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A Novel Three Phase Series-Parallel Resonant Converter Fed DC-Drive System

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ABSTRACT

This paper presents the application of a single phase AC-to-DC converter using a three-phase series parallel (SPRC) resonant converter to variable speed dc-drive. The improved power quality converter gives the input power factor unity over a wide speed range, reduces the total harmonic distortion (THD) of ac input supply current, and makes very low ripples in the armature current and voltage waveform. This soft-switching converter not only possesses the advantages of achieving high switching frequencies with practically zero switching losses but also provides full ranges of voltage conversion and load variation. The proposed drive system is the most appropriate solution to preserve the present separately excited dc motors in industry compared with the use of variable frequency ac drive technology. The simulation and experimental results are presented for variable load torque conditions. The variable frequency control scheme is implemented using a DSP- TMS320LF2402. This control reduces the switching losses and current ripples, eliminates the EMI and improves the efficiency of the drive system. Experimental results confirm the consistency of the proposed approach.

Keywords: AC-to-DC converter, DC motor, Digital signal processor (DSP), PID, Power quality, Resonant converter, Zero voltage switching (ZVS)

1. Introduction

Interest in power quality has increased significantly over the last few years. In order to comply with regulations on the quality of input power, recent research has focused on the applications of resonant techniques to the converters that are proposing and developing inherently clean new power converter topologies. These converters operate at unity power factor and inject very low harmonic content into supply, and produce relatively high converter efficiencies.

The performance of converters in terms of efficiency and power density is very important; the power density can be improved with increased switching frequency [1-3].

Several single phase as well as three phase converter topologies such as series resonant converter (SRC), parallel resonant converter (PRC), series-parallel resonant converter (SPRC), higher order resonant converter (HORC), modified series-parallel resonant converter (MSPRC) topologies have been proposed [1-7].

It has been reported that the resonant converter topologies offer many advantages [5-11] such as lower

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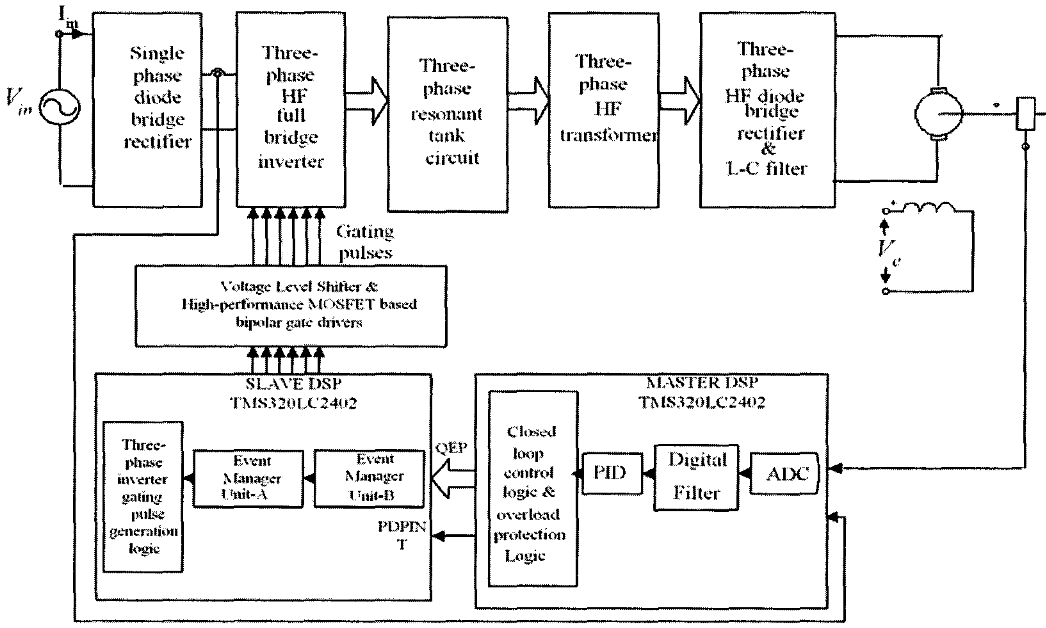


Fig. 1 Block schematic of proposed AC-DC resonant converter fed dc-motor drive

inverter output current, lower voltage stresses, and reduction in size of magnetic components, narrow variation in switching frequency for the output voltage regulation and better power conversion efficiency due to reduced conduction losses.

All these topologies deal only with passive loads. Unfortunately there is limited literature available that reports the performance of converters with active loads like dc-motors. Application of this resonant converter in dc motor drives has not been yet reported. The dc motor plays a significant role in modern industrial drives. The bulk of the separately excited dc-motors are used with power converters for constant torque and constant horse power applications in the manufacturing industry, due to the heavier upfront cost of ac drive control than that of dc drive. At present there is a need to enhance performance of existing dc drives in many industrial applications including machine tool drive, paper mills, waste water treatment and steel plants, [12-14]. The conventional dc-drive continues to take a considerable share of the variable speed drive market. This share is expected to decline only slowly with ac drives. Efforts are being made to reduce the cost and enhance the reliability of dc drive systems. Due to advent of modern power semiconductor devices, the power circuit of the general dc drive has changed drastically.

The main aim of this paper is to present a soft-switching converter (SPRC) for applications to dc motor drives, and therefore to investigate behavior of three-phase resonant converter driving dc-motor load. In this paper an attempt is made to use a high efficiency and low loss improved power quality converter to enhance the performance of dc motor drive. This new topology automatically draws sinusoidal input current from a single-phase supply with reduced switching losses [11]. The proposed converter in dc-drive systems has less component complexity as compared to the topology suggested in [15], operates in lagging power factor (P.F) mode for the entire load ensuring zero voltage switching (ZVS) operation [2-3] for all inverter switches, and gives superior motor control compared with the conventional one.

2. Principle of Operation

Fig. 1 shows the block schematic of the proposed AC-DC converter using a three- phase series-parallel resonant converter (SPRC) feeding dc-motor. It consists of a three-phase HF inverter fed through a single-phase bridge rectifier connected to a small capacitive filter. The small capacitor (C_{in}) acts as a HF bypass resulting inherently in high power factor without any active control of the ac line

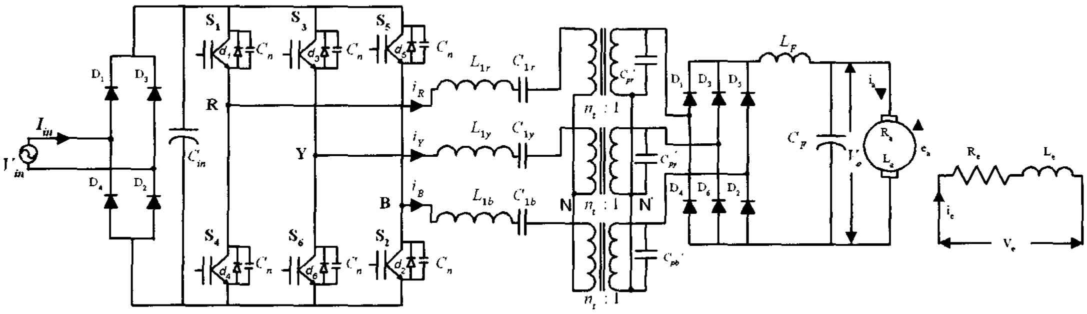


Fig. 2 Power circuit of proposed three-phase resonant converter fed dc-motor drive

current. The output of the HF inverter is fed to a three-phase *LCC* type resonant tank circuit consisting of the elements L_1, C_1 and C_p. These components in three phases have been denoted by subscripts r, y, and b. A three-phase HF transformer (Y-Y) is used for isolation as well as for voltage transformation. The capacitor C_p is placed on the secondary side of HF transformer. This helps to utilize the leakage inductance of the HF transformer as a useful part of the resonant tank component. Filter components L_F and C_F are designed for the specified current and voltage ripple contents in the output respectively. In a dc motor with separately excited field winding, speed of the motor governed by the following expression:

$\omega_m = \frac{1}{k_e\Phi}\left(V_a - \frac{R_a}{k_t\Phi_e}T_{em}\right)$ shows that armature voltage (V_a) and field flux (Φ_e) can be controlled to yield desired torque and speed. Here armature control is preferred because of high efficiency, good transient response and good speed regulation. Also armature control has the advantage of controlling the armature current swiftly.

The power circuit of the proposed drive system is shown in Fig.2 under variable frequency mode operation of the three phase HF full bridge resonant inverter, practically gating pulse widths are less than 180^0 due to dead time requirement of the devices for avoiding shoot through fault. Fig.3 shows the typical idealized operating waveforms of the HF inverter stage. This gating pattern maintains a lagging power factor mode over an entire load range confirming zero voltage switching (ZVS) operation for all inverter switches. Owing to ZVS operation the turn on losses are practically eliminated.

3. Load Modeling

The output of the three-phase series parallel resonant converter (SPRC) is fed to the armature circuit of the dc motor. This type of load is typically a second order system whose mathematical model is given in this section. Fig.4 shows the schematic diagram of a separately excited dc motor. The armature of the dc motor is modeled as a series combination of resistance, inductance and back emf. The field current determines the excitation flux, Φ_e in the machine. The armature current determines the armature flux, Φ_a in the machine.

The defining equations of the dc motor are

Field circuit equation; $V_e = R_e \cdot i_e + L_e \frac{di_e}{dt}$ (1)

Armature circuit equation;

$$V_a = V_o = R_a \cdot i_a + L_a \frac{di_a}{dt} + e_a \tag{2}$$

Mechanical system equation;

$$T_d = J\frac{d\omega}{dt} + B\omega + T_L \tag{3}$$

The mechanical subsystem is modeled using torque current analogy

Back emf equation $e_a = k_1 i_e \omega$ (4)

Torque equation $T_d = k_2 i_e \cdot i_a$ (5)

Where, $V_a = V_o$ = converter output voltage, volts

R_a = armature circuit résistance, ohms

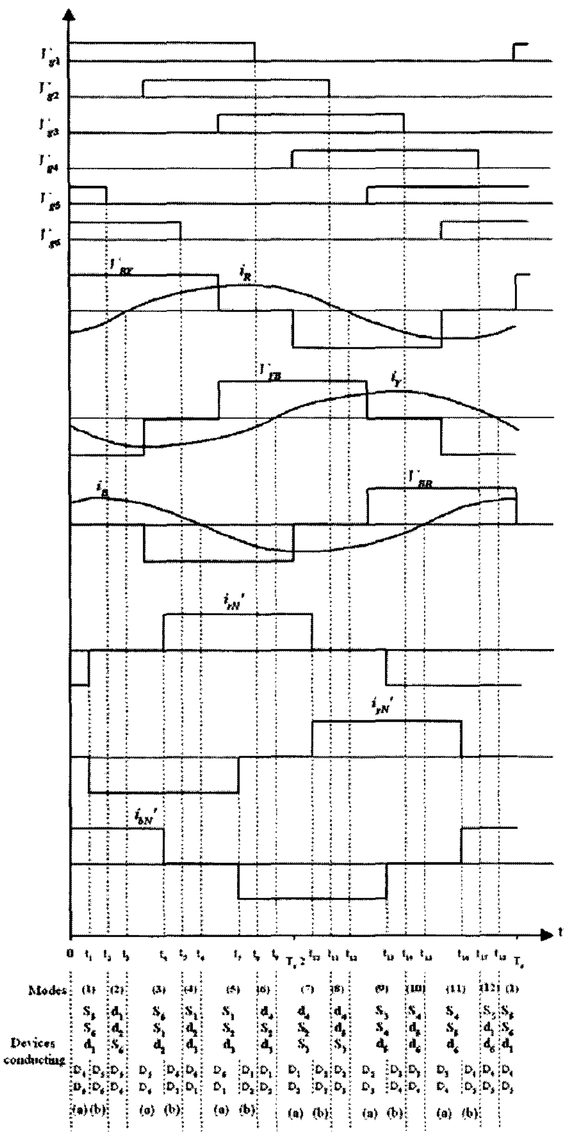


Fig. 3 Typical operating waveforms of the practical three-phase series-parallel resonant converter operating with gating pulse widths lesser than180-degree (Typical value of the gating pulse width is 150^0)

L_a = armature circuit inductance, henries

i_a = armature current, amp

e_a = back emf, volts

R_e = field circuit resistance, ohms

L_e = field circuit inductance, henries

J = moment of inertia, kg.m^2

B = viscous friction constant, Nm/rad/sec

T_L = load torque, N.m

ω = angular speed of the motor, rad/sec

T_d = developed torque, N.m

Comparing the back emf and the torque equation, we get $k_1 = k_2$, because the electrical input ($e_a\, i_a$) will be equal to the mechanical output ($T_d\, .\omega$)., the system may be modeled in the following form

$$\tau_e \frac{di_e}{dt} = \frac{V_e}{R_e} - i_e \tag{6}$$

$$\tau_a \frac{di_a}{dt} = \frac{V_a}{R_a} - i_a - \frac{e_a}{R_a} \tag{7}$$

$$\tau_m \frac{d\omega}{dt} = -\frac{T_L}{B} - \omega + \frac{T_d}{B} \tag{8}$$

Where, τ_e = the field electrical time constant,

τ_a = armature electrical time constant,

τ_m = mechanical time constant of the system.

The field circuit equation (1) is an independent equation whereas the armature circuit equation (2) and the mechanical system equations (3)-(5) are coupled to each other through the back emf and torque relationships under steady state, the time derivatives in these equations are zero. Under steady state, the above dynamic relationships reduce to

$$V_e = I_e R_e \tag{9}$$

$$V_a = KI_e\omega + I_a R_a \tag{10}$$

$$T_d = T_L + B\omega \tag{11}$$

The power developed is

$$P_d = Td \cdot \omega \tag{12}$$

In the armature control of the separately excited dc motor, the field current (I_e) is kept constant at the rated value. The armature voltage (V_a) applied to the motor is varied to control the speed. Under such control, a combination of armature and mechanical system equations gives the following

$$KI_e \frac{V_a}{R_a} - K^2 I_e^2 \frac{\omega}{R_a} = T_L + B\omega \tag{13}$$

$$aV_a - b\omega = T_L \tag{14}$$

Where, $a = \dfrac{KI_e}{R_a}$ and $b = \dfrac{K^2 I_e^{\,2}}{R_a}$

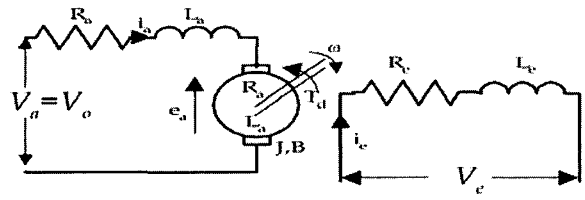


Fig. 4 The schematic diagram of a separately excited dc motor

The equation (14) defines the relationship between load torque and motor speed for any given armature input voltage (V_a).

To obtain per phase equivalent circuit model of the three phase series parallel resonant converter (SPRC) feeding motor load shown in Fig.5, all the components on the secondary side of the HF transformer are referred to its primary.

The per phase ac equivalent impedance Z_{ac} replaces the three phase HF transformer, three phase diode bridge rectifier along with its LC filter and motor load (R-L-E). Output voltage of three-phase rectifier,

$$V_o = \frac{3\sqrt{3}}{\pi} V_m \tag{15}$$

where V_m is the peak value of the input phase voltage to the three-phase diode bridge rectifier.
Voltage input to the motor is

$$V_o - E_b = \frac{3\sqrt{3}\sqrt{2}}{\pi} V_{Zac} \tag{16}$$

Assuming loss less rectifier circuit and the HF transformer; the power balance equation is as follows;

$$\frac{(V_o - E_b)^2}{Z_L} = \frac{3\left(\dfrac{\pi}{3\sqrt{6}} . V_o\right)^2}{Z_{ac}} \tag{17}$$

Where, $Z_L = R_a$

$$Z_{ac} = \frac{\pi^2}{18} \frac{V_o^{\,2}}{(V_o - E_b)^2} R_a \tag{18}$$

The per phase equivalent impedance Z_{eq} calculated from Fig. 5 is given below

$$Z_{eq} = \left[(jX_{L1} + jX_{C1}) + \frac{-jX_{CP} Z_{ac}}{Z_{ac} - jX_{CP}} \right] \tag{19}$$

Thus, the current through the resonant tank circuit I_{L1}

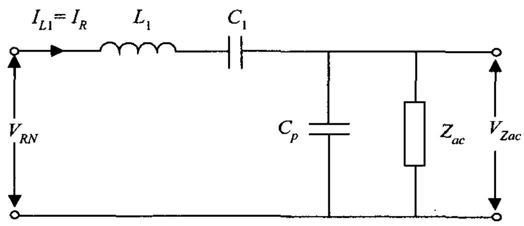


Fig. 5 Per phase equivalent circuit model for three phase SPRC feeding dc motor

$$\text{Is } I_{L1} = \frac{4V_{dcin}/\sqrt{2}\pi}{Z_{eq}} = \frac{4V_{dcin}/\sqrt{2}\pi}{\left[(jX_{L1} + jX_{C1}) + \dfrac{-jX_{CP} Z_{ac}}{Z_{ac} - jX_{CP}} \right]} \tag{20}$$

$$I_{L1} = \frac{4V_{dcin}/\sqrt{2}\pi}{\left[(jX_{L1} + jX_{C1}) + \dfrac{-jX_{CP} \dfrac{\pi^2}{18} \dfrac{V_o^2}{(V_o - E_b)^2} R_a}{\dfrac{\pi^2}{18} \dfrac{V_o^2}{(V_o - E_b)^2} R_a - jX_{CP}} \right]} \tag{21}$$

4. Design of Converter

The converter is designed for following specifications:

Input voltage:	V_{in} = 230 V (RMS)
Supply frequency:	f_{in} = 50 Hz
Output Voltage:	V_O = 230 V
Output Power:	P_O = 2500 W
Resonant frequency:	f_r = 140 KHz
Minimum switching frequency:	f_s = 150 KHz

Design of converter is based on sine wave approximation for the waveforms, in which only fundamental components are considered. This approximation is valid since the resonant tank circuit acts as a tuned low pass filter allowing only the fundamental frequency component of current to pass through the tank circuit.

The major design issues are choice of C_1/C_p ratio and selection of quality factor-Q of the resonant tank circuit. For C_1/C_p ratio greater than 1, the frequency variation required for output voltage regulation is large [8]. The peak current through the inverter switching devices does not decrease with the load current for smaller ratios of C_1/C_p It is therefore necessary to choose compromised value of C_1/C_p to be equal to 1 [9-10]. The KVA rating of the resonant tank circuit decreases as the quality factor-Q decreases for a given f_s/f_r ratio. This dictates the lower value choice for quality factor-Q. However it is observed that as the value of full load Q is increased, there is a decrease in the peak inverter output current with a larger decrease in load current. However, this decrease is small for the values of quality factor - Q greater than 4. Quality factor (Q) of the resonant tank circuit is

$$Q = \frac{\omega_r L_1}{Z_L'}$$

Where, $\omega_r = \frac{1}{\sqrt{L_1 C_1}}$ and Z_L' = Equivalent Load resistance referred to the primary of the transformer.

Considering the above constraints, the following compromised design values [9] are chosen for capacitor ratio C_1/C_p and full load quality factor Q, as $C_1/C_p = 1$ and $Q = 4$.

At full load condition the converter operated with switching frequency (f_s) of 150 KHz. For this condition the component value of resonant tank circuit are calculated as below:

$L_1 = 106.66\ \mu H$, $C_1 = 0.0121\mu f$

$C_p = 0.0121\mu f$ $C_{p'} = n_t^2\, C_p = 0.0133\mu f$

Where, n_t is turn ratio of HF Transformer primary to secondary.

4.1 Output Filter Design

To suppress the switching frequency component of the rectified resonant capacitor voltage, the output filter component L_F is used and a very small value of filter capacitor C_F is required. This switching frequency component has a minimum ripple frequency of six times the inverter switching frequency that gives the small filter size. The magnitude of its harmonic component is calculated using

$$V_{rn} = \frac{2V_o}{4n^2 - 1} \tag{22}$$

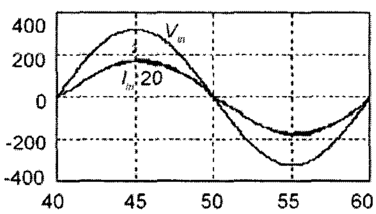
Where n is the multiple of $2^* f_s$

The dominant component of this voltage is calculated using peak-to-peak ripple specification of the output current. The filter components for the specified ripple contents in the output voltage and current are designed from the following equations.

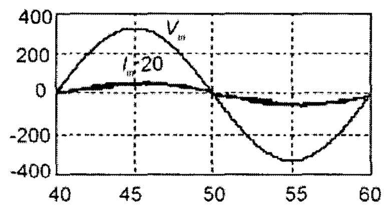
$$L_F = \frac{V_o}{105\pi \cdot f_s \cdot I_{p-p}}, \quad C_F = \frac{I_o}{105\pi \cdot f_{line} \cdot V_{p-p}} \tag{23}$$

5. Simulation Results

The converter designed in section 4 is simulated using PSIM. The behavior of the converter feeding motor load is studied at different load conditions. Fig. 6-9 show the simulation results at full load, 50% of full load and 25% of full load conditions respectively.



(a)



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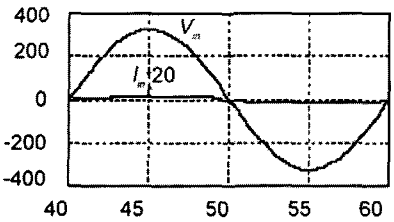
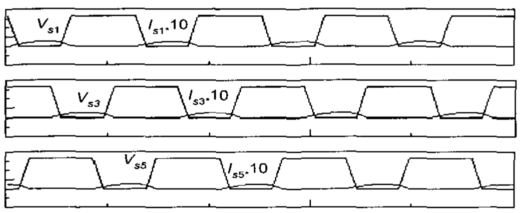


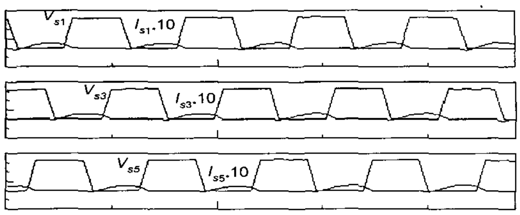
(c)

Fig. 6 Simulation results for the converter input voltage (V_{in}) and input line current (I_{in}) (a) at full load (b) at 50% load (c) at 25% load. Time scale in msec



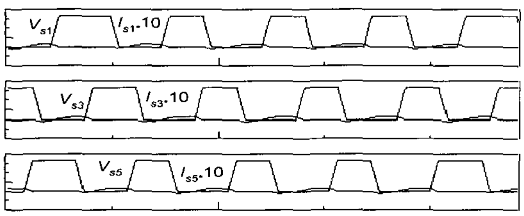
time (μs)

(a)



time (μs)

(b)



time (μs)

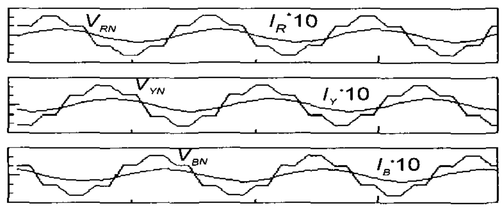
(c)

Fig. 7 Inverter switch voltages and currents, voltage: 100V/division (a) at full load condition (b) at 50% load condition (c) at 25% load condition Scale: time 5.35 μs/division

Fig.6 shows the results for the converter input voltage (V_{in}) and input line current (I_{in}). Fig. 7 shows HF inverter switch voltages and currents. Fig.8 shows HF inverter output phase voltages and phase currents and Fig. 9 shows dc-motor armature voltage and current. It seems from simulation results that input line current maintains a very high power factor and the converter operates in ZVS mode at all loads.

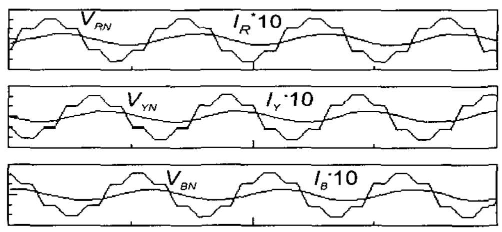
6. Experimental Results

The converter designed in section 4 is fabricated for verifying its performance, Three-phase IGBT bridge module BSM25GD120DN2 (EUPEC make), HF diodes DESI 60-12A, L_1 = 107.23μH, $C_1 = 0.01\mu f$, $C_p = 0.01\mu f$, L_F



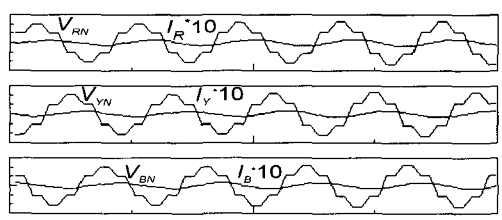
time (μs)

(a)



time (μs)

(b)



time (μs)

(c)

Fig. 8 Inverter output phase voltages and phase currents, voltage: 100V/division (a) at full load condition (b) at 50% load condition (c) at 25% load condition Scale: time 5.47 μs/division

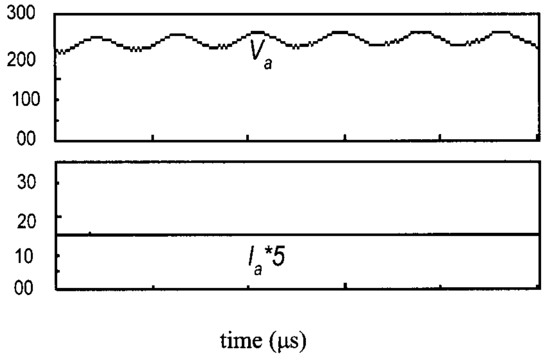
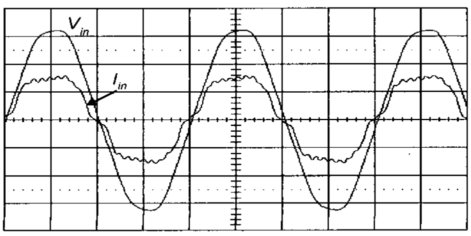


Fig. 9 Motor armature voltage (V_{aa}) and current (I_{aa}) at full load condition Scale: time 20.34 μs/division

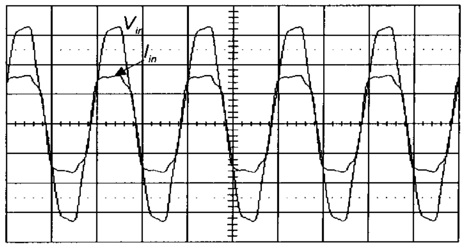
= 4.07μH. The control scheme is implemented using digital signal processor (DSP) TMS320LF2402. Variable frequency control is used for controlling the output voltage of the resonant converter for maintaining the motor speed constant at different load conditions. The resonant tank inductor L_1 of 107.23 μH is wound on a P 43x30 ferrite core with 18 turns. The HF transformer built on E55x28x20 core sections, made up of N87ferrite material.

Fig.10-12, show the experimental performance of the converter at various load conditions. It is seen from Fig.10, waveforms for input voltage and input line current that maintain very high power factor through out the entire loading conditions with very low input current THD (3.1% at full load, 3.8% at 50% of full load and 7.2% at 25% of full load). Fig.11 shows waveforms for phase voltage and phase current for three phase HF inverter levels that the converter operates in ZVS mode at all loads, eliminating the turn on losses.

Fig.12, show waveforms of motor armature voltage and armature current are reasonably



(a)



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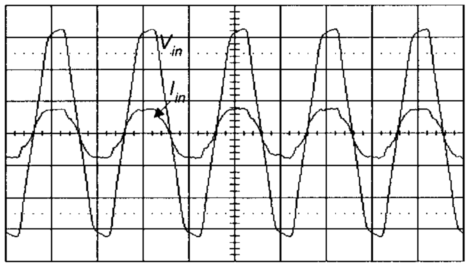

(c)

Fig. 10 Experimental waveforms for input voltage and input line current for the converter, Voltage Scale: 100V/division
(a) Full load - current scale: 20A/division : time 5 ms/division
(b) 50% load - current scale: 10A/division : time 10 ms/division
(c)25% load - current scale: 10A/division : time 10 ms/division

ripple free. It can be found that experimental waveforms agree well with the simulation waveforms.

To maintain the motor speed constant for different load torques, the armature voltage control method is used to control the speed of the dc motor. The armature voltage applied through the three phase SPRC needs to be adjusted.

Using variable frequency control strategy in the closed loop accomplishes this. With increase in load torque, the speed drops. The switching frequency (f_s) of the HF inverter stage is decreased such that the increase in output voltage of the converter restores the motor speed to the original value.

The motor normally passes through a transient period until the developed torque (T_d) equals the load torque (T_L). The closed loop control, for maintaining the speed of motor constant at all loads, is built using a DSPs TMS320LF2402 with an inbuilt PID controller. The popularity of PID controllers is due to their functional simplicity and reliability. They provide robust and reliable performance for most systems if the PID parameters are determined or tuned to ensure a satisfactory closed-loop

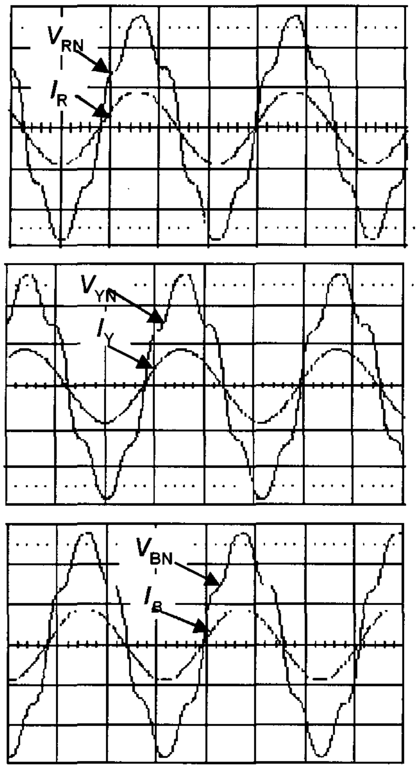

Fig. 11 Experimental waveforms for phase voltage and phase current for three phase HF inverter at full load Scales: time 2 μs /division, voltage: 100V/division, current: 20A/division

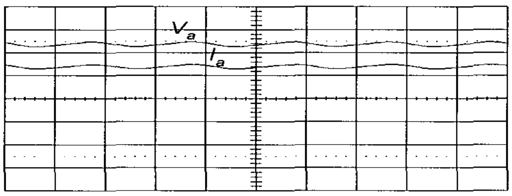

Fig. 12 Experimental waveforms for motor armature voltage and armature current at full load condition Scales: 100V/division, 20A/division, 2 μs /division

performance. This improves the accuracy, dynamic response and motor response to the load disturbances. The closed loop response of the motor speed due to a step change in load torque is shown in Fig.13 and Fig.14. This figure show that the motor speed will return to its initial value after less than 2 seconds due to a change in load by 50%. Fig.13 shows the response of the motor for a load disturbance from full load to 50% of full load. The response to a step change in the load from 50% of full load to full load is shown in Fig.14. The experimental results verify that the use of a three-phase series parallel resonant converter (SPRC) for feeding dc motor drives can provide a high efficiency over the major operating range due to the improved power quality converter, the input power factor unity over a wide operating shaft speed range, the power density and nearly ripple free output voltage and current.

The overall performance of the ac-to-dc converter using the three-phase SPRC feeding dc-motor drive under variable load conditions for variable frequency control is given in Table 1. At half full load the efficiency is at its maximum as the average value of the current is lower as compared to the full load; hence switching losses and conduction losses are reduced. Lower current switching stresses on the device are also reduced.

7. Conclusion

DC-motors play a significant role in modern industrial drives today. Separately excited dc-motors are used mainly with power converters for constant torque applications in the manufacturing industry. In the present scenario, there is a need to enhance performance of existing dc drives in many industrial applications including machine tool drive,

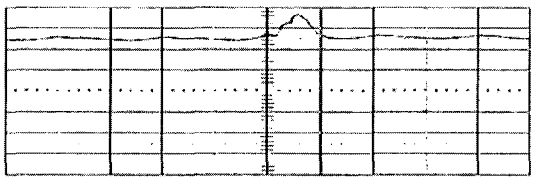

Fig. 13 The closed loop system response of motor for a load disturbance from full load to 50% of full load Scales: 100V/div., 2 s /div

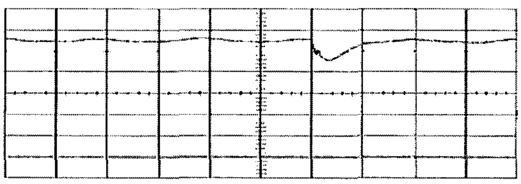

Fig. 14 The closed loop system response to step change in the load from 50% of full load to full load Scales: 100V/div., 2 s /div

Table 1 Experimental performance of the ac-to-dc converter using three-phase SPRC feeding dc-drive under variable frequency control

Parameter	Full load	50% Load	25% Load
P. F	1.00	1.00	0.99
THD (%)	3.1	3.8	7.2
Efficiency	91.5	94.8	85

paper mills, waste water treatment and steel plants. The conventional dc motor drive continues to take a considerable share of the variable speed drive market.

A three-phase series parallel (SPRC) resonant converter is a popular choice for medium to high power level variable speed dc-motor drives, due to the improved power quality converter and the input power factor unity over a wide operating speed range. This drive system is the most appropriate solution to preserve the present separately excited dc-motors in industry compared with the use of variable frequency ac drive technology.

Application of this converter in dc drives has been investigated in this paper. Design of three phase SPRC that inherently operates at a very high power factor with low THD has been outlined and the load as a separately excited dc-motor is modeled using the dynamic equation. Simulations have been carried out using PSIM for variable torque condition of the dc motor to investigate the control to output characteristic of three phase SPRC. An experimental model for three phase SPRC has been built to validate the design procedure and to verify the simulation results. A variable frequency (VF) control strategy has been implemented for closed loop operation using DSP TMS320LF2402 with inbuilt PID controller. The dynamic performances of the dc-drive showed improvement on account of the closed loop operation on three phase SPRC.

The performance on the ac line side has also been improved in terms of quality of power drawn by the drive system. The output voltage and current are reasonably ripple free. The best responses for load change can be obtained with suitable values of controller parameters. This SPRC fed dc-motor drive takes definite advantage of aspects of the PWM converter fed dc motor drive such as the reduction in current ripples and significant improvement input power quality, hence improvement in efficiency. The comparison between simulation and experimental results proved beneficial.

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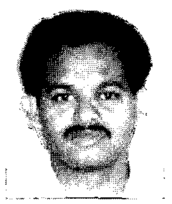

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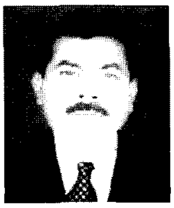

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