

Utilization of a Series Resonant DC Link for a DC Motor Drive

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Abstract—In this paper a high frequency series resonant dc link converter is utilized as a dc motor drive. This system generates a resonant current in a series link and switching is done at zero current instants, reducing switching losses to a minimal value. A Pulse Density Modulation (PDM) strategy, utilizing a current regulator loop and an external motor speed feedback loop, controls the resonant converter. A sinusoidal input fundamental current and nearly unity input power factor can be observed in different load conditions. The overall characteristics of the system, including such variables as maximum power, input current, start up, and transient responses are presented by digital simulation and verified on an actual prototype system.

I. INTRODUCTION

HIGH power converters can be broadly classified accordingly to their switching modes as:

- a) hard switching and
- b) soft switching

In a hard switching converter type, semiconductor devices are switched at the point of relatively large values of current or voltage, while in a soft-switching type the switching of the devices is made at zero voltage or zero current instants. It is apparent that soft switched converters not only have high power density but also possess very low switching losses. In general, switching schemes for resonant converters can be classified according to their resonant ac link and resonant dc link modes of operation. The resonant ac circuits utilize a parallel or series resonant link, impressing both polarities of ac voltage and current on the link, thus requiring bidirectional switches in the input and output converters [1]–[3]. The resonant dc circuits can also utilize a parallel or series resonant link. The majority of resonant dc link converters reported in the past have been restricted to parallel resonant types [4]–[5].

The series dc link circuits realize pulsating dc currents in the link by adding dc offsets to the ac resonant currents. A high frequency series resonant dc link, ac-to-ac power converter is proposed in [6] utilizing only 12 unidirectional switches and is the dual of the parallel resonant dc link system [4]. As

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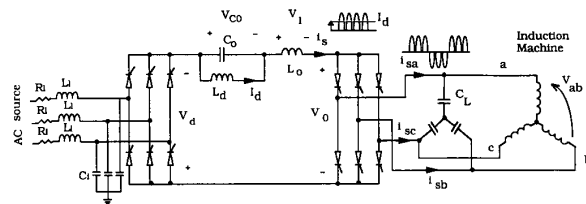


Fig. 1. AC/AC drive system utilizing the series resonant link.

shown in Fig. 1, the capacitor C_0 and inductor L_0 cause a resonant high frequency current i_s to flow from the input ac source to the load while the inductance L_d provides a dc bias, I_d , to the resonant current i_s . The total current then becomes unidirectional thereby allowing the utilization of high power thyristors as the switching element. Four thyristors conducting in series in the two bridges turn on and off at zero current instants, reducing switching losses significantly.

In [6] proportional and derivative current loop feedback and a damping series R-C circuit were utilized to minimize current pulse fluctuation and system instability of the proposed resonant system. Introduction of losses and load dependency are disadvantages of this solution. More recently a current pulse control strategy for this high frequency resonant scheme is proposed in [7]. Through adequate regulation of the current delivered to the load the output voltage error can be minimized. A circulating thyristor TCRC, is utilized to avoid over excitation of capacitors C_0 and C_L .

A variety of dc drive controls, such as phase control, integral cycle control, and chopper control [8], are well-known methods for obtaining a controllable dc voltage from a dc or ac source. In phase control the input voltage is applied to the dc motor during intervals of each half-cycle, while in the integral mode different number of half-cycles are applied to the motor during a certain period. In the chopper control mode, with or without the rectification stage, the control of the amplitude of the output voltage is obtained varying the duration of the on cycle of the switches [9].

Single and three-phase thyristor phase-controlled converter circuits utilized as a dc motor drive can be classified as half-wave, semi-converter, full-converter or dual-converter (where two full-converters are connected in inverse-parallel), offering 1, 2, or 4 quadrant operation in the voltage and current plane. Switching losses, due to the hard switching process, high ripple frequency (varying from the supply frequency, f_s , to $6f_s$) limits control of the input power factor and harmonic content, and can be cited as disadvantages of the existing systems.

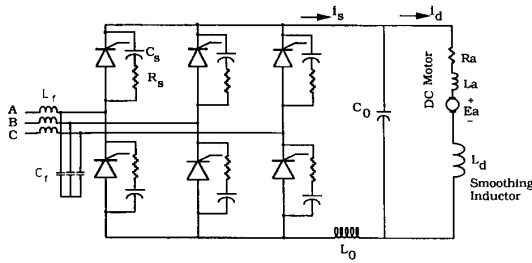


Fig. 2. Series resonant dc link converter as dc motor drive.

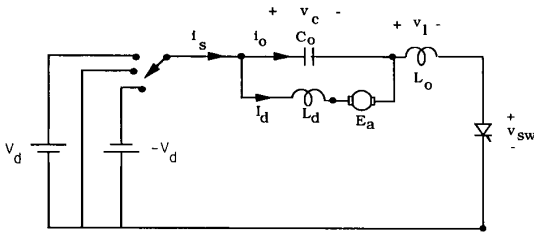


Fig. 3. Monophase model of the resonant current link ac/dc drive system.

II. SERIES RESONANT LINK FOR DC MOTOR DRIVE

The series resonant link converter can be used as a power conditioning system for Super Conducting Magnetic Energy Storage (SMES) coils [10]. The coil can be energized during low energy demand periods and when there is high demand, stored energy can be withdrawn. By proper control of the resonant link system, unity power factor can be obtained at the input terminals during the charging and discharging periods of the SMES coil.

In this paper it is shown that the series resonant link can also be utilized as a dc motor drive as seen in Fig. 2. In this figure, L_0 and C_0 are the resonant elements. Since the switching of the thyristors are done at the zero crossings of the current pulses, switching losses are reduced dramatically, and therefore switching frequencies about 30 kHz are possible. In practice, however, since after each pulse a minimum time is required for the switches to recover voltage blocking ability, the switching frequency is slightly lower than the resonant frequency.

Distribution of the pulses into the input phases by proper choice of PDM methods makes it possible to have sinusoidal input currents, unity input factor and low input harmonic content. Disadvantages of hard switching schemes seem to be eliminated by utilization of the series resonant link converter. This paper introduces digital simulation results as well as experimental results for different operating conditions. The control strategies required are also presented in the paper.

III. MONOPHASE MODEL

An understanding of how the resonant dc current link circuit operates can be gained by establishing and studying its monophase model as set forth in [11]. This circuit is shown in Fig. 3. Since the ac input voltage frequency is much smaller than the resonant frequency, the converter output voltage can be assumed constant during the period that is investigated here

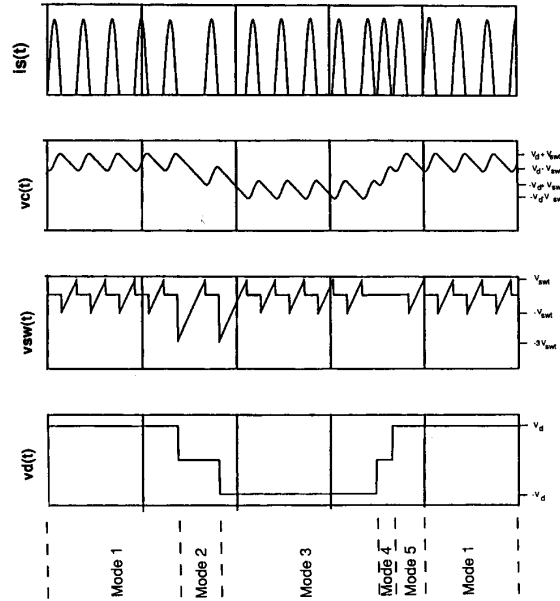


Fig. 4. Resonant current, capacitor voltage, switch voltage, and converter output voltage waveforms of the monophase model.

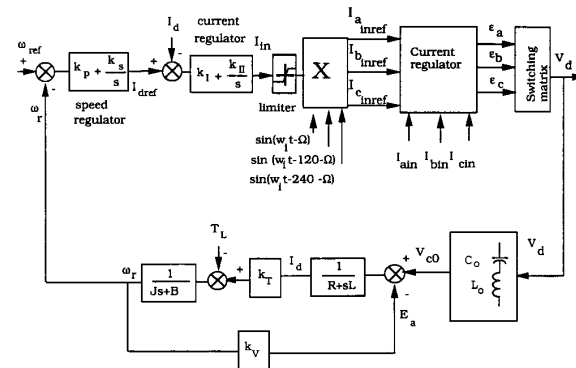


Fig. 5. Closed loop control of the drive system.

($v_d(t) = V_d, 0$, or $-V_d$). Also, the dc link current can be assumed to be constant ($i_d = I_d$).

The switch in Fig. 3 represents two oncoming switches in the converter and is turned on when the voltage across these switches ($v_{sw}(t)$) is equal to a specific positive value, V_{swt} . This value is a fraction of the peak converter output voltage V_d . The waveforms that illustrate operation of the circuit are shown in Fig. 4. Five possible modes of operation are explained below. The figure and the equations are for the case of $V_{swt} = 0.5V_d$.

Mode 1: In this mode a positive converter output voltage is required. When $v_{sw}(t) = V_{swt}$ the switch begins conducting and the following equations define circuit operation

$$\begin{aligned}
 v_c(0) &= V_d - V_{swt} \\
 i_s(0) &= 0 \\
 v_c(t) &= V_d - V_{swt} \cos \omega_0 t - Z_0 I_d \sin \omega_0 t \\
 i_s(t) &= I_d + \frac{V_{swt}}{Z_0} \sin \omega_0 t - I_d \cos \omega_0 t.
 \end{aligned}
 \tag{1}$$

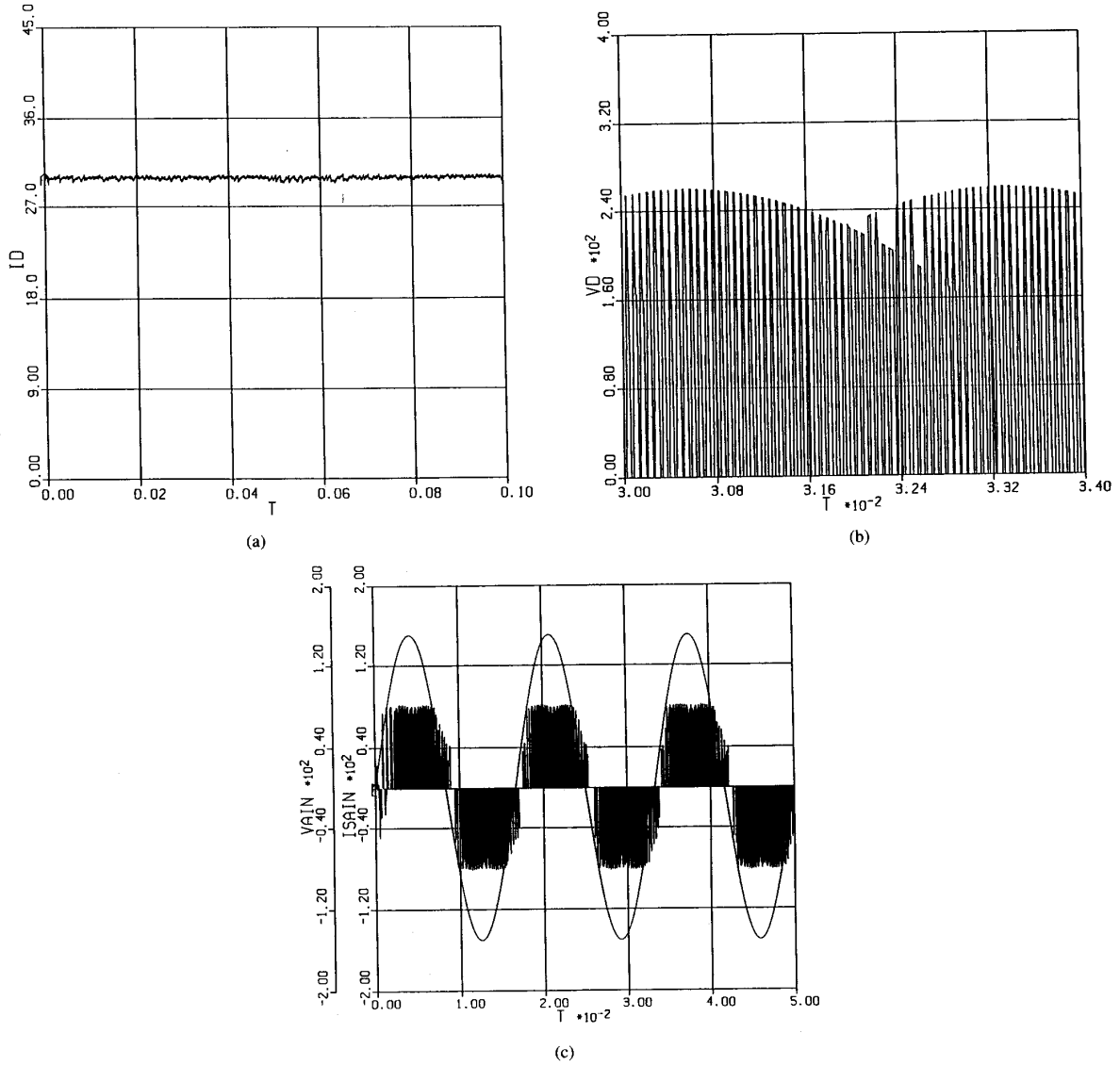


Fig. 6. Simulation results: (a) motor current, (b) converter output voltage, and (c) AC input voltage with unfiltered line current. For TL = 35 Nm, at the steady state.

where ω_0 and Z_0 are defined as follows

$$\omega_0 = \frac{1}{L_0 C_0} \quad Z_0 = \sqrt{\frac{L_0}{C_0}} \quad (2)$$

When the current pulse reaches zero, the switch turns off and the capacitor starts getting discharged by I_d

$$v_c(t) = -\frac{I_d}{C_0}t + V_d + V_{swt} \quad (3)$$

and when $v_{sw}(t) = V_{swt}$, the switch is turned on again and another current pulse is obtained.

Mode 2: If i_d is greater than the reference current, a negative converter output voltage must be applied to the link so that the resonant current can be reduced. Since an adjacent

states method is used [12], a null state must be chosen first. This corresponds to choosing the both switches on the same leg of the converter for conduction. For this mode

$$v_{sw}(t) = -v_c(t) \quad (4)$$

and, when $v_{sw}(t) = V_{swt}$ the switch is turned on. The equations are

$$\begin{aligned} v_c(0) &= -V_{swt} \\ i_s(0) &= 0 \\ v_c(t) &= -V_{swt} \cos \omega_0 t - Z_0 I_d \sin \omega_0 t \\ i_s(t) &= I_d + \frac{V_{swt}}{Z_0} \sin \omega_0 t - I_d \cos \omega_0 t. \end{aligned} \quad (5)$$

Mode 2 persists for one pulse period, and then Mode 3 begins.

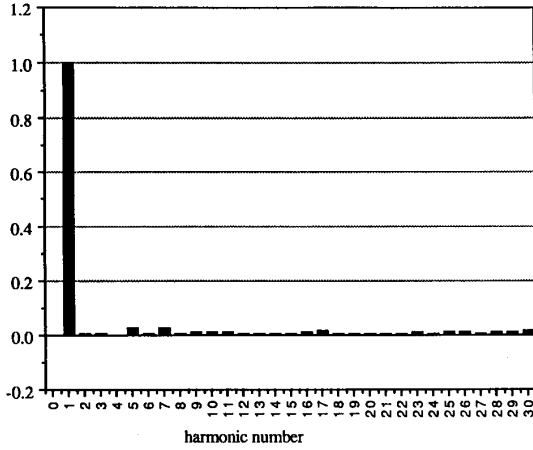


Fig. 7. Harmonic content of the input current as percentage of the fundamental harmonic. TL = 35 Nm, at the steady state.

Mode 3: The required negative voltage is now applied to the link. For this mode

$$v_{sw}(t) = V_d - v_c(t). \quad (6)$$

Until $v_{sw}(t) = V_{swt}$, the capacitor continues charging in the reverse direction. After this point, the following equations are valid

$$\begin{aligned} v_c(0) &= -V_d - V_{swt} \\ i_s(0) &= 0 \\ v_c(t) &= -V_d - V_{swt} \cos \omega_0 t - Z_0 I_d \sin \omega_0 t \\ i_s(t) &= I_d + \frac{V_{swt}}{Z_0} \sin \omega_0 t - I_d \cos \omega_0 t. \end{aligned} \quad (7)$$

The circuit remains in Mode 3 as long as a negative voltage is required for turn off of the devices. When i_d is below the reference current value, a positive voltage must be applied to the link. At this point, again, a null state must be chosen first.

Mode 4: At the end of Mode 3, when the current goes to zero, the switch voltage is already equal to V_{swt} , and therefore there is no charging or discharging time for the capacitor, and the switch is turned on immediately. The following equations define the operation of Mode 3

$$\begin{aligned} v_c(0) &= -V_d + V_{swt} \\ i_s(0) &= 0 \\ v_c(t) &= -(V_d - V_{swt}) \cos \omega_0 t - Z_0 I_d \sin \omega_0 t \\ i_s(t) &= I_d + \frac{V_d - V_{swt}}{Z_0} \sin \omega_0 t - I_d \cos \omega_0 t. \end{aligned} \quad (8)$$

When $i_s = 0$ Mode 4 ends, Mode 5 immediately begins.

Mode 5: In Mode 5, a positive voltage is applied to the link. Basically this mode is equivalent to Mode 1 except for the very first pulse. Again, since at the end of Mode 4 $v_{sw}(t) = V_{swt}$, the switch is turned on immediately. Equations for this mode are

$$\begin{aligned} v_c(0) &= V_{swt} \\ i_s(0) &= 0 \\ v_c(t) &= V_d - (V_d - V_{swt}) \cos \omega_0 t - Z_0 I_d \sin \omega_0 t \\ i_s(t) &= I_d + \frac{V_d - V_{swt}}{Z_0} \sin \omega_0 t - I_d \cos \omega_0 t. \end{aligned} \quad (9)$$

After the first pulse, Mode 5 ends and Mode 1 again begins.

IV. VOLTAGE AND CURRENT STRESSES

For given I_d and V_d values, the peak values of the resonant current, capacitor voltage and switch voltage are functions of V_{swt} . If a constant 'k' is defined as

$$k = \frac{V_{swt}}{V_d}$$

then the peak values for different modes are given with the following equations

$$i_{speak} = I_d + \sqrt{I_d^2 + k^2 \left(\frac{V_d}{Z_0}\right)^2} \quad \text{for Mode 1, 2, 3}$$

$$i_{speak} = I_d + \sqrt{I_d^2 + (1-k)^2 \left(\frac{V_d}{Z_0}\right)^2} \quad \text{for Mode 4, 5}$$

$$v_{cpeak} = V_d + \sqrt{k^2 V_d^2 + Z_0^2 I_d^2} \quad \text{for Mode 1, 3}$$

$$v_{cpeak} = \sqrt{k^2 V_d^2 + Z_0^2 I_d^2} \quad \text{for Mode 2}$$

$$v_{cpeak} = \sqrt{(1-k)^2 V_d^2 + Z_0^2 I_d^2} \quad \text{for Mode 4}$$

$$v_{cpeak} = V_d + \sqrt{(1-k)^2 V_d^2 + Z_0^2 I_d^2} \quad \text{for Mode 5.}$$

For the switch voltage $v_{swt}(t)$, the peak value is V_{swt} except when there is a transition from positive to null state, and from null state to negative voltage state. At these times the peak value is equal to

$$v_{swpeak} = \frac{1+k}{k} V_d.$$

Note that for $k = 0.5$ all the peak current values are equal so that this is the minimum overall voltage condition for rating of the solid-state switching devices.

V. CONTROL STRATEGY

Fig. 5 shows the closed loop control schematic for the series resonant link driving a dc motor used in this study. The speed feedback is the outer most loop and its error, through a *PI* controller yields the reference current. The second loop is for the dc motor current control. DC current control is very important since without current regulation, resonant pulses would not have a zero crossing and, therefore, zero current switching would not be possible. The current error is used to generate an ac input current reference. Finally, ac input current references are compared to real currents and three error signals, $\varepsilon_a, \varepsilon_b, \varepsilon_c$, are generated. These signals are used to decide which switches should be turned on at the next switching instant. For this stage, the adjacent state method is used [12]. The switches of the converter are triggered in such a manner that the output voltage may only have the possible most positive, the possible most negative and zero voltages depending upon the error signals. Transition from a positive energy pulse to a negative energy pulse is done after having selected first a null state to limit the voltage variations. Selection of the thyristors that will be turned on is based on the following criteria:

- the thyristor in the phase having the larger error out of the two phases of the same polarity is chosen to be triggered, and
- the phase corresponding to the error with the opposite polarity error is selected as the other triggering phase.

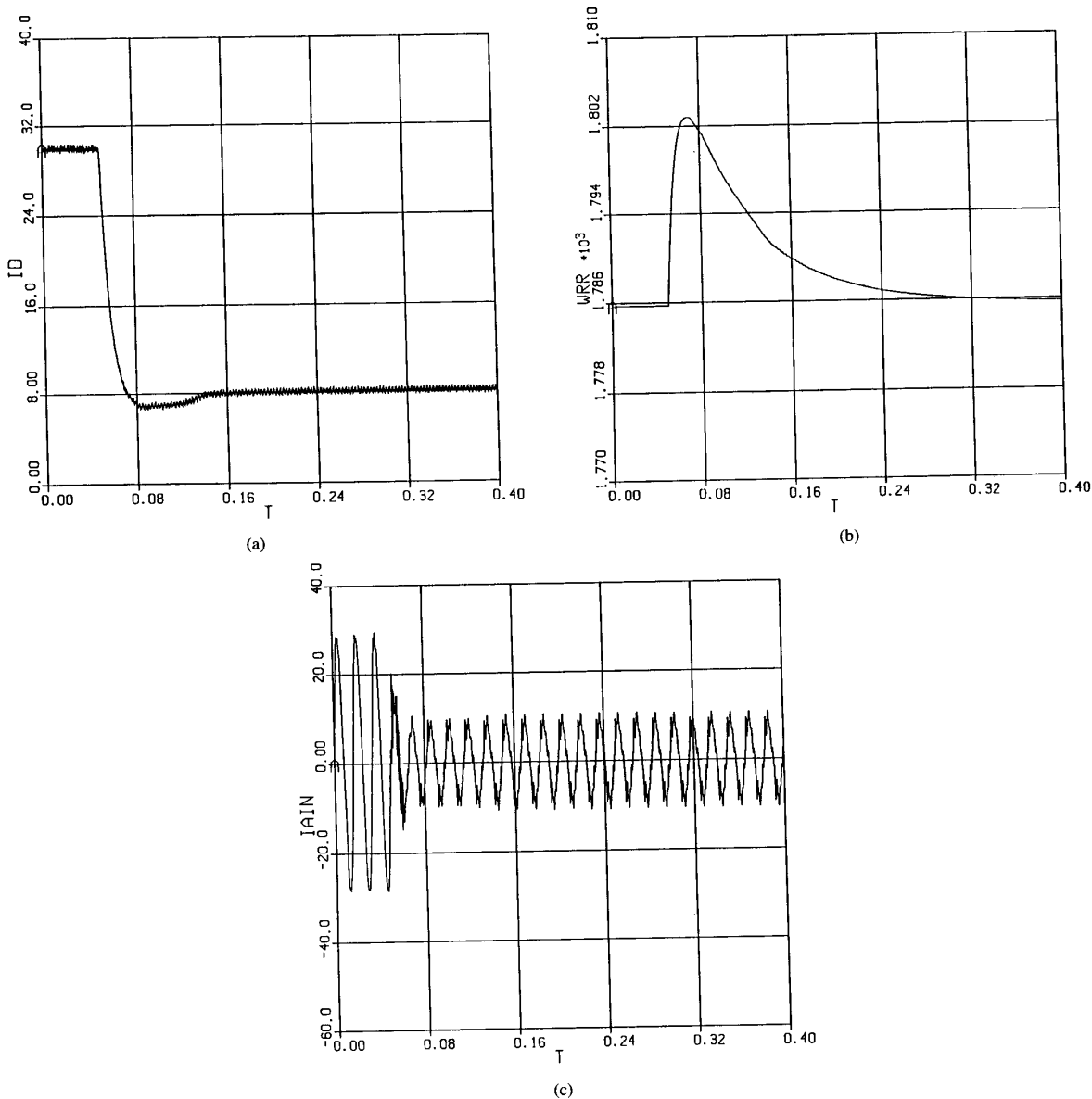


Fig. 8. (a) Motor current, (b) speed, and (c) ac input current under a load change from 35 Nm to 10 Nm.

VI. SIMULATION RESULTS

The ACSL simulation language was used to simulate the series resonant dc link circuit driving a dc motor with the following parameters:

- Rated power: 10 hp
- $L_a = 4.3$ mH
- $R_a = 0.57$ ohm
- $J = 0.0881$ kgm/rad/s²
- $B = 0.02$ kgm/rad/s
- $\omega_{rated} = 1750$ rpm = 183.26 rad/s

- $K_v = 1.17$ V/s
- $K_t = 1.17$ Nm/A
- smoothing inductor = 40 mH

Fig. 6 shows typical operating waveforms for a load torque of 35 Nm at the steady state. Fig. 6(a) shows the motor current. As can be seen, the ripple is very low. The effect of the adjacent state method can be seen in Fig. 6(b), which shows the converter output voltage. Fig. 6(c) shows the ac input voltage and ac line current pulses. It is obvious that the pulses are distributed to provide unity power factor and sinusoidal input currents.

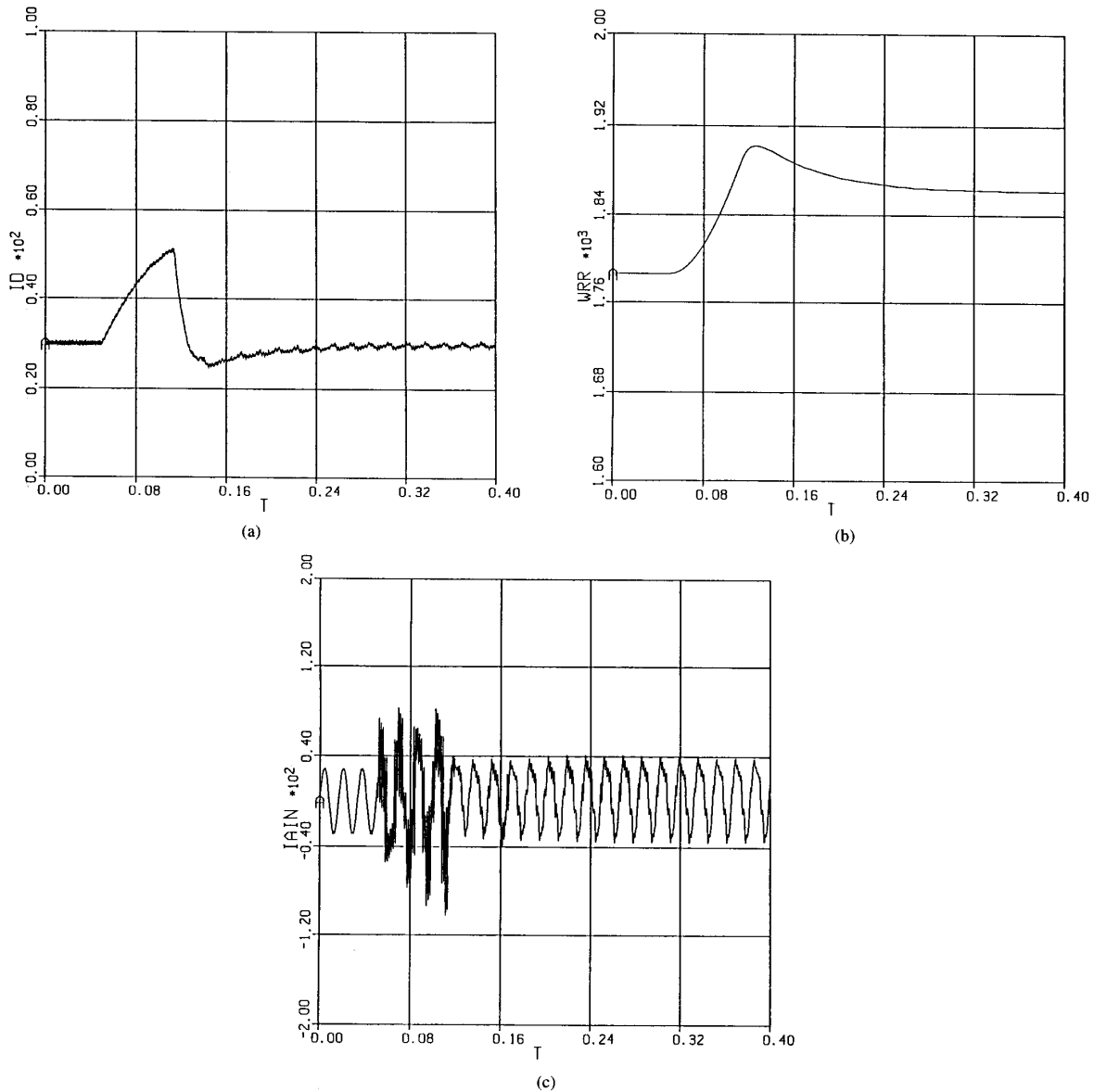


Fig. 9. (a) Motor current, (b) speed, and (c) ac input current under a speed reference change from 177 rpm to 185 rpm.

As an expected result of the proposed control, input current harmonics are very low. This is shown in Fig. 7. All the harmonics are shown as the percentage of the fundamental harmonic. Figs. 8 and 9 show the waveforms when there is a step change either at speed or load torque. Fig. 8 shows (a) the motor current, (b) the motor speed, and (c) the ac input current when the load torque is reduced from 35 Nm to 10 Nm. Similarly, Fig. 9 shows the same variables when the speed reference is increased from 170 rpm to 177 rpm.

VII. EXPERIMENTAL RESULTS

The system that has been shown in Fig. 2 has been constructed in our laboratory. The component values that were

utilized are:

$$C_0 = 0.9 \mu\text{F}$$

$$L_0 = 60 \mu\text{H}$$

$$L_d = 30 \text{ mH}$$

The ac input voltage is 115 V and IR85RDT GTO's were used with a series and an anti-parallel diode in order to realize a fast switch as seen in the figure. The dc motor is a 3/4 hp permanent magnet machine. A V_{swt} value of 75 V ($0.46 V_d$) was used. The system was controlled by a Motorola 56000 DSP microprocessor.

Two different control schemes were used for controlling the system. The first scheme is a bang-bang method in which

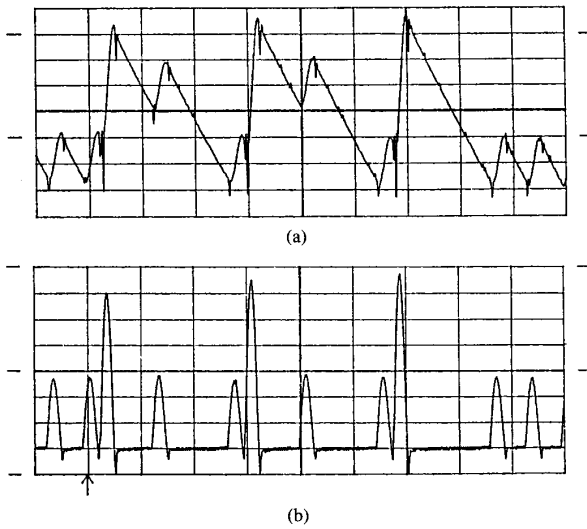


Fig. 10. Experimental results with the bang-bang method: (a) capacitor voltage, 100 V/div and (b) resonant current, 5 A/div. (0.1 ms/div).

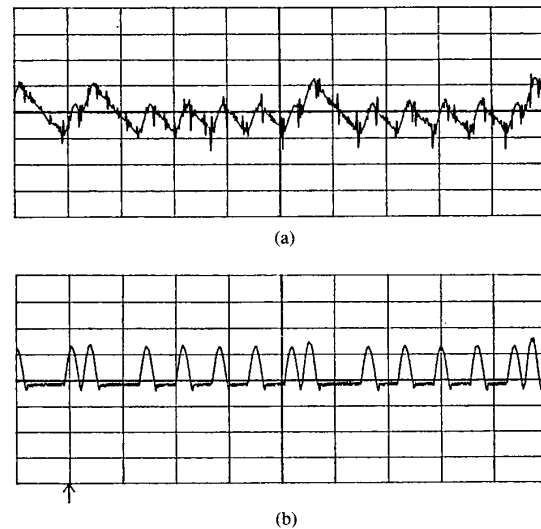


Fig. 12. (a) Capacitor voltage, 100 V/div and (b) resonant current, 5 A/div (0.1 ms/div).

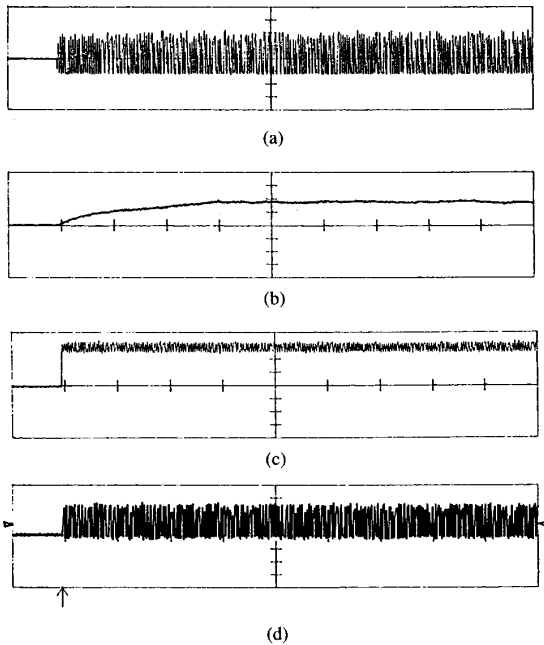


Fig. 11. Starting of the motor with bang-bang method: (a) v_d 100 V/div, (b) speed, 250 rpm/div, (c) motor current, 0.5 A/div, and (d) resonant current, 5 A/div ($i_{dref} = 2$ A, 0.2 s/div).

the switches are turned on or off to obtain the most positive or most negative voltages depending on the polarity of the error signals. In this case, as expected, high voltage stresses and non-uniform current pulses occur. The second method switches only between the most positive voltage and null state. Therefore stresses and current peaks are not as high as those of the previous method and is clearly preferred.

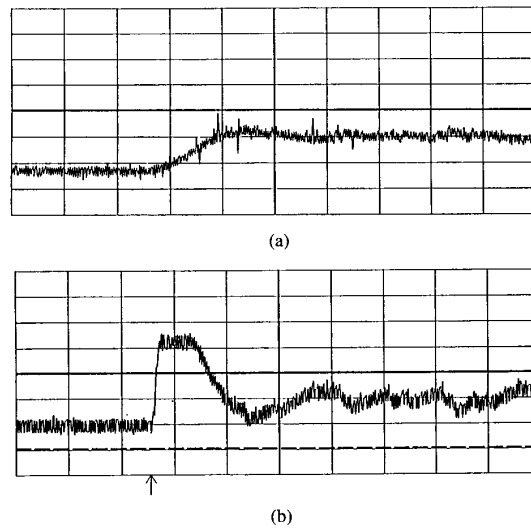
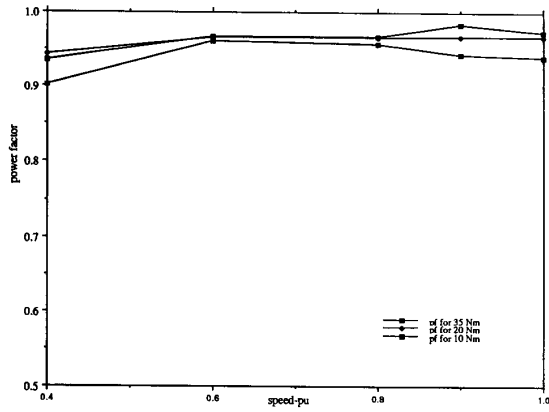


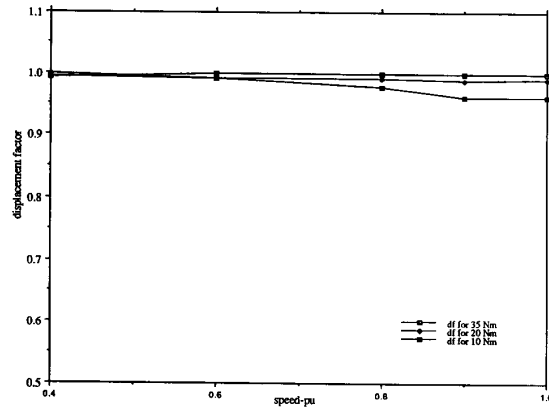
Fig. 13. Response to speed reference change: (a) speed, 250 rpm/div and (b) motor current, 0.5 A/div (0.1ms/div).

Fig. 10 shows the capacitor voltage and resonant current when the bang-bang method is used. As explained previously, large current peaks occur when a transition from negative maximum voltage to positive voltage is required. Fig. 11 illustrates the starting waveforms of the system when utilizing the bang-bang method.

Figs. 12 and 13 show waveforms when the adjacent state method is used. Fig. 12 shows the capacitor voltage and the resonant current pulses at the steady state. The current pulses are, in this case, more uniform due to the lower voltage changes as explained earlier. The response of the system to an increase at the speed reference is given in Fig. 13.



(a)



(b)

Fig. 14. Variation of (a) power factor and (b) displacement factor with respect to the motor speed.

VIII. CONVERTER PERFORMANCE

Fig. 14 shows the variation of power factor and displacement factor with respect to the motor speed, for the resonant dc link converter utilized as a dc motor drive. Three different curves are presented in each case, for motor load torque of 35 Nm, 20 Nm and 10 Nm, with a smoothing reactor of 30 mH. The input power factor curve shows an improved result in relation to conventional dc motor drive system, since its dependency of the firing angle has been eliminated in the resonant scheme. Values of 0.9 and above were obtained for the input power factor for a large speed range. The displacement factor is approximately 1 for all speed range, since that was one of the main objectives of the implemented control.

Fig. 15 demonstrates the effect of the variation of the series inductor (L_d) in the peak current values of the motor current. As the series inductor value decreases, the ripple in the motor current increases. As can be observed, for a 10 hp dc motor drive, a series inductance of 5 mH gives reasonable ripple factor.

IX. ALTERNATIVE RESONANT TOPOLOGY

In an attempt to reduce losses over the resonant inductor L_0 , an alternative topology, shown in Fig. 16, can be utilized

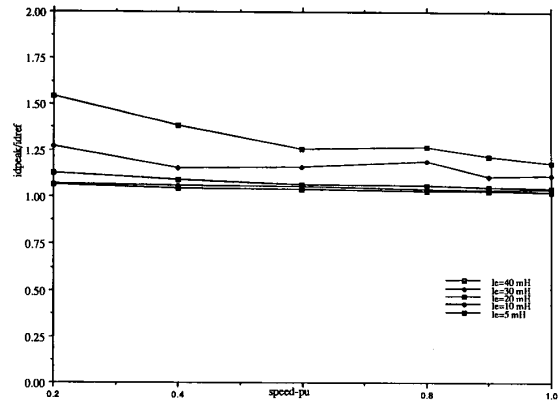


Fig. 15. Ratio idpeak/idref ratio as a function of speed for two different value of smoothing inductor.

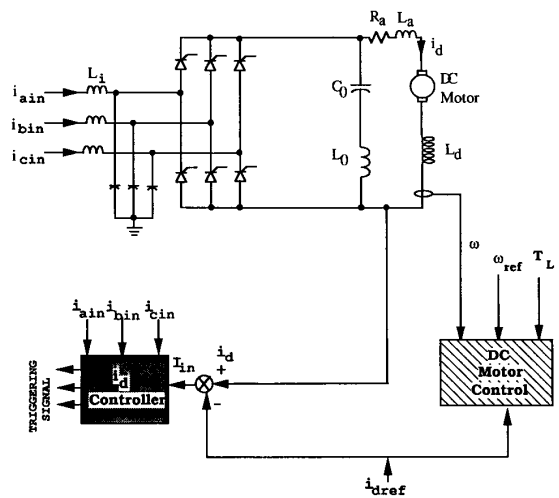


Fig. 16. Alternative topology for the resonant drive system.

where the resonant inductor is in parallel with the bias inductor and dc motor. In this configuration the current I_d does not circulate through the inductor L_0 during the duration of the resonant pulse. Similar waveforms were obtained for the motor current, voltage and input current in the alternative resonant topology.

The conventional resonant topology presents a better power factor at higher speed while the alternative topology has a better power factor at low speed. At light loads both topologies have the same power factor. Both topologies have very good displacement factor at different load condition and speed. At higher speeds the alternative resonant topology demonstrates a better harmonic factor than the conventional resonant topology. At heavy loads, the alternative resonant topology has a better ripple factor, while at light loads the conventional resonant topology presented a better ripple performance. The alternative resonant topology has a better overall performance for smaller values of the external inductance (L_d) if compared with the conventional resonant one.

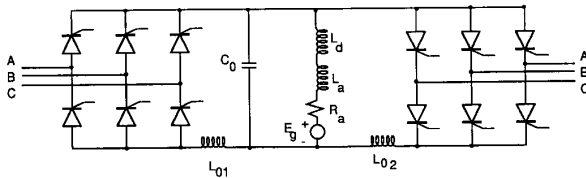


Fig. 17. Dual resonant converter dc motor drive.

X. DEVICE UTILIZATION

Different devices have been proposed to be utilized with this system, like GTO's in the gate assisted turn-off mode with a series fast diode and an antiparallel diode, for the purpose of decreasing the necessary turn-off time. Resonant frequencies of the order of 32 kHz were obtained in the experimental model. Since the resonant pulses reaches zero naturally, SCR's constitute an ideal choice for the resonant converter switches. Resonant frequencies of 27 kHz were obtained in the model if the device is kept at room temperature. An increase in the device's temperature requires a large turn-off time, which can cause failures in the switching process.

Considering that the switching losses are extremely reduced, operations at low temperature (80°F) should be possible, enabling sufficient turn-off time for the device turn-off process. Resonant peak currents of 120 A, and Id currents of 45 A with a turn-off time varying from 3–5 μ s were obtained in the experimental system. Switches known as Zero Turn-off Thyristors (ZTO) constitute the ideal switch for this application, considering that the turn-off time requirement for these devices is extremely reduced.

XI. DUAL CONVERTER

If operational conditions requires that the dc motor operates in forward and reverse direction, a mechanical contactor could be utilized to invert the armature polarity. Considering the volume of the arc extinction chamber and the losses introduced by the contactor, two resonant converters in opposite directions connected in parallel can be utilized for this purpose, allowing the flow of the motor current in any desired direction. With this topology four-quadrant operation is obtained. Fig. 17 shows the resonant converter in a dual-converter topology.

In general, switching frequencies as high as 30 kHz were used in experiments. All of the waveforms given here are for approximately a 20-kHz switching frequency.

XII. CONCLUSION

This paper has presented the theory, control, and application of a new type of dc motor drive based upon a resonant dc link principle. Since the resonant pulses that flow in the link reach zero at a high frequency rate (10 kHz or more) high frequency modulation is possible with only conventional thyristors. This new type of system allows for very high switching frequency of a conventional thyristor bridge thereby making possible ac input current waveforms of much lower distortion than previously possible with the same filtering kVA. In addition the converter can be controlled so as to always maintain unity power factor. It is expected that the approach could find utility in future applications where input current waveform is of concern.

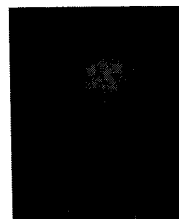
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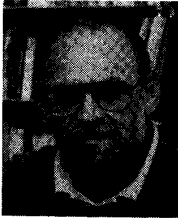
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