Apparatus and method for heat-run test on high-power PWM converters with low energy expenditure

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MS received 13 June 2011; revised 1 January 2013; accepted 5 April 2013

Abstract. Before installation, a voltage source converter is usually subjected to heat-run test to verify its thermal design and performance under load. For heat-run test, the converter needs to be operated at rated voltage and rated current for a substantial length of time. Hence, such tests consume huge amount of energy in case of high-power converters. Also, the capacities of the source and loads available in the research and development (R&D) centre or the production facility could be inadequate to conduct such tests. This paper proposes a method to conduct heat-run tests on high-power, pulse width modulated (PWM) converters with low energy consumption. The experimental set-up consists of the converter under test and another converter (of similar or higher rating), both connected in parallel on the ac side and open on the dc side. Vector-control or synchronous reference frame control is employed to control the converters such that one draws certain amount of reactive power and the other supplies the same; only the system losses are drawn from the mains. The performance of the controller is validated through simulation and experiments. Experimental results, pertaining to heat-run tests on a high-power PWM converter, are presented at power levels of 25 kVA to 150 kVA.

Keywords. Voltage source converter; pulse width modulated converter; pulse width modulation; high-power converter; heat-run test; energy conservation.

1. Introduction

A three-phase pulse-width modulated (PWM) voltage source converter (VSC) is shown schematically in figure 1. In this converter, the power could flow either from the dc side to the ac side, or in the opposite direction (i.e., ac to dc). Due to its bi-directional power flow capability, this is used in a wide variety of applications, such as un-interruptible power supplies (UPS), static

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Figure 1. Schematic diagram of a voltage source converter (VSC).

compensators (STATCOM), active rectifiers or front-end converters (FEC), distributed generation from renewable energy sources, and motor drives (Rashid 2001).

In UPS and motor drive applications, the three-phase load (or motor) is connected on the ac side. In case of UPS, the VSC converts dc into fixed frequency ac (50 or 60 Hz). In case of motor drives, dc voltage is converted into a variable-voltage, variable-frequency ac supply (Rashid 2001).

Conventionally, diode bridges or thyristor converters are used for rectification. In this case, the line current drawn from the mains is highly distorted, and the power factor is poor. A VSC can be used as an active PWM rectifier (figure 2) for ac to dc conversion, ensuring high power factor and low current distortion. The ac terminals of the converter are connected to the grid through filter inductors as shown in figure 2. The principle of operation of a unity power factor (UPF) rectifier is illustrated by the phasor diagram in figure 3a.

Figure 2. Schematic diagram of a front-end converter (FEC).

Figure 3. Phasor diagram of a grid-connected VSC (a) operation at unity power factor; (b) operation at an arbitrary power factor angle.

If the converter has to draw certain amplitude of fundamental current (I_{R1}) at UPF from the mains, then the fundamental component of the converter terminal voltage (V_{R1N}) should be as indicated in figure 3a. The harmonic components of the terminal voltage of the converter are filtered by the line inductors, and hence the line current is nearly sinusoidal (Siva Prasad *et al* 2008).

More generally, the converter can draw line currents at any power factor through suitable control of the terminal voltage V_{R1N} as illustrated in figure 3b. The line current is made to lead the supply voltage by 90° to supply reactive power to the mains in case of static VAR compensator (STATCOM) (Sensarma *et al* 2001). If the current is out of phase ($\theta = 180^\circ$) with the supply voltage, then the power flow is from the dc to the ac side. This mode is useful in grid-connected photovoltaic converters (Kadri *et al* 2011).

In grid-connected wind energy applications, two back-to-back PWM converters with a common dc bus are generally used (Portillo *et al* 2006). The machine-side converter converts the variable-frequency, variable-voltage ac generated by the wind turbine driven alternator into dc voltage. The dc voltage is converted back to line frequency ac by the second converter, and power is fed into the grid. Similar topology is useful in four-quadrant motor drives, where the power flow is from the grid to the machine during motoring mode and from the machine to the grid during regeneration (Rashid 2001).

On account of a wide range of applications, use of VSC is widespread in power electronic industry. Also, the available ratings of insulated gate bi-polar junction transistors (IGBT) are fast increasing (Krug *et al* 2007). Currently IGBTs with blocking voltages of up to 6.5 kV are available (Hiller *et al* 2010). With these high-voltage IGBTs, the VSC topology can be used up to a few tens of megawatt (Hiller *et al* 2010).

In a PWM converter, there are losses in both IGBT and its anti-parallel diode due to conduction as well as switching action. The heat generated has to be transferred safely to the surroundings

through heat-sinks so that the semiconductor junction temperatures do not exceed their safe maximum values. To increase the rate of heat transfer, forced air cooling or liquid cooling is adopted at high power levels. The thermal system of the converter is designed to ensure that the junction temperatures are within safe limits under worst operating conditions (Bruckner & Bernet 2007).

To prevent device failures on account of thermal run-away (over-temperature), it is necessary to verify the thermal design of the converter at the development stage or before installation. Hence, the converters are usually subjected to heat-run tests. At higher power level, this requires a high-capacity source along with a huge resistor bank as load. These may not always be available at the development center or the production facility. Further, as the converter needs to be tested at rated voltage and rated current for a good length of time (several times the thermal time constant of the converter), enormous amount of energy is expended during the test, significantly adding to the development or production cost.

Numerous attempts have been made to reduce the energy consumed during such tests. A survey on existing heat-run test methods for high-power converters is presented in section 2. In section 3, the proposed test set-up and its advantages are presented. A control scheme for the two converters in the experimental set-up is proposed and validated in sections 4 and 5, respectively.

A few results pertaining to the control scheme have been presented in an earlier conference publication (Siva Prasad & Narayanan 2010). Extensive results pertaining to heat-run test on high-power converter (section 8) are presented in this paper. Further, the usefulness of this test method for space-vector modulated inverters is demonstrated here (section 6). Furthermore, the influence of carrier interleaving on the quality of the input current waveform is also studied (section 7).

2. Existing test methods

Numerous attempts have been made to reduce the energy consumption during the testing phase of PWM converters (Siva Prasad & Narayanan 2010; Forest *et al* 2006; Oates 1999; Bryant *et al* 2007; Huang & Pai 2002; Lin *et al* 1997; Tsai 2000; Zhou *et al* 2008; She *et al* 2009; Ghosh 2007). The basic approach is to connect a tested converter (or a set of converters) with the converter under test in a back-to-back fashion; the two (sets of) converters are controlled such that rated power circulates through them with only the losses being drawn from the power source. As shown in figure 4, the two back-to-back connected converters can be fed from a dc supply, which itself may be derived from ac mains through rectification (Bruckner & Bernet 2007; Forest *et al* 2006; Oates 1999; Bryant *et al* 2007). Alternatively, the two (sets of) converters could be fed directly from the ac mains as illustrated in figure 5 (Huang & Pai 2002; Lin *et al* 1997; Tsai 2000; Zhou *et al* 2008; She *et al* 2009; Ghosh 2007).

Figure 4. Back-to-back connected converters fed from a dc supply.

Figure 5. Back-to-back connected converters fed from ac mains.

In the testing arrangement shown in figure 4, line voltage of required magnitude and frequency is produced by modulating converter 1 in an open loop fashion. An opposing voltage is produced by converter 2, which is controlled in a closed loop fashion. Hence this method is sometimes referred to as ‘opposition method’ (Forest *et al* 2006). The amplitude and phase of the opposing voltage are chosen to ensure that the line currents have the desired amplitude and phase as illustrated in figure 3b.

In this method, the entire dc bus voltage gets applied across the line inductor (L) during certain switching intervals. The line-inductors being low valued, particularly at high current ratings, the current shoots up very fast and needs to be limited. Therefore specially designed control strategies are required for testing the converters (Bruckner & Bernet 2007; Forest *et al* 2006; Oates 1999; Bryant *et al* 2007).

On the other hand, standard and well-known control methods for grid-connected converters can be adopted for the testing arrangement shown in figure 5. The term ‘converter system’ is used here to include line reactors and transformer, if any. One of the two converters in figure 5 is the PWM converter under test (say converter system 1). Since this is essentially a three-phase ac–dc boost converter, the dc bus voltage is higher than the peak line–line voltage of the mains (Siva Prasad *et al* 2008). The other converter system may consist of a diode bridge rectifier, whose dc bus voltage is less than the peak line–line voltage (Siva Prasad *et al* 2008). Hence, to meet the dc voltage requirement, the diode bridge is either fed from a step-up transformer (Huang & Pai 2002; Lin *et al* 1997) or is followed by a dc–dc boost stage (Tsai 2000). Alternatively, the second converter could also be a PWM converter with rating similar to or higher than that of the converter under test (Ghosh 2007).

The converter system 2 can simply be another PWM converter of rating similar to or higher than that of converter system 1 (Ghosh 2007). While a large amount of real and/or reactive power (close to the rated kVA of the converters) is circulated between the converters, only a small amount of power corresponding to the losses in the system is drawn from the mains. As the converters are connected on both ac and dc sides, circulating currents flow between them (Ghosh 2007).

The issue of zero-sequence circulating currents between the converters has been studied extensively in the context of parallel-connected ac–dc PWM converters with a common dc bus, sharing a dc load (Ye *et al* 2002; Pan & Liao 2007; Neacsu *et al* 2008; Chen 2009; Cai *et al* 2008; Asiminoaei *et al* 2008; Mao *et al* 2011; Zhang *et al* 2010, 2011). The similarity between the parallel-connected converters and back-to-back connected converters can be seen easily. In both the cases, the converters are connected on the ac as well as on the dc sides. Hence, the study on circulating currents and its mitigation in the context of parallel-connected converters is equally applicable to the back-to-back connected converters and *vice versa*.

Converters are increasingly operated with space vector PWM (SVPWM) due to better utilization of dc bus than sine-triangle PWM (SPWM) (Zhou & Wang 2002). For a given dc bus voltage,

Figure 6. Schematic diagram of the proposed experimental set-up.

SVPWM can produce a higher line-side voltage than SPWM. Conversely, for a given line-side voltage, the converter could be operated with a reduced dc bus voltage in case of SVPWM. The losses in the converter are functions of both the PWM method employed and dc bus voltage. Hence, if a converter is intended to be operated with SVPWM at a particular dc bus voltage, then its thermal testing should be carried out with the same modulation method and at the same dc bus voltage for the results to be realistic.

SVPWM involves addition of common-mode voltage to the three-phase sinusoidal modulating waves (Zhou & Wang 2002; Varma & Narayanan 2006). If the converters in figure 5 are modulated using this method, then excessive circulating currents would flow between them owing to the common-mode voltage injected. Hence, the configuration in figure 5 is not suitable for conducting heat-run test on converters operated with SVPWM (Ghosh 2007).

In the following section, a new test method is proposed for conducting heat-run test on high-power voltage source converters including those modulated with SVPWM.

3. Proposed testing method

The proposed testing arrangement is shown in figure 6. Two high-power PWM converters are connected in parallel on the ac side to the grid (similar to figure 5). But on the dc side, their dc buses are separated and open. The converters are controlled such that the line current drawn by one of them lags the supply voltage by 90° , while the other converter draws an equal amount of current, leading the supply voltage by 90° . As the two currents are in phase opposition, they cancel each other (figure 7); the net current drawn from the grid is ideally zero. In practice, the net current is very low, and accounts mainly for the losses in the system. Hence, the energy consumed to conduct this test is very nominal.

With this method, both the converters can be tested (though only at zero power factor) at their rated voltages and currents, despite the absence of high-capacity source and load arrangements. Standard control methods for such line-side converters can be adopted (explained in section 4). Also, converters operated with modulation methods involving common-mode voltage injection such as SVPWM can be tested. Due to the above-mentioned advantages, the proposed test procedure is very much useful for practicing engineers responsible for testing and certifying high-power converters. Using the above mentioned test procedure, the performance of a 150 kVA line-side PWM converter is verified in this paper.

Figure 7. Phasor diagram of the test set-up.

converted into three-phase modulating signals. These modulating signals are compared with a common triangular carrier to generate gate pulses as shown.

A detailed discussion on modelling of the converter in the dq reference frame, unit-vector generation, and selection of controller parameters for the voltage and current loops is presented in Siva Prasad *et al* (2008).

The control structure of converter 2 is similar to that of converter 1. As the aim of the controller is to circulate reactive power between the converters, the reactive power reference command I_{sd1}^* of converter 1 is negated and fed as the reference for converter 2. This ensures that the entire reactive current circulates between the two converters with only the system losses being drawn from the grid. Further, the dc bus voltage is maintained to be equal in both the converters by providing the same reference signal (V_{dc1}^*) for both of them.

The control scheme for the two converters in figure 6 is validated through simulation and experiments in the following section.

5. Validation of control scheme

The experimental set-up (figure 6) consists of two identical PWM converters, connected in parallel on the ac side. In both the converters, two 300A, 1200 V IGBT half-bridge modules (SKM300GB124D) (http://www.semikron.com/products/data/cur/assets/skm300GB124D_2890280.pdf) are connected in parallel (as shown in figure 6). The per-phase inductance (L) of the filter inductor is 260 μ H, and the dc bus capacitance (C_0) is 6750 μ F.

The converters are fed from a 100 kVA, 50 Hz autotransformer (figure 6). While the nominal rms line voltage on the primary is 415 V, the converters are fed from the 90% tap (on the secondary) of the autotransformer. The dc bus voltage is maintained at 700 V. A common carrier wave of 5 kHz frequency is used for both the converters and sine-triangle PWM is employed.

Figure 9. Supply voltage and line currents of the two converters at a reactive power reference of 150 kVA (simulation result).

Figure 10. Transient response of the converter for a step change in reactive power reference from 50 kVA to 125 kVA (simulation result).

5.1 Simulation results

The proposed set-up is simulated using MATLAB/ Simulink. Figure 9 shows the grid voltage (v_{RN}), and the converter currents i_{R1} and i_{R2} at a reactive power reference of 150 kVA. It can be observed that i_{R1} and i_{R2} are out of phase. Also, i_{R1} lags the supply voltage by 90° while i_{R2} leads it by 90° . The net current drawn from mains (i_{R-sec}) is ideally zero as mentioned earlier.

In figure 10, the transient response of the converter is shown. The reactive power reference is changed from 50 kVA to 125 kVA in step. The reactive component of current (i_{s1d}) reaches its final value very quickly, while the real power component of current (i_{s1q}) is practically undisturbed around zero. This shows the decoupling between the real and reactive power components. The dc bus voltage remains unchanged.

Figure 11. Supply voltage (v_{RN}) and line currents (i_{R1} and i_{R2}) of the two converters at a reactive power reference of 150 kVA (experimental result). Ch1- i_{R1} and Ch2- i_{R2} , 250 A/div., Ch3: v_{RN} , 240 V/div.

Figure 12. Transient response for a step change in reactive power reference from 50 kVA to 125 kVA (experimental result). Ch1- i_{s1q} , Ch2 - i_{s1d} , Ch3 - i_{R1} , Ch4 - V_{dc} . Ch1, Ch2: 210 A/div., Ch3: 250 A/div., Ch4: 450 V/div.

5.2 Experimental results

Both the converters in figure 6 are controlled from a single digital controller based on TMS320LF2407 DSP processor (Texas Instruments 1990). The experimental results are presented in figures 11, 12 and 13.

In figure 11, the traces Ch1 and Ch2 indicate the converter currents i_{R1} and i_{R2} , respectively, for a reactive power reference of 150 kVA. The trace Ch3 shows the supply voltage v_{RN} , suitably scaled for control purpose. It can be observed that i_{R1} lags v_{RN} by 90° , while i_{R2} leads the same by 90° ; the converter currents i_{R1} and i_{R2} are out of phase.

Figure 12 shows the transient response of the system for a step change in reactive power reference (I_{s1d}^*) from 50 kVA to 125 kVA. Ch2 shows the signal proportional to the reactive power handled by converter 1 (i_{s1d}), while Ch3 presents the line current i_{R1} . Both these signals attain their final values quickly. Ch1 and Ch4 represent the real power signal (i_{s1q}) and the dc bus voltage (V_{dc}), respectively, of converter 1. Both these signals are unaffected by the change in the reactive power reference, confirming the decoupling between the real and reactive power flows.

Figure 13. Measured grid current (i_{R-sec}) with a common carrier wave for the two converters at 150 kVA reactive power circulation (50 A/div.) (experimental result).

Figure 13 shows the transformer secondary current (i_{R-sec}) at 150 kVA reactive power circulation. Though the current is highly distorted, its fundamental component is quite small, as compared to the line current of the individual converters.

6. Modulation with SVPWM

As mentioned earlier, SVPWM is increasingly employed for VSC topology because of its advantages in terms of dc bus utilization and line current distortion (Zhou & Wang 2002). Hence, the proposed heat-run test method should be capable of testing PWM converters operated with SVPWM.

When reactive power is circulated between the two converters, their terminal voltages (V_{R1N} and V_{R2N}) are in-phase with the grid voltage as shown by the phasor diagram in figure 7. Hence the modulating waves (of a particular phase) of the two converters are in-phase with each other. However, the modulating waves differ in amplitude since the corresponding terminal voltages (V_{R1N} and V_{R2N}) have different amplitudes. Hence, the common-mode voltages of the two converters with SVPWM are different (Zhou & Wang 2002).

The difference in injected common-mode voltages of the two converters causes large circulating currents in the back-to-back test set-up shown in figure 5 (Ghosh 2007; Chen 2009). In the back-to-back connected converters, the sum of the six converter currents (i.e., $i_{R1} + i_{Y1} + i_{B1} + i_{R2} + i_{Y2} + i_{B2}$) is equal to zero. But the sum of the three line-currents of an individual converter (either $i_{R1} + i_{Y1} + i_{B1}$ or $i_{R2} + i_{Y2} + i_{B2}$) is not equal to zero. Hence, zero-sequence currents exist in this experimental set-up.

However, in the proposed experimental set-up (figure 6), the converters are separated on the dc side; the sum of the three line currents of both converters ($i_{R1} + i_{Y1} + i_{B1}$) and ($i_{R2} + i_{Y2} + i_{B2}$) are equal to zero. Hence, the zero-sequence circulating currents are absent in this test set-up. Hence, SVPWM can be employed, even with injection of different common-mode voltages on the two converters, as demonstrated by the following experimental results.

Figure 14 shows the R-phase modulating wave (Ch1), R-phase current (Ch2), Y-phase current (Ch3) and the sum of the three phase currents ($i_{R1} + i_{Y1} + i_{B1}$) (Ch4), all pertaining to converter 1, for 150 kVA reactive power circulation. Sine-triangle PWM is employed. As should be

Figure 14. Measured current waveforms with Sine-triangle PWM, at 150 kVA reactive power circulation Ch1 – modulating wave (v_1^*), Ch2 – i_{R1} , 250 A/div., Ch3 – i_{Y1} , 250 A/div., Ch4 – ($i_{R1} + i_{Y1} + i_{B1}$), 10 A/div.

Figure 15. Measured current waveforms with SVPWM, at 150 kVA reactive power circulation. Ch1-modulating wave (v_1^*), Ch2 - i_{R1} , 250 A/div., Ch3- i_{Y1} , 250 A/div., Ch4 - $(i_{R1} + i_{Y1} + i_{B1})$, 10 A/div.

expected, no significant zero-sequence circulating currents are observed (Varma & Narayanan 2006).

Similar results are shown in figure 15, with SVPWM. In this case also the zero-sequence currents are insignificant.

Thus, the circulating currents are insignificant even if different common-mode voltages are injected in the two converters. Hence, the proposed test set-up can be used for heat-run test on space vector modulated converters without any problem of circulating currents.

7. Carrier interleaving

In this section, the effect of carrier interleaving or phase-shifting of the carriers of the two converters on the quality of the input current waveform drawn by the test set-up is studied.

In the proposed experimental set-up, the grid current is equal to the sum of the two individual converter currents, as can be observed from figure 6. The converter currents can be divided into the fundamental and high-frequency ripple components. If losses are ignored, the fundamental components of the two converter currents are out of phase; hence they cancel each other. But,

Figure 16. Interleaving of carrier signals.

Figure 17. Measured grid current (i_{R-sec}) with interleaved carriers at 150 kVA reactive power circulation (50 A/div.).

the sum of the ripple components flow from the mains; this is seen as high frequency distortion in the grid current waveform in figure 13.

The waveform in figure 13 is obtained when the same carrier is used for both the converters. But the carrier signals of the two converters can be phase shifted by certain angle λ (at the carrier frequency) as illustrated in figure 16 (Asiminoaei *et al* 2008; Mao *et al* 2011; Zhang *et al* 2010, 2011). When $\lambda = 180^\circ$, i.e., when the two carriers are out of phase (Asiminoaei *et al* 2008; Mao *et al* 2011; Zhang *et al* 2010, 2011), the grid current waveform drawn by the experimental set-up is as shown in figure 17. As seen from figures 13 and 17, the high frequency component of the input current waveform reduces on account of phase shifting the carrier signals by 180° . This can be attributed to cancellation, to certain extent, of the switching-frequency harmonic currents drawn by the two converters. However, interleaving does not have any appreciable effect on the waveform quality of the individual converter currents as can be observed from figures 15 and 18.

Even with interleaving, the grid current waveform is considerably distorted with lower order harmonic components. This can be attributed to the source voltage being polluted with lower order harmonics (see figure 11). The source current can be improved by employing suitable control methods that take mains voltage distortion into account (Suh & Lipo 2006; Xiao *et al* 2008).

Figure 18. Measured supply voltage (v_{RN}) and line currents of the two converters at 150 kVA reactive power circulation. Ch1- i_{R1} and Ch2- i_{R2} , 250 A/div., Ch3: v_{RN} , 240 V/div. (Interleaved with $\lambda = 180^\circ$).

8. Heat-run test

Using the above mentioned control and modulation methods, heat-run test is carried out on the converters considered. The two parallel-connected IGBT modules in each phase are mounted on separate heat sinks, shown by the dashed lines in figure 6. Hence, there are six heat-sinks present in each converter. Figure 19 shows the measured heat-sink temperature of one of the two converters at 150 kVA reactive power circulation. ΔT indicates the rise in temperature above ambient. T-type thermo couples are used for temperature measurement (<http://www.fluke.com/fluke/us/en/Electrical-Test-Tools/Thermometer/Fluke-50-series-II.htm>). In the figure, (R1, R2) indicate the temperatures of the two-parallel connected modules of R-phase. Similarly Y1, Y2 and B1, B2 indicate the heat-sink temperatures of the other two phases. It can be observed that the temperatures reach to a steady value. The difference in steady state temperatures of B1 and B2, for instance, is indicative of the difference in current sharing between the parallel-connected modules.

Table 1 shows the average of the six-heat sink temperatures at different kVA levels along with the power input to the set-up, measured at the secondary side of the autotransformer (figure 6). As kVA level increases, losses in the system and hence the heat-sink temperature increases.

Evidently, no resistive loading arrangement is required as the converters are open on the dc side. Also, the converters are loaded up to 150 kVA, which is more than the rating of the autotransformer (100 kVA) in figure 6. Hence, with the proposed test method, the converters can be tested at a kVA level higher than the rating of the upstream power equipment.

Further, the total power input to the system is quite low, as compared with the loading on the individual converters. For example, the total power consumed for heat-run test on converters at a power level of 150 kVA is less than 7.5 kW as seen from table 1. Thus the energy consumed for a 90 min heat-run test (figure 19) is only around 11.3 kWh. On the other hand, if a single converter were tested at 150 kW with resistive loading, the energy consumed would have been around 225 kWh for a 90 min heat-run test. Thus, the proposed test procedure results in significant energy saving.

Figure 19. Measured heat-sink temperatures at 150 kVA reactive power circulation.

Table 1. Power input and heat-sink temperature at different kVA.

kVA	25	50	75	100	125	150
ΔT ($^{\circ}\text{C}$)	4.2	6.9	9.2	13.2	16.8	20.2
P_{input} (W)	1730	2485	3450	4618	5844	7343

9. Conclusion

An apparatus and method to conduct heat-run test on line-side PWM converters with minimum energy-consumption from the mains are presented. With the proposed test procedure, the converters can be tested at a kVA level higher than the power source capacity. The test does not require any high capacity loads (e. g., resistor banks) either.

The test method does not have the problem of circulating currents, commonly associated with circulating power or regenerative test methods for power converters. Hence as demonstrated experimentally, the test method is also useful for conducting heat-run tests on converters operated with PWM methods involving common-mode injection such as space vector modulation.

The proposed test method is much useful for practicing engineers charged with testing and qualifying high-power converters.

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