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New Series Resonant Converter for Variable Reluctance Motor Drive

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Abstract

In this paper, a new resonant type converter for a variable reluctance motor drive is proposed to perform zero current switching of all solid state devices. The resonant circuit is partially resonant rather than fully resonant from the view point of the capacitor. The power circuit of the converter can be realized with unidirectional naturally commutated switches. The converter has many advantages such as low voltage stress or improved voltage utilization, low loss and no stress in the switching instant. The operating principle and current regulation method are introduced and verified through simulation.

Introduction

In recent years, the variable reluctance motor (VRM) drive has become an attractive candidate for replacing conventional ac and dc drives in many industrial applications because the VRM has many advantages such as simple and low cost construction, absence of shoot-through faults, simplified power converter, high torque/ inertia ratio, etc [1-3]. The disadvantages of the VRM are concerned with higher torque ripple and acoustic noise level than other motors. However, quiet VRM drives with low torque ripple appear to have been developed by several researchers [4]. In addition, singly salient, synchronous reluctance motors have begun to emerge which have some of the same advantages as VRMs (simplified power converter for example) but without the noise and torque pulsation difficulties [5,6].

The cost and performance of the VRM drive is highly dependent on the converter topology used to drive the VRM. The converter for the VRM drive must meet the following requirements; very fast accurate current control for good drive performance, low VA rating for low cost, reliability and robustness, a small number of switches, and high efficiency.

Numerous converter topologies have been developed and used in a variety of applications. These configurations include the asymmetric bridge converter, the bifilar winding configuration, the split supply configuration, the H-bridge configuration, the common switch configuration and the C-dump converter and the modified C-dump converter [1,7,8]. However, all converters have their drawbacks such as numerous switches, high voltage rating (or poor utilization of the voltage), requirement of auxiliary winding or special configuration of VRM. In addition, all of these converters have greater switching loss and stress because the converters are all operated with hard switching.

In order to achieve widespread acceptance of VRM drives, it is clearly necessary to both improve the drive performance and achieve lower cost. These goals can be achieved either by improvements in machine design or innovation in the converter configuration. The ultimate answer in converter design may be a resonant type converter topology. This paper deals with the application of a particular type of resonant converter for the VRM drive.

Description of the Proposed Converter

The basic power circuit of the proposed converter is shown in Fig. 1. Note that the circuit can be realized with simple unidirectional and naturally commutated switches In this configuration, T_1 T_2 and T_3 are used as main switches for motor windings. The switches S_1, S_2, D_1 and D_2 are used for application of either positive or negative voltages to the main switches or to provide freewheeling paths. The inductor L_{dr} and capacitor C_r make up the resonant circuit. However, this circuit is partially resonant rather than fully resonant from the view point of the capacitor C_r. The capacitor voltage swings near the dc supply voltage while the inductor current resonates from zero to full current during every resonant cycle. The purpose of the series resonance is used to enalble zero current switching of the main thyristors.

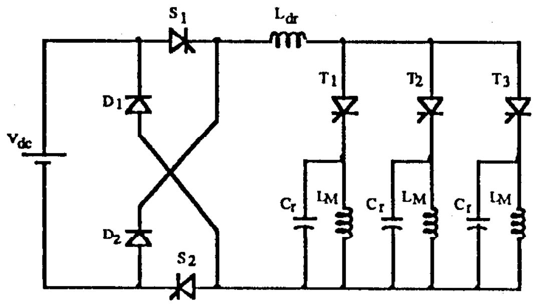

Fig. 1 Zero current switching resonant converter for SRM Drive.

Operating Principle

Before describing three phase operation, single phase operation is illustrated in Fig. 2 to clearly show the principle of the proposed concept. The operation of the proposed converter can be easily explained by analysis of the four modes of operation.

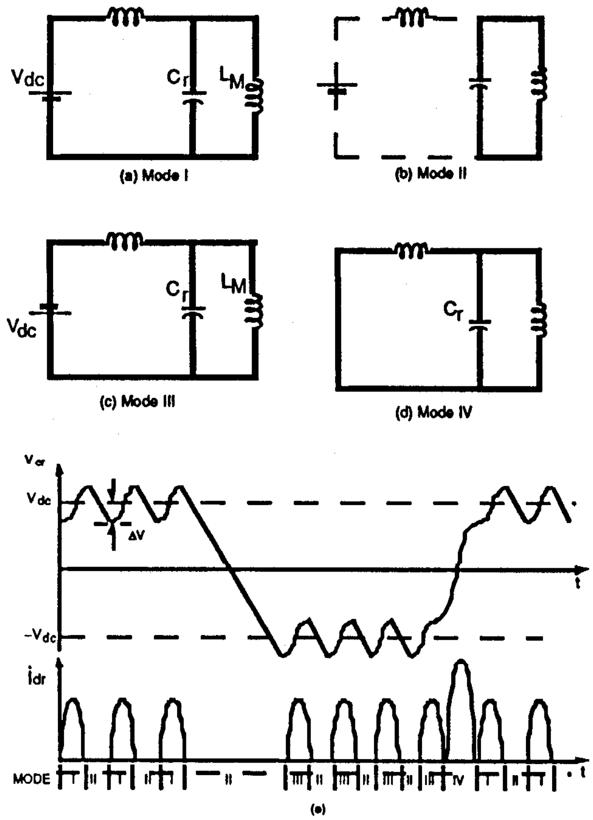

Fig. 2 Simplified circuit diagram of the proposed converter. (a) Mode I, (b) Mode II, (c) Mode III, (d) Mode IV and (e) resonant inductor current and capacitor voltage during each mode operating interval.

1) Powering mode (mode I)

It is assumed that the capacitor has been charged with initial voltage V_{co}^+ which must be less than V_{dc}. The powering mode starts with the thyristors S_1, S_2 and T_1 turned on. The resonant link current i_{dr}, flowing through L_{dr} and T_1 due to the difference between V_{dc} and V_{co}^+, begins to charge the capacitor C_r. In this mode, the simplified equivalent circuit is shown in Fig. 2 (a) and the resonant current i_{dr} and the capacitor voltage C_r are shown in Fig. 2 (c) in the steady state. The resonant inductor current and the capacitor voltage during this mode are given by

$$i_d = \frac{\Delta V}{Z_d} sin\ \omega_d t + i_s(1\text{-}cos\ \omega_d t)$$

$$v_{cr} = V_{dc} - \Delta V cos\ \omega_d t - Z_d i_s sin\ \omega_d t$$

where i_s is the load current and $\Delta V = V_{dc} - V_{dc}^+$, $\omega_d = 1/\sqrt{L_{dr}C_r}$ and $Z_d = 1/\sqrt{L_{dr}C_r}$.

After a finite period of time t_1, the powering mode is terminated when the resonant current i_{dr} reaches zero. The interval t_1 is determined both by the resonant circuit parameters and by the load current. The interval can be roughly estimated by

$$t_1 \cong \frac{\pi}{\omega_d} .$$

2) Discharging Mode (mode II)

At the end of mode I, the switches are turned off due to the condition of zero current in the switches, and the current loop remains as shown in Fig. 2 (b). In this mode, the capacitor is almost linearly discharged because the winding inductance of VRM is much larger than that of the resonant inductance L_{dr} and thus the load current is maintained almost constant during this short interval. This mode continues until the capacitor voltage v_c reaches around the initial voltage V_{co}^+ of mode I. The length of t_2 depends almost entirely on the load condition and is approximately expressed by

$$t_2 \cong \frac{2\Delta V C_r}{i_s} \qquad (5)$$

3) Recovery Mode (mode III)

It is again assumed that the resonant capacitor is initially charged with V_{co}^- where V_{co}^- must become less than $-V_{dc}$. If the thyristor T_1 is turned on while S_1 and S_2 remained in the turn-off state, The resonant current flows through diodes D_1 and D_2. In this mode, the simplified equivalent circuit according to current path can be drawn as shown in Fig, 2 (c). The inductor current and the capacitor voltage are analytically expressed by

$$i_d = \frac{\Delta V}{Z_d} sin\ \omega_d t + i_s(1\text{-}cos\ \omega_d t)$$

$$v_{cr} = \text{-}V_{dc} - \Delta V cos\ \omega_d t - Z_d i_s sin\ \omega_d t$$

where $\Delta V = V_{dc} - V_{co}^-$. At the end of this mode, the circuit operates again in the discharging mode (mode II) similar to the end mode I.

4) Reversing Mode (mode IV)

When the recovery mode is directly switched to the powering mode, the resonant inductor current and the capacitor voltage can be greatly increased. However, if the capacitor voltage is reversed to its opposite polarity whenever both modes are switched, the peak in the capacitor voltage and the resonant inductor current can be greatly reduced. This action can be obtained by turning on S_1 and T_1 while S_2 is maintained in the turn-off state continuously. In this mode, the current and the voltage wave forms are expressed by;

$$i_{dr} = \frac{V_{co}}{Z_d} sin\ \omega_d t + i_s(1\text{-}cos\ \omega_d t)$$

$$v_c = -V_{co}\cos \omega_d t - Z_d i_s \sin \omega_d t$$

Current Regulation Method

The current regulation method using the resonant converter is similar to the conventional CRPWM (current regulation pulse width modulation) method. The proposed current regulation method can be illustrated as shown in Fig. 3. Figure 4 shows idealized waveforms for the the current command i_s^*, the actual current i_s and the command voltage v_s^*. In the proposed system, a voltage regulation process is required for control of the converter because the converter is operated under a soft switching condition instead of hard switching. The process in which the output voltage is regulated according to the command value, and the process in which the output current is regulated can be easily understood by examining Fig. 4.

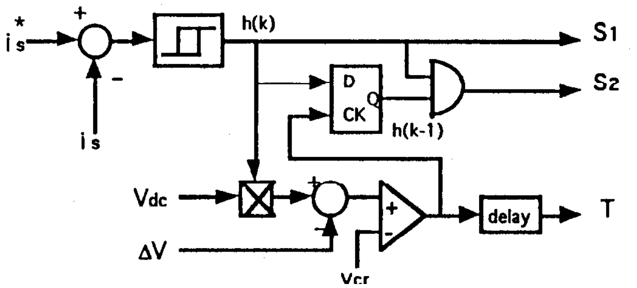

Fig. 3 Implementation of current controller of series resonant converter.

In a typical VRM, a positive current command is applied to each phase and corresponds to the positive sloped portion of the corresponding inductance profile in motoring operation as shown in Fig. 5 (a) and (b), respectively. When the stator current is commutated from A to B phase, a long commutation time is required because the inductance of A phase is very large at this time. For this reason, the actual currents are delayed and overlapped during the commutation interval as shown in Fig. 5 (c).

During the overlapping interval of the A and B phase current, the converter is operated in the powering mode for phase B while it is operated in the discharging mode for phase A as shown in Figs. 5 (d) and (e), respectively. In next switching interval, the converter is operated in the recovery mode for the A phase and in the discharging mode for the B phase, simultaneously. By means of the above control method, the stator currents are well regulated even while the currents are in the overlapping interval.

Converter Utilizing 3-Phase Input

In some applications the VRM must be often operated in regenerating mode for long period. In this case, the converter must have a capability for dynamic braking or regenerative operation. This requirement is often indispensable for safe operation in a high power drive system. In such application areas, the input side converter must change must be modified from that of Fig. 1. Figure 6 is a modified configuration of the proposed converter. The input side converter can now return the regenerating energy to the 3ϕ ac source so that the VRM can be operated continuously in regenerating action (or operate permanently as a generator). In this case the positive (or negative) maximum rectified voltage can be obtained by way of controlling the firing angle of the input side converter which is controlled with $\alpha = 0°$ (or $\alpha = 180°$). For the reversing mode, one leg of the input side converter conducts, then the resonant current can freewheel through the resonant inductor to reverse the capacitor voltage. Of course, the input side converter can be also switched during the condition of the zero current switching in much the same manner as the

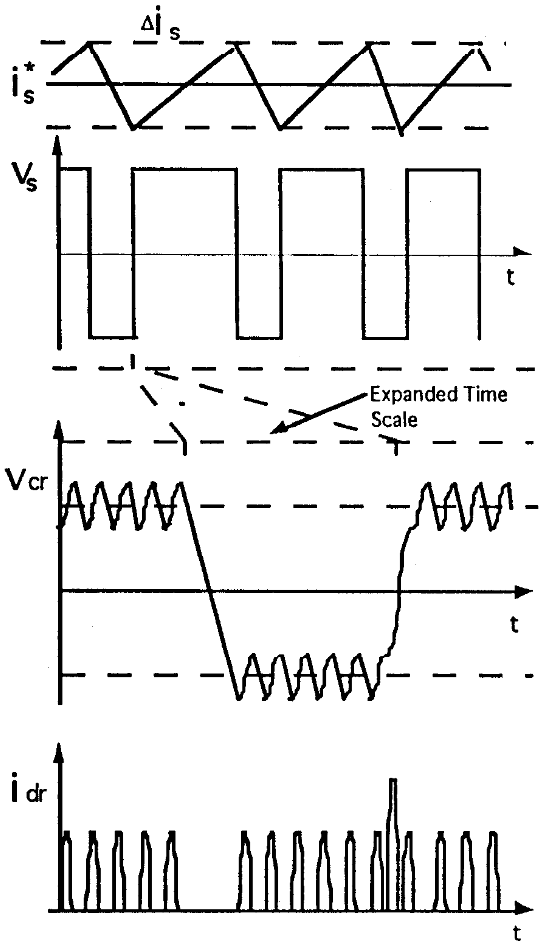

Fig. 4 Stator Current Control Waveforms.

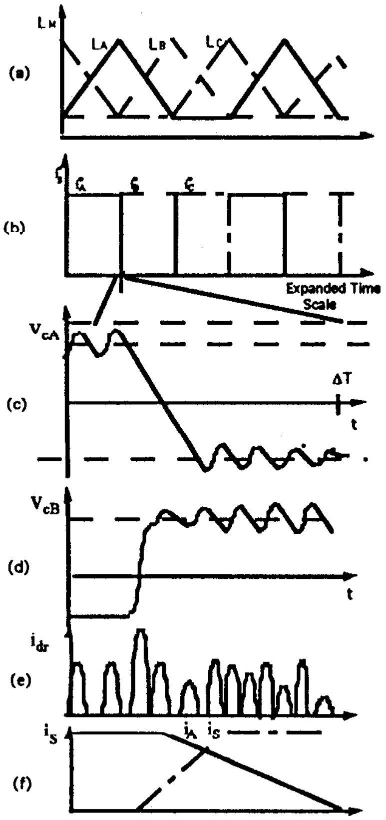

Fig. 5 Illustrating current regulation during the overlap interval.

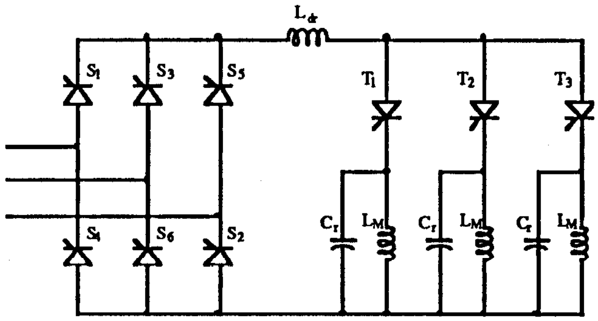

Fig. 6 Series resonant converter for VR motor having three phase input.

switches in series with the winding of the VRM so that force commutated input side converter switches are not required.

Simulation Results

In order to demonstrate the feasibility of the proposed circuit, a simulation was performed for a three phase VRM in which the parameter values are $L_{max} = 35$ mH, $L_{min} = 5$ mH, $R_s = 0.5\ \Omega$, $C_r = 1.0\ \mu F$ and $L_{dr} = 50\ \mu H$. Figure 7 shows the simulation results when the VRM is operated in the motoring region at 50 Hz. In this case, the error boundary of the current hysteresis control was fixed to ± 1.0 A . Figure 7 (a) demonstrates that the stator current is well regulated within the given error bounded except when the winding current is commutated from one phase to another. The capacitor voltage v_c has a swing between $V_{dc} - \Delta V$ and $-V_{dc} - \Delta V$ and the inductor current pulsates as shown in Figs. 7 (c) and (d), respectively.

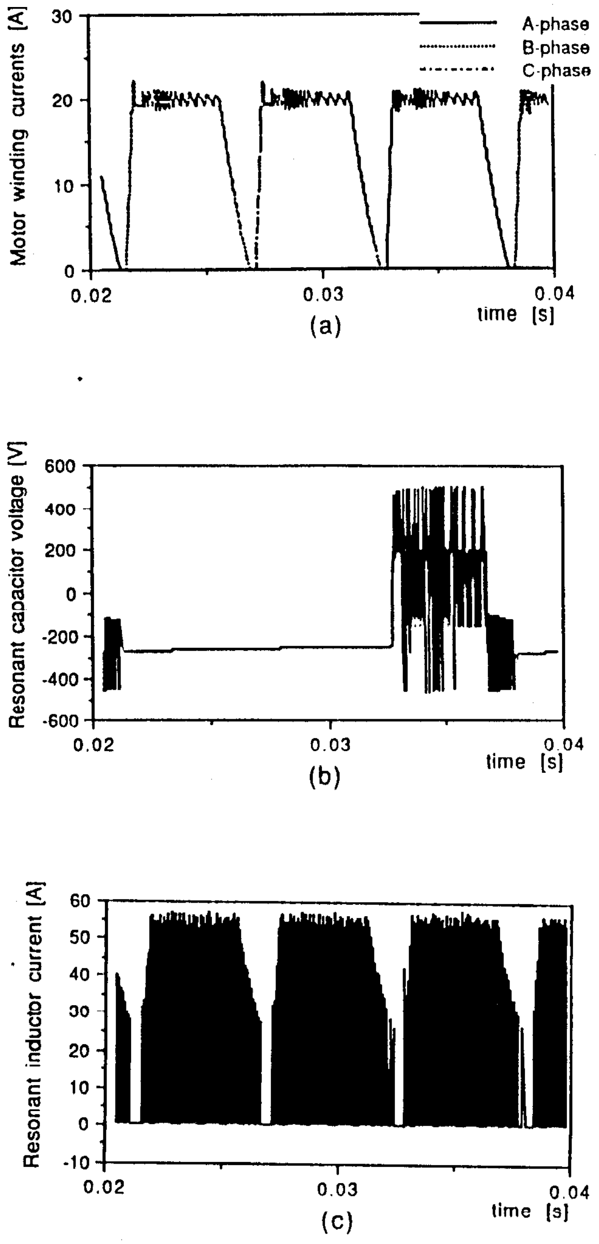

Fig. 7 Simulaltion results when the VRM is operated in the motoring region at 50 Hz.

Figure 8 shows the simulation results when the VRM operates on the motoring region at 100 Hz. Figure 9 shows the simulation results when the VRM is operated in the regenerating region at 100 Hz. Note that ample time for commutation is still available for eithe case. The simulation results clearly show that the current regulation of the proposed converter can be achieved while maintaining low voltage stress a zero switching losses in the semiconductor devices.

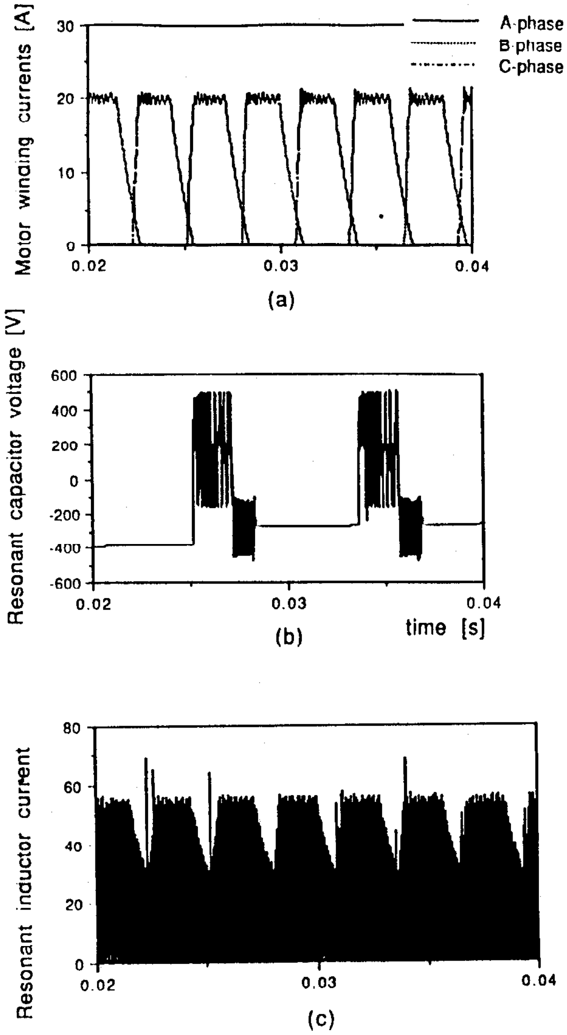

Fig. 8 System Behavior when the SRM is operating in the motor region at 100 Hz.

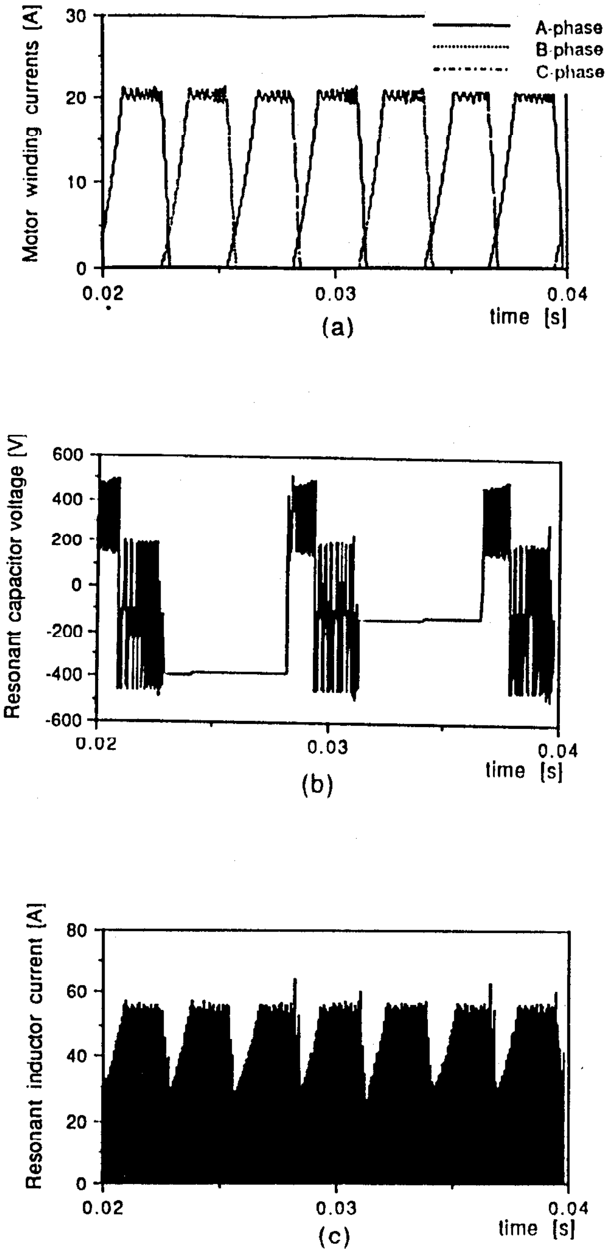

Fig. 9 Simulation traces for the case where the SRM is operated in the regenerating region at 100 Hz.

Conclusion

This paper has present new converter circuit topology for variable reluctance motor drive. The circuit is basically operated by series resonant action such that zero current switching is possible. Realization of such desirable characteristics is obtained with only unidirectional and naturally commutated switches. The operating principle has been illustrated through an analysis of each operating mode and a new current regulation method, especially concerned with the overlapping interval of two phase current has been explained. A converter having a 3-phase input instead of a dc-source was proposed for high power applications. The operating principle and advantages have been verified through simulation. The converter proposed has many advantages such as a simple configuration, low voltage stress, improved utilization of the supply voltage, and, in particular, no stresses or losses in the switching interval. Work on a hardware implementation of this topology is presently in progress.

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