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Achieving Perfect Tracking in Presence of Saturationplant model and Model Uncertainty in Current Regulators for Voltage Source Inverters

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***Abstract*—Model Predictive Control (MPC) has attracted significant attention from the power electronics community. In this body of work it is usual to utilize a linear load model. Also, to deal with computational complexity, it is common to deploy horizon 1 optimization. A major difficulty with these standard approaches is that modelling errors, e.g. those arising from non-linear effects such as magnetic saturation, lead to poor tracking performance. In this paper we show that perfect tracking is achievable provided that MPC controller is augmented by a suitable observer. This claim is supported by extensive simulation results.**

I. Introduction

Model Predictive Control (MPC) has become widely used by the power electronics community. Following its early implementations for current control [2], [9] and torque control [4] of inverters, MPC has been applied to a variety of converter topologies and power electronic devices [8]. Some advantages of the MPC approach include the associated clear and intuitive algorithms, the ability to handle complex and non-linear situations [6], and a distributed harmonic spectrum which can be shaped in different ways [5], [7].

The most common situation in drive control and grid connected applications is to have a pulse-width modulated (PWM) voltage source inverter (VSI) driving a three phase load, for example, an induction motor. Figure 1 illustrates this situation. It also shows how the voltage and current waveforms typically look like in the case of a two level inverter. In the majority of inverter applications the load is inductive and, therefore, a linear RL load model approximation is typically assumed. This can be expressed as:

$$L\frac{di(t)}{dt} + Ri(t) = v(t) \tag{1}$$

We recall that PWM, by using different voltages from a limited set in one modulation cycle, very closely approximates a constant voltage of a desired magnitude. A well-known solution of equation (1) at the end of interval Δ, if a constant voltage $V = V(0)$ is applied for the entire interval Δ, is given by:

$$i(\Delta) = i(0)e^{-\frac{\Delta}{\tau_p}} + \frac{1}{R}V(0)\left(1 - e^{-\frac{\Delta}{\tau_p}}\right) \tag{2}$$

where $i(0)$ and $i(\Delta)$ are current at the beginning and at the end of interval Δ, respectively; and $\tau_p = L/R$ is the plant time constant.

This can be written in the following discrete form

$$\left(1 - az^{-1}\right) i(z) = bz^{-1}v(z) \tag{3}$$

where $a = e^{-\frac{\Delta}{\tau_p}}$; $b = \frac{1}{R}\left(1 - e^{-\frac{\Delta}{\tau_p}}\right)$; $\tau_p = L/R$ and z^{-1} is the standard shift operator.

Unfortunately, non-linear effects are often present in practical power electronics circuits. Such non-linear effects include magnetic saturation and non-linear switching delays. These non-linear effects are hard to model and can significantly impact the performance of standard MPC control laws. In particular, reference tracking of desired quality is not achieved if nonlinearities are encountered.

The objective of this paper is to develop an improved inverter control strategy, in MPC framework, to mitigate the effects of non-linear load model on reference tracking. We argue that the impact of nonlinearities on tracking performance can be treated by augmenting a standard MPC structure by a suitable observer. This leads to a modified MPC structure which includes a nearest neighbour quantizer and a special feedback filter around it. Such a structure, to the best of authors' knowledge, has never been applied in power electronics.

In the traditional vector control of VSI (see, for example, [10]) the effect of the main flux saturation is either neglected (resulting in tracking errors) or an attempt is made to estimate the saturation and incorporate it into the load model.

Our approach is fundamentally different in that it does not require an accurate load model. We show in this paper that even with a very approximate (linear) load model and using an appropriately designed observer, it is possible to completely eliminate the effects of the load model inaccuracies and non-linearities on reference tracking, at given frequencies. We validate the proposed MPC structure by extensive simulations and show that it does indeed give zero tracking errors, at the frequencies of interest, in the presence of nonlinear effects.

The layout of the remainder of the paper is as follows. In section II we develop a polynomial form of a steady state Kalman filter for a linear system with periodic disturbance. Section III will explain how disturbance rejection and perfect tracking can be achieved using horizon 1 MPC. Sections IV shows how the ideas can be applied to a volt-

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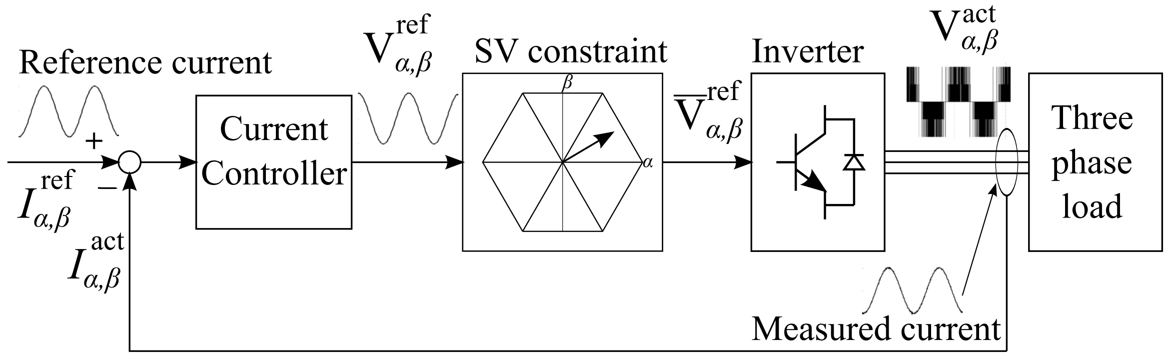

Figure 1: Three phase current controlled VSI

age source inverter. Section V presents results of detailed simulations and their discussion. Section VI summarizes the main findings and conclusions of the paper.

II. Polynomial Form of Linear Steady State Kalman Filter

Consider the following linear system having Gaussian white noise and periodic disturbance:

$$x_{k+1} = A_o x_k + B_o u_k + n_k \tag{4}$$

$$d_{k+1} = A_d d_k + \omega_k \tag{5}$$

$$y'_k = C_o x_k + C_d d_k + \nu_k \tag{6}$$

where x, u and y are state, input and output vectors, respectively; n, ω and ν are Gaussian white noise sequences; d is periodic disturbance; A_o, A_d, C_o, C_d and B_o are matrices of appropriate dimensions.

Given the model (4) to (6), the corresponding steady state Kalman filter takes the form:

$$\hat{x}_{k+1} = A_o \hat{x}_k + B u_k + J_o \left(y_k - C_o x_k - C_d \hat{d}_k \right) \tag{7}$$

$$\hat{d}_{k+1} = A_d \hat{d}_k + J_d \left(y_k - C_o x_k - C_d \hat{d}_k \right) \tag{8}$$

Using the above filter, the system output can be expressed in innovation form as

$$y'_{k+1} = C_o \hat{x}_k + C_d \hat{d}_k + \varepsilon_k \tag{9}$$

where ε_k is a white noise sequence.

It can be easily shown that the innovation model (9) can be written in polynomial form as

$$A(z)D(z)y'_k = B(z)D^{'}(z)u_k + C(z)\varepsilon_k \tag{10}$$

where

$$\frac{B(z)}{A(z)} = C_o \left(zI - A_o\right)^{-1} B_o \tag{11}$$

$$A(z) = \det\left(zI - A_o\right) \tag{12}$$

$$D(z) = \det\left(zI - A_d\right) \tag{13}$$

$$C(z) = \det \begin{bmatrix} zI - A_o & J_o C_d \\ J_d C_o & zI - A_d + J_d C_d \end{bmatrix} \tag{14}$$

As an example, say that the disturbance d_k is a single frequency disturbance at frequency ω_o, then

$$D^{'}(z) = 1 + 2\cos\left(\omega_o \Delta\right) z^{-1} + z^{-2} \tag{15}$$

where Δ is the sampling period.

III. Disturbance rejection using MPC horizon one

The basic principle to be used here is to bring the predicted output to a desired value $\{y_k^*\}$. Note that $\{y_k^*\}$ can be converted to an output disturbance by defining the tracking error as

$$y_k = y'_k - y_k^* \tag{16}$$

Then the model (10) can be expressed in terms of the tracking error as

$$A(z)D(z)y_k = B(z)D(z)u_k + C(z)\varepsilon_k \tag{17}$$

where $D(z)$ now includes the model for both disturbance and reference. Typically in power electronics applications this will be a sinewave of a given frequency. We can always make $a_0 = 1$; $d_0 = 1$ and $c_0 = 1$ by dividing both sides by $a_0 d_0$ and including a constant into ε_k if necessary.

We assume that the plant has p units delay. Thus we can write

$$B(z) = z^{-p} b_0 + z^{-p-1} b_1 + \ldots = z^{-p} B'(z)$$

with $b_0 \neq 0$. Now using the division algorithm of polynomial algebra we can write

$$\frac{C}{AD} = F + \frac{z^{-p} G}{AD} \tag{18}$$

where $F = f_0 + f_1 z^{-1} + \ldots + f_{p-1} z^{1-p}$; $G = g_0 + g_1 z^{-1} + \ldots$; and $f_0 = 1$.

Then

$$FAD = C - z^{-p} G$$

which allows one to transform equation (17) into

$$C\left(y_{k+p} - F\varepsilon_{k+p}\right) = G y_k + B' F D u_k$$

Notice that $F\varepsilon_{k+p}$ denotes future "white noise", which is unpredictable at time k. Hence the p step ahead prediction can be expressed as

$$C \hat{y}_{k+p} = \alpha y_k + \beta u_k \tag{19}$$

where $\alpha(z) = G(z)$ and $\beta(z) = B'(z)F(z)D(z)$.

We divide both sides of (19) by polynomial $C(z)$ and then we factor $\beta(z)/C(z)$ as: