New Modulation Encoding Techniques for Indirect Rotor Position Sensing in Switched Reluctance Motors

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***Abstract*— Rotor position information is essential in switched reluctance machines (SRMs) in order to generate the phase excitation pulses for optimum control of the motor. In this paper, two new methods of indirect rotor position sensing are presented. These methods are based on the modulation techniques which are commonly used in communication systems. The instantaneous phase inductance of a non-conducting phase is extracted in an encoded form using a modulator consisting of electronic circuitry. The signal is then decoded to obtain the instantaneous inductance from which the shaft position is determined. An analysis of these modulation techniques along with the implementation scheme using a microcontroller is presented in this paper.**

I. INTRODUCTION

THE DOUBLY salient switched reluctance machines (SRMs) are receiving increasing attention for their advantages in specific adjustable speed motor drive applications [1]. The SRMs are easily operable in four quadrants and provide excellent speed-torque characteristics through appropriate positioning of the phase excitation pulses with respect to the rotor position. Furthermore, the simpler power electronics requirement is a definite advantage for SRMs in many consumer product applications. The wide speed range of operation also make it suitable for a variety of unique applications.

The SRM operates on the basis of varying reluctance in its magnetic circuit. A schematic diagram of the SR motor with six stator poles and four rotor poles is shown in Fig. 1. The stator and rotor poles appear in pairs but are usually of unequal numbers. The stator windings on diametrically opposite poles are connected in series to form one phase. The rotor does not have any winding and is made of magnetic steel laminations. Four phase 8-6 or a three phase 6-4 structures are the most common SRM configurations. When a stator phase is energized, the nearest rotor pole pair is attracted towards the energized stator phase thus minimizing the reluctance of the magnetic path. Therefore, by energizing consecutive stator phases in succession it is possible to develop constant average torque in either direction of rotation. The developed torque is proportional to the rate of change of stator phase inductance.

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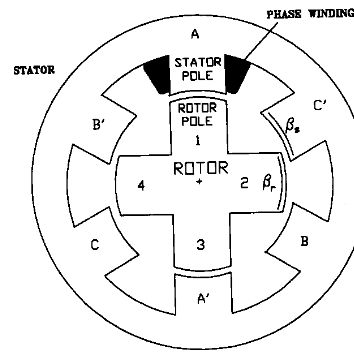

Fig. 1. Cross-section of a three phase 6-4 SRM.

The phase inductance of an SRM varies from a maximum when a rotor pole is aligned with the corresponding stator pole and is minimum when the corresponding poles are unaligned. For optimum torque production, the phase current is switched on prior to the rising inductance period and is switched off before the decreasing inductance period to allow the current to decay completely so that no negative torque is produced. For precise control of this phase magnetization, an accurate measurement of the rotor position is necessary.

The rotor position can be measured directly by using mechanical devices, such as Hall sensors or opto-interrupters with slotted disk [2, 3]. However, due to poor resolution in the above schemes resolvers or encoders are used to derive the shaft position in more sophisticated drives. The difficulty in using these devices in space restricted applications and the additional cost involved have led to the research in developing indirect rotor position sensing schemes. Several such indirect schemes have already been reported in the literature [4, 5, 6, 7, 8]. All these indirect sensing techniques are based on measurements of the periodically varying phase inductance. Acarnley et al. [4] first suggested monitoring current waveforms in stepper and switched reluctance motors and successfully implemented the scheme. MacMinn et al. [6] and Harris et al. [7] also used similar techniques to develop sensorless schemes. These techniques are based on injecting diagnostic pulses to an unenergized phase from the main converter and the rate of rise of current gives a measure of the instantaneous inductance. The rotor position is obtained from the phase inductance information using an inverse function or a conversion table. Since the phases are pulsed from the main converter, the current may be significant enough to generate a negative torque. Also the current takes a longer time to decay and the next sensing pulse cannot be applied until the previous one has fallen to negligible levels. This decreases the frequency of sampling and hence the resolution of position sensing. A method of using an external small signal circuitry to pulse the unenergized phases has been described in [8]. A more sophisticated method of using a state observer which models the machine characteristics and estimates the rotor position as a state variable has been described by Lumsdaine et al. [9]. This method is computation intensive and requires accurate knowledge of machine and load parameters.

Some of the methods mentioned above add complexity to the controller particularly if good starting and running performance are to be achieved with a wide range of load torque and inertia. Moreover, some of these methods are not always robust and may be susceptible to switching noise. In this paper, two new methods of indirect position sensing based on modulation techniques, that provide accurate estimates in the presence of switching noise, are presented.

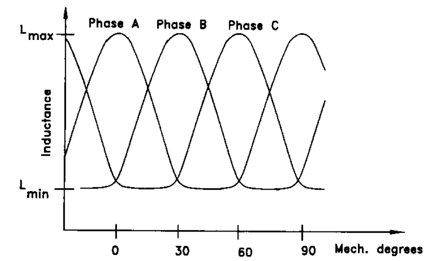

Fig. 2. Inductance profiles for a three phase SRM.

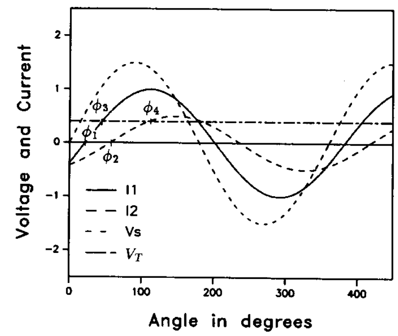

Fig. 3. Sinusoidal voltage and currents corresponding to $L_{\min}$ and $L_{\max}$.

II. Modulation Encoding Techniques

The modulation encoding techniques of rotor position detection are based on extracting the periodically varying phase inductance in an encoded form by applying a high frequency carrier signal. The signal containing the information, in this case, is the phase inductance $L(\theta)$ which is assumed to have a much smaller frequency of variation compared to the carrier frequency. The encoded inductance information is decoded using a suitable demodulation technique.

In an SRM, the phase inductance varies significantly with respect to the rotor angular position. Fig. 2 shows the inductance profile of a three phase 6-4 SRM for one electrical cycle. In three phase SRMs typically only one phase is energized to produce torque at any instant and there is always at least one phase which is switched off and not carrying any current. Hence if phase A is conducting, phase B and phase C are off and can be driven from an external oscillator to obtain the phase inductance in an encoded form. The signal level of this oscillator is negligible, in comparison to the voltage level of the converter, and no significant torque is generated. The method of frequency modulation (FM) encoder technique is one way of extracting the phase inductance using an external oscillator and the details of the method appear in [8]. The basic concept of the FM encoder technique is to generate a square wave whose frequency is linearly proportional to the instantaneous inductance of a phase. A simple L-F converter which maintains a linear relationship between inductance and time period was used for the purpose. Two other modulation techniques used in communication systems, namely the phase modulation and the amplitude modulation, can also be used to extract the phase inductance information in an SRM. These can have specific advantages over the FM method.

The phase modulation (PM) and the amplitude modulation (AM) techniques are based on the phase and amplitude variations, respectively, of the phase coil current due to the time varying inductance when a sinusoidal voltage is applied to the phase coil in series with a resistance R [10]. The current flowing through the circuit in response to the alternating voltage is a function of the circuit impedance. Since the coil inductance is varying periodically, the phase angle between the current and the applied voltage also varies in a periodic manner. Fig. 3(a) shows the input sinusoidal voltage and the current waveforms $i_1(t)$ and $i_2(t)$ corresponding to the minimum and maximum phase inductances respectively. Angles ϕ_1 and ϕ_2 are the corresponding phase angles by which the phase current lags the input voltage. Furthermore, the current has the maximum peak for the waveform of $i_1(t)$ and the minimum peak for the waveform of $i_2(t)$. The PM encoder technique measures the instantaneous phase angle on a continuous basis while the AM encoder technique measures the peak amplitude. These instantaneous measurements contain the phase inductance information which can be obtained after passing the signals through a demodulator. The demodulator generates a signal that represents the phase inductance as a function of the rotor position. Using an inverse function or a conversion table the rotor angles can be estimated for the controller section. The block diagram of the scheme is given in Fig. 4. The implementation schemes for the two methods are described below.

A. PM Encoder Technique

A sinusoidal carrier voltage signal is chosen to have a frequency which is much higher than the frequency of phase inductance variation. Thus, the transient variation of the current phase will contain information about the dynamic motor winding inductance. It can be shown that this phase variation is a one-to-one function of the inductance.

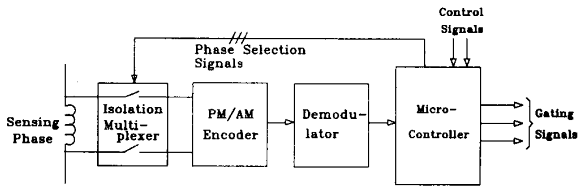

Fig. 4. Block diagram of the modulation encoder based indirect sensing scheme.

The encoded inductance information can then be decoded by using zero crossing detectors for the voltage and current. The input voltage is taken as the reference and the demodulator generates a square wave signal with the help of zero crossing detectors whose pulse width has a maximum variation of $\phi_1 - \phi_2$ as shown in Fig. 3. This pulse width varying square wave signal represents the phase inductance variation and can be calibrated to give the instantaneous rotor position.

B. AM Encoder Technique

Under the same conditions as in the above section, the amplitude of the current is a one-to-one function of the motor winding dynamic inductance. A demodulation method for the AM encoder technique would be to detect the envelope of the modulated current signal which represents the phase inductance variation. This decoded signal can be calibrated to obtain the instantaneous rotor position.

An alternative method of decoding would be to measure the amplitude in terms of angles using a level crossing detector. Referring to Fig. 3, the amplitude variations represented in terms of angles for the aligned and unaligned rotor positions are $\phi_3 - \phi_1$ and $\phi_4 - \phi_2$ respectively.

C. Modified PM Encoder Technique

Exact mathematical derivation of the instantaneous dependence of the phase and amplitude of current on a dynamic inductance is difficult. However, mathematical analysis and simulation will show that phase variation is more sensitive for lower values of inductance while amplitude variation, represented in terms of angles, is more sensitive at higher inductance values. This readily suggests that a combination of the two methods would result in a better sensitivity, i.e., a higher change in the decoded inductance function for the same change in the rotor angle. To achieve this better sensitivity without sacrificing simplicity a level crossing detector can be used instead of a zero crossing detector in the PM encoder circuit to obtain the square wave representation of the phase inductance. The level crossing detector is set to a threshold value at V_T as shown in Fig. 3. The square wave outputs of the level crossing detector corresponding to the input voltage and the maximum and the minimum currents are shown in Fig. 5. The phase angle variation now corresponds to $\phi_4 - \phi_3$ for the same change in phase inductance from the aligned position to the unaligned position in Fig. 3. Fig. 5d shows the ANDed signal of the voltage and the current detector signals. The dotted position represents the variation in the signal that has the encoded information of the phase inductance and hence the shaft position. The implementation of the PM encoder circuit has been carried out

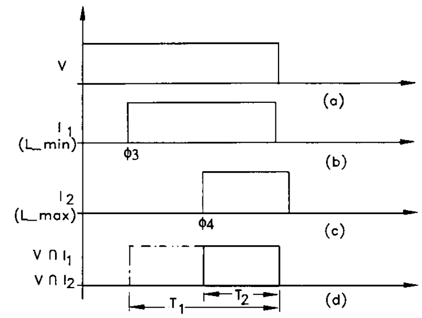

Fig. 5. Sensing circuit voltage and currents.

using this modified technique of measuring the phase at a level other than zero. The design procedure and the experimental results are given based on this modified PM encoder technique.

III. PM Encoder Circuit Optimization

The phase modulated square wave signal is fed to a microcontroller where it is digitized by measuring the pulse width using a timer clocked at a very high frequency. The wider the spread corresponding to the maximum and the minimum phase angles, the more will be the sensitivity of the rotor angle detection. Therefore, to maximize the phase angle difference $\phi_4 - \phi_3$, the series resistance R in the sensing circuit and the threshold level V_T in the level crossing detector must be optimized.

A. Optimization of R

The phasor diagrams of the sensing circuit corresponding to the aligned and unaligned rotor position are given in Fig. 6. Referring to Fig. 6

$$\phi_2 = \phi_1 + \alpha \tag{1}$$

where α is the phase angle variation for the change in phase inductance from the aligned position to the unaligned position. In the aligned position, we have

$$\tan(\phi_2) = \frac{X_{L2}}{R} \tag{2}$$

and in the unaligned position

$$\tan(\phi_1) = \frac{X_{L1}}{R} \tag{3}$$

where X_{L1} and X_{L2} are the inductive reactances in the unaligned and the aligned positions respectively. Putting Eq. (1) in Eq. (2) and using trigonometric identities, we get

$$\frac{\tan(\phi_1) + \tan(\alpha)}{1 - \tan(\alpha)\tan(\phi_1)} = \frac{X_{L2}}{R}. \tag{4}$$

Substituting for $\tan(\phi_1)$ from Eq. (3) and simplifying we get

$$\tan(\alpha) = \frac{R(X_{L2} - X_{L1})}{R^2 + X_{L2}X_{L1}}. \tag{5}$$

The objective here is to maximize the phase angle difference α. The value of resistance R for which α is maximum can be obtained by differentiating α with respect to R. Differentiating Eq. (5) with respect to R, we get

$$\sec^2\alpha \frac{d\alpha}{dR} = \frac{(X_{L2} - X_{L1})(X_{L1}X_{L2} - R^2)}{(R^2 + X_{L1}X_{L2})^2}. \tag{6}$$

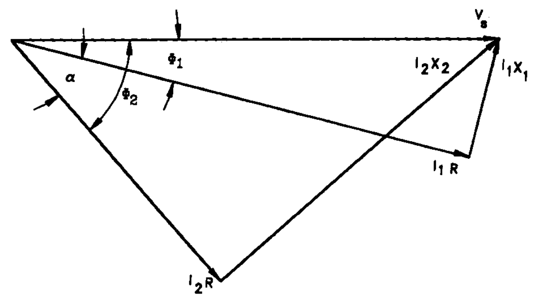

Fig. 6. Vector diagram of the sensing circuit at direct and quadrature positions.

Equating $\frac{d\alpha}{dR}$ to zero, the value of R for which α is maximized can be obtained. The optimized series resistance R is given by

$$R = \sqrt{X_{L1}X_{L2}}. \tag{7}$$

The maximum phase angle difference that can be obtained with this value of R is

$$(\phi_2 - \phi_1)\big|_{max} = \tan^{-1}\left(\frac{X_{L2} - X_{L1}}{2R}\right). \tag{8}$$

B. Threshold Level for PM

In order to derive maximum phase angle variation using a level crossing detector, let us consider the two current waveforms $i_1(t)$ and $i_2(t)$ of Fig. 3 corresponding to the minimum and the maximum phase inductances respectively. Assuming that at these two positions the inductances have fixed values the equations governing the current waveforms are given by

$$i_1(t) = \frac{V}{\sqrt{R^2 + (\omega L_1)^2}}\sin(\omega t - \phi_1) \tag{9}$$

and

$$i_2(t) = \frac{V}{\sqrt{R^2 + (\omega L_2)^2}}\sin(\omega t - \phi_2). \tag{10}$$

From Fig. 3 it is obvious that phase angle difference of the phase modulating signal of Fig. 5d is maximized when the threshold is set at the peak amplitude of $i_2(t)$. The current waveform $i_1(t)$ attains this value at a phase angle $\omega t = \phi_3$ and hence we can write

$$\frac{V}{\sqrt{R^2 + (\omega L_2)^2}} = \frac{V}{\sqrt{R^2 + (\omega L_1)^2}}\sin(\phi_3 - \phi_1). \tag{11}$$

Using $R = \sqrt{X_{L1}X_{L2}}$ from Eq. (7) and solving for ϕ_3 we get

$$\phi_3 = \sin^{-1}\sqrt{\left(\frac{L_1}{L_2}\right)} + \phi_1. \tag{12}$$

Also we have

$$\phi_4 = \phi_2 + 90^\circ. \tag{13}$$

The phase angle variation of the modulated signal using a level crossing detector now becomes

$$\phi_4 - \phi_3 = (\phi_2 - \phi_1) + \left(90^\circ - \sin^{-1}\sqrt{\frac{L_1}{L_2}}\right). \tag{14}$$

Note that the phase angle variation has increased by a factor of $\left(90^\circ - \sin^{-1}\sqrt{\frac{L_1}{L_2}}\right)$ compared to $\phi_2 - \phi_1$ by using a level crossing detector instead of a zero crossing detector. This increase in sensitivity is shown graphically in Fig. 7. The variation of the modulated signals of the PM, AM and modified PM encoder techniques as a function of varying inductance is plotted using equations (3), (7) and (12). The modified PM offers the best sensitivity among the three techniques through the entire range of inductance variation from the unaligned position to the aligned position.

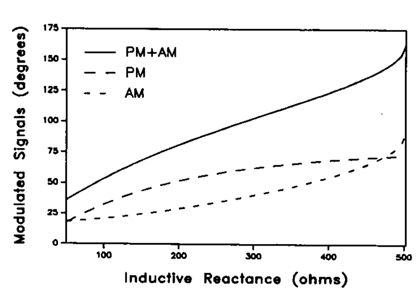

Fig. 7. Sensitivity of modulating signals to inductive reactance.

IV. The Complete Drive System

The classic bridge converter, having two switches and two diodes per phase, has been used to drive the 5 HP three phase SR motor for the experimental setup. This configuration has the advantage over other SRM converters in that it does not require bifilar winding, a split power supply or an even number of phases. The phase independence, the flexibility of control and the regenerative capability are among the other advantages of the classic converter. The block diagram of the complete SRM drive including the different controller segments is shown in Fig. 8. IGBTs are used as the power semiconductor devices and are driven by high voltage integrated circuit gate drivers IR2110. These drivers have a floating supply to drive the upper switches of each phase leg without using any opto-coupler or pulse transformer. A 16-bit microcontroller (Intel 80C196KR) having a clock rate of 16 MHz has been used to implement the control algorithm. The microcontroller is configured to take the decoded phase modulated signal and generate the gate switching signals for the three phases. The microcontroller also generates a fixed frequency PWM signal that regulates the motor phase currents. The torque is controlled in closed loop by varying the duty cycle of the PWM signal for the power switches S1–S6. The remainder of the drive system consists of the PM encoder circuit which is interfaced with the phase coils through a multiplexer, MUX and feeds the microcontroller at the other end. The sensing circuit must be isolated from the power circuit when the phase is being energized by the main converter for the magnetization process. This is accomplished by using the multiplexer, MUX consisting of photovoltaic bi-directional MOSFET switches, as an interface between the SRM phase winding and the sensing circuitry.

The PM encoder circuit, i.e., the sensing circuit basically comprises of a 5 kHz sine wave generator, a low pass filter and a level crossing detector. The sinusoidal input voltage is generated by the Intersil waveform generator ICL8038 and its current drive capability has been improved by a power amplifier stage. The output of the power amplifier gives a stiff sinusoidal signal that can provide a current of up to .5A at 10 V peak amplitude. To detect the rotor position, the previously conducting phase is connected to the sensing circuit through the multiplexer (MUX). For example if phase B is conducting the motor energization current, phase A is connected to the position sensing circuit. The MUX receives its signals from the microcontroller through the sense signals. The phase modulated signal becomes somewhat corrupted by the high frequency switching in an adjacent phase that causes mutually induced voltages. Therefore, the signal containing the information is passed through a low pass filter to eliminate the high frequency switching noise. A level crossing detector converts this encoded signal to a square wave while a zero crossing detector generates a square wave output of the input sinusoidal voltage to be used as a reference. The two waves are ANDed to obtain the decoded PM signal. This PM signal is fed to one of the timer inputs of the microcontroller. The timer uses a 4 MHz clock and gives a digital representation of phase coil inductance and hence the rotor position. The control algorithm in the microcontroller routinely measures the rotor angular position using this decoded signal and generates the optimum phase excitation pulses.

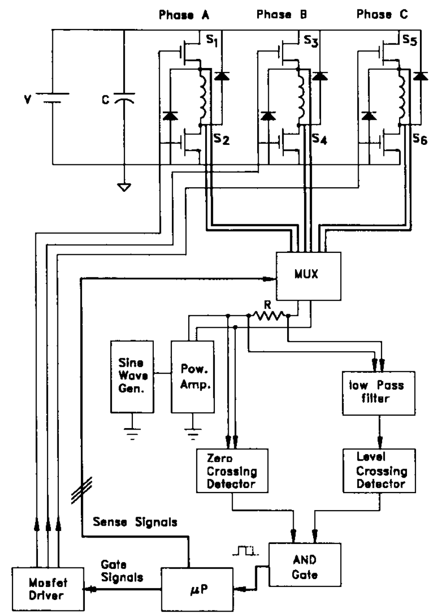

Fig. 8. Drive circuit and control block diagram using PM encoder technique.

V. Resolution and Error Analysis

The accuracy of the rotor position information determines the efficiency and smoothness of the drive. The resolution and fundamental accuracy achievable by the PM encoder technique will be discussed in this section.

A. Commutation Error Determination

For the 3 phase SRM with 6 stator poles and 4 rotor poles, the phase inductance has a period of 90 mechanical degrees. This means that

$$\theta_{L\max} - \theta_{L\min} = 45^\circ\text{Mech.} \tag{15}$$

where $\theta_{L\max}$ and $\theta_{L\min}$ are the rotor angles corresponding to the maximum and minimum phase inductances respectively. All the angles defined in this section are in mechanical degrees. Commutation threshold detection comes from comparing samples of rotor position with a reference angle. Therefore, commutation error comes from error in rotor samples. These errors are caused by sampling rate and quantization error in each sample. The worst case commutation angle error is given by [11]

$$\Delta\theta_{\text{comm}} = \Delta\theta_{\text{sr}} + \Delta\theta_{\text{sq}} \tag{16}$$

where

$$\Delta\theta_{\text{comm}} = \text{Commutation angle error,}$$
$$\Delta\theta_{\text{sr}} = \text{Error due to sampling rate,}$$
$$\Delta\theta_{\text{sq}} = \text{Quantization error in each sample.}$$

The variation of inductance from maximum to minimum is a non linear function of the rotor position with varying sensitivity. Therefore, only the average sample error in the digitizing process can be calculated. Let the digital equivalent values of the PM signal corresponding to $L_{\max}$ and $L_{\min}$ be d_1 and d_2 respectively. $L_{\max}$ is the phase inductance corresponding to the aligned position of stator and rotor and $L_{\min}$ is the phase inductance corresponding to the unaligned position.

Assuming linear variation of inductance the average sample quantization error can be calculated as

$$\Delta\theta_{\text{sq}} = \frac{180}{P_r(d_2 - d_1)}$$

where P_r = number of rotor poles = 4.

Also the frequency of sampling the phase inductance is another measure of rotor position resolution. The sampling rate error depends on the motor speed N in r/min and the frequency of sinusoidal carrier frequency injected from the sensing circuit. This error is [11]

$$\Delta\theta_{\text{sr}} = \frac{6N}{f_s}.$$

In the PM encoder circuit used in the experimental set-up, the digital counts d_1 and d_2, corresponding to $L_{\max}$ and $L_{\min}$ respectively, were 198 Hex and 310 Hex. This lead to a quantization error of

$$\Delta\theta_{\text{sq}} = 0.45^\circ\ \text{Mech.}$$

in each sample. The experimental SRM drive was designed for adjustable speeds ranging up to 2500 r/min. Let us calculate the sampling error for one operating point at 1000 r/min. For a sinusoidal carrier frequency of 5 kHz in the PM encoder circuit the error due to the sampling rate is

$$\Delta\theta_{\text{sr}} = 1.2^\circ\ \text{Mech.}$$

Therefore, the commutation angle error at 1000 r/min from Eq. (16) is

$$\Delta\theta_{\text{comm}} = .45^\circ + 1.2^\circ = 1.65^\circ\text{Mech.}$$

The motor was run at various speeds within the range of 250 r/min and 2500 r/min and the PM encoder technique was found to keep excellent track of rotor position for commutation control for the entire range. Note that better rotor position accuracy can be obtained by changing the design parameters of the PM encoder. The above numerical case was given for illustration.

B. Errors Due to Filter Delay

A low pass filter is used to eliminate the noise in the PM current signal due to the noise generated by the PWM switching in the active phase. The PWM frequency is 22 kHz and the PM signal frequency is 5 kHz and hence the noise is easily eliminated by a simple low pass RC filter.

The filter will introduce a dynamic error in the current phase and amplitude modulation. The amount of this error, in degrees, will vary as the current changes from ϕ_3 to ϕ_4 states in Fig. 3. The functional dependence of this phase error on time and instantaneous stator phase inductance can be obtained theoretically or by numerical simulation. This function depends on the current frequency and filter component values. However, a worst case error, $\Delta\phi$, can be found. The rotor angle error due to this worst case phase error can be calculated as follows. The microcomputer count error due to the worst case phase reading error is

$$\Delta d = \frac{d_2 - d_1}{\phi_4 - \phi_3}(\Delta\phi)$$

where d_1 and d_2 are defined as before and ϕ_3 and ϕ_4 are defined as shown in Fig. 3. The mathematical rotor angle error due this count error is then given by

$$\Delta\theta_\phi = \frac{180}{P_r(d_2 - d_1}).\frac{(d_2 - d_1)}{\phi_4 - \phi_3}(\Delta\phi)$$
$$= \frac{180}{P_r(\phi_4 - \phi_3)}(\Delta\phi)$$

where $\Delta\theta_\phi$ is the rotor angle error due to the filter.

VI. EXPERIMENTAL RESULTS

The performance of the indirect position sensing scheme using the PM encoder was evaluated on the previously mentioned 5 HP, three phase SRM with 6 stator poles and 4 rotor poles. The DC bus voltage was fixed at 40 V and the PWM switching frequency of the converter was set at 22 kHz. The results of the experiments are given in this section.

Test results presented here are for an SRM running at 2500 r/min under lightly loaded conditions. The waveforms of the PM encoder circuit appear in Fig. 9. The oscillograph shows the input sinusoidal voltage in the sensing circuit and the unfiltered and filtered lagging current drawn by the coil inductance. The notches in the current waveform are due to the high frequency PWM switching in the adjacent active phase. The current waveform in phase B is shown in the oscillograph of Fig. 10. Note that this is an unregulated current with constant PWM duty cycle for this experiment. Each cycle of the motor phase current waveform consists of three modes of operation. The phase coil carries the main energization current for one-third of the period while in the previous one-third period it carries the diagnostic sensing current. The phase is completely non-conducting for the remaining one-third portion of the cycle, as seen in Fig. 10. The maximum sensing current

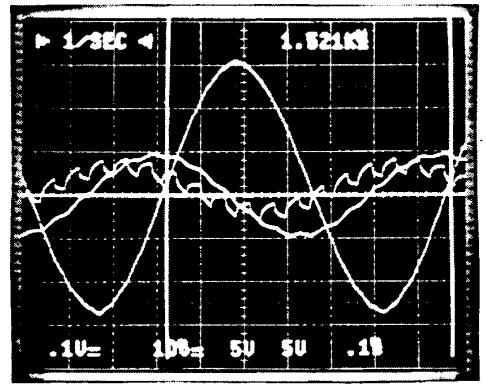

Fig. 9. The input sinusoidal voltage of the PM encoder circuit and the unfiltered and filtered currents drawn by the phase coil.

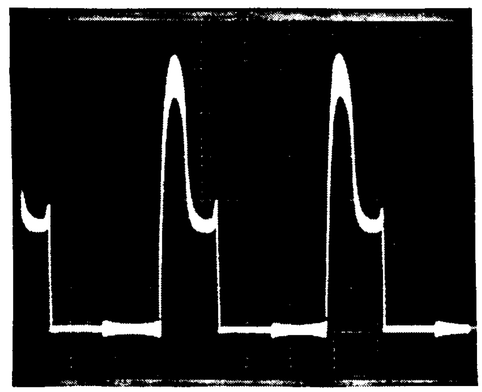

Fig. 10. The motor current of phase B. Scale: .5 A/div., 10 ms/div.

is about 100 mA and does not have any significant effect in the torque production.

VII. CONCLUSION

Two new indirect SRM rotor position sensing schemes based on winding current amplitude and phase modulation are presented in this paper. The techniques use the basic properties of amplitude modulation and phase modulation, commonly used in communication systems, in a simple circuit configuration. An off phase is supplied with a stiff sinusoidal carrier voltage from the sensing circuit which is modulated by a rotor position dependent phase inductance function. The modulated signal is decoded to obtain the signal containing the information and is calibrated to obtain the rotor position angle continuously.

A combination of phase modulation and amplitude modulation encoder technique was designed and tested on a 5 HP three phase 6-4 SRM over a wide range of speed. The position detection scheme kept excellent track of the rotor angle continuously both under load and no-load conditions. This scheme is extremely robust to switching noise that is present in the sensing phase due to mutually induced voltages. A study of the resolution and position accuracy is also presented in this paper. This study will help in designing the modulation circuitry to meet the desired rotor position accuracy.

REFERENCES

[1] T. J. E. Miller, *Brushless Permanent Magnet and Reluctance Motor Drives*, Oxford, U.K.: Oxford University Press, 1989.
[2] T. J. E. Miller, P. G. Bower, R. C. Becerra and M. Ehsani, "Four-quadrant brushless reluctance motor drive," *IEE Conf. Power Electron.*, CP-291, pp. 273–276, July 1988.
[3] B. K. Bose, T. J. E. Miller, P. M. Szczesny and W. H. Bicknell, "Microcomputer control of switched reluctance motor," *IEEE Trans. Industry Applications*, vol. IA-22, no. 4, pp. 708–715, July/Aug. 1986.
[4] P. P. Acarnley, R. J. Hill and C. W. Hooper, "Detection of rotor position in stepping and switched reluctance motors by monitoring of current waveforms," *IEEE Trans. Ind. Elec.*, vol. IE-32, no. 3, pp. 215–222, Aug. 1985.
[5] J. T. Bass, M. Ehsani and T. J. E. Miller, "Robust torque control of switched reluctance motors without a shaft position sensor," *IEEE Trans. Ind. Elec.*, vol. IE-33, no. 3, pp. 212–216, Aug. 1986.
[6] S. R. MacMinn, P. M. Szczesny, W. J. Rzesos and T. M. Jahns, "Application of sensor integration techniques to switched reluctance motor drives," *IEEE Trans. Industry Applications*, vol. IA-28, no. 6, pp. 1339–1344, Nov./Dec. 1992.
[7] W. D. Harris and J. H. Lang, "A simple motion estimator for variable reluctance motors," *IEEE Trans. Industry Applications*, vol. IA-26, no. 2, pp. 237–243, Mar./Apr. 1990.
[8] M. Ehsani, I. Husain and A. B. Kulkarni, "Elimination of discrete position sensor and current sensor in switched reluctance motor drives," *IEEE Trans. Industry Applications*, vol. IA-28, no. 1, pp. 128–135, Jan./Feb. 1992.
[9] A. Lumsdaine and J. H. Lang, "State observers for variable reluctance motors," *IEEE Trans. Ind. Elect.*, vol. IE-37, no. 2, pp. 133–142, April 1990.
[10] M. Ehsani, "Method of indirect rotor angle detection in variable reluctance motor drives by phase and amplitude signal modulation," U.S. patent pending.
[11] M. Ehsani, I. Husain and K. R. Ramani, "An analysis of the error in indirect rotor position sensing of Switched Reluctance Motors," *IEEE-IECON Conf. Rec.*, pp. 295–300, Oct. 1991.

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