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Novel Approach of Constant Switching-Frequency Inverter Control with Optimum Current Transient Response

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***Abstract*--Power converters with carrier based pulsewidth modulation (PWM) are currently the most widespread solution in industrial drive applications. However, the average dead time of the current controller affect the transient behavior of the whole drive. Therefore hysteresis and predictive current control can be seen as an alternative to PWM, providing enhanced transient response with the major drawback of variable inverter switching-frequency. In this paper a novel predictive current controller is proposed, using a penal weight in switching state selection for controlling the switching-frequency. Consequently inverter switching-frequency will no longer dependent on current controller sampling time facilitating sampling frequency to be increased up to some 100kHz at an inverter switching-frequency of just a few kHz. Hence – compared to classical predictive current control – a further reduction of the transient response time can be accomplished.**

***Index Terms*-- Control of Drive, Converter control, Modulation strategy, Variable speed drive**

I. Introduction

Current control strategy for a voltage as well as a current source inverter system is one of the most important parts of today's power electronic circuits. Since the current control loop is the innermost one of a cascaded control, it determines the dynamic behavior of the drive. Investigations showed that the current control loop properties affect the dynamic performance of the superposed loops by the factor of two or four [1]. It is thus necessary to achieve best possible transient response of this controller.

In conventional applications linear current control methods, such as PI-controllers in combination with pulsewidth modulation (PWM) are dominant and well established. PWM techniques obviously have several advantages compared to other control approaches like hysteresis or predictive current control. A marginal output ripple current, a constant inverter switching-frequency and related with that, a well defined load current spectrum, to mention some of the benefits, leads this control technique to be the most suitable for most drive applications. Under a dynamic point of view PWM technique has one major drawback, as mentioned in [1], because the sampling instant and the reference value are not statistically correlated. This defines an average dead time of the current controller of half a period of the used carrier signal. If not compensated, this effect even leads to undesired steady state deviations of the current space vectors, deteriorating the flux and torque decoupling in modern field orientated control (FOC) schemes.

Though the integral part in the controller eliminates this disadvantage at stationary operation, the transient response will still be strongly defined by the mentioned dead time. Increasing the carrier signal frequency leads to enhanced dynamic behavior of the control, however, with the side effect of an augmented thermal stress of the power semiconductors due to the increased switching-frequency.

Usually, switching losses and the associated junction temperature rise [3]-[5] characterize the upper limit of the allowable switching-frequency of an inverter. This maximum frequency can either be calculated out of a worst case thermal design, or directly measured as shown in [2]. An alternative way to improve the transient response is given in [6] introducing a strategy for dynamically changing the switching-frequency and the bandwidth by using a Dynamic Frequency Scaling (DFS) technique in combination with a PWM modulator. This allows the average switching-frequency to be held constant, however, with the drawback of a significantly increased demand on calculation power.

In this paper a new approach is proposed to optimize switching-frequency and transient response of the current control loop. In a first step a survey of hysteresis current control is given and in a second step a novel discrete time predictive current control technique is presented. These two control schemes provide best possible transient response but the classical approaches suffer from an undesired variable inverter switching-frequency [7] [8]. Main focus of attention hence will be on techniques eliminating this drawback by permitting constant or at least reduced switching-frequency, while maintaining the response speed of conventional hysteresis and predictive current controllers.

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II. Hysteresis Current Control With Constant Inverter Switching-frequency

Hysteresis band current control strategy is conceptual easy, both in theory as well as in hard or software implementation still delivering excellent performance. Marginal computational complexity and distinguished transient behavior is linked with a variable inverter switching-frequency and a not strictly limited current error. When realizing three independent hysteresis controllers, one for each phase, interaction may lead to a resulting current error in one phase with a magnitude equal twice the width of the hysteresis band. To overcome this major drawback many efforts have been made in the past. A space vector hysteresis current control avoiding undesired interactions between the phases and in addition permitting the use of zero voltage vectors for inverter switching-frequency reduction is described in [9]. Generally there are two approaches to achieve an adaptive hysteresis band and thus a constant inverter switching-frequency:

- Analytical approaches, describing the width of the hysteresis band as a function of switching-frequency, back EMF, the inverter output voltage, and other system parameters [10][11].

- Approaches for synchronizing the commutations of the hysteresis to a carrier signal [12]-[15].

Hysteresis current control today enjoys great popularity in Direct Torque Control (DTC) schemes. Two independent hysteresis controllers adjust, separated from each other, the demanded torque and flux. Almost all investigations in this field promising constant switching-frequency generally deal with the reduction of torque ripple. Therefore just the switching of the torque hysteresis controller is considered, trying to keep it at a constant frequency. Inverter switching-frequency on the contrary is a function of both, switching-frequency of torque and flux hysteresis control, which is analytically analyzed in [8]. Hence inverter switching-frequency will still show a distinct variation over motor speed.

Two strategies of solving the drawback of variable (torque) switching-frequency can be distinguished [16]: Variable torque hysteresis bands [20]-[22] and approaches to achieve switching at regular time intervals [16]-[19]. The second category generally subdivides the switching period into two or more states even with the help of a modulator in order to minimize torque ripple. The introduction of such a modulator however, leads to an increased averaged controller step response time.

III. Discrete Time Predictive Current Control With Constant Inverter Switching-frequency

Because of today's powerful microprocessor systems used in drive applications more complex control strategies such as predictive ones become of interest. The predictive current control approach has recently been studied intensively because it delivers a great degree of freedom in controlling multiple system parameters at once by introducing different quality functions in the switching state selection process [23]-[30].

Discrete predictive current control is typically based on a minimization process of a cost-function. In general a description of the switching state dependent current deviation at the next time step is given and the optimum and hence next switching state is calculated. This classical approach can be expanded by an additional penal weight k ($k \geq 0$) making switching transitions to another state less attractive. To obtain a switching-frequency reduction the penal weight is added to the predicted current error of all inverter switching states except the current one. Remaining in the current state thus gets more attractive. The quality function for a three phase inverter system providing seven different output voltages (six active and two zero voltage space vectors) therefore can be written in scaled values as:

$$q_{0..7} = \left| \underline{\Delta i}_{demand} - \underline{\Delta i}_{pred,0..7} \right|^2 + \boldsymbol{A} \cdot k \,, \qquad (1)$$

where $\underline{\Delta i}_{demand}$ equals the demanded current phasor to adjust the reference value, $\underline{\Delta i}_{pred,0..7}$ represents the values of the predicted current change phasors caused by each of the possible switching states and k denotes the penal weight, scaled by one or zero by matrix $\boldsymbol{A}$.

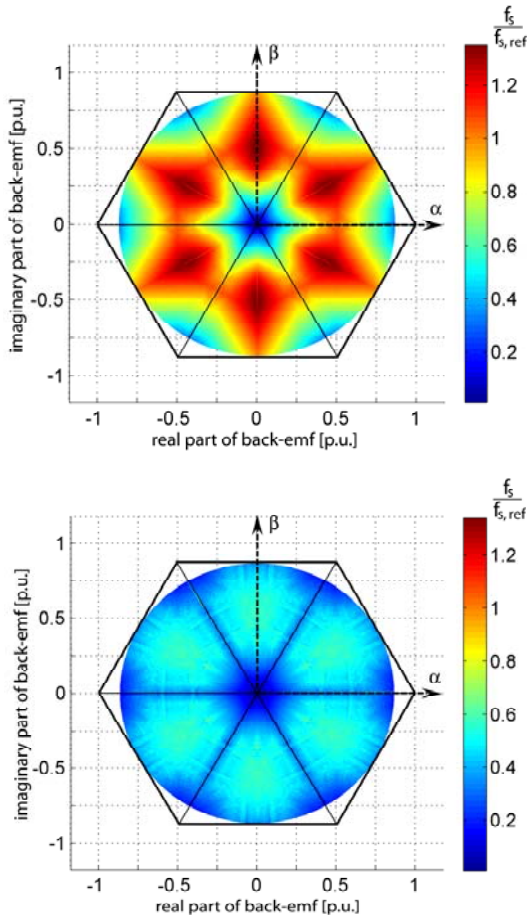

Fig. 1: Comparison of inverter switching-frequency using classical predictive current control (upper) and the proposed structure (lower) depicted by (1) as a function of back emf. (steady state operation)

The scaled inverter switching-frequency of the fixed step predictive current control is shown in Fig. 1. In the upper diagram the penal weight k is set to zero corresponding to the classical approach. In the lower diagram the modified controller, as proposed in this paper is shown using a scaled penal weight of $k=1.0$. The figure represents ideal reactive steady state operation with constant current phasor magnitude. The value of the back emf is changed in magnitude (0-0.866) and angular position (0-2π). The different values of the switching-frequency are marked with colors.

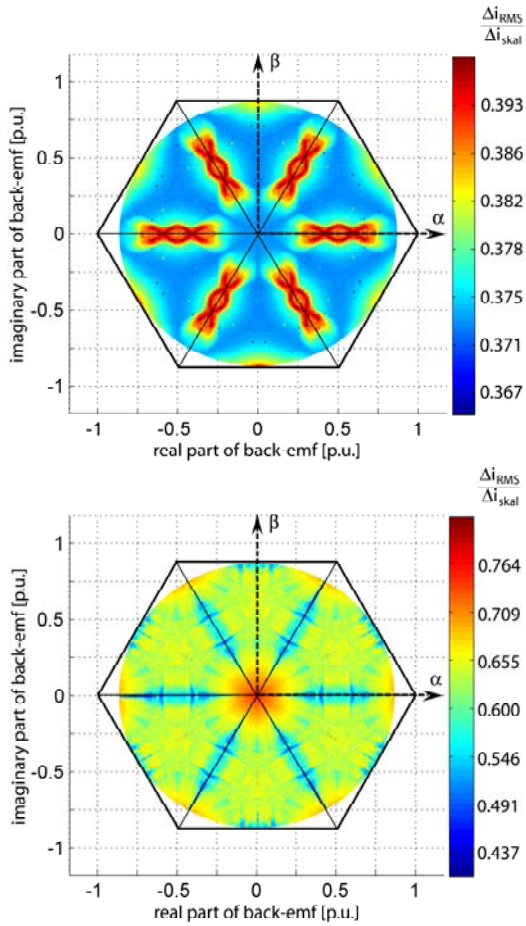

Fig. 2: Comparison of current ripple RMS using classical predictive current control (upper) and the proposed structure (lower) depicted by (1) as a function of back emf. (steady state operation)

As can be seen, the inverter switching-frequency is a function of the back emf, both in magnitude as well as in angle. For the classical approach of the predictive current controller the switching-frequency has very distinct maxima in the centers of each sector. Using the proposed penal weight approach clearly reduces these maxima and smoothes the overall inverter switching-frequency.

Obviously reducing inverter switching-frequency by using the mentioned penal weight will increase the current ripple RMS, additionally depicted in Fig. 2. A comparison to Fig. 1 shows, that back emf areas of high switching-frequency are characterized by a lower current ripple than those of low frequencies. However, the modified approach reduces current ripple RMS compared to classical predictive current control, considering same inverter switching-frequency, by oversampling [31]-[33]. Because switching-frequency no longer depends on current controller sampling time. Thus high sampling frequencies shorten lead time of possible switching transitions. Hence, switching operations temporally can be performed close to their optimum instant.

Considering inverter switching-frequency to be scaled and averaged over one fundamental period of the back emf the function only depends on the magnitude of back emf (indicated as modulation index m) and the penal weight:

$$f_S(m,k) = \frac{1}{2\pi} \cdot \int_0^{2\pi} f_S(m,\gamma_{EMK},k)\,d\gamma_{EMK} \qquad (2)$$

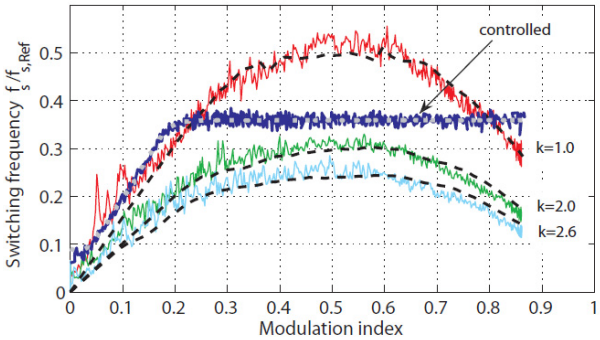

Fig. 3: Averaged inverter switching-frequency as a function of modulation index for different constant penal weights k (simulation (black dashed); measurement (solid)) and a switching-frequency controller (demanded (grey dotted); measurement (solid))

Fig. 3 illustrates the switching-frequency characteristics over modulation index using a switching-frequency controller as well as the traces for three different constant values of penal weight k. The switching-frequency controller is used to adjust the actual measured switching-frequency (blue solid) to a reference trace (gray dotted) using the proposed penal weight [30]. In addition, simulation results (dashed black traces) as well as measured data (solid traces, red $k=1.0$, green $k=2.0$ and magenta $k=2.6$) from a 220kW industrial high precision load dynamometer (induction machine, 8 poles, 220kW, 500V, 304A, 233Hz) are depicted.

In addition, Fig. 4 (lower) shows the traces of the penal weight k, adjusted by the switching-frequency controller (SFC), for three different inverter switching-frequency characteristics. Therefore the modified predictive current controller with a sampling time of 10µs has been used. As can be seen, the penal weight is close to zero at small modulation indices to maximize the switching-frequency. Enlarging the modulation index, e.g. during acceleration of the drive, would result in an increasing inverter switching-frequency up to more than 19kHz at a modulation index of approximately $m=0.5$. To limit switching-frequency to 3.3kHz (1), 5kHz (2) and 8.3kHz (3), the penal weight has to be increased up to the point of maximum switching-frequency. At higher

modulation indexes the value of k has to be reduced to hold inverter switching-frequency almost constant.

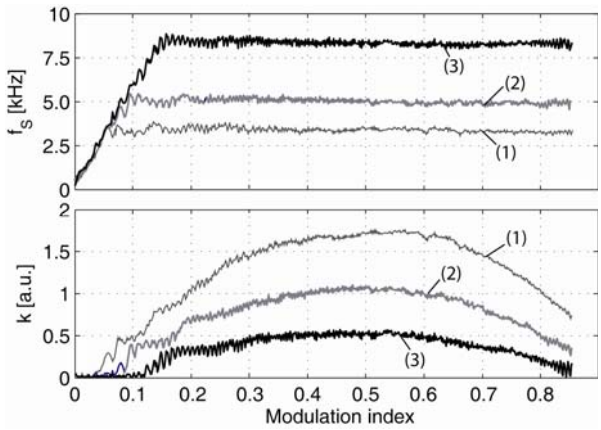

Fig. 4: Almost constant inverter switching frequencies as a function of modulation index m (upper diagram) as well as the adjusted penal weights k (lower diagram) to control the inverter switching-frequency.

Besides the capabilities of controlling inverter switching-frequency the modified predictive current controller improves the output voltage spectrum caused by the inverter. To clarify this improvement the voltage spectra for three different current control strategies have been simulated and are depicted in Fig. 5. The drive was operated at 650rpm at comparable inverter switching frequencies respectively.

In the upper diagram of Fig. 5 shows (based on a 2kHz carrier-based space-vector PWM with a PI-current controller) that the harmonic content of the output voltage spectrum is concentrated around the carrier frequency and multiples of it. The middle diagram illustrates the spectrum by using classical predictive current control. To obtain an averaged inverter switching-frequency at this speed (proportional to the modulation index in base speed range) according to eq. (2) of approximately 2kHz, the current controller sampling time was adjusted to $T_S=500\mu s$. As can be seen, the spectral lines are spread over the whole frequency range with a slight distinctive maximum around the average inverter switching-frequency f_S. A possible explanation for this can be found in the upper diagram of Fig. 1. Considering a constant speed or more precisely a constant modulation index m, inverter switching-frequency over a whole revolution of the back-emf is characterized by very distinct maxima and minima. This involves a broad distribution of the switching-frequency around its averaged value leading to the spectrum depicted in the middle diagram of Fig. 5. As shown in the lower diagram of Fig. 1, the modified approach minimizes the minima and maxima of switching-frequency over one revolution of the back-emf. Furthermore the switching-frequency controller (SFC) additionally reduces the distribution of the inverter switching-frequency by adjusting it to the reference value. This can also be seen in the output voltage spectrum depicted in the lower diagram of Fig. 5.

The introduced modified predictive current controller also reduces total harmonic distortion of the output voltage (THDv). Compared to the applied PWM strategy the THDv can be reduced by 8.87%. Compared to the classic predictive current control strategy it even can be reduced by 14.63%.

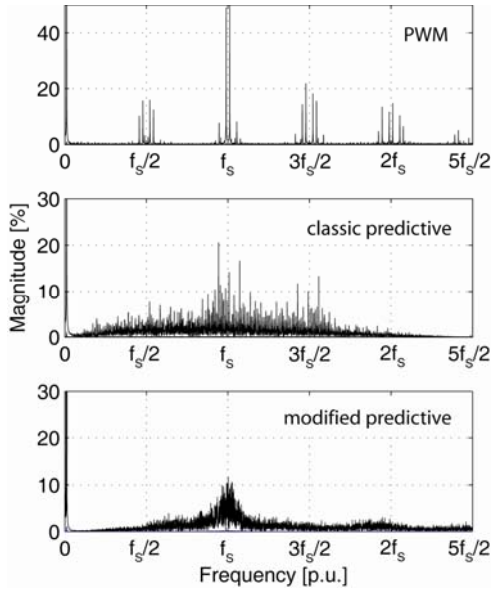

Fig. 5: Simulated load voltage spectrum for three different current control strategies at a speed of 650rpm when the inverter is operated at comparable switching frequencies respectively. (carrier based PWM – upper diagram: T_S=500µs; classic predictive – middle diagram: T_S=100µs; modified predictive – lower diagram: T_S=10µs)

The ability to control the switching-frequency to a demanded reference trace over the whole output frequency range has two major advantages:

- Enhanced thermal exploitation of the inverter according to the possibility of controlling inverter switching-frequency dependent on load parameters (dynamic overload).

- Improved transient response even compared to classical predictive current control. In the classical approach, the controller task time determines the maximum inverter switching-frequency. Especially in high power applications switching-frequency is limited to a few kHz leading to a minimum task time of several 100µs and the corresponding controller dead time. Using the proposed structure enables task times of only a few µs, drastically reducing control deadtime while still keeping switching-frequency below the limit of a few kHz.

In addition, to illustrate the improved transient current response of the introduced control scheme Fig. 6 shows an externally impressed load step at time instance t=0.01. As the current controller sampling time in this case was 10µs (100 kHz) the average current controller deadtime is 5µs corresponding to half the task time. Nevertheless, the switching-frequency controller limits the frequency to approximately 5 kHz. To realize the same transient behavior using a classical predictive current controller the switching-frequency would be increased to more than 15 kHz corresponding to a factor of 3-4. Using a standard

PWM modulation technique switching-frequency would be 100 kHz corresponding to a factor of 20.

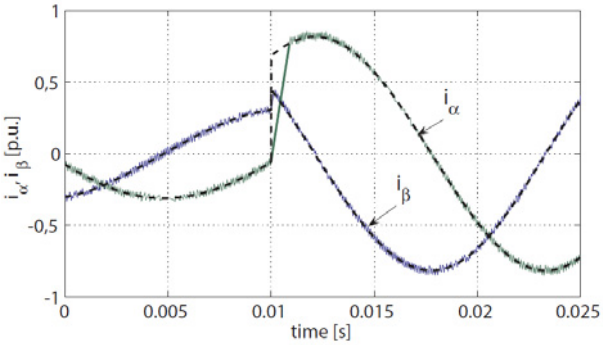

Fig. 6: Transient current response of the proposed current controller to an external step from no load to 80% rated torque at t=0.01 s at a speed of 650 rpm (demanded (black dashed); measured (solid))

IV. Conclusions

Using a penal weight in the switching state selection as proposed in this paper for predictive current control allows a very effective control of the inverter switching-frequency. As a result the inverter switching-frequency no longer depends on current controller sampling frequency. Transient current response as well as output voltage spectrum can hence be improved by increasing the sampling frequency while keeping inverter switching-frequency below the maximum tolerable level. Consequently, thermal stress of the power semiconductor will not be increased, although current controller response rates of a few microseconds can be achieved. In addition, a dynamic overload current can be realized by switching-frequency reduction, what reduces the switching losses as a trade for the higher conduction losses. This tradeoff can keep the thermal losses of the power semiconductors below the specified limits.

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