A GENETIC ALGORITHM DESIGN METHOD FOR A CURRENT CONTROLLER EMPLOYING “TWO AHEAD” PREDICTION

Milijana Odavic
University of Nottingham
University park
Nottingham NG7 2RD
UK
elxmo1@nottingham.ac.uk

Pericle Zanchetta
University of Nottingham
University park
Nottingham NG7 2RD
UK
Pericle.Zanchetta@nottingham.ac.uk

Mark Sumner
University of Nottingham
University park
Nottingham NG7 2RD
UK
Mark.Sumner@nottingham.ac.uk

Abstract – This paper introduces an improved predictive control for the control of the current in a three-phase active shunt power filter (ASF). The proposed control system compensates for unavoidable digital controller delays, and provides an accurate prediction of the ASF behaviour in the next sampling period. A “Two Ahead” current reference prediction is proposed which estimates the filter current two samples in advance. The predictor is based on a second order extrapolation method, and can also account for sudden changes in the reference waveform. A Genetic Algorithm is used for parameter optimisation in the reference prediction/extrapolation formula. In addition to good tracking performance, the proposed controller is insensitive to measurement noise. The effectiveness of the proposed method is verified through experiment.

I. INTRODUCTION

The reduction of the higher order harmonics in a power network is receiving more attention due to strict legalisation concerning power quality issues [1] and because of increased communication and protection equipment failures. The conventional approach taken to deal with harmonic currents is to use tuned passive filters. Modern active power filters are superior in filtering performance, but introduce a challenging control task in the field of power electronics research and development.

The control structure of the active shunt filter (ASF) includes three key elements: the dc-link voltage control, the current control system and the method to determine the current references from the sensed currents of the harmonic producing load. The main focus of this paper is the development of an efficient current control design.

In ASF applications, the current reference consists of the harmonic components at frequency much higher than the fundamental. This complicates the design of the current controller. Demands concerning a current controller design are: minimisation of the phase and amplitude errors for a wide range of reference trajectories; fast dynamic response; robustness to model and parameter uncertainties, and noise rejection. The classical approach (the proportional plus integral or PI controller) applied to the ASFs fails to accomplish these requirements, and therefore advanced control structures are starting to replace them.

In the literature different control strategies have been proposed under the title “predictive controller” [2]. The task of the predictive controller is to predict a future behaviour of the plant according to the plant model and the accessible past and present plant inputs and outputs. The prediction techniques published so far can be divided into two main categories according to the processing power required. A less computational intensive approach incorporates prediction methods concerning the next few sampling periods [3]-[10]. Alternatively a prediction up to a specified finite horizon demands intensive processor time, which can be a major restriction in applications with high frequency reference trajectories such as ASFs. Both of these approaches are model based control structures incorporating a linear plant model as a part of the controller structure.

This paper presents a new predictive controller, which incorporates a method for predicting variables two sample periods ahead of their appearance. This allows the predictive control to work in the presence of microprocessor, and actuation based delays. The proposed controller is a model-based controller and therefore the knowledge of system parameters is essential for satisfactory performance. A system stability analysis concerning parameter uncertainties is presented in the paper and confirms system robustness to parameter mismatch.

The proposed controller introduces a minimal phase error by predicting the current reference two sampling instants ahead. This prediction is based on the polynomial extrapolation technique, designed using the Genetic Algorithm (GA) optimisation tool. It also incorporates transient conditions into the proposed prediction algorithm.

A serious disadvantage of a very high bandwidth control loops is a high sensitivity on the noise that can appear on the experimental rig, and therefore a practical implementation can be very difficult in an environment where the noise cannot be sufficiently reduced. A comparison of the noise rejection of the proposed predictive controller with that of a deadbeat controller is also presented in this paper.

II. SYSTEM MODELLING

Predictive control is a model-based control method and therefore an appropriate model of the system is an essential part for the controller design. An ASF connected to a power network is illustrated in Fig.1. Fig. 2 shows an equivalent single-phase simplified ASF model. The equivalent circuit can be described with the following linear first order differential equation:

$$\frac{di}{dt} + \frac{R}{L}i = \frac{e - v}{L},$$

where $e$ is the supply voltage, $i$ the active filter current, $v$ the active filter voltage and $R$, $L$ resistance and inductance of input inductors.
From (1), for the sampling period between the time instants \( k \) and \( k+1 \), assuming that the supply voltage over one sampling period is constant, the discrete linear model of the ASF connected to the network is given by the following equations:

\[
i(k + 1) = i(k) - a + (E(k) - V(k)) \cdot b
\]  

\[a = e^{-\frac{R}{L} T_s} = 1 - \frac{R}{L} T_s
\]  

\[b = \frac{1 - e^{-T_s/\tau}}{R} = \frac{1 - \frac{R}{L}T_s}{R} = \frac{T_s}{L}
\]  

and the coefficients \( a \) and \( b \) are approximated by a Taylor series.

III. CURRENT CONTROLLER

Digital controllers introduce a delay caused by the microprocessor computation time, and an additional delay caused by the sample/hold nature of the actuation signal (i.e. PWM). To simplify the control law, the delay caused by the computation for the ASF voltage reference calculation, is kept constant and equal to one sampling period. The aim of this controller is to calculate the ASF voltage reference for the next sampling period to eliminate current error at the time instant \( k+2 \), as shown in Fig 3, achieving the deadbeat condition \( \Delta i(k + 2) = 0 \).

The discretized ASF model for the sampling period between the time instances \( k+1 \) and \( k+2 \), can be rewritten from (2) in the following form:

\[
i(k + 2) = i(k + 1) \cdot a + (E(k + 1) - V(k + 1)) \cdot b
\]  

By introducing \( i(k+2) = i^*(k + 2) - \Delta i(k + 2) \):

\[V(k + 1) = E(k + 1) - \frac{1}{b} [i^*(k + 2) - i(k + 1) \cdot a]
\]

It should be noted that at the instant of measurement \( k \), current and voltage values for the next sampling instant are not available and need to be predicted or estimated based on the system model. The value of the current at the instant \( k+1 \) can be estimated from the linearized system model (2).

For the following analysis the current reference prediction error is going to be neglected, i.e. it is assumed that the reference value at the instant \( k+2 \) is available. The influence of the input resistance is also neglected. The supply voltage is approximated by \( E(k+1)=E(k) \). Taking into account a difference between the modelled input inductance \( L_m \) and its real value \( L \), the closed current loop can be defined with the following three discrete equations in the \( z \) domain. The first two equations describe the behaviour of the current controller while the third one presents the real plant.

\[
V(k) \cdot z = E(k) - \frac{L_m}{T_s} \left[ i^*(k) \cdot z^2 - i(k + 1) \right]
\]

\[
i(k+1) = E(k) \cdot \frac{T_s}{L_m} - V(k) \cdot L_m + i(k)
\]

\[
V(k) \cdot z^{-1} = E(k) \cdot z^{-1} - \frac{L}{T_s} \left[ i(k) - i(k) \cdot z^{-1} \right]
\]

From (7), (8) and (9) considering the supply voltage as a disturbance, the closed loop transfer function of the current loop is:

\[
\frac{i(k) \cdot z^2}{i^*(k)} = \frac{L_m / L \cdot z^2}{(z^2 + L_m / L - 1)}
\]

From (10), ignoring the parameters mismatch (i.e \( L_m/L=1 \)), a deadbeat behaviour of the control loop is seen with both poles at the origin. Ideally, the transient is achieved without any overshoot and within one sampling period, i.e. with no phase delay.
The main disadvantage of the deadbeat control is its high sensitivity to the high frequency noise that can appear under real conditions. To achieve better noise rejection, the closed loop poles need to be moved from the origin.

In order to move the current loop closed loop poles from the origin, a new prediction method for the ASF current at the instant $k+1$ is here proposed:

$$i_{pred}(k+1) = i(k) + (i'(k+1) - i'(k)) - error$$

$$error = i_{pred}(k+1) - i'(k+1)$$

$$i_{pred}(k+1) = i'(k+1) - 0.5 \cdot i'(k) + 0.5 \cdot i(k)$$

This prediction method uses only the current reference values at the instant $k$ and $k+1$, excluding the supply voltage and the ASF voltage values. In comparison, the estimation method (8) uses a value of the ASF voltage at the instant of measurement and the ASF voltage values. In comparison, the estimation method (8) uses a value of the ASF voltage at the instant of measurement and the ASF voltage values.

From (7), (11) and (9) the transfer function of the closed loop current control is:

$$i(k) - L_m/L \cdot z^2 - L_m/L \cdot z + L_m/L \cdot 0.5$$

$$\frac{z^2 - z + 0.5 \cdot L_m/L}{z^2 - z + 0.5 \cdot L_m/L}$$

From (12) and from the root locus plot for the current controller as shown in Fig 4a, it can be seen that the closed loop poles are moved from the origin. The noise rejection is analysed in the section VII. The stability of the closed loop current control with respect to parameter mismatch is ensured until the input inductance is overestimated by 100%, as shown in Fig 4b, confirming a good robustness to the parameter inaccuracy. Underestimation of the input inductance is not critical.

IV. REFERENCE CURRENT PREDICTION USING GA OPTIMISATION

The ASF current reference calculation is not the subject of this work. After the ASF current reference has been determined, a two-ahead prediction of the current reference needs to be applied to compensate for the phase error of the proposed current control method.

Generally active filter current references consist of the reactive and higher harmonic components with step changes in amplitude and frequency caused by random load connection or disconnection. This makes the prediction of the current reference of the ASF difficult. To overcome this problem, a polynomial extrapolation formula is proposed. Generally, extrapolation techniques use values from a few previous sampling instants, according to the order of the extrapolation polynomial used, to approximate a value in one or more sampling instants ahead. Extrapolation techniques use past data to predict future behaviour of the signal, and when a step change occurs a big error is introduced to the prediction for the next few sampling instants. To decrease the prediction error at the instant of a sudden change in the current reference a modified method of reference prediction is proposed. When a reference change is detected, the prediction over the next few sampling instants is frozen; the reference current value at the instant of measurement can be used instead. For the proposed predictive current control method in particular, a two sampling instants ahead prediction of the reference current is required and a second order extrapolation polynomial is used. This the prediction at the instant of change in the reference for the next two sampling instants must be frozen. In the case in which a third order polynomial is used instead, this delay is prolonged to three sampling instants. The proposed modified prediction method includes a pre-adjustment required for the reference change detection. This detection includes a prediction error evaluation; when the error is greater than some predetermined tolerance value for the nominal reference current it means that a reference change has occurred and instead of the predicted values the actual reference current values at the instant of measurement $k$ are used for the next few samples.

As mentioned earlier a second-order two-step-ahead extrapolation technique is applied resulting in the following prediction form:

$$i^*_p(k+2) = a_h \cdot i^*(k) + b_h \cdot i^*(k-1) + c_h \cdot i^*(k-2)$$

where $a_h$, $b_h$ and $c_h$ are the polynomial coefficients to be
determined. Eq 13 is applied to each particular harmonic \( h \) present in the active filter reference current and previously identified with the chosen reference calculation algorithm. There are therefore 12 coefficients to be calculated if a compensation of 5\(^{\text{th}}\), 7\(^{\text{th}}\), 11\(^{\text{th}}\) and 13\(^{\text{th}}\) harmonics is sought. Those parameters have been selected using a Genetic Algorithm (GA) optimisation routine based on the minimization of the following fitness function:

\[
\text{error} = \int \left| i_p^*(k) - i_r^*(k) \right|^2 \]

(14)

where \( i_p^*(k) \) denotes a predicted current reference at the instant \( k \). The GA search results are shown in Table I.

Equation (13) can be applied for a specific harmonic prediction. Therefore after the current reference is calculated the particular harmonics need to be extracted. The reference prediction error, using the sampling frequency of 5 kHz and second order extrapolation type prediction (13), is negligible (a factor of \( \approx 10^{-4} \)).

<table>
<thead>
<tr>
<th>( h )</th>
<th>5(^{\text{th}})</th>
<th>7(^{\text{th}})</th>
<th>11(^{\text{th}})</th>
<th>13(^{\text{th}})</th>
</tr>
</thead>
<tbody>
<tr>
<td>( a_h )</td>
<td>5.3781</td>
<td>-0.2438</td>
<td>11.4677</td>
<td>1.3648</td>
</tr>
<tr>
<td>( b_h )</td>
<td>-7.1520</td>
<td>2.7483</td>
<td>18.2495</td>
<td>1.6967</td>
</tr>
<tr>
<td>( c_h )</td>
<td>2.7600</td>
<td>-2.5187</td>
<td>12.8425</td>
<td>2.2393</td>
</tr>
</tbody>
</table>

V. SUPPLY VOLTAGE PREDICTION

It should be noted that the supply voltage commonly includes low order harmonics. To include the more realistic condition of a distorted supply voltage, a prediction of an arbitrary voltage waveform should be considered. The prediction of the supply voltage at the time instant \( k+1 \), using available voltage values at the instant of measurement and at the previous sampling instant, is made using a linear-type prediction:

\[
E(k + 1) = 2 \cdot E(k) - E(k - 1)
\]

(15)

To take into account changes in the supply voltage during one sampling period, an integral of the supply voltage between the time instants \( k+1 \) and \( k+2 \), in the solution of system differential equation (2), is modelled by the average value of the voltage assuming it changes linearly. The supply voltage at the time instant \( k+2 \) can be derived directly from (15):

\[
E(k + 2) = 3 \cdot E(k) - 2E(k - 1)
\]

(16)

VI. OVERALL CONTROL STRUCTURE

The system control structure is cascaded, with the current control as inner and the voltage control as outer control loop. The time constant of the voltage control loop is at least ten times higher than the current control loop so the design of these two loops can be independent. The output of the voltage PI controller presents the active component of the active power filter current to cover losses in the switching devices and parasitic resistance in the circuit. The active component needs to be added to the harmonic reference. Traditional three phase system controllers are usually derived in the equivalent two axes reference frame to reduce the computation effort. Restricting the analysis only to balanced systems, the control signals of the proposed predictive current controller can be directly calculated for two phases while the measurement of the third phase current and of the third supply voltage is redundant. The proposed ASF current controller is derived in the fixed a-b-c reference frame to maintain flexibility for future work on an unbalanced supply; to add an active current component reference to the harmonic current components reference, the supply voltage angle needs to be calculated as shown in Fig 5.

VII. NOISE REJECTION

A high bandwidth control loop is very sensitive to measurement noise that can appear in the practical implementation. To cope with this problem, the proposed controller moves the closed loop poles from the origin by introducing the linear type prediction (11) of the current at the time instant \( k+1 \) rather than estimating the one using the plant model and sensed values (8). This approach can also find its application in drives; not only for the noise rejection but also to achieve better current tracking in the presence of emf voltage estimation errors.

A sensitivity of the proposed predictive current controller to measurement noise is compared with the deadbeat controller. Simulation results, Fig.6 - Fig.10, show the control signals as the input to the PWM unit in the presence of the ASF current and supply voltage measurement noise. The current reference consists of the fundamental and 5\(^{\text{th}}\) harmonic components. The noise penetrating the control and PWM references is clearly lower for the predictive approach described in this paper.
Fig. 8 Predictive controller output signals in the presence of 3% voltage and 5% current measurement noise (d_c-deadbeat controller, p_c-predictive controller)

Fig. 9 Deadbeat controller output signals in the presence of 8% voltage and 15% current measurement noise

Fig. 10 Predictive controller output signals in the presence of 8% voltage and 15% current measurement noise

The experimental results of Fig. 11, confirm the high noise rejection of the proposed predictive controller.

Fig. 11 Experimental results of the three-phase output control signals of the predictive current controller

Fig. 12 Experimental rig of the fully digital ASF

VIII. EXPERIMENTAL RESULTS

For the purpose of experimental validation of the proposed predictive current control algorithm, a 20kVA experimental rig has been constructed (Fig. 12). The developed active shunt filter consists of a three-phase voltage source converter connected to the power supply via three inductors (L=3.75 mH, R=0.3Ω) with a capacitor (C=1000 µF) on the dc side. The control hardware includes a floating point PowerPC 603e microprocessor for the control algorithm processing which coordinates an operation of a slave fixed point DSP TMS320F240 with a three phase PWM unit for PWM signals generation and the operation of an A/D unit for the data sampling. The A/D unit includes 4 independent fast 12-bit A/D converters for the supply voltage and current sampling.
and one of four 16-bit multiplexed channels for the dc link voltage sampling. Using the fast A/D unit for the measured plant signals sampling, an additional delay concerning A/D conversion is avoided and only a single delay of one sampling time period for the reference voltage generation needs to be accounted for. The data sampling and control algorithm processing are synchronised with the PWM unit by the PWM interrupt signal that is sent from the slave DSP to the master at the centre of each PWM period. It must be assured that the execution time of the control algorithm processing does not exceed one sampling period. A sinusoidal PWM modulation with fixed switching frequency of $f_s=5$kHz was used. The sampling frequency was chosen to be the same as the switching frequency.

The experimental results of Fig. 13, show a comparison between the arbitrary current reference including fundamental, 5$^{th}$ and 7$^{th}$ harmonics, and the real current generated from the ASF. The results confirmed very good current reference tracking performance of the proposed predictive controller. The influence of the power supply voltage distortion is not included in the analysis; a programmable ac power supply was in fact used ensuring a purely sinusoidal three-phase supply voltage.

![a) Experimental results for the fundamental and 5$^{th}$ harmonic components](image1)

![a) Experimental results for the fundamental and 7$^{th}$ harmonic components](image2)

Fig 13 Comparison of the current reference and the real current (one phase)

IX. CONCLUSION

An improved predictive current control, specifically designed for ASF applications, has been proposed in this paper. Prediction parameters are optimised using a Genetic Algorithm. The overall performance of the controller has been tested both at fundamental frequency and in the case of harmonic current references, particularly the 5$^{th}$ and the 7$^{th}$ harmonics. The produced experimental results show excellent tracking performance in steady state and dynamic conditions.

The sensitivity on the noise is significantly reduced by moving the closed loop poles from the origin keeping still satisfactory minimization of the current tracking error that presents an essential issue in the current controller design for ASF

The performance of the proposed controller is based on the system model accuracy. A good system stability concerning model mismatches has also been verified through analytical analysis.

Experimental and theoretical results show that the analysed and designed controller is suitable for the precise compensation of the unwanted higher harmonics and reactive component of the load current.

X. REFERENCES


