Design And Development Of 3-Phase Variable Frequency Drive

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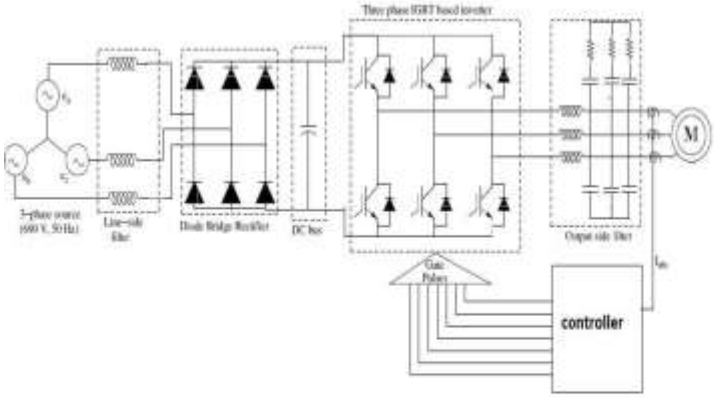
Abstract : With the change in scenario of continuous thrust for the improvement of efficiency of the overall system, motors are being coupled with variable frequency drives (VFD) to achieve energy savings, reduced wear and tear on motors, reduction of maintenance cost of mechanical flow controls etc. To meet the market requirements in the low voltage segment, 1MW, 690V ac drive has been developed. The system was successfully tested for V/f control, sensor-less vector control, power on ride through and catch on fly. Various issues related to testing of the machine with open loop v/f control and sensor less vector control were addressed successfully. Load testing of the 1MW VFD also completed with back to back test facility. This paper presents the development and testing of the developed 1MW VFD with the experimental results**.**

I. Introduction

The simple structure and robustness of the induction machine has made it the most preferred machine for industrial applications. However, when directly operated from the 50Hz AC grid this machine operates only at speeds near the synchronous speed. For applications requiring the variable speed, variable frequency drive is invariably required. It is also advantageous to use Variable Voltage Variable Frequency (VVVF) drives for greater efficiency, as compared to schemes which use mechanical gates and dampers for controlling the machine speed. Now a days a VVVF drive is commonly realized using IGBT based three-phase voltage source inverter (VSI) whose output voltage & frequency is controlled appropriately by controlling the switching pulses of IGBTs.To meet the market requirements in the low voltage segment, 1MW, 690V ac drive has been developed. The single line diagram of developed 1MW LV VFD is shown in Fig. 1. The power circuit comprises of input filter, three phase diode bridge rectifier, DC link, three phase inverter and output filter. System studies, design of IGBT based Voltage Source Inverter, input & out filter have been completed and verified with simulation studies using MATLAB / Simulink. Control algorithm for V/f control, sensor-less vector control, catch on fly and power on ride through have been developed and verified using simulation studies.

II. DEVELOPMENT OF 1MWVFD

Fig. 1: Single line Diagram of 1MW LV VFD



The power structure of the drive used is as shown in Fig. 1. The three phase balanced supply from the transformer is rectified by a three phase diode bridge rectifier to obtain the required DC bus voltage. A precharging circuit consisting of a resistor in parallel with a relay or contactor is used to prevent large inrush currents that may damage the DC bus capacitor. The relay bypasses the resistor from the circuit when the capacitor voltage has built up to a large enough value. The rectifier normally draws distorted and peaky currents from the supply and hence an inductance is used in series with the line to make the currents smoother. The rectified output is then given to the inverter whose switching signals are controlled suitably to obtain the required voltage. Wide bandwidth, high precision, Hall Effect sensors are used to sense the instantaneous line currents for feedback. The DC bus voltage is also sensed by using a precision Hall Effect sensor. The controller is a Digital Signal processer (DSP) from Texas Instruments. The control algorithm running on the processor senses the feedback signals through the built-in Analog-Digital Converter (ADC) on the controller and processes these information to generate the required switching pulses. The converter is switched at a frequency of 1.5 kHz. The inverter output voltage is again filtered by using an output LC filter with a split capacitor topology. Active damping technique is used to provide lossless damping to the resonant frequency oscillations [1]. Thermal design was carried out for 1MW

AC drive using semisel software. Solid state relays were used for the implementation of start-stop logic of the drive. Developed AC VFD can be seen in Fig.2.

Fig. 2: Developed Panel of 1MW VFD

Fig. 3 Inside View of 1MW VFD

Fig3. Gives the inside view of 1MW VFD. All the control electronics along with power supplies, Solid state relays, TB’s can be seen here.

I. IMPLEMENTATION OF SENSORLESS CONTROL

A. *Theory of Sensor less Vector control*

Vector Control or Field Oriented Control of induction machines refers to a high performance drive control scheme where decoupled torque and flux control similar to a separately producing current (armature current) are supplied to two separate windings from independent sources. As a result, the speed control of DC machines can be easily achieved by controlling these two components independently. However, in induction machines, there is only a single three phase winding excited from a three phase source. Hence it is difficult to control the two sources independently. In vector control technique, the torque producing and flux producing components of the input current are identified and controlled separately in a similar fashion to that of a separately excited DC machine. This can be achieved by transforming the instantaneous currents to a rotating frame of reference fixed to the rotor flux. This requires a knowledge of the instantaneous position of the rotor flux in the machine. This is obtained by using a shaft encoder and the machine model equations in sensored vector control. In a sensor less control, the rotor flux position is estimated from the knowledge of line currents and voltages only thereby avoiding the encoder [2], [3].

The basic theory of vector control can be appreciated by considering the dynamic model of the machine in the rotor flux frame of reference. The equations of the induction machine in this reference frame are given as:

$$\begin{aligned} i_{sd} &= i_{mr} + \tau_r \frac{di_{mr}}{dt} \\ \omega_{mr} &= \omega + \frac{i_{sq}}{\tau_r i_{mr}} \\ M_d &= K_t i_{mr} i_{sq} \end{aligned}$$

Where

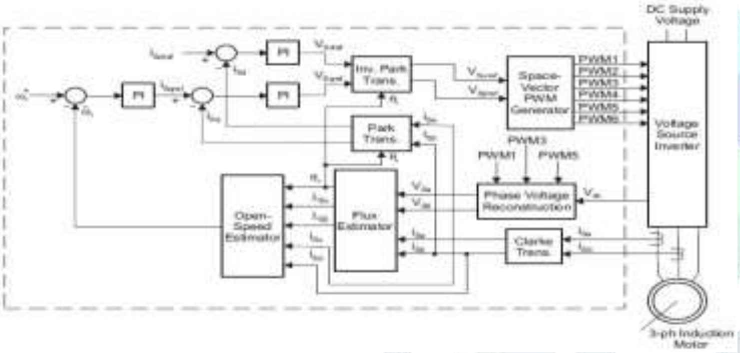
isd - instantaneous d-axis stator current isq - instantaneous q-axis stator current

imr - instantaneous rotor magnetizing current wmr - instantaneous rotor flux speed in elect.rad/s w - Instantaneous rotor speed in elect.rad/s

tr - rotor time constant

Md - instantaneous developed electromagnetic torque Kt - a constant of the machine

B. *Controller for sensor-less vector control*



C. *Estimations for sensor-less vector control*

Accurate information about the rotor flux position is critical for proper decoupled control. Also speed control of the drive requires a feedback of the instantaneous speed of the rotor. In a typical Sensored vector control scheme, an encoder is used to obtain these information. However, the use of an encoder has certain drawbacks as discussed. Hence a sensorless scheme is preferred.

In a sensorless FOC scheme, the speed and the flux position needs to be estimated. The dynamic performance of the sensorless vector controlled drive depends on the accuracy of the estimators. The most basic approach towards estimating the rotor flux is by integrating the stator voltage to estimate the stator flux and then calculating the rotor flux from the stator flux.

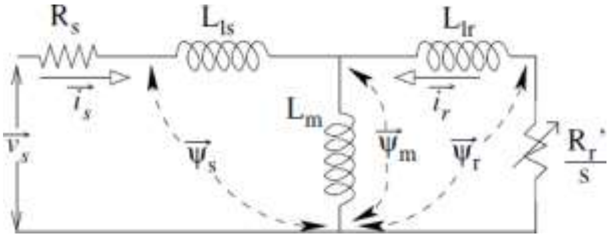


Fig. 5: Induction Machine equivalent circuit

From the equivalent circuit, the stator voltage can be written as

$$\vec{v_s} = R_s\vec{i_s} + \frac{d\vec{\psi_s}}{dt}$$

Therefore the stator flux space vector can be obtained by simply integrating the back emf. This is represented as

$$\vec{\psi_s} = \int (\vec{v_s} - R_s\vec{i_s})dt$$

Also from the equivalent circuit,

$$\begin{aligned} \vec{\psi_s} &= L_s\vec{i_s} + L_m\vec{i_r} \\ \vec{\psi_r} &= L_r\vec{i_r} + L_m\vec{i_s} \end{aligned}$$

$$\vec{\psi_r} = \frac{L_r}{L_m}\vec{\psi_s} - \frac{L_sL_r - L_m^2}{L_m}\vec{i_s}$$

This equation can be written in terms of Alpha-beta components.

$$\begin{aligned} \vec{\psi_{r\alpha}} &= \frac{L_r}{L_m}\vec{\psi_{s\alpha}} - \frac{L_sL_r - L_m^2}{L_m}\vec{i_{s\alpha}} \\ \vec{\psi_{r\beta}} &= \frac{L_r}{L_m}\vec{\psi_{s\beta}} - \frac{L_sL_r - L_m^2}{L_m}\vec{i_{s\beta}} \end{aligned}$$

Knowing the alpha and beta components of the rotor flux, the instantaneous angular position and the flux speed can be estimated as

$$\begin{aligned} \rho(t) &= \tan^{-1}\left(\frac{\psi_{r\beta}}{\psi_{r\alpha}}\right) \\ \omega_{mr} &= \frac{d\rho(t)}{dt} \end{aligned}$$

Once the flux speed is calculated, rotor speed can be calculated by using the slip speed computation. A block diagram of the basic structure of the flux estimator is shown in the Fig.4.

Although the structure of the estimator looks relatively simple, there are some problems associated with a practical implementation. Any practical implementation of an integrator will have some offset at the input which causes the output to drift away with time. This problem is mitigated to some extent by using a low pass filter with a very low cutoff frequency, instead of a pure integrator. This scheme works well at high speeds, but at low speeds the low pass filter causes significant phase and magnitude errors. Also if there is a slight error in the stator resistance value used in the estimator, then at low frequencies the error in estimated resistive drop becomes comparable to the voltage applied to the motor. This causes significant errors in the estimated flux magnitude and position at low frequencies. Also a direct differentiation of the estimated angle causes large spikes in the estimated speed due to the presence of noise in the estimated value. Thus the basic estimator scheme needs to be modified to achieve good performance over a wide speed range.

In flux estimation scheme, the pure integrator is replaced by an integrator with a small amount of negative feedback. Without any negative feedback, the estimated flux obtained by pure integrator can create destabilization due to a variety of errors that may become particularly detrimental at low stator frequencies. These include measurement noise, parameter detuning and input dc offsets. The effect of the feedback is to stabilize the integrator output against these errors, particularly for DC offset errors which can lead to integrator saturation. The modified scheme can be shown to be essentially equivalent to a pure integrator followed by a high pass filter. The structure thus has a low pass filter characteristic with the pole close to zero. This scheme provides good accuracy and stability for speeds from rated and down to about 2 Hz stator frequency. Below 2 Hz, the phase and magnitude errors introduced by the low pass filter becomes significant and degrades the performance.

D. Estimations for sensor-less vector control

The information about the instantaneous speed of the machine is essential for doing speed control. In sensorless vector control, the speed is estimated using the dynamic model equations of the induction machine in the rotor flux frame of reference. The synchronous speed (wmr), is estimated from the estimated values of sinβ and cosβ obtained from the flux estimation scheme, thereby avoiding the need for an inverse tan operation and direct differentiation of the angle β(t).

The relevant equations are given as,

$$\begin{aligned}\omega_{mr} &= \frac{d\rho}{dt} = \frac{d\rho}{dt}[\cos^2\rho + \sin^2\rho] \\ &= \cos\rho\Big(\cos\rho\frac{d\rho}{dt}\Big) + \sin\rho\Big(\sin\rho\frac{d\rho}{dt}\Big) \\ &= \cos\rho\frac{d}{dt}[\sin\rho] - \sin\rho\frac{d}{dt}[\cos\rho]\end{aligned}$$

Then the rotor speed in electrical rad/sec is given as

$$\omega = \omega_{mr} - \omega slip = \omega mr - \frac{isq}{\tau_r i_{mr}}$$

It should be noted that the speed estimation still involves a digital differentiation of the quantities, sinϸ and cosϸ. Thus it may be required to filter the estimated values of sinϸ and cosϸ using low pass filters of appropriate cut off frequency so as to reduce the effect of differentiating the noise. Also the estimated speed might need low pass filtering before giving it as a speed feedback, depending on the bandwidth of the speed controller. Digital filters of cut-off frequency 500 Hz are recommended for filtering the estimated values of sinϸ and cosϸ and 5 Hz for filtering the estimated speed.

II. POWER ON RIDE THROUGH WITH CATCH ON FLY OPERATION

AC drive applications used in critical industrial processes often require the drive to run smoothly inspite of voltage dips or momentary power interruptions. The drive should also be able to continue operation as soon as the power supply is restored. This feature of the drive is known as ’ride-through with on-the-fly start’ capability. When the drive is in operation, there is some amount of electrical energy stored in the DC bus capacitor and some mechanical energy stored in the inertia of the rotating shaft. When a supply interruption occurs this stored energy is used in feeding the load torque and the losses in the system. This causes the DC bus voltage as well as the shaft speed to fall with time. If the capacitor voltage falls below the undervoltage limits, the capacitor precharging circuit comes into picture to prevent large inrush currents, and on-the-fly starting is then not possible. The maximum time of ride through at a given load, depends on the losses occurring in the system. Several schemes for ride through by energy storage using large capacitor banks and super capacitors have been discussed in literature. However, these methods can provide short duration ride through only such as that for a short duration voltage sag. In vector controlled drives, the DC bus can be kept charged even during a power interruption by regeneration.

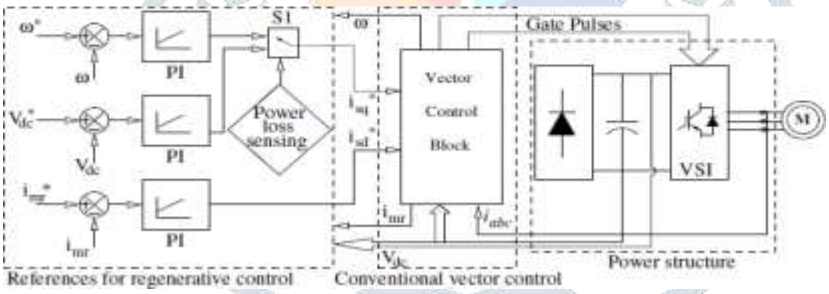


Fig. 6: Power on Ride through with catch on fly

In this method [6], a small amount of energy from the rotating shaft is recovered by regeneration, to maintain the DC bus voltage. The developed torque, Md of the machine can be expressed as Md = Kt*imr*isq, where imr is the rotor magnetizing current, isq is the q-axis stator current and Kt is the torque constant of the machine. Thus the machine can be made to regenerate by making the value of isq negative, thereby feeding power into the DC bus terminals of the inverter, to maintain the capacitor voltage. Therefore, during supply interruptions, isq can be made negative in a controlled manner by using a PI controller that monitors the DC bus voltage and controls the reference value of the q-axis current. Fig.6 shows the control block diagram of the drive. S1 is a selection switch. During normal operation, the position of the switch is so as to select the output from the speed controller as the q axis reference current. When a supply interruption occurs the q-axis reference will change to the DC bus voltage controller output.

III. TESTING OF 1MW LV VFD WITH 25HP INDUCTION MOTOR

The assembled 1MW VFD was tested for continuity checks for power and control hardware. All the protections were verified, then 1MW VFD was tested with 25HP Induction Machine using V/f Control as well as Sensorless Vector control.

A. *V/f Control*

1MW VFD was successfully tested with 25HP induction

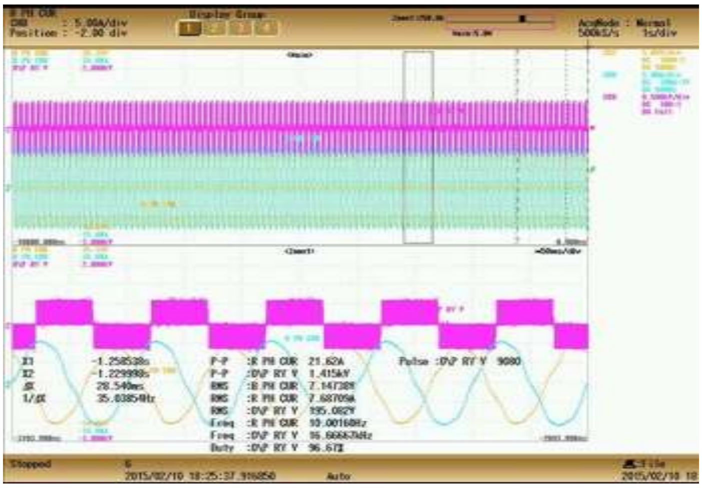


Fig.7: V/f control at 10Hz with 25HP motor

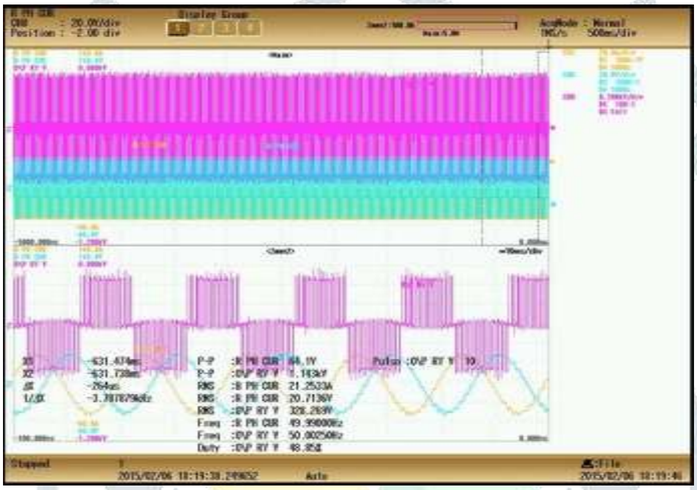


Fig. 8: V/f control at 50Hz with 25HP motor

B. *Sensor less Vector Control*

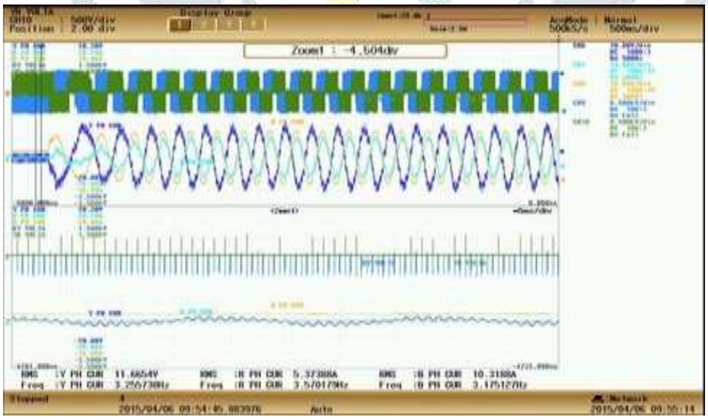


Fig.9: Starting transient of the machine from zero speed with 25HP motor

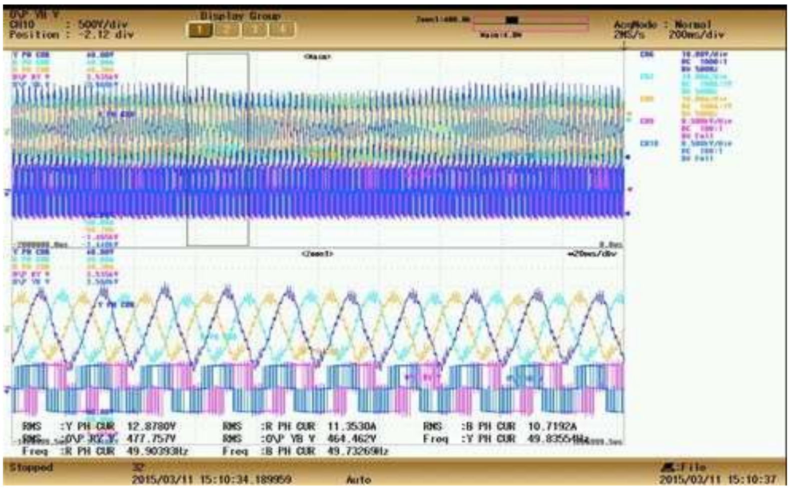


Fig.10: With the Speed Reference as -1.0 pu with 25HP motor

C. *Power on Ride Through with on the fly start*

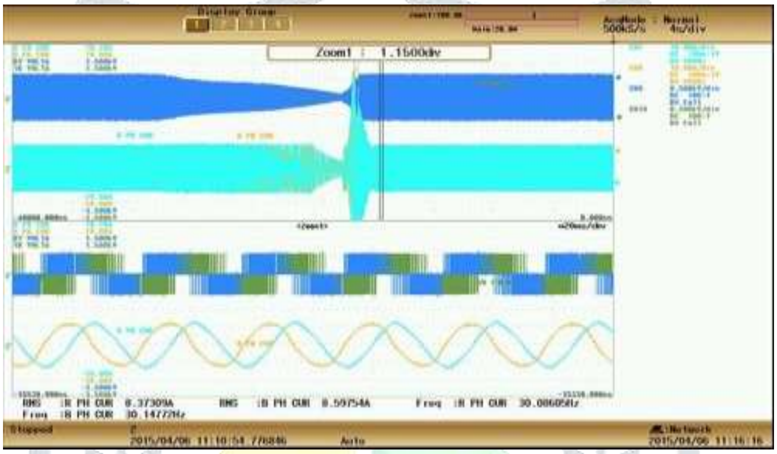


Fig.11: Ride through with catch on fly with 25HP motor

IV. TESTING OF 1MW LV VFD WITH 875KW INDUCTION MACHINE

Developed 1MW VFD was also tested with 875KW induction motor which is also developed by BHEL. Initially, machine was tested on no-load for both v/f and sensorless vector control.

A. *Sub-Harmonic Oscillations with V/f Control*

While testing with 875KW induction machine on no-load with V/f control, it was observed that current waveform is distorted. Test conditions are input voltage 690V, switching frequency 1.5 KHz, and Dead band as 12 us. Refer to Fig. 12 & 13 for the practical waveforms.

This is because, in a practical voltage source inverter (VSI), the top and bottom transistors in a leg are not switched in an exact complementary fashion. A dead-time duration, where the gettinng signals to both the top and bottom devices are low, is introduced for safe commutation. This dead-time results in harmonic distortion and certain change in the fundamental voltage [4]. This has also been reported to cause sub-harmonic oscillations (in the motor current, torque and speed) in certain motors under light-load conditions [4].

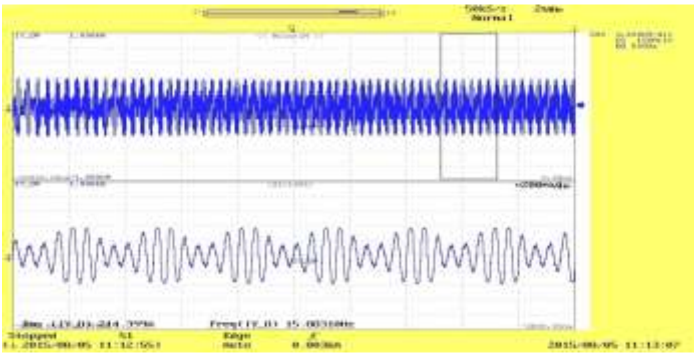


Fig.12: Motor current on no-load with v/f control- frequency 15Hz

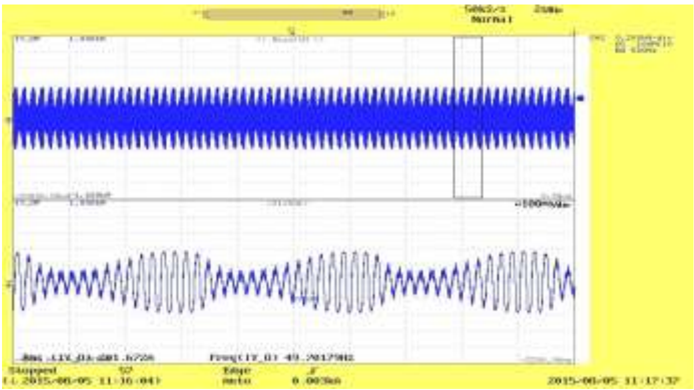


Fig.13: Motor current on no-load with v/f control- frequency 50Hz

From 1Hz-50Hz, sub-harmonic currents were observed and the effect is much worse at lower frequencies.

To solve this problem, Lower v/f ratio (0.8), reduced dead band (3.2 us) was adopted.

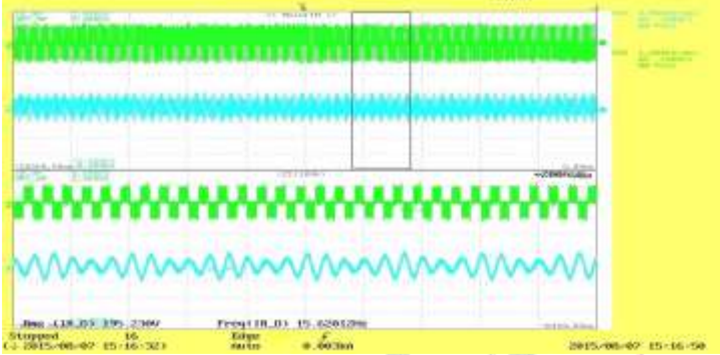


Fig.14: Motor current with reduced dead band- frequency 15Hz

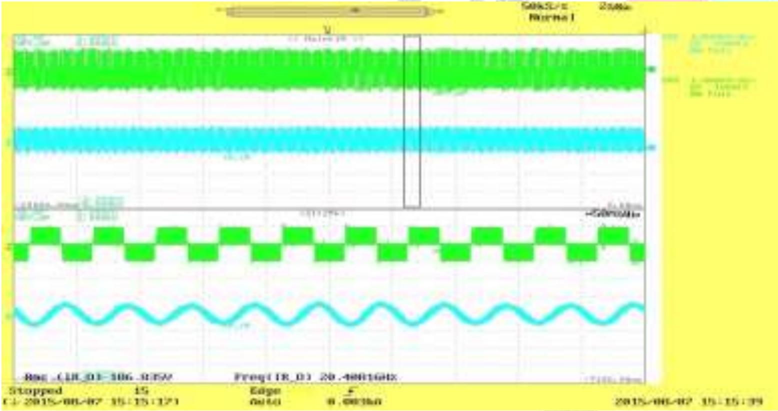


Fig.15: Motor current with reduced dead band- frequency 30Hz

It can be observed that current waveform is pure sinusoidal at 30Hz. To solve the problem at lower frequencies, dead band compensation schemes is adopted.

B. Sensor-less Vector control

Sensor-less vector control of 875KW machine on no-load is also completed.

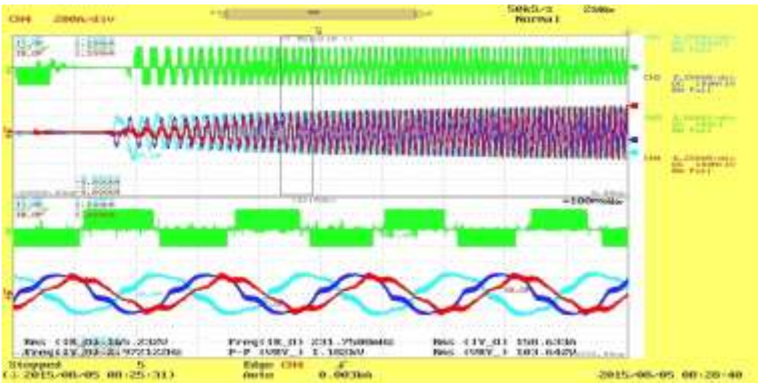


Fig.16: Starting Transient with sensorless vector control

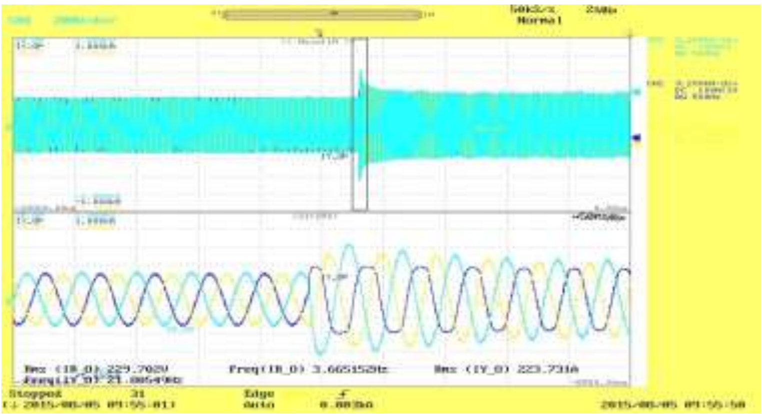


Fig.17: Motor currents with speed reference from 500 to 1000 RPM

C. *Back to Back testing for 700KW loading*

As per the testing arrangement mentioned in Fig.18, Induction machine was loaded upto 700KW.

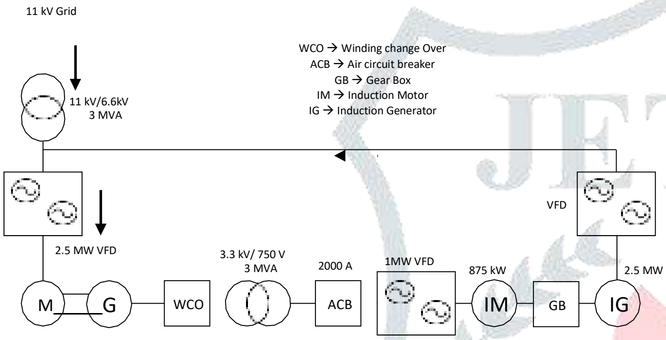


Fig.18: Back to back testing arrangement

Initially load motor was started and once load motor picked the full speed, source motor rotor also runs at the same speed of load motor because of the common shaft. Then source motor was energized with VFD at rated frequency. At this stage, both machines are running at rated speed but without loading. Then load motor, supply voltage frequency is reduced in the step by step manner to progressively load the machine. By doing this, 700KW loading was achieved.

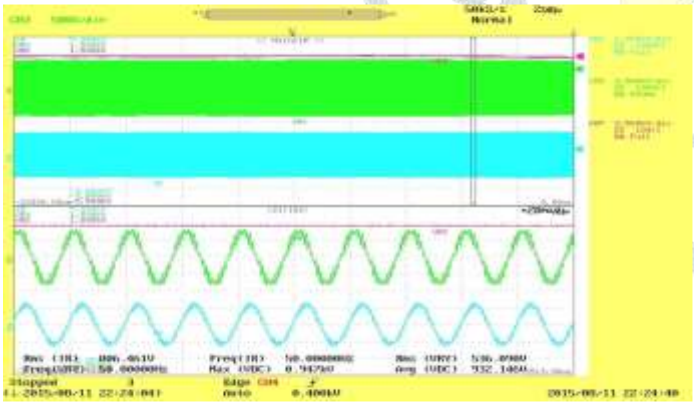


Fig.19: Back to back testing with 800A loading

Beyond 700KW loading, source voltage was fluctuating very widely since there was no AVR. Hence could not proceed beyond 700KW. Refer to Fig.20 for the source voltage fluctuation.

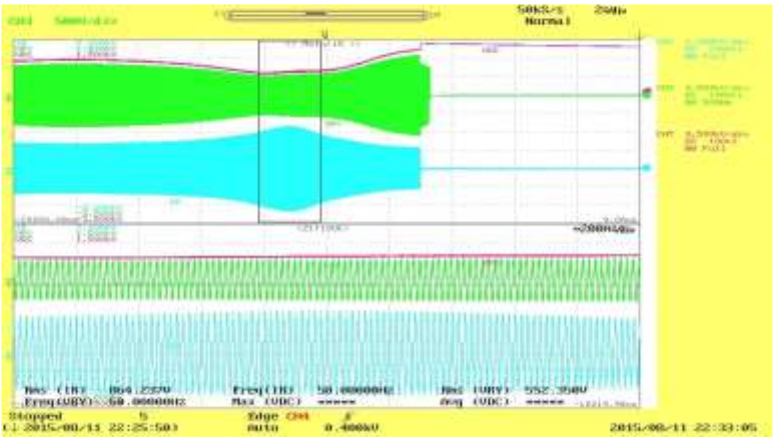


Fig.20: Back to back testing with 864A loading

V. CONCLUSIONS

In this paper, development and testing of 1MW VFD is discussed in detail. This paper has reported various issues related to high power testing of induction motor drives. Issues like sub-harmonic oscillations, back to back testing of drives were also discussed.

REFERENCES

1. K. Hatua, A.K. Jain, Banerjee. D, Ranganathan.V. T., Active Damping of Output LC Filter Resonance for Vector-Controlled VSI-Fed AC Motor Drives. Industrial Electronics, IEEE Transactions on, vol.59, no.1, pp.334, 342, Jan. 2012.
2. V.T. Ranganathan, Course notes on Electric Drives , IISc Bangalore
3. R. Krishnan. Electric motor drives: modeling, analysis, and control. Prentice Hall PTR, 2001.