A Novel Solution for Phase Current Sensing in PWM-VSI Based AC Drives

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Abstract
In this paper a novel solution for phase current sensing in PWM–VSI based AC drives is presented, based on a resistive sensors put in series to the lower switches of the inverter legs. It allows current sensing during each modulation cycle within the whole operating area of the inverter. Inside the overmodulation area a simple but effective modification of the modulation pattern is proposed which does not increase the complexity of the implementation almost at all. Encumbrance, complexity and cost resemble the solution using a single sensor on the DC link, but almost no operating limitations are introduced. Experimental results are presented in order to show the performance and effectiveness of the proposed method.

Introduction
Ac drives are widely used for high performance servo drives. Fundamental to the successful operation of all servo drives is the ability to quickly and accurately control the motor torque, necessitating precise closed loop control of the motor phase currents. The feedback for the current control loop is usually obtained by sensing the instantaneous currents in at least two of the motor phases by means of Hall-effect based current transducers. This solution causes several disadvantages from the standpoint of drive cost, encumbrance and non-linearity, especially for low power and low cost drives.

In the last few years the desire to reduce the sensor requirements of ac drives has led to the development of strategies capable to derive motor phase current signals from resistive shunt sensors properly arranged inside the power converter [1]-[11]. This approach has the advantage of a relatively low-cost sensing but introduces a series of disadvantages and trade-offs that must be taken into account when designing a motor drive system. Particularly the problems related to the derivation of phase currents in ac drives from resistive sensors can be resumed as follows:

- the current flowing in the sensors is the motor phase current. This produces two side effects: the modification of the equivalent series reactance of the system and the modification of the resistance value of the sensor due to the Joule effect;
- low equivalent series inductance resistors are needed due to high frequency switching and dv/dt of the power converter stage;
- the amplitude of the output signals of the current measurement circuitry is very small (about some tenths of mV) leading to small values of signal to noise ratio (SNR);
- the signal of interest is a small differential value but its common-mode components could hundreds of volts depending on the sensor placement inside the power converter;
- the isolation between the control and the power stages of the drive is lost unless isolated op-amps are adopted;
• depending on the operating condition and on the adopted sensors placement inside the power converter, the relation between the motor phase currents and signals from those sensors is not well established or could in some cases be lost.

One of the most known solution is the one which uses only one shunt resistive sensor on the negative dc bus of the inverter [2][5][9]. It is easy to demonstrate that the current flowing in the sense resistor is equal to one of the motor phase current or its opposite when each particular configuration of the inverter switches is considered [4][9]. Unfortunately the behaviour of a feedback control system does not assure that each phase current always flows in the negative dc bus for a sufficient amount of time. In fact there exist some particular operating conditions in which the relation between the shunt sensor signal and the motor phase current is lost. Thus it is very important to define alternative strategies that permit to overcome those limitations. Some of the proposed strategies are based upon additional hypothesis which are made on the particular drive and cannot be generalised. In [3] the adopted brushless dc control technique, where only two of the motor phases are conducting at the same time while the third one is open, always allows the correct reconstruction of motor phase currents from a single dc link current measurement. In [2] the information provided by the dc link sensor is used to reconstruct the amplitude of the motor phase current and not the instantaneous value, thus being useless for motor control purposes. Some of the strategies have been proposed recently in literature [2][6] which introduce modifications to the modulation algorithm in order to guarantee the reliability of the measurements from the resistive sensors in all the operating conditions. Unfortunately the modification to the modulation pattern causes distortion on the output voltage and consequently on output phase currents. Moreover some of those techniques introduce modifications during more than one single modulation cycle, thus worsening the dynamical response of the control system. Some interesting approaches are based on the estimation of the motor phase currents when the reliability of the dc link measurements is not guaranteed [7][8]. They normally makes use of prediction-correction algorithms, thus introducing additional computational burden to the drive system. Other approaches to phase current sensing are based on three shunt resistive sensors on each low-side transistor emitter. As long as the measurement circuit is referenced to the dc link common, this approach eliminates the common-mode voltage problem. However, a second problem arises: the measured current is no longer motor phase current, but half-bridge current. If the low side switch is conducting (through either the transistor or freewheeling diode) then the current is equal to that motor phase current. This certainly occurs periodically throughout the modulation cycle. In order to sample all three motor phase currents simultaneously (for zero phase-shift), all three low side switches must be conducting. This only happens at a zero vector state in which all three low side switches are on. Even for conventional modulation methods like sinusoidal or space vector modulation [12][13], the width of the pulses to be measured can become very narrow, placing an increasing performance burden on the sample an hold circuit. When the modulation index exceeds a certain value, however, the current pulses disappear altogether. This condition is met when overmodulation methods are employed or when the non ideal behaviour of the system is taken into account. Those situations require a different and more complex strategy for motor current reconstruction if this method of sensing is used.

Finally, the application of motor phase current sensing strategies based on the information provided by resistive sensors properly arranged inside the power converter has highly been conditioned by the impossibility to operate in the whole operating area of the inverter. These also relates to aspects involving the non ideal behaviour of different parts of the system (presence of dead time, rise and settling time of the current, sampling time of the A/D conversion module) [10]. Depending of the adopted sensing strategy and the system configuration, different approaches have been proposed in order to minimise those effects and extend the operating area inside which reliable sensing of phase current is guaranteed.

In this paper a novel solution for phase current reconstruction based on three shunts resistors is presented which is capable to overcome all the limitations described in [10]. The proposed approach allows to accurately determine motor phase currents during each modulation cycle within the whole
operating area of the inverter. Inside the overmodulation area a simple but effective modification of the modulation pattern is proposed. Encumbrance, complexity and cost effectiveness resemble the solution using a single sensor on the DC link [9], but almost no operating limitations are introduced. The performance and effectiveness of the proposed current sensing solution has been verified experimentally by means of a three phase induction motor test drive based on a digitally controlled voltage source inverter.

**Phase Current Measurement Based on Shunt Resistors**

The inverter topology which has been adopted to develop and test the proposed phase current sensing methodology is shown in Fig. 1a. A low resistance and low equivalent-series-inductance (ESL) shunt resistive sensor has been added on each low-side transistor emitter. The analogical signal across each sensor is conditioned by means of a simple op-amp stage. Its output provides the input to the A/D conversion module of the µC DSP which has been used for the experiments. It is clear that, as long as the measurement circuit is referenced to the dc link common, this approach is not affected by any common-mode voltage problem as in the case of the resistive sensors placed in series with the motor phases. However, a second problem arises: the measured current is no longer motor phase current, but half-bridge current. Nevertheless it is possible to demonstrate that the measured current relates to the motor phase current during certain portions of the modulation cycle and it is a function of the applied voltage vector and the system non ideal behaviour. It is then important to analytically identify the influence of that behaviour and to define the proper sampling instants of the signals across each resistive sensor which guarantee reliable motor phase current samples.

**Sampling on zero vector state**

The simpler principle of the phase current measurement based on the information provided by the resistive sensors on the low-side transistor emitter is resumed in Fig. 1b. If the low side switch is conducting (through either the transistor or freewheeling diode) then the current is equal to that motor phase current. This certainly occurs periodically throughout the modulation cycle. In order to sample all three motor phase currents simultaneously (for zero phase-shift), all three low side switches must be conducting. This only happens at a zero vector state in which all three low side switches are on, that is when vector $v_0$ is applied (refer to Fig. 1b). In conventional PWM methods like sinusoidal or space vector modulation, the width of the pulses to be measured can become very narrow, placing an increasing performance burden on the sample and hold circuitry. Moreover, the sampling instant should be arranged to occur at the middle of the resulting zero vector state in order to ensure that the resultant samples reflect, as closely as possible, the average phase currents. When the modulation index exceeds a certain value, however, the current pulses disappear altogether. This condition is met when overmodulation methods are employed. Moreover the non ideal behaviour of the system must be taken into account in order to assure reliable and precise current sampling. Those situations require a different and more complex strategy for motor current reconstruction, if this method of sensing is used.

![Fig. 1: a) adopted inverter topology; b) configuration of the inverter when the state $v_0$ is used.](image-url)
Fig. 2: Comparison between a) ideal and b) actual situations for sampling phase currents.

The basic problems are recalled in the following.
Let us refer to Fig. 2b and suppose $t_0 = t_0'$. In this case system non ideal behaviour has been considered, that is the following quantities have been introduced:
- $t_{dt}$ is the time required to prevent a shoot-through condition across the dc supply when switching between upper and lower switch of one inverter leg (dead time);
- $t_{rs}$ is the time needed for the phase current to reach a well established value after switches commutation (rise and settling time);
- $t_{sh}$ is the time needed for the sample and hold circuitry to acquire the correct current sample (sampling time).

Then, taking into account $t_{sh}$, $t_{dt}$, $t_{rs}$ and considering that $t_0' = t_0 = t_{null}/2$ [12][13], a minimum time $t_{0,min}$ for the zero vector state $v_0$ should be guaranteed, given by:

$$ t_{0,min} = t_{dt} + t_{rs} + 2 \cdot t_{sh} \quad (1) $$

The sampling instant should be modified from the ideal situation $\tau = 0$ to the actual one:

$$ \tau_i = \frac{t_0/2 - (t_0'/2 - t_{null}/2)}{2} + t_{driver} \quad (2) $$

The term $t_{driver}$ has been introduced taking into account the overall delay which is introduced on the commutation instants by the power switches driving circuitry.

The minimum time $t_{0,min}$ for the zero vector state $v_0$ results in an upper limitation on the amplitude of the voltage vector as shown in Fig. 3a, i.e. to a limitation of the maximum modulation index.

Fig. 3: Resistive sensors method limitations (shaded area) – a) standard method; b) novel solution.
Extending the measurement range

The possibility to extend the measurement range to almost all the shaded area in Fig. 3a too, relies on the following considerations:

- an active vector is applied for a significant amount of the modulation cycle when high modulation index is used;
- according to the adopted modulation strategy (Adjacent Vectors–Space Vector Pulse Width Modulation, AV–SVPWM, [12][13]) the null vector states \(v_0\) and \(v_7\), an active vector with one upper switch on \((v_1, v_2, v_4)\) and an active vector with two upper switches on \((v_3, v_5, v_6)\) are always employed within each modulation cycle;
- two significant phase currents can be measured on the corresponding resistive sensor when an active vector with one upper switch on \((v_1, v_2, v_4)\) is applied;
- only one significant phase current can be measured on the corresponding resistive sensor when an active vector with two upper switches on \((v_3, v_5, v_6)\) is applied.

Thus, inside the shaded area we do not have to sample anymore when \(v_0\) is applied, but we have to sample when the active vector of the modulation pattern with one upper switch on is applied (either \(v_1\) or \(v_2\) or \(v_4\)). In this case two significant currents are obtained and the third one can easily be calculated.

As for AV–SVPWM, a simplifying condition occurs being that active vector always the first one to be applied after \(v_0\). Thus the sampling instant does not depend on the sector the reference voltage vector belongs to but is a function only of the application time of the zero vector state and the previously defined system parameters. It is possible to demonstrate that the two significant currents remain the same for 120° degrees while the third one could not be reliably sampled. The situation is resumed in Fig. 3b, the last condition being synthetically indicated by \(i_{\text{shunt}} = 0\).

As in the case of sampling on zero vector state, the sampling instant \(\tau_i\) should be arranged in order to ensure that the resultant samples reflect, as closely as possible, the average phase currents. We can certainly assume that this condition is met when the current is sampled at the beginning of the first active vector. In fact this could be considered a good approximation because of the small value of zero vector state application time inside the shaded area of Fig. 3a. Thus the sampling instant \(\tau_i\) can be simply chosen as follows:

\[
\tau_i = t_0/2 + t_{ds} + t_{rs} + t_{\text{driver}}.
\]

Some problems are still present when the voltage vector passes an active vector with two upper switches on. In this case that vector is being applied for a significant amount of the modulation cycle while the application time of the active vector with only one upper switch on is too small to permit reliable current samples to be taken. The area where this condition appears is very small and corresponds to the shaded polygons in Fig. 3b. Such a problem could be satisfactory solved by introducing a local perturbation of the reference voltage vector when it belongs to those shaded polygons.

Two strategies are proposed (Fig. 4) which require only simple modifications of the modulation algorithm and do not increase the complexity of the implementation almost at all. The idea is to force the reference vector to stay on those boundaries of the shaded polygons which are inside the limit voltage hexagon. The perturbation on the reference voltage vector is performed either by placing a lower limit to the time the active vector with one switch on is applied for \((t_4 = cost = t_{\text{min}} \ \text{locus, Fig. 4a})\) or by reducing its amplitude in order to allow a proper sampling on the zero vector state \((t_0 = cost = t_{0,\text{min}} \ \text{locus, Fig. 4b})\). Notice that the shaded polygon in Fig. 4 has been intentionally exaggerated for clarity though the amount of the needed perturbation (which however depends on systems parameters (dead time, etc.)) is limited to small values.
Fig. 4: Limitations on the reference voltage vector.

Some examples of perturbed reference voltage vector loci in the αβ reference frame are shown in Fig. 5, as obtained through the experimental system based on the space vector algorithm commutation instants.

Fig. 5: Perturbed voltage reference loci: a), b), c) $t_{\text{min}} = 4\mu s$ ($V_{\text{ref}} = 0.57, 0.6, 0.7$); d), e) $t_0 = t_{0,\text{min}}$ ($t_{\text{min}} = 20\mu s$ and $4\mu s$).

Finally it is interesting to point out a problem that comes out when the reference voltage vector passes the sector commutation zone around one of the shaded polygons in Fig. 3b and the application time of the zero state vector approaches to the limit (1). In fact, referring to Fig. 6, the modulation pattern becomes unsymmetrical due to the sector change and the application time of the vector $v_4$ (100) becomes a half of the time needed to obtain a reliable phase C current sample. As it is clear, the same thing does not occur for phase B.

The problem can easily be solved by maintaining the previous current values in the sampling period immediately following the sector change. This solution does not cause any problem from the current control point of view as it is adopted only three times for each electrical period.
Implementation Issues

The performance and effectiveness of the proposed current sensing solution has been verified experimentally by means of a three phase induction motor test drive based on a digitally controlled voltage source inverter. The control hardware is based on a Texas Instruments TMS320F240 µC Digital Signal Processor control board dedicated to control of electrical drives which also integrates an IGBT based Intelligent Power Module. The rotating reference voltage is arranged in the two phase fixed $\alpha\beta$ reference frame and impressed to the motor by means of the AV–SVPWM. Both the amplitude and the frequency of the reference voltage vector can be changed independently.

The implementation of the solution illustrated in the previous section relies on the possibility to simply synchronise the A/D conversion module operations to the PWM pattern. Referring to Fig. 7a, it is possible to see that the A/D start of conversion (SOC) signal could be provided by the compare logic associated to the same timer used to generate PWM signals without recourse to any external hardware. This is done by writing the proper compare value depending on the sampling instant $\tau_i$ as a function of the operating condition. Also the selection of the two channels being converted is done by means of the two 8 channels multiplexers which provide the inputs to the A/D converters (Fig. 7b). One of the two currents ($i_{\text{shuntA}}$) is sent to both the multiplexers to allow the simultaneous sampling of each combination of the three currents.

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**Fig. 6:** PWM modulation patterns across one sector commutation.

**Fig. 7:** a) µC DSP subsystems used to synchronise current sampling; b) Block diagram of the A/D conversion module.
At software level, the control period is synchronised with the modulation carrier underflow. The reconstruction of phase current is done at the beginning of each modulation cycle after the two current samples have been converted. On the contrary the selection of the two channels being converted at the beginning of the present modulation cycle is performed in the previous one once the new reference voltage vector has been calculated. Also the need for local perturbation of the reference voltage vector is checked and performed by acting on the commutation instants synthesised by the space vector modulation algorithm. Each modification to the PWM registers take place only at the beginning of the next modulation cycle.

**Experimental Results**

A preliminary analysis was carried out on the experimental set-up in order to identify the values of those parameters which were considered to model the non ideal behaviour of the drive system. The obtained results has been reported in a previous paper [10] and will simply be resumed here: $t_{\text{driver}} = 400\,\text{ns}$, $t_{\text{th}} = 1.0\,\mu\text{s}$, $t_{\text{rs}} = 400\,\text{ns}$, $t_{\text{sh}} = 1.0\,\mu\text{s}$.

Fig. 8 is intended to show the shape of one resistive sensor current over a modulation cycle together with the corresponding modulation pattern. One can notice that the current is non zero only during those patterns where the corresponding lower switch is on. Moreover the sampling instant (rising edge of ADC-SOC signal) has been chosen to occur in the middle of the resultant current as provided by (2).

![Fig. 8: Resistive sensor current a) over a modulation cycle and b) when the reference voltage vector is in the overmodulation area (when sampling always on zero state vector).](image)

In Fig. 9 the problem occurring when the reference voltage vector passes the sector commutation zone around one of the shaded polygons in Fig. 3b and the application time of the zero state vector approaches to the limit (1) is highlighted.

![Fig. 9: Resistive sensor current sampling error when the reference passing the sector commutation zone around one of the shaded polygons: a) no action and b) sample maintained.](image)
The falling edge of the *ADC-SOC* signal is synchronised with the modulation carrier underflow (start of control period), while the rising edge defines the A/D start of conversion which is adjusted at each cycle depending of the operating conditions. The D/A output is updated before the end of the control period (about 50µs after the start of sampling period). As evident in Fig. 9a, the current value \(i_{\text{DAC}}\) in the control period after the sector commutation zone is wrong as the current \(i_{\text{shuntC}}\) flowing in the sense resistor is unsymmetrical with respect to the sector change. As it was said in a previous section, the problem has been solved by maintaining the previous current values in the sampling period immediately following the sector change. In Fig. 9b the signal \(\text{hold}_{\text{DAC}}\) has been reported in order to show the control period when this occurs. Fig. 9 also allows to point out a problem which affects the reliability of the signal across the resistive sensor when the high side duty cycle of the corresponding leg approaches unity. The problem appears due to the adopted power switches driving electronics, making use of bootstrap capacitor for the high side. In fact it is possible to demonstrate that, when the low side switch is closed, the current \(i_{\text{shuntX}}\) in the sense resistor is the difference between the motor phase current \(i_X\) and the bootstrap capacitor charging current \(i_{\text{boot}}\). The situation is highlighted in Fig. 10 where, even in the case the phase current is different from zero (Fig. 10a), the sense resistor signal is lost. Particularly, in Fig. 10b, the bootstrap capacitor charging current is shown, its shape clearly approaching that of an RC circuit. The reported problem is however present, also when sampling on the null state vector only.

Finally, Fig. 11 is presented to show the effectiveness of the proposed current sensing solution.

![Fig. 10: Bootstrap capacitor charging current affects the reliability of sense resistor signals.](image1)

![Fig. 11: Actual and reconstructed phase current in two different operating conditions: a) voltage vector inside inner hexagon (see Fig. 3a); b) voltage vector outside inner hexagon.](image2)

**Conclusions**

In this paper a novel solution for phase current sensing in PWM-VSI based AC drives has been presented, based on a resistive sensors put in series to the lower switches of the inverter legs.
Differently from the approaches based on a single sensor on dc link, the method allows current sensing during each modulation cycle within the whole operating area of the inverter. Inside the overmodulation area a simple but effective modification of the modulation pattern is proposed which does not increase the complexity of the implementation almost at all. The performance and effectiveness of the proposed current sensing solution has been verified experimentally by means of a three phase induction motor test drive based on a digitally controlled voltage source inverter. Encumbrance, complexity and cost resemble the solution using a single sensor on the DC link, but almost no operating limitations are introduced. The main problems of the proposed current sensing solution appear to be both offset and gain errors in the measurement circuitry. The former can be solved on line by means of an offset self calibrating procedure. The latter can be significantly reduced by a proper design of the amplifying circuitry and the choice of low tolerance components with low temperature drift. Moreover an off-line self calibrating procedure by means of reference current signal injection can be adopted. Finally, a problem heavily affecting the reliability of the proposed current sensing method has been pointed out. It appears when bootstrap-based power switches driving circuitry is adopted for each high side of the power converter and its duty cycle approaches unity. It has been demonstrated by experiments that the bootstrap capacitor charging current influences the reliability of the signals out of resitive sensors, also in the case measurements are taken on the null state vector only. That problem must then be accurately taken into account when designing the drive system and the power switches driving circuitry.

References


