Power electronic transformer with adaptive PLL technique for voltage-disturbance ride through

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Abstract The power electronic transformer (PET) has recently emerged as a type of power converter. It features the basic functions of power conversion and isolation as well as additional functions related to power quality control. A novel PET for a distribution grid called a flexible power distribution unit is proposed in this paper, and the energy exchange mechanism between the network and the load is revealed. A 30 kW 600 VAC/220 VAC/110 VDC medium-frequency isolated prototype is developed and demonstrated. This paper also presents key control strategies of the PET for electrical distribution grid applications, especially under grid voltage disturbance conditions. Moreover, stability issues related to the grid-connected three-phase PET are discussed and verified with an impedance-based analysis. The PET prototype is tested, and it passes the voltage-disturbance ride-through function. The experimental results verify the power quality control abilities of the PET.

Keywords Power electronic transformer, Medium-frequency isolation, Voltage-disturbance ride through, Power quality, Stability

1 Introduction

A distribution transformer is the most important and common equipment in a power distribution network, which is responsible for voltage transformation and voltage isolation. A traditional distribution transformer is very reliable; however, it is bulky and cumbersome. The harmonics between the primary and secondary sides cannot be isolated, and extra equipment is needed to monitor and protect for possible breakdown issues. Nowadays, these drawbacks are real concerns in academia and the industry. Therefore, power-electronics-based transformers called power electronic transformers, intelligent universal transformers, solid-state transformers, smart transformers, energy routers, and others have gradually become an emerging topic

CrossCheck date: 22 August 2017

Received: 6 July 2016 / Accepted: 22 August 2017 / Published online: 5 January 2018

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over the last 10 years, especially for aerospace, railway traction, smart grid, and Energy Internet applications [1–8]. Their initial use may be in special applications where cost and efficiency are secondary to the size and weight [1].

Recent advances in solid-state semiconductors, passive component materials, and microelectronics technologies coupled with the growing need for high power density, low footprint space, and reduced weight without compromising the efficiency, cost, and reliability have provided the impetus for aircraft 115 VAC/400 Hz (or 360–800 Hz) high-frequency-link power-conversion systems as well as telecommunication power supply applications. Similar work has been carried out for traction applications by ABB, Alstom, Bombardier, and Siemens. A pilot installation was completed by ABB in mid-2011, and the Swiss Federal Office for Transport (FOT) homologated it by the end of the year [6].

Moreover, partly because the existing 50/60 Hz power system is more complicated than the 16.67 Hz traction electric system, scientists and engineers working on projects including the Advanced Power Converters for Universal and Flexible Power Management in Future Electricity Networks (UNIFLEX-PM), the Future Renewable Electric Energy Delivery and Management (FREEDM), MEGA Cube, and the Highly Efficient And Reliable smart Transformer (HEART), a new Heart for the Electric Distribution System as well as other projects led by leading universities and companies are still continuously investigating various issues related to PETs for the smart grid and Energy Internet. These issues include the modularity, efficiency, stability, reliability, cost, DC connectivity, active/passive component selection, modulation and control, power flow, and power quality [9–17]. The key characteristics of SST systems designed for smart-grid applications are demonstrated in [10, 11]. The overall efficiency of these systems ranges from 84% to 88%. Systematic optimization of the key medium-frequency transformer for different optimization targets is presented in [12]. Reference [13] prefers soft-switching dual active-bridge DC/DC isolation to cycloconverter AC/AC isolation with a lower efficiency in a symmetrical topology. SiC devices are adopted in [14] for a high-frequency-link AC solid-state transformer. The advanced components allow it to achieve a maximum efficiency of 96.0%. The series resonant converter (SRC) operated in the half-cycle discontinuous conduction mode (HC-DCM) is a highly attractive choice for an isolated DC/DC converter because of its high efficiency; however, control is not possible, and the system basically acts as a “DC transformer” [15]. The unbalanced-load correction capability of two H-bridge-based three-phase three-stage modular PET topologies, the separate phase connection (SPC), and the cross-phase connection (CPC) are analyzed and compared. It is found

that the SPC is suitable for dealing with a full range of unbalanced loads under the condition where the input-stage current stress increases. Nonlinear and intelligent controllers such as an internal model controller, a sliding mode controller, and a neurofuzzy controller are adopted in [18–20] to improve PET performance.

The keynote presentation in [17] points out that an SST is not a 1:1 replacement for a conventional distribution transformer, and it will not replace all conventional distribution transformers (even in the midterm). An SST offers high functionality but has several weaknesses and limitations. Further, this presentation summarizes 10 key existing SST realization/application challenges, which cover most scenarios that the scientists and engineers have been working on in recent years.

It is known that many grid codes have been released to regulate the power quality and integrate new energy systems within the distributed grid [21]. However, there are few reports on grid codes for PETs. A design criterion for an SST under no-load conditions has been proposed in order to avoid instabilities using an impedance-based analysis, but only analytical and simulation results were provided [22]. The main purpose of this paper is to discuss the key issues of the voltage-disturbance ride-through operation of the Gen-I PET project for distribution power systems, entitled “a flexible power distribution unit for a future distribution system,” which has been completed by our group.

First, a novel PET structure for the Gen-I PET project is proposed and briefly described. Then, some key control strategies for the PET are proposed and explained in detail, especially under voltage-disturbance conditions. Moreover, an impedance-based stability analysis is also presented and verified. The hardware design and implementation considerations are also presented. Finally, the PET prototype is tested, and it passes the voltage-disturbance ride-through function.

2 Structure and specifications of PET

A 30 kW 600 V/220 V three-phase four-wire PET prototype is built as shown in Fig. 1, which provides generic building blocks for power conversion, regulation, and distribution. The distributed control system comprises a front-end pulse width modulation (PWM) rectifier as Fig. 2 illustrated, a medium-frequency open-loop isolated DC/DC converter operated under DC transformer conditions, and a downstream three-phase combined inverter using the same single-phase inverter.

A fixed switching frequency open-loop control method is adopted for the multi-winding medium-frequency isolated DC/DC converter. It is referred to as a DC

transformer and provides an unregulated output voltage. By reducing the regulation requirements and narrowing the input voltage ranges, the DC transformer can achieve a higher efficiency and greater power output than the standard regulated transformer, even if the filter choke is eliminated.

Three-phase inverters are made up of three identical modular single-phase full-bridge H4 inverters, which have an excellent inner unbalanced-load correction capability, or other extra control methods should be added to the three-phase inverter [23]. The AC output voltage is regulated with double-loop controllers, where the outer loop is set to

regulate the RMS value of the voltage, while the inner loop regulates the instantaneous value of the voltage. In addition, a bipolar SPWM control strategy helps to support the reactive power.

Moreover, another interleaved Buck converter with the same half-bridge branch as that of the single-phase inverters is embedded in the 350 VDC bus terminal u_{dcL3} , which provides a 110 VDC bus for local DC load usage. It is regulated by an instantaneous-voltage outer loop and instantaneous-inductor-current inner loop. Then, the PET simultaneously provides an AC and DC hybrid distributed grid.

Fig. 1 Power electronic transformer for distribution system

Fig. 2 Control diagram of front-end three-phase PWM rectifier

The main parameters of the system are as follows:

- 1) Input stage: the rated line voltage is 600 VAC, the rated line frequency is 50 Hz, the input inductance is 1.5 mH, the high-voltage DC link capacitors have a capacitance of 2160 μ F, a switching frequency of 4.8 kHz is selected considering the thermal issues for the adopted device having a voltage rating of 1700 V, and the semiconductor switches are SKM400GB176D switches.
- 2) Isolation stage: the switching frequency is 2 kHz, the primary–secondary ratio of the transformer is 3:1:1:1, the low-voltage DC link capacitors have a capacitance of 3000 μ F, the primary semiconductor switches are SKM400GB176D switches, and the secondary switches are SKM300GB128D switches. A middle frequency of 2 kHz is selected rather than a higher frequency because the Gen-I PET project is developed for the next-generation 10 kV PET in preparation for high-voltage IGBT tests at 3300 V and 6500 V in the near future.
- 3) Output stage: the output filter inductance is 0.4 mH, the output filter capacitors have a capacitance of 50 μ F, the switching frequency is 10 kHz, and all switches are SKM300GB128D switches.

In Fig. 1, the proposed PET topology has an inner high-voltage ($u_{dc} = 1050$ VDC) bus and a low-voltage ($u_{dcL1,2,3} = 350$ VDC) bus (a u_{dc} voltage command could be set by operators, e.g., $u_{dc} = 1200$ VDC). Thus, voltage and load-disturbance isolation would be possible with the help of DC link buffer capacitors. Further efforts should focus on the key control strategies, especially the front-end rectifier.

3 Key strategies for PET for voltage-disturbance ride through

For a PET operated with a grid voltage disturbance, observability and controllability are essential. The accurate and fast detection of the frequency and phase angle of the grid voltage is essential to ensure the correct generation of reference signals and to cope with the utility codes, especially for those operated under common utility distortions such as harmonics, voltage sags, frequency variations, and phase jumps [21]. The dynamic change in the grid voltage should be considered for fast control concerns. Therefore, two key strategies for the PET have been investigated and are separately presented in this section, including the phase-locked loop (PLL) design methods, control principles, and small-signal model of the three-phase PWM rectifier. The stability issues related to the grid-connected three-phase PET are also discussed.

3.1 Phase-locked loop design

In this paper, an adaptive three-phase PLL based on the synchronization reference frame is proposed, as shown in Fig. 3.

An orthogonal voltage is generated using the second-order generalized integrator (SOGI) method, as shown in Fig. 4. The closed-loop transfer functions of the orthogonal voltage to the grid voltage are given as follows:

$$D(s) = \frac{v_{2\alpha,\beta}(s)}{v_{\alpha,\beta}(s)} = \frac{k\omega_0 s}{s^2 + k\omega_0 s + \omega_0^2} \quad (1)$$

$$Q(s) = \frac{qv_{2\alpha,\beta}(s)}{v_{\alpha,\beta}(s)} = \frac{k\omega_0^2}{s^2 + k\omega_0 s + \omega_0^2} \quad (2)$$

where $q = e^{-j\frac{\pi}{2}}$; $v_{\alpha,\beta}$ and $v_{2\alpha,\beta}$ are artificial sinusoidal signals; ω_0 is the fundamental frequency.

Furthermore, equations (1) and (2) can be rewritten as:

Fig. 3 General structure of three-phase PLL

$$v_{2\alpha,\beta}(\omega) = Dv_{\alpha,\beta}(\omega) \quad (3)$$

$$qv_{2\alpha,\beta}(\omega) = Qv_{\alpha,\beta}(\omega) \quad (4)$$

where $|D| = \frac{k\omega\omega_0}{\sqrt{(k\omega\omega_0)^2 + (\omega^2 - \omega_0^2)^2}}$, $\angle D = \tan^{-1}(\frac{\omega_0^2 - \omega^2}{k\omega\omega_0})$, $|Q| = \frac{\omega_0}{\omega}|D|$, $\angle Q = \angle D - \frac{\pi}{2}$.

Equation (4) indicates that no matter how ω_0 , ω and k change, $v_{\alpha,\beta}$ and $v_{2\alpha,\beta}$ have a precise 90° phase difference.

In addition, at ω_0 , equations (3) and (4) are simplified as follows:

$$v_{2\alpha,\beta}(\omega_0) = Dv_{\alpha,\beta}(\omega_0) \quad (5)$$

$$qv_{2\alpha,\beta}(\omega_0) = Qv_{\alpha,\beta}(\omega_0) \quad (6)$$

where $|D| = 1$, $\angle D = 0$, $|Q| = 1$, $\angle Q = -\frac{\pi}{2}$. This means that the generated orthogonal system is filtered without any delay at ω_0 owing to its resonance at ω_0 .

In addition, the PLL output frequency is fed back to the SOGI part, as shown in Figs. 3 and 4. Equations (1) and (2) indicate that no matter how ω_0 changes, the bandwidth of the filter is only determined by the given coefficient k . Therefore, it is an adaptive PLL that is theoretically not affected by the variation in the line frequency.

Figures 5, 6, 7 and 8 show the performance of the adaptive SOGI-SPLL. Accurate and fast detection of the frequency and phase angle of the grid voltage is achieved, even if utility distortions such as harmonics, voltage sags, and frequency variations occur.

3.2 Control principles and small-signal model of three-phase PWM rectifier

As shown in Fig. 1, the front-end converter of the PET is a three-phase PWM rectifier. In the dq framework, the three-phase PWM rectifier can be modeled as a DC system as:

$$\begin{bmatrix} e_{sd} \\ e_{sq} \end{bmatrix} = L \frac{d}{dt} \begin{bmatrix} i_{gd} \\ i_{gq} \end{bmatrix} + \begin{bmatrix} 0 & -L\omega \\ L\omega & 0 \end{bmatrix} \begin{bmatrix} i_{gd} \\ i_{gq} \end{bmatrix} + \begin{bmatrix} e_{ud} \\ e_{uq} \end{bmatrix} \quad (7)$$

Fig. 4 SOGI method for constructing orthogonal component

$$\frac{3}{2}(e_{ud}i_{gd} + e_{uq}i_{gq}) = v_{dc}i_{dc} \quad (8)$$

where θ is the output power grid voltage vector angle of the phase-locked loop and $\theta = \omega t$; ω is the angular frequency of the grid voltage, the rated ω is 100π rad/s; e_{sd} and e_{sq} are the d and q -axis components of the power grid voltage; i_{gd} and i_{gq} are d and q -axis components of the power grid current; v_{dc} is the reference value of the DC bus voltage v_{dc} ; i_{sd}^* and i_{sq}^* are the reference values of i_{sd} and i_{sq} ; R is the equivalent resistor load.

Therefore, the constant DC operation point can be obtained with (7) and (8) with the DC values defined as:

$$\begin{bmatrix} E_{sd} \\ E_{sq} \end{bmatrix} = \begin{bmatrix} 0 & -L\omega \\ L\omega & 0 \end{bmatrix} \begin{bmatrix} I_{gd} \\ I_{gq} \end{bmatrix} + \begin{bmatrix} E_{ud} \\ E_{uq} \end{bmatrix} \quad (9)$$

$$I_{dc} = \frac{V_{dc}}{R} \quad (10)$$

$$\frac{3}{2}(E_{ud}I_{gd} + E_{uq}I_{gq}) = V_{dc}I_{dc} \quad (11)$$

From (9)–(11), the detailed values of the DC operation point are:

$$\begin{cases} E_{ud} = E_{sd} \\ E_{uq} = -\omega LI_{gd} \\ I_{gd} = 2V_{dc}^2 / (3E_{sd}R) \\ I_{gq} = 0 \end{cases} \quad (12)$$

Furthermore, with a small-signal disturbance applied to the DC operation point, the small-signal-model transfer functions in Fig. 9 are as:

Fig. 5 Simulation results for SOGI-SPLL under conditions with three-phase voltage imbalance

Fig. 6 Simulation results for SOGI-SPLL under conditions with input voltage drop

Fig. 8 Simulation results for SOGI-SPLL with grid-frequency fluctuations

Fig. 7 Simulation results for SOGI-SPLL with harmonic voltage input

$$\hat{i}_{gd}(s) = \frac{1}{sL} (\hat{e}_{sd}(s) + \omega L \hat{i}_{gq}(s) - \hat{e}_{ud}(s)) \quad (13)$$

$$\hat{i}_{gq}(s) = \frac{1}{sL} (\hat{e}_{sq}(s) + \omega L \hat{i}_{gd}(s) - \hat{e}_{uq}(s)) \quad (14)$$

$$\hat{v}_{dc}(s) = G_p(s) (I_{gd} \hat{e}_{ud}(s) + E_{ud} \hat{i}_{gd}(s) + I_{qd} \hat{e}_{uq}(s) + E_{uq} \hat{i}_{gq}(s)) \quad (15)$$

$$G_p(s) = \frac{3}{2sV_{dc}C_{dc} + I_{dc}} \quad (16)$$

$$\hat{e}_{ud}(s) = (\hat{e}_{sd}(s) + \omega L \hat{i}_{gq}(s) - (\hat{i}_{gd}^*(s) - \hat{i}_{gd}(s)) H_{c1}(s)) K_n \quad (17)$$

$$\hat{e}_{uq}(s) = (\hat{e}_{sq}(s) - \omega L \hat{i}_{gd}(s) - (\hat{i}_{gq}^*(s) - \hat{i}_{gq}(s)) H_{c1}(s)) K_n \quad (18)$$

$$\hat{i}_d^*(s) = -\hat{v}_{dc}(s) H_{v1}(s) \quad (19)$$

where $K_n = K_{pwm}/(0.5T_s + 1)$ for time delay control concerns, T_s is the switching period, and K_{pwm} is the equivalent gain of the main circuitry.

3.3 Stability issues of grid-connected three-phase PET

The inclusion of a PET with a high number of power electronics converters in an existing AC grid introduces a number of technical issues that have not been previously encountered. One concern is the potential instability caused by PET interactions. This instability may take the form of a harmonic resonance induced by the interaction between the input impedance and the source output impedance, as illustrated in Fig. 10, where the grid is emulated with an ideal voltage source and its output impedance, whereas the PET is modeled with the input impedance Z_i or input admittance Y_i [24, 25].

The input admittance matrix is expressed as:

$$\begin{bmatrix} i_{gd} \\ i_{gq} \end{bmatrix} = \begin{bmatrix} Y_{dd} & Y_{dq} \\ Y_{qd} & Y_{qq} \end{bmatrix} \begin{bmatrix} e_{gd} \\ e_{gq} \end{bmatrix} = \mathbf{Y}_{in} \begin{bmatrix} e_{gd} \\ e_{gq} \end{bmatrix} \quad (20)$$

The input admittances are defined as follows:

$$\begin{cases} Y_{dd} = \frac{i_{gd}}{v_{sd}} \Big|_{v_{sq}=0} \\ Y_{qd} = \frac{i_{gq}}{v_{sd}} \Big|_{v_{sq}=0} \\ Y_{qq} = \frac{i_{gq}}{v_{sq}} \Big|_{v_{sd}=0} \\ Y_{dq} = \frac{i_{gd}}{v_{sq}} \Big|_{v_{sd}=0} \end{cases} \quad (21)$$

$$Y_{dd}(s) = \frac{1}{a_3 - a_1 \omega sL + sL} \quad (22)$$

$$Y_{qd}(s) = \frac{1}{a_3 - a_1 \omega L + sL} \quad (23)$$

$$Y_{qq} = \frac{1}{b_3 + b_1 \omega L + sL} \quad (24)$$

$$Y_{dq} = \frac{b_1}{b_3 + b_1 \omega L + sL} \quad (25)$$

where

$$a_1 = \frac{K_n - K_{pwm}}{K_n H_{c1}(s) + sL K_{pwm}} \omega L \quad (26)$$

$$a_2 = \frac{-K_n(H_{c1}(s) + sL)}{K_n H_{c1}(s) + sL K_{pwm}} \omega L \quad (27)$$

$$a_3 = \frac{-K_n(sL + H_{c1}(s)(1 + H_{v1}(s)G_p(s)(a_1 E_{uq} + E_{ud}))}{(K_n - K_{pwm}) + K_{pwm} H_{c1}(s)H_{v1}(s)G_p(s)I_{gd}} \omega L \quad (28)$$

$$b_1 = \frac{-K_n H_{c1}(s)H_{v1}(s)G_p(s)(E_{uq} + \omega L I_{gd}) + (K_{pwm} - K_n)\omega_1 L}{K_n H_{c1}(s) + sL K_{pwm} + K_n H_{c1}(s)H_{v1}(s)G_p(s)(E_{ud} - sL I_{gd})} \quad (29)$$

Fig. 9 Small-signal control block of PWM rectifier ($H_{v1}(s)$ and $H_{c1}(s)$ are the corresponding PI controllers shown in Fig. 2)

Fig. 10 Impedance model of PET system

$$b_3 = \frac{-K_n(H_{c1}(s) + sL)}{K_n - K_{pwm}} \quad (30)$$

The cross-coupling terms Y_{dq} and Y_{qd} are very small because of the decoupling loops introduced in the control scheme. To simplify the model, the time delay is ignored; then, $K_n = K_{pwm}$, and $Y_{qd} = 0$ and $Y_{qq} = 0$ would be obtained owing to ideal decoupling and feedforward control. Therefore, the dominant Y_{dd} is adopted for further analysis of stability.

Assuming that the source voltage V_s is stable, from the general Nyquist criterion [24], the cascaded system stability is determined by $L(s) = Y_i Z_o$ if and only if the net sum of anticlockwise encirclements of the critical point $(-1, j0)$ by the set of characteristic loci of $L(s)$ is equal to the total number of right-half plane poles of $L(s)$.

$$V_i = V_s \frac{Z_i}{Z_o + Z_i} = V_s \frac{1}{I + Y_i Z_o} \quad (31)$$

Figure 11 indicates that for a grid emulator using a PWM voltage source inverter with a line inductance of 8 mH, the system is unstable. When the grid emulator uses a line inductance of 1.5 mH as shown in Fig. 12, the general Nyquist plot and time-domain simulation results show that the system is stable.

4 Hardware design and implementation considerations

4.1 Active half-bridge standard module design

Figure 1 shows that the active half-bridge single-phase standard module can be quickly and easily configured to address a wide range of applications such as AC–DC, DC–DC, or DC–AC converters to provide a platform for the rapid development of multiphase high-power converters and systems and to provide the ability to rapidly develop new AC–AC PET systems.

Typical 62 mm package IGBT modules and core PWM gate driver boards from SEMIKRON, Ltd. are adopted.

Fig. 11 Unstable system simulations (with a line inductance of 8 mH)

Cycle-by-cycle protection functions are embedded in the driver. Fiber-optic cables are used for reliable isolated drive design. In addition, the auxiliary power interface, control interface, and power interface are connected through hardwired terminals, and some forced-air-cooling heat sinks and laminated bus bars need slight modifications due to mechanical and structural issues.

Fig. 12 Stable system simulations (with a line inductance of 1.5 mH)

4.2 Digital control platform design

A power electronics universal control platform is designed and implemented here, which covers AC–DC, DC–DC, and DC–AC converters. Figure 13 shows that the control platform provides a sufficient number of peripheral interfaces to the commercially available digital signal processor (DSP; TI 2808), including voltage, current, and temperature sensors and A/D and D/A conversion of sensor signal circuits.

4.3 Passive component design

DC power storage and a filter are needed in the input of the isolation stage. Capacitors with a large capacitance are usually used. The capacitors mainly have two functions. One is to filter the DC voltage ripple caused by high-frequency switching. The other is to maintain the DC voltage fluctuation inside the qualified range within the inertial delay time of the transformer when the loads change.

During the dynamic process, the amplitude of the voltage fluctuation caused by changes in the loads at the moment t_0 can be expressed as:

$$\Delta u_{dc} = \frac{1}{C} \int_0^{0+t_s} [i_0(t) - i_1(t)] dt \quad (32)$$

where t_s is the settling time; $i_0(t)$ and $i_1(t)$ are the load and output currents. t_s is related to the response speed of the voltage loop. The value of the DC capacitor can be calculated according to the energy balance code. Suppose that $\Delta P_{\max}$ is the maximum variation in the load power, T_{imax} is the maximum inertial time of the rectifier, and $\Delta u_{dc\max}$ is the maximum voltage fluctuation. Then, the maximum energy provided by a DC capacitor during the dynamic process can be calculated as:

$$\Delta W = \Delta P_{\max} T_{imax} / 2 \quad (33)$$

With (32) and (33), the capacitance of the DC capacitor can be calculated as:

$$C \geq \Delta P_{\max} T_{imax} / (2u_{dc}\Delta u_{dc\max}) \quad (34)$$

Setting $\Delta P_{\max} = 5 \text{ kW}$, $T_{imax} = 1.2 \text{ ms}$, $\Delta u_{dc\max} = 10 \text{ V}$, and $u_{dc} = 1050 \text{ V}$, the minimum capacitance can be calculated as $286 \text{ } \mu\text{F}$ with (12). To reduce the hardware cost and the equivalent series resistance of the capacitors, several capacitors in parallel are adopted instead of one capacitor with a large capacitance. Eight $1600 \text{ V}/270 \text{ } \mu\text{F}$ film capacitors in parallel are chosen in the DC stage of the isolation part. Large capacitances are used for possible heavy load usage.

In the output of the isolation part, the DC voltage of each phase is 350 V. With (34), the minimum capacitance

Fig. 13 Control platform layers

can be calculated as 857 μF . Six 700 V/500 μF film capacitors in parallel are chosen.

5 Experimental verification

In order to verify the key control strategies of the PET under voltage-disturbance conditions, another PWM inverter that is the same as the front-end converter of the PET operates as a disturbance-voltage source to emulate the grid in the field. Owing to the limited loads, the maximum power is achieved at 30 kW, and the rated system power is 100 kW.

5.1 Steady-state performance tests

Figure 14 shows the experimental waveforms of the PET. Figure 14a shows the three-phase input line voltage waveform measured by a Fluke 434 power quality

Fig. 14 Experimental waveforms

Fig. 15 Voltage-disturbance ride through of the PET with a three-phase 60% balance voltage sag and phase jump

analyzer, where u_{ab} is 597.1 V, u_{bc} is 595.6 V, and u_{ca} is 593.5 V. Figure 14b shows the system operation parameters. The load power is 14.83 kW, and the system power factor is 0.95, which demonstrates that the input operates at a high power factor. Figure 14c and 14(d) show the primary- and secondary-side voltages of the medium-frequency isolated transformer. The single-phase full-bridge rectifier converts the square AC wave into a 350 V direct current. Figure 14e shows the extra DC output terminal of the PET—a 110 VDC bus for the local DC load. Figure 14f shows the output three-phase AC voltage whose waveforms are symmetrical and sinusoidal, and the RMS voltage is 221 V. Compared with the reference 220 V, the error is only 0.4%. Figure 14f shows the corresponding three-phase load current. Figure 14g shows that the THD of the output A-phase voltage is only 2.5%, which conforms to the national standard that the THD must be within 5%. The primary- and secondary-side currents of one winding of the transformer are shown in Figs. 14i and 14j, respectively. A double-output frequency variation of 100 Hz is observed in Fig. 14j because the unregulated DC

Fig. 16 Voltage-disturbance ride through of the PET with a three-phase 60% balance voltage sag, a frequency variation of 50/40 Hz, and a phase jump

transformer topology is adopted [26]. When the C-phase load is suddenly removed, an acceptable overshoot occurs at the high-voltage DC bus, and the DC voltage is regulated after 400 ms. In addition, the output C-phase voltage is regulated well during the transition.

5.2 Voltage-disturbance ride-through tests

Owing to limitations on the length the manuscript, some critical tests results are selected and shown in Figs. 15 and 16, where some utility distortions occur simultaneously. V_{ab} is the PWM rectifier line voltage, i_a is the PWM rectifier input current, V_a is the single-phase inverter output voltage. The voltage-disturbance ride through of the PET with a three-phase 60% balance voltage sag and phase jump are shown in Fig. 15, and the voltage-disturbance ride through of the PET with three-phase 60% balance voltage sags, a frequency variation of 50/40 Hz, and a phase jump are shown in Fig. 16. It is observed that no matter how the voltage changes, the input current and

Fig. 17 High-voltage DC bus variation during voltage-disturbance ride through

output voltage are both well-regulated with a fast dynamic response. The input power quality is only affected for two or three grid periods. The output voltage changes are not observed owing to the DC link buffer. Figure 17 further illustrates the small-voltage variation in the high-voltage DC bus during voltage-disturbance ride through with the help of a feedforward control method.

5.3 Grid-connected instability tests

Sections 5.2 and 5.3 discuss the steady-state and transient performance of the stable PET system. However, owing to an uncertain grid impedance in the distribution power system, unstable oscillations would also be observed with the impedance-based stability analysis in Sections 3.2 and 3.3, as shown in Fig. 18, which helps to understand the harmonic resonance in power-electronics-based power systems using a PET. Finally, the output single-phase inverters of the PET still maintain normal operation without being affected by the input oscillation.

Fig. 18 Unstable waveforms

6 Conclusion

A novel PET for a distribution grid called a flexible power distribution unit is proposed in this paper. DC/DC isolation for the three-phase inverters is implemented through one compact multiwinding transformer, which reduces the system complexity.

Focusing on the grid code issues of the PET, such as the voltage-disturbance ride through and harmonic resonance, which have not been previously encountered, this manuscript presents the key PLL design methods under distorted grid conditions, the control principles, a small-signal model, and the input admittance of the three-phase PWM rectifier in detail. This helps understand the harmonic resonance in power-electronics-based power systems using a PET.

Although the cost, volume, and weight of the PET are presently much higher than those of conventional power transformers, the future of the PET is still promising, as it can play many different but important roles in future smart grid and Energy Internet applications.

Acknowledgements This work was supported in part by the National Basic Research Program of China (No. 2016YFB0900404), the National Natural Science Foundation of China (No. 51477030, No. 51207023), the Cooperative Innovation Fund of Jiangsu Province—the Prospective and Joint Research Project (No. BY2015070-18), the Basic and Prospective Science and Technology Project of State Grid Corporation of China (No. PD71-17-024), and the Fundamental Research Funds for the Central Universities (No. 2242017K40159).

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