A Separate Double-winding 12-phase Brushless DC Motor Drive Fed from Individual H-Bridge Inverters

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***Abstract* -- This paper deals with a separate double-winding 12-phase BLDC motor fed from individual H-bridge inverters. The system is fault tolerant in that continuous operation is possible under fault conditions. Every phase is electrically independent from each other and doubly wound to intensify the redundancy of the system. Furthermore, each phase is asymmetrically 15 degrees apart to have the same torque ripple reduction effect as 24-phase counterpart. Two partial windings in each phase are taken control of by one inverter module. One module consists of two single phase H-bridge inverters and can be detached from other modules to be repaired or replaced while it is out of operation. In terms of motor drive, sinusoidal wave currents are applied to the stators for minimizing torque ripples and the operating efficiency at low rotational speed of motor is improved by series connection of the partial windings. Simulations and experimental studies are conducted to evaluate the performance of the system based on a prototype motor.**

***Index Terms*-- separate double-winding, H-bridge inverter fed BLDC, fault tolerance.**

I. INTRODUCTION

Multi-phase variable-speed motor drives dates back to the 1960s, when inverter-fed ac drives were in its early stage [1]. At that time, as six-step mode operation of three-phase inverter was generally employed instead of PWM, there existed low frequency ripple in output torque. To handle this problem, five-phase and six-phase variable-speed drives were developed [2]-[3]. In the mean time, after PWM technology enabled control of the inverter output voltage harmonics, three-phase motor with vector control was mainly hired.

Over the last 2 decades, thanks to the developments in the areas where power density, efficiency and fault tolerance were of great importance such as electric propulsion applications, multi-phase motor drives obtained recognition again. For a given output power, multi-phase motor drives can reduce the stator current per phase, which leads to the usage of semiconductor switches with lower rating and by increasing the number of phases, the torque per ampere for the same volume machine also becomes higher [4]. Additionally, in terms of fault tolerance, multi-phase motor drives are able to continue operation after unexpected accidents such as winding open/short circuit or inverter switch open/short circuit.

Especially, multi-phase BLDC motors are increasingly utilized in traction or propulsion applications due to the higher efficiency, power density and relatively easy control [5]-[6]. This paper describes a separate double-winding 12-phase BLDC motor drive for electric propulsion applications. Comprehensive analyses regarding design aspects and system configuration are carried out. Experimental results as well as simulations are presented to complement the works.

II. SYSTEM REQUIREMENT

A. Fault-tolerance

The system dealt in this paper is designed especially for reliability-critical applications. In other words, motor structure has to be implemented not to compromise the drive performance even in fault conditions. For these purposes, modular designs are adopted in both stator windings and supply system to minimize the coupling between each phase. It gives electrical isolation of stators and no neutral point exists. Each phase is driven by individual H-bridge (single phase) inverters and one has no effect on others when any fault occurs. With these structures, fault tolerance is ensured since healthy phases can continue operation under fault conditions and inverters out of order can be disconnected to be repaired or replaced.

Furthermore, the winding of the motor is made up of 12 phases with 2 electrically identical single phase windings each to enhance redundancy of the system. Each of these partial single phase windings is connected to the H-bridge inverter terminals and can be provided with power by two different sources to prepare for source faults.

The conceptual layout considering the aforementioned requirements is depicted in Fig. 1. Switch for series connection of the partial windings for higher efficiency operation at low speed – high efficiency mode – is also shown in the figure and its operation is described in part C.

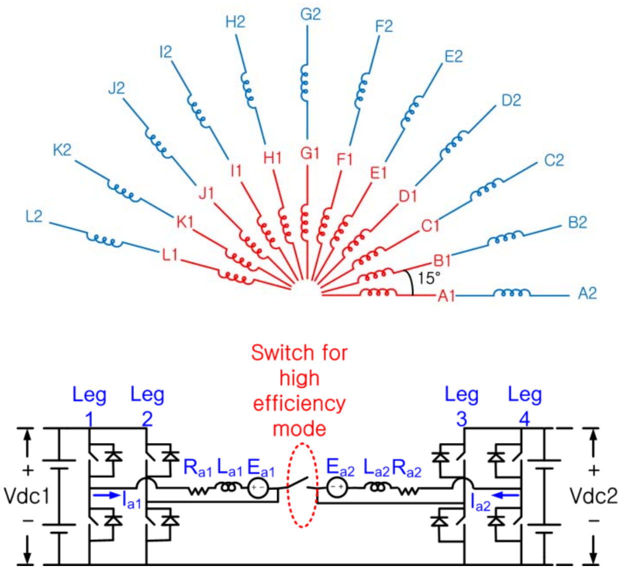

Fig. 1. Conceptual electrical layout of motor and inverter connection

B. Torque Ripple Reduction

BLDC motors have trapezoidal back EMF waveforms and are generally fed with rectangular stator currents to produce constant torque. Unfortunately, in practice, non-uniformity of magnetic material and design trade-offs make it hard to induce the desired trapezoidal back EMF. Besides, cogging torque from the interaction between the permanent magnets of the rotor and stator slots also causes torque pulsations.

On top of these problems, BLDC motors suffer from so-called commutation torque ripple. According to a two-phase feeding method in the event of 3-phase drive, it is required to apply a rectangular current in the flat portion of the ideal trapezoidal back EMF, thus the commutation occurs every electrical 60 degrees. However, as the motor windings are inductive, the commutation takes finite time and it leads to six order harmonics of torque.

To improve torque performance in this 12 phase motor, the spatial displacement between any two consecutive stator phases in the system is designed to be $\pi/12$ which leads to an asymmetrical distribution of magnetic winding axes in the cross section of the machine. In this structure, if the stators are collected in 4 groups of 3 phases, such as (A1, E1, I1), (B1, F1, J1), (C1, G1, K1) and (D1, H1, L1) in Fig. 1, the torque ripple from each group is electrically 15 degrees apart and compensates each other's six order harmonic torques.

Simulation results are presented for comprehension. Fig. 2 shows the real back EMF waveforms of double winding 12-phases and they are reflected in the simulation. PI controlled rectangular phase currents are injected and the torque ripple reduction from the increase of operating phase number is demonstrated. Fundamental components in torque ripple come from the non-ideal trapezoidal back EMF.

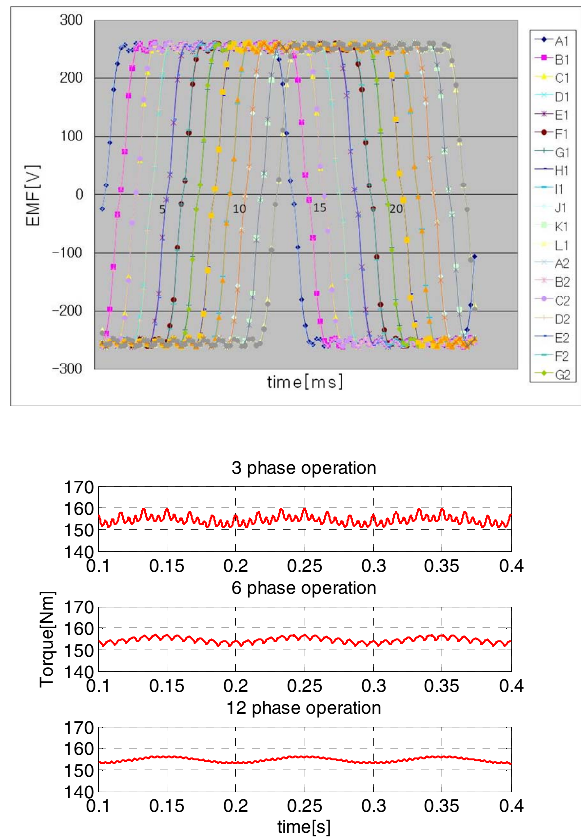

Fig. 2. Waveforms of back EMF and torque ripples of 3, 6 and 12 phase operation – computer simulation

C. High Efficiency Mode Operation

In propulsion drives such as electric ship applications, efficiency is one of the major concerns to save operating costs. To take this into account, phases in pair, for example phases A1 and A2 are connected in series at low speed region by turning on the switch between the phases for high efficiency mode shown in Fig. 1. In this mode, the switches of leg 2 and leg 3 are turned off and a new H-bridge set is composed with leg 1 and leg4, which contributes reduction of switching loss. Additionally, since impedances and EMFs of 2 phases are added, the voltage reference increases at low speed and voltage syntheses error due to the dead time effect can be relatively decreased as well.

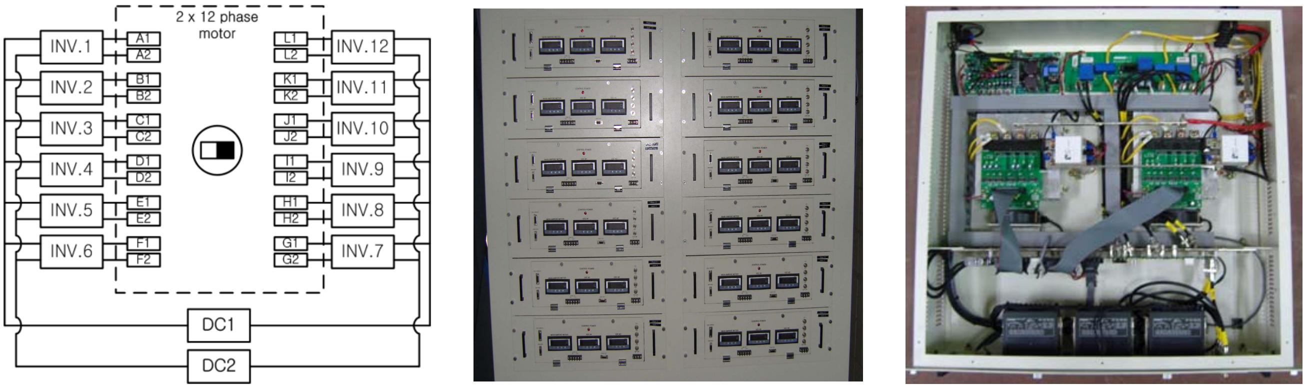

Fig. 3. Inverter structures and experimental realization

III. SYSTEM CONFIGURATION

Fig. 3 illustrates the structure of the entire inverter modules and real experimental inverter sets. 12 inverter modules drive 24 phases of the separate double winding 12 phase BLDC motor.

One control board is mounted in each inverter module and takes control of 2 phases. As individual inverter modules do not share A/D inputs, one inverter module has its own phases' information only. As a result, currents/voltages references should be derived with these restrictions to produce constant torque.

IV. MOTOR DRIVE

Generally, a three phase BLDC motor with trapezoidal back EMFs is operated by square wave currents to produce constant torque. For the separate double-winding 12-phase motor dealt in this paper, however, although sinusoidal currents are fed, torque ripple is expected to reduce by compensating each 3 phase groups' ripple components. Fig. 4 shows the output torque between the cases of square wave currents and sinusoidal wave currents, where real back EMFs are reflected by offline measurement. It indicates that they have very similar performance in terms of torque ripple and therefore sinusoidal currents can be a choice for easy implementation.

To apply sinusoidal currents without phase delay or steady state errors, synchronous frame PI controllers are commonly employed. However, as this system consists of 24 single phases fed by 12 H-bridge inverters, synchronous frame controllers are difficult to implement. Accordingly, stationary reference controllers equivalent to synchronous frame ones have to be used [7] and it is called PR (Proportional and Resonant) controller. The transfer function of PR controller is given in (1) and it is applied to all 24 single phases.

$$G_{DC}(s) = K_P + \frac{K_I}{s} \Leftrightarrow G_{AC}(s) = K_P + \frac{2K_I s}{s^2 + \omega_c^2} \quad (1)$$

where ω_c is the operating frequency which has to be regulated.

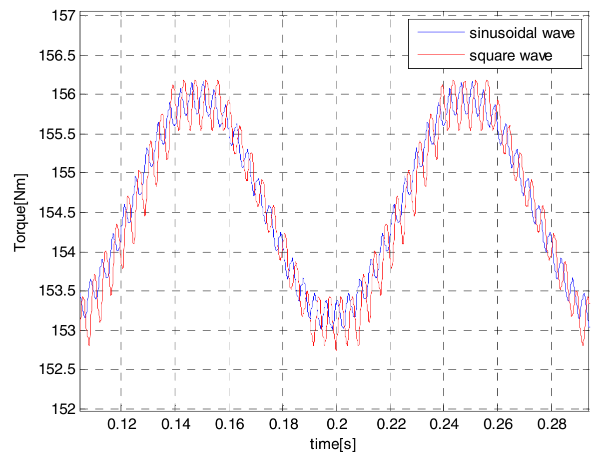

Fig. 4. Output torque from two different currents injection

Unfortunately, as a result of current control in 24 phases, there exists a lot of harmonics in phase currents as shown in Fig. 5. The harmonic components appear when the number of operating phases increase or operating speed rises, and the analysis on the reasons are followed.

A. *Influence of back EMF*

Fig. 6 and Fig. 7 show the current control of a phase when the motor is at a standstill and driven at constant speed by a load machine, respectively. At a standstill, when there is no back EMF, the current control shows reasonable performance whereas more harmonic components are included in the operation with back EMFs. In short, the back EMF components of motor have significant effects on the current control performance.

The back EMF information has been obtained by measuring terminal voltages of each phase with no current. As shown in Fig. 8, high order harmonics in back EMF causes problems when feed-forwarding in digital current controller. Digital controller inevitably suffers from digital delay and the delay in angle becomes larger as the rotating speed becomes higher though the delay in time is fixed. For example, in 10 kHz sampling controller as same as this system, since 340 Hz component makes 18.36 degrees delay, perfect back EMF compensation is impossible even though exact back EMF information is known. Besides the high order harmonics in back EMF, the real back EMF can be changed by distorted linkage flux through the self and mutual coupling of phases when current is flowing in all phases. Hence, harmonics in phase currents are unavoidable with only fundamental frequency PR controllers.

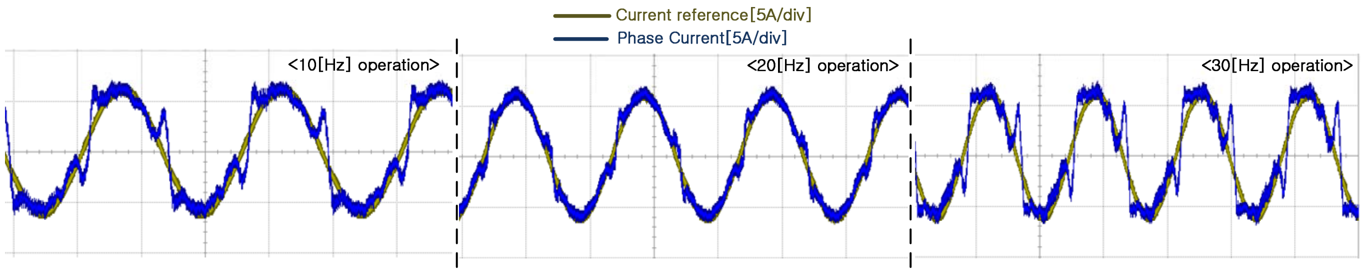

Fig. 5. Current control of variable speed operation of 24 phases

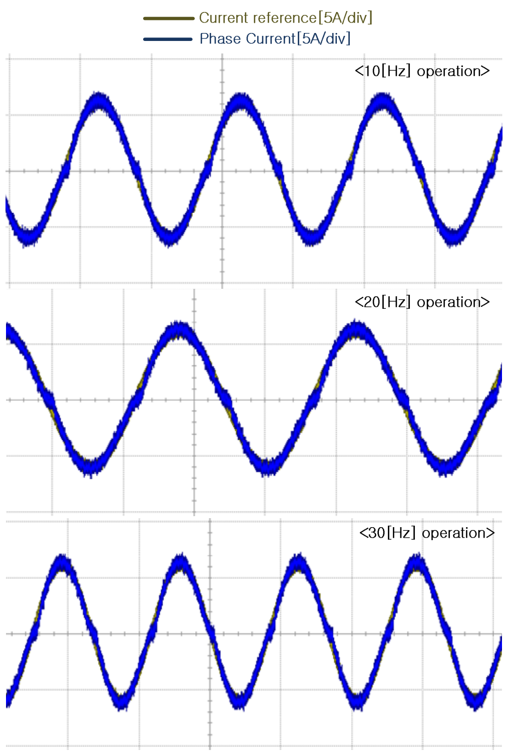

Fig. 6 Current control of a phase at a standstill

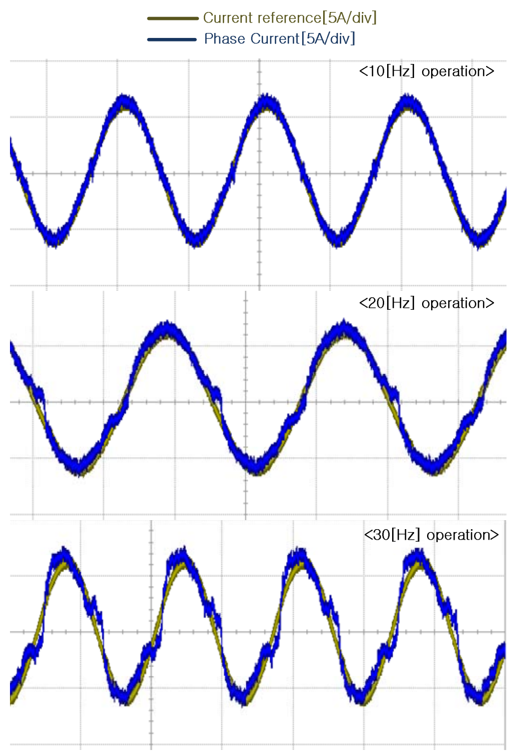

Fig. 7 Current control of a phase run by a load machine

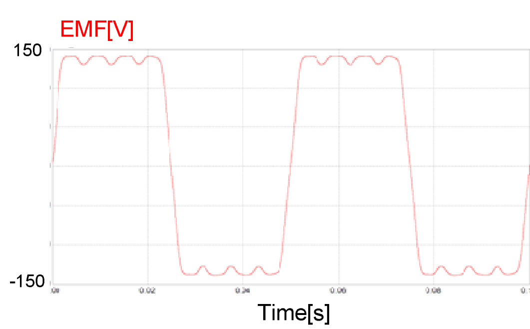

Fig. 8 Measured back EMF waveform

B. *Influence of Operating Phases' Number*

The voltage equation of phase A in this system is given as (2).

$$V_a = Ri_a + \frac{d\lambda_a}{dt} = Ri_a + L\frac{di_a}{dt} + \lambda_{linkage} \qquad (2)$$

where $\lambda_{linkage}$ changes according to the number of operating phases. If $\lambda_{linkage}$ contains only fundamental frequency component, sinusoidal phase current can be obtained with fundamental frequency voltage. However, as mentioned in the previous section, high order harmonics from the deficient back EMF feed-forward are included in other phases and the linkage flux to phase A also incorporates harmonics. As a result, couplings with other phases changes current waveforms. Taking three phase and six phase systems as an example, phase current can be analyzed as follows.

i. Three phase system

For convenience of calculation, it is assumed that two phases have 3rd, 5th and 7th harmonics and every phase is symmetrical. Then the currents of two phases with the harmonics can be expressed as (3) and (4).

$$\begin{aligned} i_2 &= A\cos(\omega_e t - 2\pi/3) + B\cos(3\omega_e t) \\ &+ C\cos(5\omega_e t + 2\pi/3) + D\cos(7\omega_e t - 2\pi/3) \end{aligned} \qquad (3)$$

$$\begin{aligned} i_3 &= A\cos(\omega_e t + 2\pi/3) + B\cos(3\omega_e t) \\ &+ C\cos(5\omega_e t - 2\pi/3) + D\cos(7\omega_e t + 2\pi/3) \end{aligned} \qquad (4)$$

By these two phases' currents, the linkage flux to the other phase can be represented as (5).

$$\begin{aligned} \lambda_{1\_3ph} &= (-\frac{1}{2}L_m)i_2 + (-\frac{1}{2}L_m)i_3 \\ &= \frac{1}{2}L_m \begin{bmatrix} A\cos(\omega_e t) - 2B\cos(3\omega_e t) \\ +C\cos(5\omega_e t) + D\cos(7\omega_e t) \end{bmatrix} \end{aligned} \qquad (5)$$

where If λ_{1_3ph} stands for flux linkage to the phase by other two phases. Equation (5) indicates that harmonics are added to the phase when other phases contain harmonic components of current.

ii. 6 phase system

Through the same process as in 3 phase system, the linkage flux to a phase by other five phases can be derived.

$$\begin{aligned} i_2 &= A\cos(\omega_e t - \pi/3) + B\cos(3\omega_e t - \pi) \\ &+ C\cos(5\omega_e t + \pi/3) + D\cos(7\omega_e t - \pi/3) \end{aligned} \qquad (6)$$

$$\begin{aligned} i_3 &= A\cos(\omega_e t - 2\pi/3) + B\cos(3\omega_e t) \\ &+ C\cos(5\omega_e t + 2\pi/3) + D\cos(7\omega_e t - 2\pi/3) \end{aligned} \qquad (7)$$

$$\begin{aligned} i_4 &= A\cos(\omega_e t - \pi) + B\cos(3\omega_e t - \pi) \\ &+ C\cos(5\omega_e t - \pi) + D\cos(7\omega_e t - \pi) \end{aligned} \qquad (8)$$

$$\begin{aligned} i_5 &= A\cos(\omega_e t + 2\pi/3) + B\cos(3\omega_e t) \\ &+ C\cos(5\omega_e t - 2\pi/3) + D\cos(7\omega_e t + 2\pi/3) \end{aligned} \qquad (9)$$

$$\begin{aligned} i_6 &= A\cos(\omega_e t + \pi/3) + B\cos(3\omega_e t - \pi) \\ &+ C\cos(5\omega_e t - \pi/3) + D\cos(7\omega_e t + \pi/3) \end{aligned} \qquad (10)$$

$$\begin{aligned} \lambda_{1\_6ph} &= (0.5L_m)i_2 + (-0.5L_m)i_3 \\ &+ (-L_m)i_4 + (-0.5L_m)i_5 + (0.5L_m)i_6 \\ &= \frac{1}{2}L_m \begin{bmatrix} 4A\cos(\omega_e t) - 2B\cos(3\omega_e t) \\ +4C\cos(5\omega_e t) + 4D\cos(7\omega_e t) \end{bmatrix} \end{aligned} \qquad (11)$$

By comparing (5) and (11), it is clear that 3rd, 5th and 7th components in the linkage flux grows absolutely as the number of operating phases increases. Since the PR controller implemented in this system takes control of only fundamental components, other harmonics are not able to be regulated.

In conclusion, when the operating phases increase, absolutely more harmonics are included in the linkage flux and this distorted linkage flux induces more harmonics in the phase current.

C. *Reduction of Harmonics in Phase Currents*

Through the reasoning in the previous sections, it is deduced that harmonics in phase currents are inherently from the motor and its drive system. Hence, to remove these undesirable harmonic components, PR controllers for each harmonic can be added in parallel. Fig. 9 depicts one of the possible structures of the controller to regulate components from 3rd to 9th harmonics.

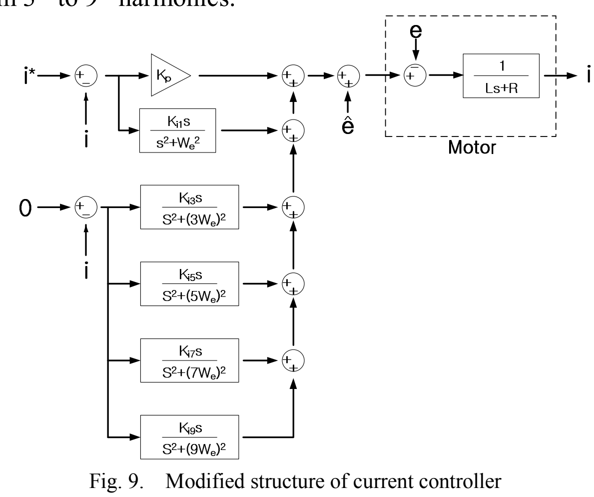

Fig. 9. Modified structure of current controller

As only fundamental component should be meaningful, reference for other harmonics is set zero. If higher order harmonics appear in the phase currents, another PR controller corresponding to the harmonic can be added. When implementing PR controllers for high order harmonics, phase delay in voltage synthesis should be considered. For example, in 30 Hz operation, phase delay of voltage for 11th order harmonic is approximately 330[Hz] x 150[us] = 17.82 [degrees] under the assumption of constant 10 kHz PWM switching operation. Therefore, phase compensators are also needed for proper voltage synthesis. Transfer function of the phase compensator is given as (s-a)/(s+a) and the variable 'a' is calculated to be 325.0646 for compensating 11^{th} order PR controller in 30 Hz operation. Fig. 10 and Fig. 11 show the block diagram of the controller with a phase compensator and its bode plot characteristics.

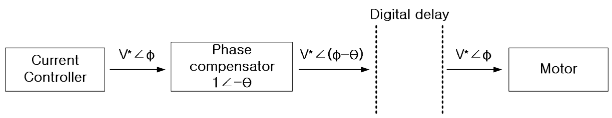

Fig. 10. Block diagram of controller with a phase compensator

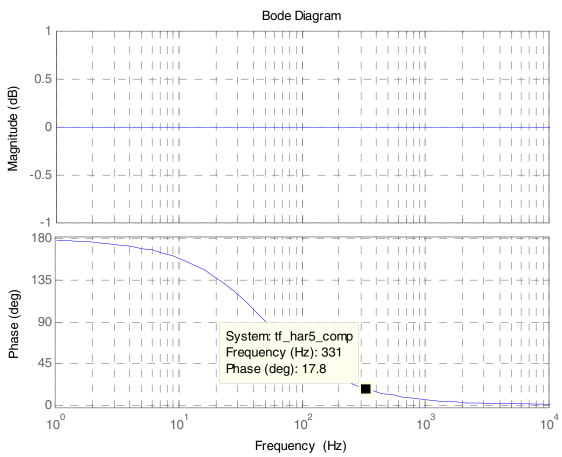

Fig. 11. Bode plot of phase compensator for 11^{th} order PR controller

By utilizing PR controllers for each harmonic and its phase compensators, sinusoidal currents are injected to all the 24 phases. Fig. 12 displays phase current of 24 phase operation at 30 Hz operation, running at 900 r/min and it is confirmed that all the harmonics except fundamental component are less than 3%.

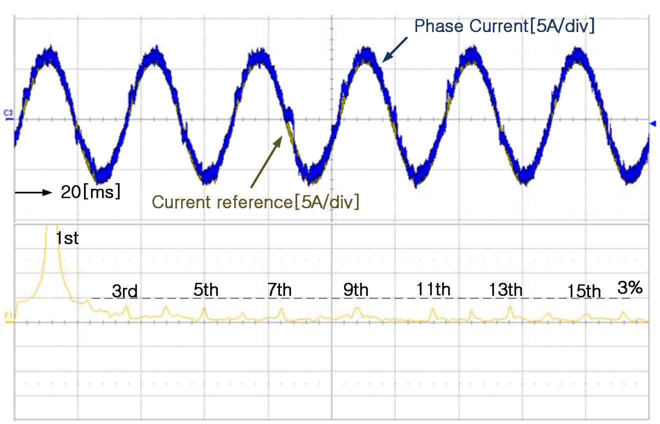

Fig. 12. Phase current in 24 phase operation at 30 Hz operation, 900 r/min

V. TORQUE RIPPLE ANALYSIS

To identify the torque ripple reduction effect of 12 phase system, comparisons between different numbers of operating phases have been made. A torque sensor has been installed to measure output torque and its analog signal is investigated based on FFT. The motor has been run at 600 r/min, equivalent fundamental frequency of 20Hz. Fig. 13 displays the experimental setup and Fig. 14 through Fig. 16 show the output signal of torque sensor for three, six and twelve phase operation, respectively. Actual value in y axis is in voltage but it reflects the torque ripple. As expected, smoother torque is produced as the operating phase numbers increase judging by the analog output of torque sensor.

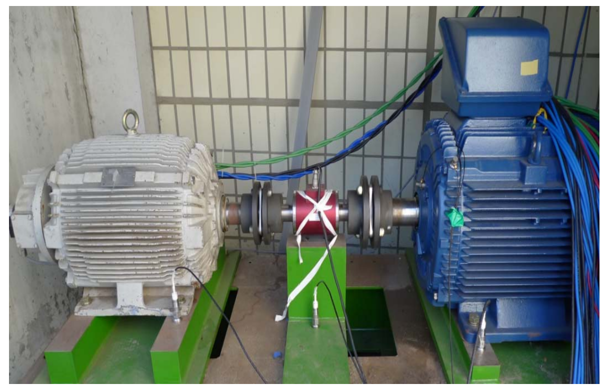

Fig. 13. Experimental setup

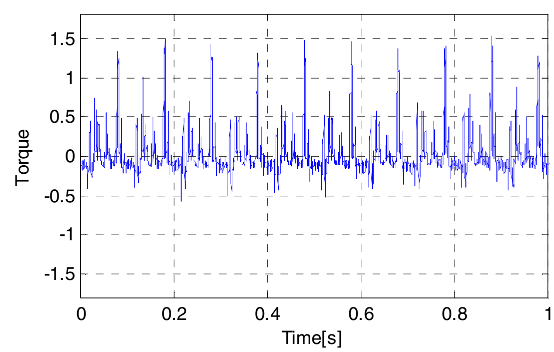

Fig. 14. Torque in three phase operation

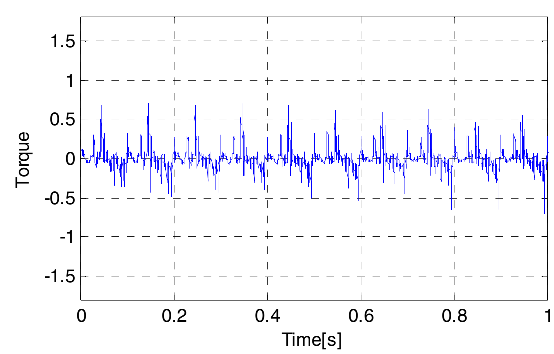

Fig. 15. Torque in six phase operation

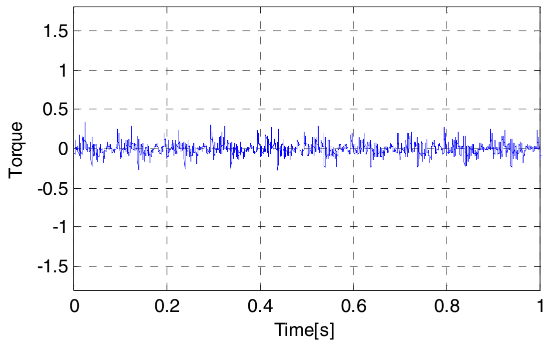

Fig. 16. Torque in twelve phase operation

VI. CONCLUSION

This paper has discussed a separate double-winding 12-phase brushless DC motor drive. No neutral point exists and each phase is isolated electrically. Stators are doubly wound for duplicate redundancy and individual H-bridge inverters are employed so that the system is still able to operate continuously with healthy phases under fault conditions. In addition, single inverter module can be detached from other modules for repair or replacement in case of any fault in the inverter module.

In terms of motor drive, sinusoidal currents are injected to all the phases, which shows similar torque performance as with square wave currents. High order harmonics generated from inherent motor characteristics are removed by additional PR controllers in parallel to the fundamental controller.

Conspicuous reduction of torque ripples has also been verified by increased number of phases with asymmetrical distribution of magnetic winding through simulations and experiments with a prototype motor.

Moreover, to improve the efficiency of the system, high efficiency mode operation is introduced at low speed region. This mode can lead to switching loss reduction from decreased number of active legs. It also improves voltage syntheses performance of inverters and therefore shows better achievement in torque production at low speed.

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