Integrated Higher Order PWM Filter-Transformer Structure for Single Phase STATCOM

D. Venkatramanan¹ and Vinod John

Department of Electrical Engineering, Indian Institute of Science, Bangalore 560012, India.
e-mail: venkat86ram@gmail.com, vjohn@ee.iisc.ernet.in

Abstract

Grid connected pulse width modulation based voltage source inverters (PWM-VSIs) are extensively used for applications such as distributed generation, power quality and uninterrupted power supplies applications. A power filter is invariably required at the output of the power converter to be connected to grid in order to reduce the harmonic distortion caused by the power converter. Many a times, a transformer is also present before the point of common coupling (PCC). Magnetic components often constitute a significant part of the overall weight, size and cost of the grid interface scheme. So, a compact inexpensive design is desirable. A higher order LCL-filter and a transformer are increasingly being considered for grid interconnection of the power converter. This paper proposes a design method based on a three-winding transformer, that generates an integrated structure that behaves as an LCL filter, with both the filter inductances and the transformer that are merged into a single electromagnetic component. A single phase full bridge power converter is being operated as a static compensator (STATCOM) for performance evaluation of the integrated filter-transformer. A resonant integrator based single phase PLL and stationary frame AC current controller are employed for grid frequency synchronisation and line current control respectively.

Keywords

Power filter, power transformer, harmonic distortion, PWM power converter, magnetic flux, voltage control, PLL, transient response

¹D. Venkatramana is presently with American Power Conversion India Pvt. Ltd. (APC by Schneider Electric), Kalyani Platina, Block 1, EPIP zone, Phase II whitefield, Bangalore 560066, India.
1 Introduction

Grid connected PWM-VSIs are becoming exceedingly popular for applications such as DG, power quality, UPS etc. A power filter is invariably required at the output of such a power converter to bring down the current and voltage distortions caused by harmonics injected by the VSI into the grid [1].

The conventional way of interface is through a simple first order L-filter. But such a filter is bulky, inefficient and cannot meet the regulatory requirements pertaining to interconnection of harmonic loads to the grid. This is evidently seen at higher power levels of operation where the switching frequency is limited to limit the switching loss occurring in the semiconductor devices. For such applications, higher order LCL filters are more attractive as they offer higher attenuation at lower switching frequency for a similar filter size [1–5]. A variety of methods for LCL-filter design are available in literature. Many a times, a power transformer is also employed after the filter stage, which provides galvanic isolation and adds an extra degree of freedom to adjust the output voltage level to that of the grid [6]. Fig. 1(a,b) shows the typical grid interface scheme of a single phase full bridge power converter.

It may be noted that copper that carries current and magnetic core that carries flux, are present in both filter and transformer. Also all the copper must be gauged for rated line current. Magnetic components such as filter inductors and transformer constitute a significant part of the overall size, weight and cost of the grid interface scheme. Hence, a compact inexpensive design is desirable. Since magnetic components are present in both filter and transformer, there is scope for their potential integration and such a magnetic integration can reduce both weight and cost of the interface scheme [6].

It is relatively simple to integrate the magnetics of a L-filter into the transformer. Since a L-filter comes in series with the transformer, the leakage inductance of the transformer, which also comes in series with the transformer, can potentially replace the filter [7]. Such a design is simple and feasible as the typical values of transformer leakage inductance and desired filter inductance are quite close on a per-unit basis. Integration of the magnetic components in the case of an LCL filter requires more complex design procedures [6–9]. Hence, it is desirable to have a simple core and winding structures so that it is easy to procure and manufacture to desired tolerances. Some of these integrated magnetic structures are designed to be used as LC filters [10] and can achieve a high frequency attenuation of only $40 \text{dB/dec}$. However a filter that is employed for grid interactive applications is of the LCL variety.
Figure 1: Grid interface scheme for a single phase STATCOM with a full bridge PWM power converter and (a) filter plus transformer stage that is constructed using, (b) discrete component LCL-filter and transformer, and (c) three-winding integrated filter-transformer structure, as discussed in [1–5].

In this paper, a three-winding transformer configuration is proposed to achieve an LCL-filter behaviour with all the magnetic components integrated in a single structure. The single compact structure would function as both filter and transformer. The leakage inductances of the transformer would perform the functions of the filter inductances. An external capacitor is used that acts as the filter capacitor. Fig. 1(a) and 1(c) show the grid interface scheme with the integrated filter-transformer.

For verifying the effectiveness of the integrated structure, a grid interactive operation of a single phase VSI has been constructed in the laboratory. Control of three phase VSIs is widely reported in literature. Conventionally, d-q control in Synchronous Reference Frame (SRF) is employed for both PLL and line current control where PI-controllers are used to track the DC references [15, 16]. Single phase systems do not have defined direct (d) and quadrature (q) axis components that are required for SRF transformation. Thus, references are AC in nature and hence usage of PI controllers cannot yield
zero steady state errors. However, proportional-resonant (PR) controllers track AC references accurately without steady-state errors [17–21].

In this work, a single phase 3kVA IGBT based full bridge power converter is being operated as a STATCOM for performance evaluation of the filter-transformer. A resonant integrator based single phase PLL and stationary frame AC current control scheme are employed for closed loop control of the system [23]. To begin, the paper explains the working principle of the multi-winding filter-transformer. An appropriate equivalent circuit for the integrated structure is next developed from which relevant transfer functions are obtained. Subsequently, filter-transformer design is explained. Generic control block diagrams for illustrating the current and voltage control strategy are next described followed by a discussion of the experimental results validating the theoretical and implementation concepts.

2 Multi-winding Filter-transformer

The multi-winding structure that is intended to emulate a higher order filter is depicted in Fig. 1(c). The structure consists of two primary windings comprising of a main winding, MW, and an auxiliary winding, AW, of equal turns and one secondary winding, SW. The three windings are magnetically coupled by the core. A capacitor, $C_f$, is externally connected to AW as shown. The primary side is connected to the inverter and the secondary winding is connected to grid. It must be noted that the two primaries are wound with opposite dot points over the core.

2.1 Working Principle

At low frequencies including power frequency, the capacitor $C_f$ acts as an open circuit due to its very large impedance and thus, AW is effectively disconnected. Hence, power frequency component of the inverter voltage will see only the MW. Subsequently energy transfer takes place to the secondary and hence to the grid.

At higher frequencies inclusive of switching frequencies, $C_f$ acts as a short circuit. Under this condition, the two primaries work in conjunction with applied voltage ($V_i$) being common to both. Due to their opposite sense of winding, the magnetic flux produced by each winding are opposite in direction, hence cancel one another. So conceptually, for high frequency components, flux gets cancelled in the
core. Thus no energy transfer takes place to the secondary due to the absence of magnetic field in the core. In other words, there is no magnetic coupling between primary and secondary at high frequencies. This would amount to filtering of all high frequency distortion components and passing through the low frequency fundamental present in the inverter output voltage. This phenomenon is referred to as internal differential mode distortion cancellation [10]. A drawback of this configuration is that the inverter is ideally going to see a dead-short across the primary side, since due to the cancellation of flux, no induced emf is produced across the magnetizing inductance. However, winding dispositions can be chosen to obtain suitable leakage inductances in the primary to limit the inverter ripple current to acceptable level in a grid interactive application where an LCL filter structure is required. The proposed design does not entail complete internal differential mode distortion cancellation in the magnetic core.

2.2 Equivalent Circuit Model

To develop the transfer function of the filter-transformer, the equivalent circuit model of a multi-winding transformer is analyzed. The notion of leakage flux is meaningful only when a pair of windings are considered. Thus, leakage inductance is always defined for a pair of windings. In case of a three-winding transformer as shown in Fig. 3(a), where there are three pairs of windings, leakage inductance must be defined for every pair since leakage flux exists for every pair.

For a three-winding transformer, leakage inductances are defined as,

\[ L_{ij} = \text{total leakage inductance of winding pair (i,j) as seen from the } i\text{th winding side, with winding } k \text{ being open.} \]

where,

\((i, j, k) \text{ vary from (1 to 3)}\)

and \(i \neq j \neq k\).

Thus, inductances \(L_{12}, L_{21}, L_{13}, L_{31}, L_{23}, \text{ and } L_{32}\) can be defined for the three-winding transformer. These leakage inductances can be measured by performing multiple short-circuit tests with pertinent windings. The equivalent circuit model of the three-winding transformer as shown in Fig. 3(b), consists of \(L_I, L_{II}\) and \(L_{III}\), which are the resultant leakage inductances seen by the respective windings. These are composite inductances formed by linear combinations of different leakage inductances \(L_{12}, L_{23}\) and
$L_{31}$, since different leakage fluxes are going to act in conjunction [11], [12]. The composite inductances are given by,

$$L_I = \frac{L_{12} + L_{13} - L_{23}}{2}$$  \hspace{1cm} (1)

$$L_{II} = \frac{L_{21} + L_{23} - L_{13}}{2}$$  \hspace{1cm} (2)

$$L_{III} = \frac{L_{31} + L_{32} - L_{21}}{2}$$  \hspace{1cm} (3)

The winding resistances have been disregarded in the equivalent circuit for simplicity.

### 2.3 Leakage Inductance Evaluation

A shell-type transformer design where the core surrounds the concentric copper windings is used for the proposed integrated filter-transformer. Shell type transformers are extensively discussed in literature [11, 12]. Analytical expression for its leakage inductance is given by Rogowski’s formula (4). Fig. 2 shows the basic construct of a shell type transformer and the corresponding leakage flux existing in it respectively.

For a given pair of windings $(i, j)$, the total leakage inductance $L_{ij}$ as seen from the $i$th winding side is given by,

$$L_{ij} = \mu_0 \frac{N^2}{H_{eqij}} \frac{2\pi}{3} \left[ \frac{1}{3} (T_i D_{wi} + T_j D_{wj}) + T_{gij} D_{wgi} \right]$$  \hspace{1cm} (4)

Where,

$$H_{eqij} = \frac{H_w}{k_{Rij}}$$  \hspace{1cm} (5)

$$k_{Rij} = 1 - \frac{1 - \epsilon - \pi H_w}{\lambda_{ij}}$$  \hspace{1cm} (6)

$$\lambda_{ij} = T_i + T_j + T_{gij}$$  \hspace{1cm} (7)

and transformer assembly parameters,

$T_i$ and $T_j$ = Winding widths of $i$th and $j$th winding respectively,

$T_{gij}$ = Inter-winding distance between $i$th and $j$th winding,

$D_{wi}$, $D_{wj}$ and $D_{wgi}$ = Mean diameter of $i$th winding, $j$th winding and inter-winding gap respectively,

$H_w$ = Height of the winding,

$k_{Rij}$ = Rogowski factor for the winding pair $(i, j)$
Figure 2: Shell-type test transformer with leakage field, corresponding MMF variation and dimensions used to obtain the equivalent circuit parameters.

\[ N_i = \text{Number of turns of } i\text{th winding.} \]

This approach is followed for every pair of windings to evaluate \( L_{12}, L_{23}, \) and \( L_{31}. \)

### 3 Transfer Function Analysis

The equivalent circuit model of the filter-transformer structure is shown in Fig. 3(b). The grid voltage acts as a short for all frequency components except the fundamental as shown in Fig. 3(a). The opposite winding sense of the auxiliary winding is conveniently modelled in the equivalent circuit by reversing the polarity of the applied inverter voltage. The cancellation of flux in the core is reflected in the equivalent circuit as cancellation of voltage across the magnetizing inductance due to the two sources. The voltage \( e(s) \) across the magnetizing inductance \( L_M \) represents the open circuit secondary voltage \( V_s(s) \) produced due to the net flux present in the core. The required voltages and currents can be obtained by applying superposition theorem. Assuming \( L_M \) is significantly larger as compared to leakage inductances, the
pertinent transfer functions are given by,

\[
\frac{e(s)}{v_i(s)} = \frac{1}{1 + (L_{III} + L_I)Cs^2} \left\{ 1 + (L_{III} - L_I)Cs^2 \right\} \tag{8}
\]

\[
\frac{I_p(s)}{v_i(s)} = \frac{1}{s(L_I + L_{III})} \left\{ 1 + \frac{(L_{III} - L_I)Cs^2}{L_pCs^2 + 1} \right\} \tag{9}
\]

\[
\frac{I_i(s)}{v_i(s)} = \frac{1}{s(L_I + L_{III})} \left\{ 1 + \frac{(L_I + 4L_{III} + L_{III})Cs^2}{L_pCs^2 + 1} \right\} \tag{10}
\]

where,

\[
L_p = \frac{L_IL_{III} + L_{III}L_{III} + L_{III}L_I}{L_I + L_{III}} \tag{11}
\]

Equation (8) represents the transfer function pertaining to voltage harmonic attenuation if the inverter were to operate in standalone mode. Equation (9) represents the transfer function relating the injected grid current and the inverter voltage. This is required to evaluate the TDD of injected grid current. The inverter current to inverter voltage transfer function, which is required for control loop design, is given by (10).

In all the transfer functions, the term outside the braces in equations (8) – (10) represent the same transfer functions under two-winding operation of the structure without AW connected. In this configu-
ration, the structure emulates a first order L-filter. The term inside the braces represents the correction factor introduced by the addition of the AW winding in each of these transfer functions.

It can be observed that (8) and (9) contain a second order zero in them as opposed to that of a discrete LCL-filter [14]. However, the control transfer function (10) is similar to that of a discrete component LCL-filter and hence the same control strategy can be adopted for closed loop control.

Hence, the integrated filter-transformer structure does not emulate a LCL-filter entirely from the output current perspective. However, by designing suitably and placing the zeros far away, LCL-filter characteristics can be achieved in a frequency range of interest where significant attenuation is required. Nevertheless, from the input current viewpoint, the structure is same as a discrete component LCL-filter.

It may be noted that adopting a bifilar winding structure for the primaries as suggested by [10] will make $L_{13} = 0$ and hence, $L_p = 0$. Using this in equation (9), it can be seen that the structure will no longer emulate higher order LCL-filter.

4 Transformer Design

A three-winding transformer test prototype was designed using conventional area-product approach [13]. The area-product is given by

$$A_p = A_{core}A_{window} = \frac{VA \times 10^{-6}}{2.22fB_mJ_kw^4}$$

(12)

The design details are furnished in Table. 1. It can be noted that the area-product, $A_p$, employed is greater than the required value. This is because it is essential to house the AW winding as well in the same window. An E-core shape was achieved by using multiple C-core structures.

4.1 Winding Attributes

- Main Winding (MW): Functions at both low and high frequencies; carries fundamental as well as switching ripple current. Hence, a thicker wire gauge is essential.

- Secondary Winding (SW): Functions mainly at low frequencies; carries fundamental and a small amount of filtered switching ripple current. Hence, a thick gauge is essential.
• Auxilliary winding (AW): Functions only at high frequencies and thus carries only switching ripple current. Hence, a smaller wire gauge is sufficient.

Table 1: Design details of three winding shell-type test transformer.

<table>
<thead>
<tr>
<th>Item</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>VA</td>
<td>3 kVA</td>
</tr>
<tr>
<td>$V_{MW}, V_{SW}$</td>
<td>240 V, 240 V</td>
</tr>
<tr>
<td>$I_{MW}, I_{SW}$</td>
<td>12.5 A, 12.5 A (SWG12)</td>
</tr>
<tr>
<td>$I_{AW}$</td>
<td>1.6 A (SWG20)</td>
</tr>
<tr>
<td>$B_m$</td>
<td>1.28 T</td>
</tr>
<tr>
<td>$j$</td>
<td>2.5 A/mm$^2$</td>
</tr>
<tr>
<td>$k_w$</td>
<td>0.656</td>
</tr>
<tr>
<td>$f$</td>
<td>50 Hz</td>
</tr>
<tr>
<td>$A_p (required)$</td>
<td>1288 cm$^4$</td>
</tr>
<tr>
<td>$A_p (employed)$</td>
<td>2425 cm$^4$</td>
</tr>
<tr>
<td>$A_{core}$</td>
<td>37.5 cm$^2$</td>
</tr>
<tr>
<td>$N_{MW}, N_{SW}, N_{AW}$</td>
<td>225, 225, 225</td>
</tr>
<tr>
<td>Core type</td>
<td>Amorphous AMCC367’s</td>
</tr>
</tbody>
</table>

The calculated and measured values of circuit parameters are shown in Table 2. Subsequently, the composite inductances were obtained. It can be noticed that one of the composite inductances is negative. This is realistic and quite possible in case of multi-winding transformers depending on winding dispositions. The leakage inductance between a pair of windings can never be negative. But the mutual effects of leakage fluxes among windings for load currents can very well be negative depending on how the leakage flux of one interlinks the turns of the other, leading to negative composite inductances [11].
Table 2: Three-winding transformer and its equivalent circuit parameters.

<table>
<thead>
<tr>
<th>Quantity</th>
<th>Calculated Value</th>
<th>Measured Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_{MW}$</td>
<td>0.31Ω</td>
<td>0.3Ω</td>
</tr>
<tr>
<td>$R_{SW}$</td>
<td>0.31Ω</td>
<td>0.3Ω</td>
</tr>
<tr>
<td>$R_{AW}$</td>
<td>3.56Ω</td>
<td>3.6Ω</td>
</tr>
<tr>
<td>$L_{12}$</td>
<td>3.33mH</td>
<td>3.77mH</td>
</tr>
<tr>
<td>$L_{23}$</td>
<td>3.1mH</td>
<td>4mH</td>
</tr>
<tr>
<td>$L_{31}$</td>
<td>8mH</td>
<td>9.32mH</td>
</tr>
<tr>
<td>$L_I$</td>
<td>4.115mH</td>
<td>4.545mH</td>
</tr>
<tr>
<td>$L_{1I}$</td>
<td>-0.785mH</td>
<td>-0.775mH</td>
</tr>
<tr>
<td>$L_{1III}$</td>
<td>3.9mH</td>
<td>4.775mH</td>
</tr>
</tbody>
</table>

4.2 Capacitor Design

Eq.(9) that governs the injected grid current can be rewritten as,

$$\frac{I_g(s)}{v_i(s)} = \frac{1}{sL_T} \left( \frac{s^2 + 1}{\omega_{zg}^2 \left( \frac{s^2}{\omega_p^2} + 1 \right)} \right)$$

(13)

where,

$$\omega_{zg} = \frac{1}{\sqrt{L_{zg} C}}$$

(14)

with $L_T = 9.32\text{mH}$, $L_{zg} = L_{1III} - L_I = 0.23\text{mH}$. External capacitor $C$ is chosen such that the anti-resonant valley caused by the zero is placed at 20kHz. This is because for a full bridge converter employing sine-triangle PWM technique with 10kHz carrier frequency ($f_{sw}$), switching frequency harmonics start occurring from 20kHz.

$$\omega_{zg}^2 = (2\pi 20000)^2 = \frac{1}{0.23\text{mH} \times C}$$

(15)

$$C = 0.25\mu\text{F}$$

(16)

Metallised polypropene AC capacitor was employed for high frequency filtering operation.
Single Phase Closed Loop Control

A single phase full bridge power converter with the proposed integrated filter-transformer was operated as a STATCOM. The power circuit of the system is shown in Fig. 1(a, c). The overall control structure to operate the power converter as a single phase STATCOM is shown in Fig. 4. For closed loop control, an outer voltage loop for DC bus voltage regulation and an inner current loop for converter line current control are required. A PI controller would suffice for outer voltage loop. For the current loop, a resonant integrator based proportional-resonant (PR) controller is being employed.

A resonant integrator is a generalised AC second order integrator with a tuned resonant frequency $\omega_0$ [17, 21]. Its transfer function is given by (17). The gain of the resonant integrator is infinite at the tuned resonant frequency.

$$H(s) = \frac{p(s)}{e(s)} = \frac{k_i s}{s^2 + \omega_0^2}$$  (17)

When $\omega_0$ is set to zero, it reduces to an integral controller. A PR controller is a AC controller that makes the system loop gain infinite at the tuned resonant frequency $\omega_0$ and thereby eliminating any steady state error at that frequency. This is based on the internal model principle [18]. For a grid connected application, $\omega_0$ could be appropriately set corresponding to grid frequency (50Hz) such that line current tracks the AC reference [20].

For grid interactive operation of the power converter, its frequency must be synchronised with that of the grid. In the present case, a resonant integrator based single phase d-q PLL shown in Fig.4 serves the purpose of grid frequency synchronisation. This scheme is based on a three phase d-q PLL [16], except for the orthogonal vector generation scheme. In the single phase case, the two orthogonal vectors must be generated from the single available grid voltage. In this paper, the required orthogonal vectors are generated with the help of a resonant integrator [21].

5.1 Control Strategy

In a conventional three phase case where d-q control is employed, $I_q$ and $I_d$ represent real current and reactive current reference respectively (in synchronous reference frame), provided the grid space vector is aligned along q-axis [15]. For brevity, same notation is being followed here. However, in this case they represent the actual peak values of real and reactive current references respectively in stationary
Figure 4: Overall control structure for single phase STATCOM showing the DC bus voltage controller, ac output current controller, feedforward compensation terms and single-phase PLL.
reference frame. The PLL that produces sinusoidal unit vectors in phase and quadrature with the grid voltage, enables decoupling of active and reactive currents and thus aids their independent control [22]. Inductive voltage drop and grid voltage amplitude $V_{gFF}$ are added as feed-forward in the current loop.

Table 3: System specifications and control loop design data

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_{\text{rated}}$</td>
<td>3kVA</td>
</tr>
<tr>
<td>$V_{\text{rated}}$</td>
<td>240V</td>
</tr>
<tr>
<td>$I_{\text{rated}}$</td>
<td>12.5A</td>
</tr>
<tr>
<td>$V_{dc-\text{bus}}$</td>
<td>400V</td>
</tr>
<tr>
<td>$f_{sw}$</td>
<td>10kHz</td>
</tr>
<tr>
<td>$f_{bw(\text{inner})}$</td>
<td>600Hz</td>
</tr>
<tr>
<td>$k_p$</td>
<td>4.68</td>
</tr>
<tr>
<td>$k_i$</td>
<td>35286</td>
</tr>
<tr>
<td>$f_{bw(\text{outer})}$</td>
<td>8Hz</td>
</tr>
<tr>
<td>$k_v$</td>
<td>2.5</td>
</tr>
<tr>
<td>$T_v$</td>
<td>120ms</td>
</tr>
</tbody>
</table>

The design details of single phase PLL, AC current controller and voltage controller are elaborately discussed in [23]. The system specifications and design data are furnished in Table 3.

6 Experimental Results

A 3kVA IGBT based single phase full bridge power converter is being operated as a STATCOM. Sine-triangle comparison PWM technique is employed for switching the IGBTs of the power converter at a carrier frequency of 10kHz. Thus switching frequency harmonics occur as sidebands around even multiples of carrier frequency.

Comparison of analytically predicted and measured frequency response of various transfer functions, based on the developed transfer function model of the filter-transformer, are presented in Fig. 5. Fig. 5(b) shows the measured open circuit secondary voltage to inverter voltage transfer function of the filter-
transformer in three-winding operation. It can be noticed that the switching frequency components are significantly attenuated. Fig. 5(d) and Fig. 5(f) show the corresponding measured inverter current and grid current transfer functions. It can be noticed that for a given switching frequency harmonic, the grid current is attenuated to a greater extent than the inverter current due to the action of the third winding. Figs. 5(a), 5(c), 5(e) show the corresponding analytically predicted transfer functions.

Measurement was done using an analog network analyzer manufactured by AP Instruments which has a frequency range from 0.01 Hz to 15 MHz, with a maximum output of 1.77V. It can noticed that the measured frequency response closely match the analytically predicted ones as far as the main resonant and anti-resonant frequencies are concerned, as can be seen at frequencies ($F_{1a}$ to $F_{6a}$) and ($F_{1m}$ to $F_{6m}$) in Fig. 5. Since winding resistances were disregarded in the transfer function analysis, the resonant and anti-resonant peaks are undamped in the analytically predicted frequency responses. Also, the low frequency behaviour of the actual system is resistive, which was ignored in the filter transfer function analysis for simplicity.

FPGA based digital platform with ALTERA CYCLONE EP1C12Q240C8 chip was employed for STATCOM control implementation. Fig. 6(a) shows the open circuit secondary voltage of the filter-transformer in two-winding configuration, without AW, under open loop operation of the power converter. It can be seen that the structure performs no filtering action to the voltage since it emulates only a L-filter in this configuration. Fig.6(b) shows the open circuit secondary voltage of the filter-transformer under three-winding operation. It can be seen that with the AW winding in place, the structure is able to filter its input voltage significantly. Fig. 6(c) and 6(e) show the inverter current and the corresponding injected grid current at 90% current reference under two-winding, 2.5kVA operation the filter-transformer. It can be noticed that the two currents are almost identical in terms of the switching ripple current that rides over the fundamental. Fig. 6(d) and 6(f) show the inverter current and the corresponding injected grid current at 90% current reference under three-winding, 2.5kVA operation the filter-transformer. It can be noticed that the two currents are significantly different in terms of switching ripple current owing to the action of the third winding.

In terms of total RMS value, the grid current is slightly smaller than inverter current due to the no-load current of the transformer which is about 10% of the rated current. It lags the input voltage by
Figure 5: Comparison of analytically predicted (a, c, e) and experimentally measured (b, d, f) frequency response of the integrated filter-transformer in three-winding configuration showing: (a), (b) open circuit secondary voltage to applied primary voltage characteristics; (c), (d) inverter side current to inverter voltage characteristics with $V_g = 0$; and (e), (f) grid side current to inverter voltage transfer functions with $V_g = 0$. 
Figure 6: Comparison of two-winding (a, c, e) and three-winding (b, d, f) operation of the filter-transformer for: (a), (b) primary voltage and open circuit secondary voltage in standalone mode; (c), (e) inverter side current leading the unit vector representing fundamental voltage; and (d), (f) grid side current and unit vector, in grid interactive STATCOM mode of operation of the power converter at 90% loading.
Figure 7: Comparison of grid current harmonics at 90% loading for (a) two-winding operation and (b) three-winding operation.

90° and hence adds out of phase with the inverter current that is leading the voltage by 90°, to form the grid current. Fig. 7 shows the frequency spectrum of grid current in two-winding and three-winding operation of filter-transformer. In three-winding configuration, the switching frequency harmonics in the grid current are highly attenuated owing to the action of the third winding. Due to the peaky magnetizing current drawn by the transformer, grid current has mild low frequency harmonic distortions. Fig. 8 shows the measured output current and grid voltage for a reactive leading reference current change from 10% to 90% in the digital controller. The inverter current starts tracking the reference within a quarter cycle.
Figure 8: Transient response of the STATCOM for a step increase from 10% to 90% in current command. Waveforms are grid voltage (CH1: blue, 200V/div.), inverter current (CH2: red, 10A/div.), Step change signal (CH3: green, 10V/div.), reference current signal (CH4: pink, 5V/div.).

7 Conclusion

An integrated filter-transformer structure for single phase grid interactive power converter has been designed and tested. A design method is proposed that employs a three-winding transformer configuration for the purpose of magnetic integration, to achieve a third order filter characteristics. The measured frequency responses of the structure matched the analytically predicted ones from the simplified equivalent circuit. The integrated structure does not emulate a LCL-filter in the entire frequency range from the output current perspective due to the presence of zeros. From input current viewpoint, the structure behaves same as a LCL-filter. The filter-transformer’s performance was found to be significantly better than a L-filter since LCL-filter characteristics was achieved in the frequency range upto 150kHz. Such an integrated design uses a standard core and economizes on copper as the third winding is of much smaller gauge. Experimental results pertaining to STATCOM mode of operation of a single phase power converter show that fast transient response of less than a quarter cycle can be achieved. Laboratory tests conducted on the STATCOM indicates the high performance that can be obtained with such a
filter-transformer, which is useful for many grid interactive inverter applications.

References


