A "TWO AHEAD" PREDICTIVE CONTROLLER FOR ACTIVE SHUNT POWER FILTERs

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***Abstract* – In this paper an improved predictive control is proposed for the control of the current in a three-phase voltage source converter. The application for this work is an active shunt power filter (ASF). The proposed control system takes into account unavoidable digital controller delays, which make it difficult to predict the ASF behaviour in the next sampling period. A "Two Ahead" current reference prediction is proposed which predicts the filter current two samples in advance. The predictor is based on a polynomial-type extrapolation method, and can also account for step changes in the reference waveform. In addition to good tracking performance, the controller has good stability with respect to the system parameters and model mismatch, and good resistance to the measurement noise. The design and implementation of the digital controller are developed in detail in this paper and the effectiveness of the proposed method is verified through experiment.**

I. INTRODUCTION

In recent years the power quality of the supply network has decreased because of the intensive use of the power converters, which do not draw a purely sinusoidal current because of their non-linear switching action. Non-sinusoidal currents can cause the supply voltage distortion introducing a number of problems in the network. Prompted from one side by strict legalisation concerning power quality issues and from the other side by greater power losses and increased communication and protection equipment failures, the solution to harmonic pollution problem is becoming more and more necessary. The conventional approach dealing with the 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 active shunt power filter control structure 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.

The classical control approach using PI controllers eliminates steady state error if reference signals are dc quantities. Therefore multiple reference frames rotating at the harmonic frequencies to be compensated need to be implemented if the application of a traditional proportional-integral control is preferred. Unfortunately the interactions among these different frames and the presence of suitable band-pass filtering stages make the design and tuning of these loops a quite complicated task.

A single dq frame controller, synchronous with the 50Hz supply voltage, is commonly found on commercial products, as it provides a good compromise between implementation simplicity, limited commissioning and control robustness and stability. The PI controller characteristic of tracking dc quantities can be used in this case also for sinusoidal current control. But in the case of active power filtering systems, reference currents consist of higher harmonics, whose d-q components in this frame are pulsating quantities at high frequency and steady state error cannot be eliminated.

Predictive current control [1]-[4] is a linear control technique suitable for ASF applications offering the advantage of precise current tracking over a wide frequency range. A predictive current controller is a model-based controller; therefore the knowledge of system parameters is essential for satisfactory performance. In particular stability problems occur in ac drive applications [5] when the back emf needs to be estimated. This drawback doesn't apply for active power filtering systems because the line voltage is available and can be measured.

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 digital control, and actuation based delays. The method is presented for active filter control, and is shown to significantly improve the control of high frequency harmonic currents.

II. MATHEMATICAL MODEL

The circuit topology of the pulse width modulation voltage-source converter (PWM-VSC) used in the application as active power filtering systems is presented in Fig. 1. The converter consists of six fully controllable switches (Sa, Sb, Sc) and it is connected to the grid via input inductors. The connection of an active power shunt filter ASF to the network is presented in Fig. 2. The output power stage of the ASF (single phase equivalent) can be described with the following linear first order differential equation:

$$\frac{di}{dt} + \frac{R}{L} \cdot i = \frac{e - v}{L}, \qquad (1)$$

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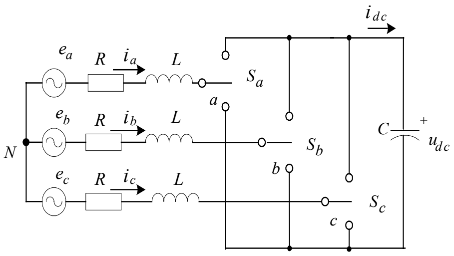


Fig.1 Main topology of the analyzed system

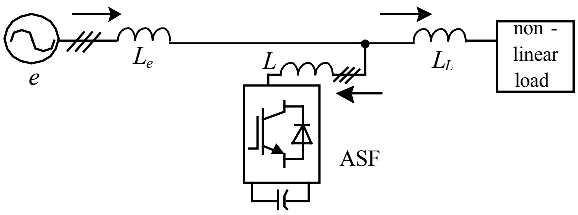


Fig. 2 Active shunt power filter ASF connected to the network

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. The solution of the system differential equation (1) is:

$$i(t) = i(t_0) \cdot e^{-(t-t_0)/\tau} + \int_{t_{in}}^{t_{end}} \frac{e-v}{L} \cdot e^{-(t-t')/\tau} \cdot dt' \qquad (2)$$

From (2), 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 model of the voltage source converter connected to the network is given by the following equations:

$$i(k+1) = i(k) \cdot a + (E(k) - V(k)) \cdot b \qquad (3)$$

$$a = e^{\frac{-R}{L} \cdot T_s} \approx 1 - \frac{R}{L} \cdot T_s \qquad (4)$$

$$b = \frac{1 - e^{-T_s/\tau}}{R} \approx \frac{1 - 1 + \frac{R}{L} \cdot T_s}{R} = \frac{T_s}{L} \qquad (5)$$

where the coefficients a and b are approximated by Taylor series, the active filter and supply voltage are assumed to be constant and equal to $v(k)=V(k)$ and $e(k)=E(k)$ respectively during one sampling time T_s. The time constant of the output stage of the ASF is denoted with $\tau = L/R$. The ASF current at time instants k and k+1 is denoted by $i(k)$ and $i(k+1)$ respectively.

III. CONTROL PROBLEM OVERVIEW

Ideally the active shunt filter at the point of common connection (Fig. 2) injects harmonic and reactive current components demanded by the load so that the power supply needs to ensure only active power to the load. To meet these requirements a good reference tracking of the active filter current controller is needed. Therefore an ASF current controller needs to be designed with a high bandwidth that introduces a minimal time delay. A time delay introduced by ASF current controller becomes a more serious phase delay the frequency of the harmonic increases. Depending on the introduced current controller delay, an ASF can cause harmonic reinforcement at some frequencies rather than elimination.

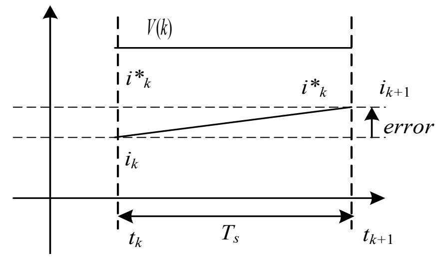


Fig. 3 Basic deadbeat control principal

Digital controllers can be formulated to produce the desired closed loop response. Among them a dead-beat controller aims for the best response possible to a reference change. This means that after a reference change and after a time period equal to the system time delay due to the sampling, the output should reach a reference value not introducing any additional delay or overshoot into the output response. At each sampling instant the controlled plant variables are sampled while at the same time a control signal is sent to the plant. Therefore it is only at the next sampling time that the consequences of the applied control action can be noted.

At first, a deadbeat controller for the ASF current controller is considered under the assumptions which imply that the current references and the supply voltages are constant during one switching period and the computational time required for the controller output signal calculation is neglected, Fig. 3. The switching frequency is also assumed to be constant.

At the instant of measurement k, the current error to be compensated over the current sampling period is $\Delta i(k) = i^*(k) - i(k)$. Assuming ideal conditions, the reference ASF average voltage to compensate for this current error can be formulated. The current error detected in the sampling instant k can't take effect on the controller output signal before the next sampling instant occurs; the time delay of the system is in fact at least one sampling period. The current value at the sampling instant k+1 can be expressed using the desired current value at this instant and the error between these two values: $i(k+1) = i^*(k) - \Delta i(k+1)$ to be included in the discretized ASF model (3). The reference ASF average voltage to eliminate the current error at the instant k+1, can then be given from (3) as:

$$V(k) = E(k) - \frac{1}{b}\left[i^*(k) - i(k) \cdot a\right] \qquad (6)$$

A more realistic situation concerning the ASF current control has the current reference changing over the sampling period. A value of the current reference at the instant k+1 is not known at the instant of measurement k and a prediction needs to be applied. Including the current reference change over the sampling period, the deadbeat control low (6) is now:

$$V(k) = E(k) - \frac{1}{b}\left[i^*(k+1) - i(k)\cdot a\right] \tag{7}$$

Also a computational time delay required for the control signal calculation must be taken into account in the ASF current controller design and will be taken into account in the design procedure for the ASF current controller in the following sections. The poor steady state and dynamic fundamental harmonic current response of the deadbeat controller (7), considering the computational delay of a single sampling period and not taking into account a current reference prediction error, is reported in [1].

IV. 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 then 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 reference to cover losses in the switching devices and parasitic resistance in the circuit. The active reference 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 4.

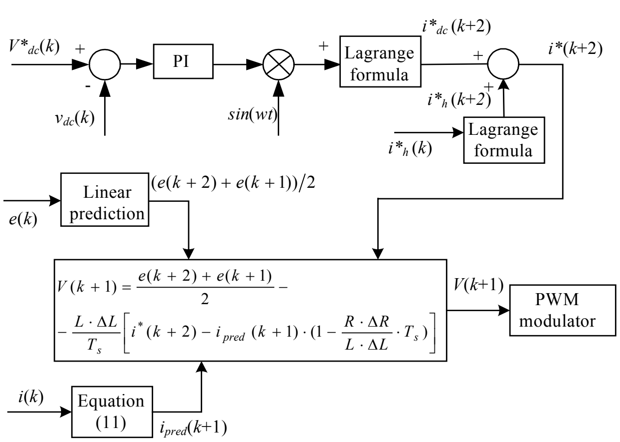


Fig. 4 The proposed control strategy (single phase representation)

V. PREDICTIVE CURRENT CONTROLLER

A predictive current controller to deal with an unavoidable system delays here is proposed. To simplify the control law, the delay caused by the computational time required for the ASF voltage reference calculation, is kept constant and equal to one sampling period. The aim of this controller is to predict the ASF voltage reference for the next sampling period to eliminate the current error at the instant k+2.

The discretized ASF model for the sampling period between the time instances k+1 and k+2, can be rewritten from (3) in the following form:

$$i(k+2) = i(k+1)\cdot a + (E(k+1) - V(k+1))\cdot b \tag{8}$$

The control law of (8) can now be modified in the following form introducing $i(k+2) = i^*(k+2) - \Delta i(k+2)$:

$$V(k+1) = E(k+1) - \frac{1}{b}\left[i^*(k+2) - i(k+1)\cdot a\right] \tag{9}$$

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.

A. Reference current prediction

Active filter current reference calculation is not the subject of this paper. After the ASF current reference has been determined, a two-ahead prediction of the current reference needs to be applied to assure a minimal phase error of the proposed current control method.

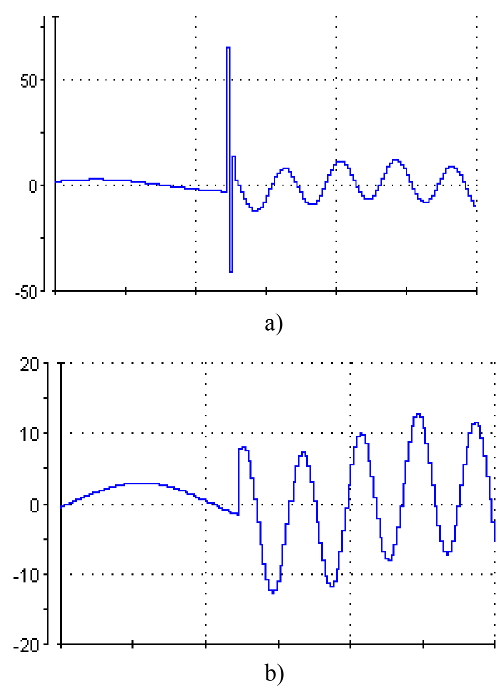


Fig. 5 Current reference prediction implemented in DSP (x axis – time in seconds (Δt=0.005s), y axis - current in amperes)
a) Lagrange formula used
b) modified reference prediction used

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 making the prediction of the current reference of the active shunt filter difficult. To overcome this problem, the Lagrange extrapolation formula is used [1]. 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 as shown in Fig. 5. To decrease the prediction error at the instant of a step change in the current reference a modified method of the reference prediction is proposed. When a reference change is detected, the prediction over the next few sampling instants is frozen; the current reference 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 third order Lagrange extrapolation polynomial is used. Therefore freezing the prediction at the instant of change in the reference for the next three sampling instants is required. In the case in which the fourth order polynomial is used instead, this delay is prolonged to four 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 then some predetermined tolerance value for the nominal reference current it means that a reference change has occurred and instead of the predicted values (10), the actual reference current values at the instant of measurement k are used for the few next samples.

As mentioned earlier the third-order two-step-ahead Lagrange extrapolation of current reference is applied resulting in the following prediction:

$$i^*(k+2) = 10\,i^*(k) - 20\,i^*(k-1) + 15\,i^*(k-2) - 4\,i^*(k-3) \tag{10}$$

B. ASF current prediction

At the instant of measurement, the value of the ASF current at the sampling instant k+1 can be estimated directly from the discretized model (3). However this estimation method uses a value of the ASF voltage at the instant k which can differ from the calculated reference value (e.g. dead time effect). Additional sensors for the ASF voltage could be considered, especially if the supply voltage estimation is included is algorithm [6].

Here is proposed a new non-model based prediction method, which includes only the predicted current reference value for the following sampling instant and the values of actual current and current reference at the instant of measurement (11). A desired value of the current at the instant k+1 is $i^*(k+1)$. The proposed prediction adds a predicted value of the current reference change over the current sampling period (between the sampling instants k and k+1) to the actual current measured at the instant k. This is basically a linear-type prediction of the ASF current. An error between the predicted and the desired current at the instant k+1 is included into the prediction method. An accuracy of the proposed ASF current prediction mainly depends on the accuracy of the reference prediction, Eq. 10.

$$i_{pred}(k+1) = i(k) + (i^*(k+1) - i^*(k)) + \Delta i(k+1) \tag{11}$$

$$\Delta i(k+1) = i^*(k+1) - i_{pred}(k+1) \tag{12}$$

From (11) and (12) the current prediction follows:

$$i_{pred}(k+1) = i^*(k+1) - 0.5 \cdot i^*(k) + 0.5 \cdot i(k) \tag{13}$$

This prediction method excludes the supply voltage and the ASF voltage values to make this approach less sensitive to the measurement noise.

C. Supply voltage prediction

It should be noted that the supply voltage commonly includes low order harmonics. To include the more realistic conditions with the distorted supply voltage, a prediction of an arbitrary voltage waveform is 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:

$$E(k+1) = 2 \cdot E(k) - E(k-1) \tag{14}$$

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 (14):

$$E(k+2) = 3 \cdot E(k) - 2E(k-1) \tag{15}$$

VI. CURRENT LOOP ANALYSIS

The block diagram of the proposed control strategy is presented in Fig. 4. The performance of the analysed predictive current controller, which is a model-based controller, depends on the accuracy of the model parameters. While all the system parameters can be measured in a pre-commissioning stage and their drift from the rated values is negligible, the supply impedance is unknown and can vary unpredictably causing control problems. Therefore the system stability according to inaccuracy of the supply impedance is examined here.

The following discrete equation describes the presented predictive current controller:

$$V(k+1)=\frac{E(k+2)+E(k+1)}{2}-$$
$$-\frac{L\cdot\Delta L}{T_s}\left[i^*(k+2)-i_{pred}(k+1)\cdot(1-\frac{R\cdot\Delta R}{L\cdot\Delta L}\cdot T_s)\right] \quad (16)$$

where the modelled input impedance is denoted by $R\Delta R$ and $L\Delta L$, where R and L present the actual values of the input impedance and ΔR and ΔL represent the inaccuracy of the input resistance and the input inductance respectively. It is to be noted that R and L represent respectively the sum of supply resistance and inductance and ASF output resistance and inductance. The prediction of the supply voltage at the time instants k+1 and k+2 is made using the linear-type prediction (14) and (15) while for the prediction of the reference current the Lagrange interpolation formula (10) is used.

The plant model is given by the following discrete equation:

$$V(k-1)=E(k-1)-\frac{L}{T_s}\left[i(k)-i(k-1)\cdot(1-\frac{R}{L}\cdot T_s)\right] \quad (17)$$

Transforming the equations (10), (13), (14), (15), (16) and (17) to the z-domain the closed loop transfer functions can be obtained. The closed-loop characteristic equation is given by:

$$z^4\cdot L+z^3\cdot(-L+R\cdot T_s)+z^2\cdot(L\cdot\Delta L\cdot 0.5-0.5\cdot R\cdot\Delta R\cdot T_s)=0 \quad (18)$$

Neglecting the two-ahead reference prediction assuming that the reference at the instant k+2 is available, the simplified transfer function of the closed current control loop is:

$$\frac{i(z)\cdot}{i^*(z)}=\frac{\Delta L\cdot z^2-\Delta L\cdot z+\Delta L\cdot 0.5}{z^2-z+0.5\cdot\Delta L} \quad (19)$$

So from (19) it can be observed that this control approach, when the plant model used for the predictive controller derivation is accurate (ΔL=1), eliminates the steady sate error and the transient inside one sampling period without oscillations.

To obtain the root locus of the system as the parameters ΔL and ΔR vary, the characteristic equation is then rewritten in the desired form:

$$1+k\cdot G_o(z)=0 \quad (20)$$

where $G_o(z)$ denotes the open loop transfer function and k is the gain of the system. Equation (18) can be re-written to get the root locus as the parameter ΔL varies according to (20):

$$1+\Delta L\frac{L\cdot 0.5\cdot z^2}{z^4\cdot L+z^3\cdot(-L+R\cdot T_s)-z^2\cdot 0.5\cdot R\cdot\Delta R\cdot T_s}=0 \quad (21)$$

A similar procedure can be performed to get the root locus as the parameter ΔR varies. Fig. 6 presents the root locus of the system varying the ΔL parameter (21). The system is stabile, as it is well known, if all roots are inside the unit circle. As can be seen from Fig. 6 and as would be expected, an overestimation of the input inductance is more critical situation. An overestimation of the input inductance by the factor 2 of the real value still produces stable poles. This stability analysis therefore shows the control has good robustness to parameter inaccuracy. The accurate knowledge

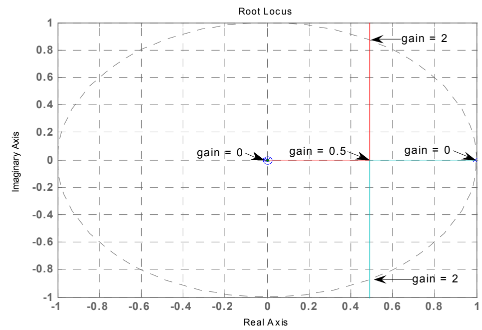


Fig. 6 Root locus of the system as the parameter ΔL (gain) varies

of the input resistance is less critical for the system stability so it is neglected in this analysis.

VII. EXPERIMENTAL RESULTS

For the purpose of experimental validation of the proposed predictive current control algorithm, a 20kVA experimental rig has been constructed (Fig. 7). 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. 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 in 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.

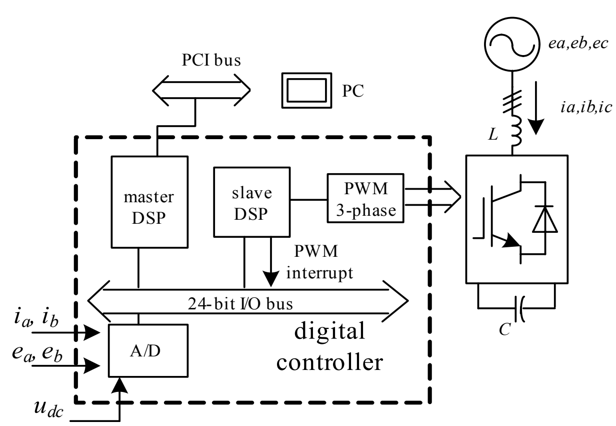
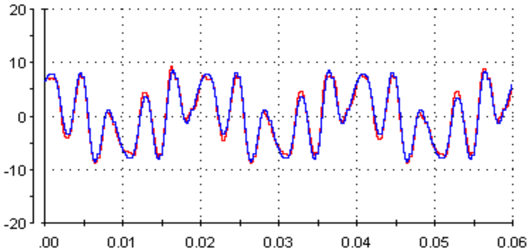
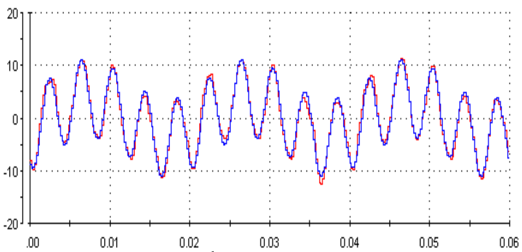


Fig. 7 Experimental rig of the fully digital ASF

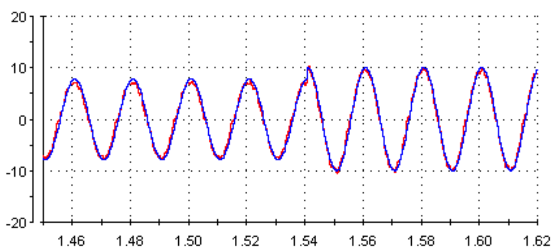


a) fundamental, 5th and 7th harmonic components

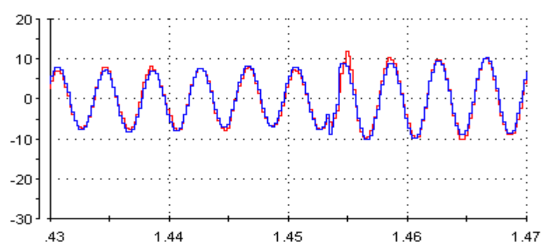


b) fundamental and 5th harmonic components

Fig. 8 Comparison of the current reference and real current for one phase (x axis – time in seconds, y axis - current in amperes)



a) fundamental harmonic(step change in amplitude from 8A to 10A)



b) 5th harmonic (step change in amplitude from 8A to 10A)

Fig. 9 Comparison of the current reference and real current for one phase (x axis – time in seconds, y axis - current in ampere)

A sinusoidal PWM modulation with fixed switching frequency of f_s=5kHz was used. The sampling frequency was chosen to be the same as the switching frequency.

Experimental results of the proposed ASF predictive control method for steady state and dynamic conditions are shown. Fig. 8 and Fig. 9 show a comparison between the arbitrary current reference including fundamental, 5th and 7th harmonics, and the real current generated from the ASF, given in discrete domain. The figures show good tracking performance without an overshoot during the transient response. 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.

VIII. CONCLUSION

An improved predictive current control, specifically designed for ASF applications, has been proposed in this paper. Its performance has been tested both at fundamental frequency and in the case of harmonics current reference, particularly the 5th and the 7th harmonics. The produced experimental results show excellent tracking performance in steady state and dynamic conditions.

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.

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