control process involves the following three steps which is summarized in the flowchart of Fig. 3.

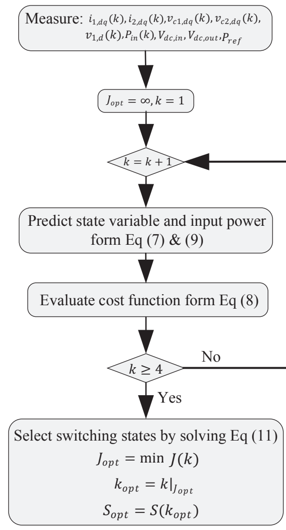

Fig. 3. Flowchart of the proposed FCS-MPC.

A. Step 1. Discrete-Time Model

To calculate the predictive model in MPC strategy, a discrete-time model of the system is required. To this end, the continuous state-space model of Eq. (3) is discretized using the forward Euler approach as (6):

$$\frac{dx(t)}{dt} = \frac{x(k+1) - x(k)}{T_s} \tag{6}$$

where $x(k)$ is defined as the state vectors at k instant, $k = nT_s, n \in Z^+$is sample instant, and $T_s = 6e^{-6}$ is the sampling time of the MPC controller. By considering the switching states operation, the nonlinear state-space representation of the system based on an average discrete-time model can be presented as follows:

$$x(k+1) = x(k) + T_s f(x(k), u(k)) \tag{7}$$

where $u_{(k)} = v_{1,d}(k) = \begin{cases} \frac{4}{\pi} \cdot V_{dc,in} & S_1/S_3 \quad ON \\ 0 & else \quad OFF \end{cases}$

In Table I, it can be seen that of the possible switching states for S1 and S3, [0 0] and [1 1] have the same operation and provide the OFF state, whereas the remaining switching operations provide the ON state. Therefore, to simplify the switching state, in d-axis and reduce the current fluctuation, there are two states (ON or OFF state), in which the ON state is when at least one of the switches S_1 or S_3 is turned on, and the OFF state is when both switches are turned on or off. Another word, the output voltage of the FB inverter on the primary side in d-axis reference frame based on switching functions can be defined as $v_{1,d}(k) = |S_1 - S_3| \cdot \frac{4}{\pi} \cdot V_{dc,in}$.

B. Step 2. Cost Function

The main objective of the proposed control design is to ensure accurate switching states for FB inverter so that the input and output power track the reference value with fast

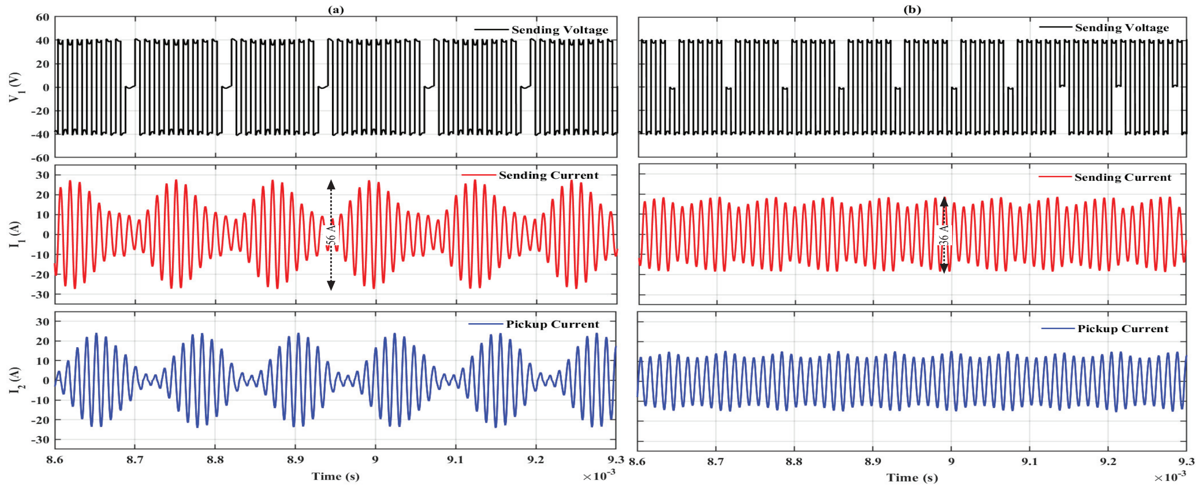

Fig. 4. Steady-state performance of the IPT system with CVL, when $P_{ref} = 0.925P_{nom}$: (a) PI-based PDM, and (b) FCS-MPC.

dynamic response and minimum steady-state error. Indeed, input power regulation accuracy can be ensured by penalized the deviation of the desired power trajectory in the cost function. Thus, the cost function of proposed method can be represented as follows:

$$J(k) = \frac{1}{N_p}\sum_{k=1}^{N_p}\left(P_{in}^p(k+1) - P_{ref}\right)^2 \tag{8}$$

where N_p is the prediction horizon, which here is set to two in order to reduce computational effort and based on operation between S_1 and S_3 at each sampling time. P_{ref} and $P_{in}^p(k+1)$ are input reference and predicted power, respectively.

As shown in Fig. 2(a), an optimal power controller is presented by predicting the input power and current on the primary side as well as adjusting the sending voltage of the proposed IPT system at each sampling time. It can be observed that in (2) by assuming $v_{1,q} = 0$, the input power is equal to $P_{in} = v_{1,d} \cdot i_{1,d}$. Since in this formulation, the voltage is adjusted based on the ON/OFF switching states, the fluctuation of the input power will be very high. To address this challenge the predicted input power is calculated from two terms, which can be represented as follow:

$$P_{in}^p(k+1) = \tau \cdot v_{1,d}^p(k+1) \cdot i_{1,d}^p(k+1) + (1-\tau) \cdot P_{in}(k) \tag{9}$$

where $v_{1,d}^p(k+1)$ is predicted value of $v_{1,d}$ which is controlled based on optimal ON/OFF states at each sampling time and $i_{1,d}^p(k+1)$ is the predicted value of the sending current. τ is the weighting coefficient of LPF which can be given by:

$$\tau = \frac{T_s}{T_f + T_s} \tag{10}$$

where T_s is the sampling time of the proposed FCS-MPC approach and T_f is the time constant of the LPF. In general, to obtain the time constant, the higher the sampling rate, the lower the time constant. In other words, if the sampling frequency is too high or the sampling time is too small, the time constant must be small to keep the narrow band of ripples. In this regard, to reduce the steady-state error and current power ripples, while ensuring a narrow band, which is set to 3% of nominal power in this paper, a weighted filter coefficient has been set to 0.03. It should be noted that the cutoff frequency of the LPF in this case is approximately 6% of the switching frequency.

C. Step 3. Optimization Problem

As shown in (7), the input control signal depends on dc input voltage based on the possible switching states. Therefore, to achieve accurate control of the power transfer, it is necessary to obtain the optimal switching states, which is achieved by solving the cost function. In other words, the output of the proposed FCS-MPC is the result of the local optimization problem, which can be obtained by minimizing the given objective function as:

$$\begin{aligned} &S(k) = \arg\min J(k) \\ \text{subject to}\quad &Eq(7),\ \text{and}\quad v_{1,d} \in \{0, \frac{4}{\pi}V_{dc,in}\} \end{aligned} \tag{11}$$

At each sampling time, the process of minimization involves comparing the cost values of the ON and OFF positions of the switch. The switching state ($S(k_{opt})$) is then determined by the cost function which represents minimum error. If the cost function of the ON position yields the lowest error, the switching state is set to the ON state and vice versa. Therefore, these ON/OFF states are applied as the output of the optimization problem to the switching pattern.

IV. Simulation Results

To confirm the effectiveness of the proposed control method, a simulation study is conducted in Matlab/Simulink. For this purpose, the proposed control method is applied to the FB inverter as the primary-side controller in the IPT system to control switching states according to Fig. 2(a). The parameters of the proposed IPT system are listed in Table II.

The proposed FCS-MPC technique is comparable to pulse-skipping modulation such as PDM, since both use skipping pulses for power regulation. However, the IPT charging system

TABLE II
ELECTRICAL AND CONTROL PARAMETERS.

Electrical Parameters of the IPT system		
Parameter	**Symbol**	**Value**
Nominal Power	P_{nom}	400 w
Nominal Sending Voltage	V_1	52 V
Nominal Receiving Voltage	V_2	50 V
Nominal Coupling Coefficient	k_c	0.2
Resonance Frequency	f_0	85 kHz
Primary/Secondary self inductances	L_1/L_2	31.24/30.85 μH
Primary/Secondary capacitances	C_1/C_2	117.83/113.64 μF
Primary/Secondary Resistances	R_1/R_2	0.0525/0.0555 Ω
Detuning factor	x_c	1.05

with the FCS-MPC method takes advantage of a predictive model to estimate the future behaviour of the system such as the state variables and input power, then by minimizing the difference between the predicted power and power reference at each sampling time, it can avoid significant current/power oscillations, and the optimal control action is obtained as an ON/OFF signal. Moreover, the proposed FCS-MPC does not require any modulator. To validate the effectiveness of the proposed FCS-MPC control method, a comparison of the IPT system performance using PI-PDM and the proposed FCS-MPC methods is presented in the following scenarios. The outcomes of these scenario provide a clearer understanding of the difference between these two strategies.

A. Scenario 1: Steady-State Analysis

In this case study, the steady-state performance of the IPT system with CVL using the PI-PDM based on delta-sigma modulator and the FCS-MPC control strategies are compared. In this regard, to verify the performance of the proposed control method, the comparison has been investigated in the worst density (i.e. $d = 0.9$), meaning that the frequency of the skipping PDM pulses is the same as the natural frequency of the system [8]. In fact, in the PI-based PDM control method, the power transfer can be controlled by adjusting the duty cycle (d), while ensuring the soft switching operation. To modify the worst density, the PI control parameters have been set as $k_p = 0.009$ and $k_I = 100$. Fig. 4(a) and (b) show the performance of sending voltage and current, as well as pickup current of the proposed IPT system using PI-PDM and FCS-MPC methods, respectively. As can be seen the current ripple of the IPT system with the proposed FCS-MPC strategy is smaller than that of the PI-PDM method. It can be verified based on the peak-to-peak sending current, which has been reduced to $36A$ through the proposed control method, compared to $56A$ with PI-PDM.

Furthermore, another important issue to be evaluated in IPT systems is to ensure the soft switching performance in the presence of the proposed control method to reduce the switching losses and increase efficiency. To validate the ZVS performance of the IPT system using the proposed FCS-MPC, the closed-loop system is tested under different power variations as seen in Fig. 5. It can be seen that the IPT system has always obtained ZVS, while ensuring that it can reduce the current/power ripple at any density.

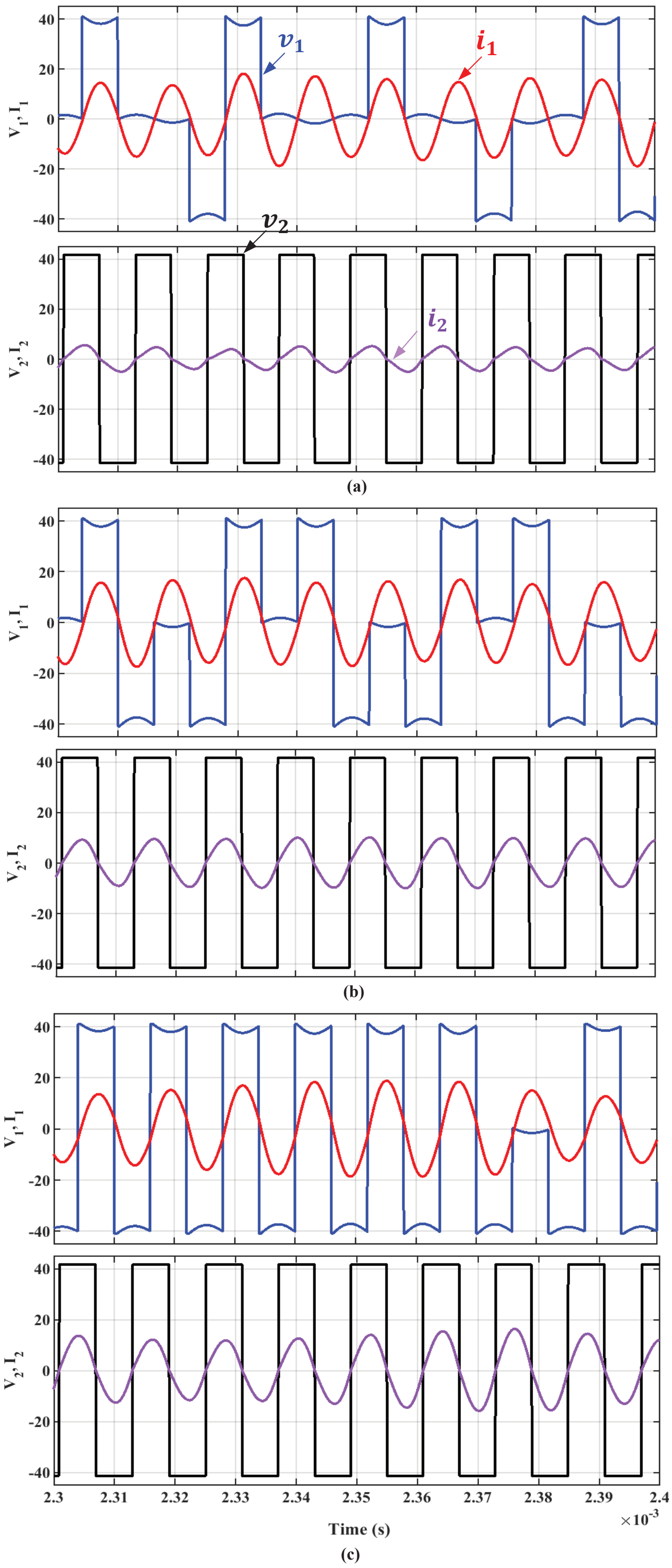

Fig. 5. ZVS operation of the IPT system using the proposed FCS-MPC under different power: (a) $0.35P_{nom}$, (b) $0.65P_{nom}$ and (c) $0.95P_{nom}$

B. Scenario 2: Dynamic Analysis

1) Transient Operation: In this case study, the transient and robustness performance of the proposed FCS-MPC control method is investigated under reference power change in comparison with the PI-PDM. It is assumed that a step change in the reference power is imposed with the initial value of $P_{ref} = 0.86P_{nom}$ to $P_{ref} = 0.65P_{nom}$ at $t = 0.01s$. Fig. 6 shows a comparison of the dynamic response of the input power and output power using the PI-PDM and FCS-MPC control methods. It is observed that the IPT system with PDM has large current/power oscillations and reaches a new steady-state condition more slowly (about $3ms$). Comparing with the PI- based PDM, it can be seen that the IPT system with the proposed FCS-MPC control method under the same conditions can reach the new steady-state very quickly only after $0.1-0.15ms$. Moreover, there are no large current/power oscillations in the dynamic process, as well as the current/power ripples of the system with the proposed control method in steady-state are also slightly smaller than the system with PI- based PDM.

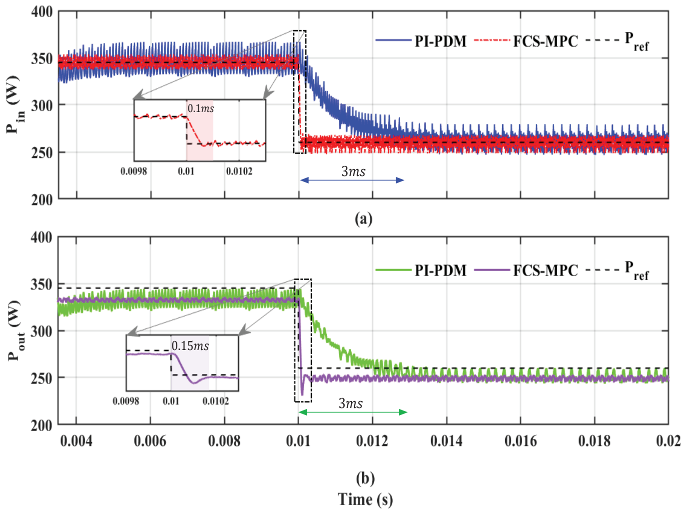

Fig. 6. Performance of the IPT system using the proposed FCS-MPC and PI-based PDM under power change: (a) comparison of input power and (b) comparison of output power.

2) Coupling Change Operation: Fig. 7 shows the dynamic performance of the IPT system when the system coupling changes from $k_c = 0.15$ to $k_c = 0.2$ at $t = 4ms$. It can be seen that the sending current decreases with increasing coupling. Moreover, as seen in Fig. 7(b) and (c), the pickup current and system power remain stable and settle to an acceptable steady-state condition after a short time (about $0.06-0.1ms$), as well as there is no current/power ripple with coupling changes. This also confirms that the IPT system with the proposed FCS-MPC control method has great dynamic performance.

3) Power Change Operation: Furthermore, the dynamic response of the proposed IPT system in the synchronously rotating dq reference frame is evaluated in the presence of the proposed FCS-MPC method. In this regard, the dynamic response of the state variables (i.e x) under step power change is shown in Fig. 8, when the power reference is changed at $t = 4ms$ from $0.77P_{nom}$ to $0.6P_{nom}$. It is observed that the

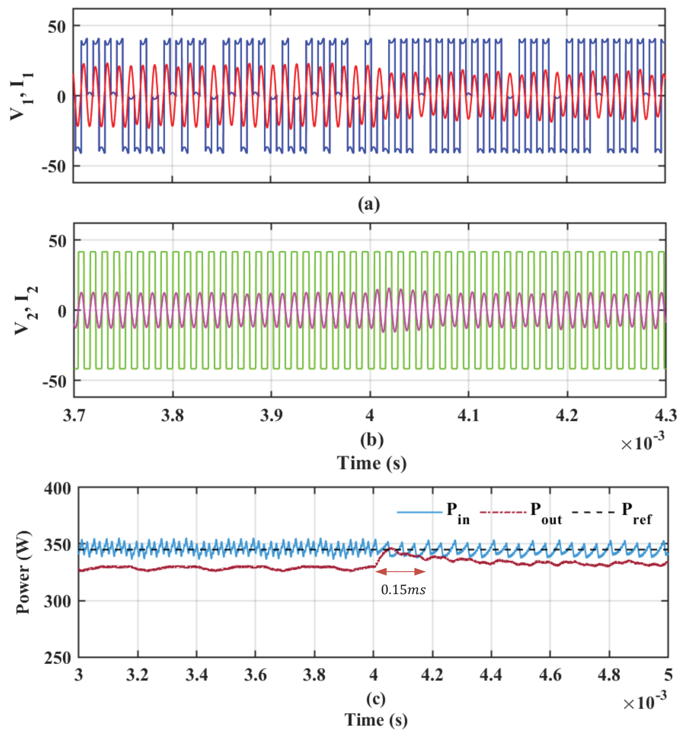

Fig. 7. Performance of the IPT system using the proposed FCS-MPC under coupling change, when $P_{ref} = 0.86P_{nom}$: (a) primary-side voltage and current, (b) secondary-side voltage and current, and (c) input and output power.

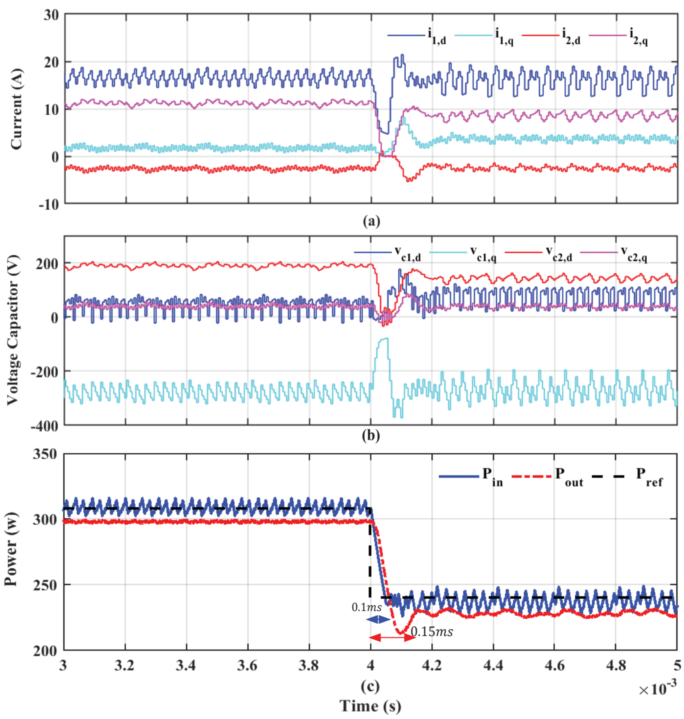

Fig. 8. Performance of the IPT system's state variable in the synchronously rotating dq reference frame using the proposed FCS-MPC under step power change from $0.77P_{nom}$ to $0.6P_{nom}$: (a) primary and secondary inductive currents, (b) primary and secondary capacitor voltages, and (c) input and output power.

state variables, the input power and output power reach the new steady-state very quickly (about $0.1-0.15ms$). Therefore, the results in this scenario verify that the proposed FCS-MPC strategy provides superior performance compared to the PI-PDM method with a fast dynamic response, and small current/power ripple.

V. Conclusion

This paper has presented an FCS-MPC based on a nonlinear model of an IPT system with CVL. The main objective of this paper is the power regulation, while ensuring a fast dynamic response with low current/power ripples and achieving full-range soft switching operation. In the process of the proposed control method, a cost function is considered based on main objective (i.e, the power regulation) and by minimizing the error between the predicted power value and power reference at each sampling time, the current/power oscillations can be suppressed. In addition, the proposed control method by solving the optimization problem can obtain the optimal switching states as ON/OFF signal for the FB inverter on the primary-side. Simulation results confirm the effectiveness of the proposed FCS-MPC in comparison with the PI-based PDM method for an SS-compensated IPT system with CVL.

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