POWER ENHANCEMENT BY ADOPTING ACTIVE POWER FILTER SCHEME WITH PI BASED ONE CYCLE CONTROL UNDER GRID DISTORTIONS

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Abstract—*Distribution line consists of a range of loads which are both linear and nonlinear, due to these nonlinear loads a harmonic current is developed at the source current which leads to detrimental disturbances, aging effect, poor power factor, lower efficiency. Using active power filter is one of the viable solutions to eliminate power line harmonic currents generated by nonlinear loads and to improve the power factor. APF is operated in dual boost convertor mode with constant switching frequency by using one cycle control method. The advantages of going for this type of control are it requires source current only and voltage as reference. And flip flops are used to generate the pulse signal to the convertor which is connected in parallel with the distribution system. The inverter will provide phase shifting current. This phase shifted current is helpful in injecting and cancelling out the harmonic current and to achieve three phase unit power factor for the current to and from the power grid. A DC capacitor is intended which acts as a source to the converter. In order to reduce the error and to produce firing signal, PI controller is used .*

Index Terms—Nonlinear loads, Active Power Filter (APF), Dual -Boost converter, One-Cycle control,

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I. INTRODUCTION

When a three phase supply is connected to a nonlinear load usually current harmonics are introduced. These current harmonics result in various problems such as Low efficiency, Power system voltage fluctuations, and Low power factor. One of the viable solutions to eliminate harmonics are power factor correction techniques such as passive power filters and active power filters. Passive filters are easy to design simple structure and economical but passive filters have many disadvantages, such as Resonance, fixed compensation character, large size, possible overload. To defeat the above disadvantages due to Passive Filters, Active Power Filters (APFs) have been proposed as a current-harmonic compensator. The Active Power Filter is connected in shunt
With nonlinear load. This approach is based on the principle
of injecting harmonic current into the ac system, of the same amplitude but of reverse phase to that of the load current harmonics. This will more result in unity power factor and sinusoidal line currents in the input power system. In this case, a minor portion of the energy is processed, which may result in overall higher power processing potential and overall higher energy efficiency [1]. These types of approaches are applicable for low-power (less than 5kVA) to high-power applications (around 100kVA). A three-phase shunt APF is generally comprises of a three-phase bridge converter and control circuitry. Most of the earlier used control approaches require sensing the load current and calculating its harmonics and reactive components in order to produce the position for scheming the current of the bridge converter [2]. Those control methods require fast and accurate calculation; therefore, high-performance A/D converters and high-speed digital microprocessors are required, which yields high cost, low stability and complexity. So here introducing a promising solution based on One-Cycle control.

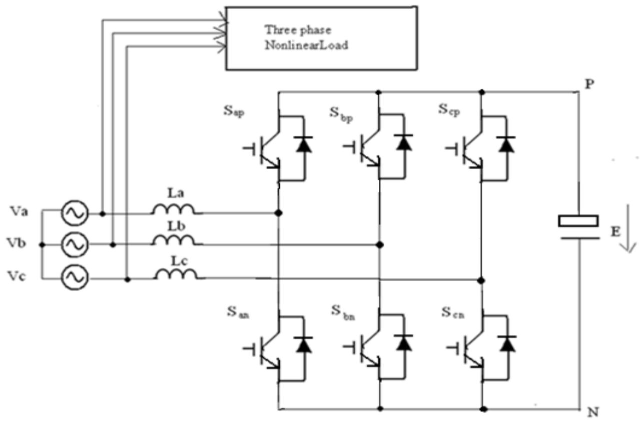
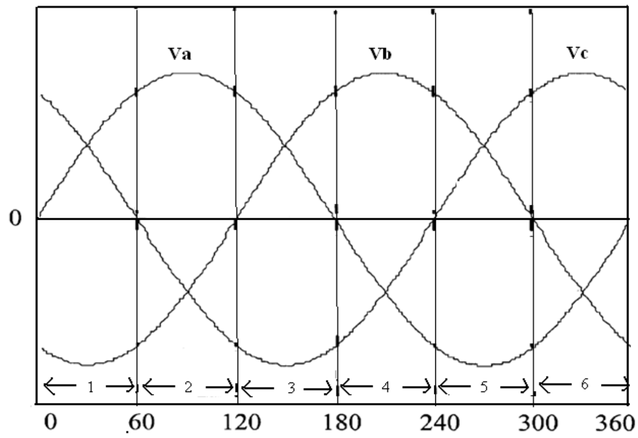


Fig.1 Power stage of the three-phase APF

The one-cycle control system do not require any use of multipliers in the control loop and the need for calculating current reference. The control circuitry is simple and more reliable. In pulse width-modulation active power filter, all the
switches are run with switching frequency; hence, the switching losses are relatively much and more than that of the vector based active power filters. In this reference, a three-phase parallel filter with six-switch bridge voltage-source
Fig.2 Normalized three-phase grid voltage waveforms converter is operated.

Fig:2 Parallel Connected Dual Converter

PRINCIPLE OF OPERATION



The waveforms of these three-phase voltages Va, Vb and Vc of the grid are shown in above Fig..2 During the each 60^o region in Fig.2, the converter in Fig.1 is decoupled as a parallel-connected dual-boost converter.

Here the total 360^o region is divided into six different regions as shown in Fig.2. In every region the two grid voltages remain either positive or negative, one voltage is either positive or negative, and depending on this condition the total 360^o region is further divided into six regions. In each respective region the voltage-source converter can be decoupled as a dual-boost converter as like explained in the further section. In the region ($0{\sim}60^o$), the phase voltage V_b is the least. In this case, switch S_{bp} is completely kept off and switch S_{bn} is kept on throughout the whole complete 60^o region, while switches in the extra two branches such as S_{an},S_{ap} and S_{cn},S_{cp} are controlled in complement way at the switching frequency. For example, during the respective switching cycle, if switch S_{an} is OFF, switch S_{ap} will be fully ON and vice versa. Here line frequency is much more lesser than the switching frequency. Here the Line frequency is 60 HZ and Switching frequency is around 50 KHZ.

Characteristics of the proposed system

For the dual-boost converter shown below in Fig.4 or 5, four different types of switching states are possible for the two switches T_p and T_n . The four types of switching states and the inductor voltages are shown in Table 1.

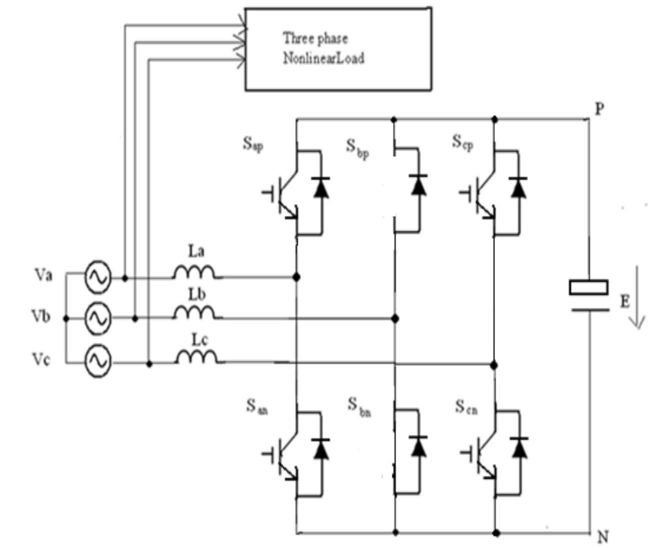


Fig.3 Power stage of a three-phase APF during $0{\sim}60^o$ regions.

Where as

$$\begin{bmatrix} V_p^* \\ V_n^* \\ V_t^* \end{bmatrix} = \begin{bmatrix} \frac{2}{3} & \frac{-1}{3} \\ \frac{-1}{3} & \frac{2}{3} \\ \frac{1}{3} & \frac{1}{3} \end{bmatrix} \cdot \begin{bmatrix} V_p \\ V_n \end{bmatrix} \quad \text{----------} \textbf{(1)}$$

For a three-phase shunt filter with a constant switching frequency, only two switching sequences are possible, i.e., I, II, IV (condition $d_p > d_n$ where d_p , d_n are the duty ratios of the switches , T_p, T_n respective) or I, III, IV (condition $d_p < d_n$) during the every switching cycle, if trailing-edge modulation is done.

State	T_p	T_n	$\overline{T}_p$	$\overline{T}_n$	V_{Lp}	V_{Ln}	V_{Lt}
I	ON	ON	OFF	OFF	V_p^*	V_n^*	V_t^*
II	ON	OFF	OFF	ON	$V_p^* + \frac{1}{3}\cdot E$	$V_n^* - \frac{2}{3}\cdot E$	$V_t^* - \frac{1}{3}\cdot E$
III	OFF	ON	ON	OFF	$V_p^* - \frac{2}{3}\cdot E$	$V_n^* + \frac{1}{3}\cdot E$	$V_t^* - \frac{1}{3}\cdot E$
IV	OFF	OFF	ON	ON	$V_p^* - \frac{1}{3}\cdot E$	$V_n^* - \frac{1}{3}\cdot E$	$V_t^* - \frac{2}{3}\cdot E$

The voltage waveforms along the inductors L_p, L_n, L_t are shown in Fig.6 .For first switching sequence (d_n<d_p). Based on the above assumption that line frequency is much lesser than the switching frequency, inductor voltage-second balance is approximately applicable, that is

$$V_P^* d_n + \left(V_P^* + \tfrac{1}{3}E\right).(d_p - d_n) + \left(V_P^* - \tfrac{1}{3}E\right).(1 - d_p) = 0$$
$$V_n^* d_n + \left(V_n^* - \tfrac{2}{3}E\right).(d_p - d_n) + \left(V_n^* - \tfrac{1}{3}E\right).(1 - d_p) = 0$$
-----(2)
$$V_t^* d_n + \left(V_t^* - \tfrac{1}{3}E\right).(d_p - d_n) + \left(V_t^* - \tfrac{2}{3}E\right).(1 - d_p) = 0$$

The following mentioned equation is valid for a symmetrical three-phase system:

$$V_p^* + V_n^* - V_t^* = 0 \qquad \text{--------------------- (3)}$$

From (2) and (3) with further simplification we will get

$$\begin{bmatrix} (1-d_p) \\ (1-d_n) \end{bmatrix} = \begin{bmatrix} 2 & 1 \\ 1 & 2 \end{bmatrix} \cdot \begin{bmatrix} \frac{V_p^*}{E} \\ \frac{V_n^*}{E} \end{bmatrix} \qquad \text{------------}$$

It has been verified that the above equation is valid for the other switching sequence I, III, and IV ($d_p<d_n$) also.

Equation (4) gives the inherent relationship between the duty cycle and the input, the output voltage for the shunt-connected dual-boost converter.

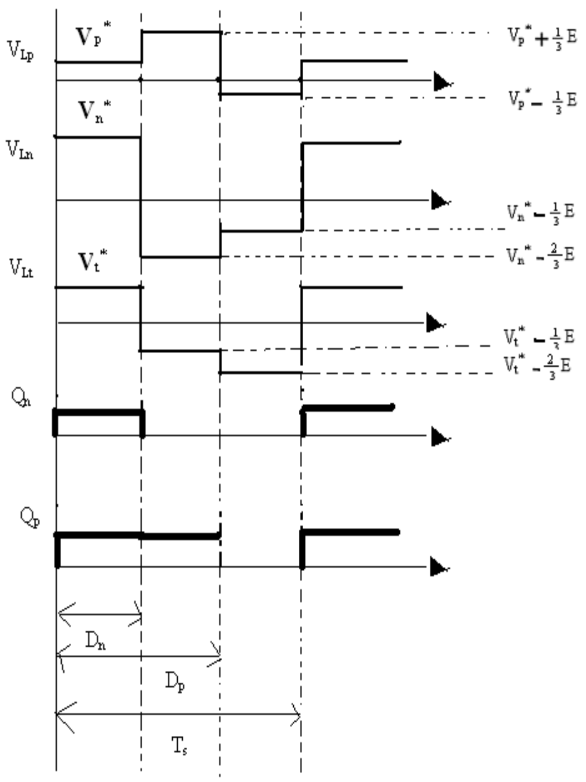


Fig.4 Inductor voltage waveforms for the converter during condition $d_p > d_n$

In Fig.6 V_{Lp}, V_{Ln}, and V_{Lt} represents the different voltages across inductors V_p , V_n , and V_t ,respectively. Q_p and Q_n are the signals for switches T_p and T_n respectively.

PROPOSED ONE-CYCLE CONTROLLER FOR THREE-PHASE APF

For the unity-power-factor of three-phase APF, the main control goal is to force the grid line current in each phase so to follow the corresponding sinusoidal phase voltage, i.e.,

$$\left.\begin{aligned} V_a &= Re \cdot i_a \\ V_b &= Re \cdot i_b \\ V_c &= Re \cdot i_c \end{aligned}\right\} \text{------------------- (5)}$$

where Re is a emulated resistance that reflects real power of the load. The control goal can be achieved by controlling the equivalent currents i_p and i_n to follow the voltages V_p^* and V_n^*. The control equation of the three-phase APF can be rewritten as below

$$\left.\begin{aligned} V_p^* &= Re \cdot i_p \\ V_n^* &= Re \cdot i_n \end{aligned}\right\} \text{---------------- (6)}$$

Substituting (6) into (4) and consider as the switch is ON for the whole 60^o region, it is achieved as

$$\left.\begin{bmatrix} (1-d_p) \\ (1-d_n) \end{bmatrix} = \frac{Re}{E\,Rs} \cdot Rs \cdot \begin{bmatrix} 2 & 1 \\ 1 & 2 \end{bmatrix} \cdot \begin{bmatrix} i_p \\ i_n \end{bmatrix}\right\}_{d_t=1} \text{------------- (7)}$$

Defining

$$Vm = \frac{E\ Rs}{Re} \qquad \text{------------ (8)}$$

Where signal V_m can be generated from output voltage feedback compensator which is used to regulate that output voltage of the capacitor E of voltage source converter consequently to its load level; Rs is its equivalent current sensing resistance and it is fixed constant. Merging of these two equations 7,8 and control key equation is obtained as

$$\left.Vm \cdot \begin{bmatrix} (1-d_p) \\ (1-d_n) \end{bmatrix} = Rs \cdot \begin{bmatrix} 2 & 1 \\ 1 & 2 \end{bmatrix} \cdot \begin{bmatrix} i_p \\ i_n \end{bmatrix}\right\}_{d_t=1} \text{ - (9)}$$

This equation indicates that three-phase power factor can be derived by controlling duty ratios of the switches so that first-order polynomial equation (9) can be satisfied. This can easily be recognized by the one-cycle control core as shown in Fig.7. The operation waveforms are shown below in Fig.8.

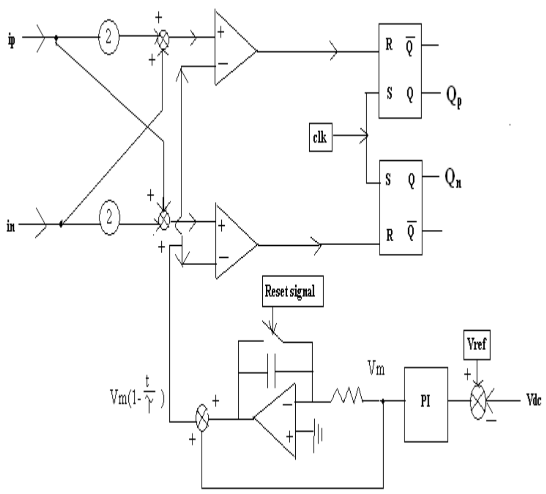


Fig.5 One- cycle control logic

In the beginning of the each cycle, clock pulse sets the two flip-flops. The currents i_n and i_p from current selection logic is linearly mixed to form an input to each of the two comparators. At further the input of the two comparators is the value of Vm and the subtracted integrated value of V_m. Signal $V_m(1\text{-}t/T_s)$ is compared with $R_s(i_p+2\ i_n)$ in lower comparator and is compared with $R_s(2\ i_p+\ i_n)$ in upper comparator as shown in Fig.7. When the two inputs of a comparator happen to meet as shown in Fig.8, the comparators change its state, which reset the corresponding flip-flop. As an outcome result, the correspondent switch is turned off. Therefore, the duty ratios d_p and d_n are determined for a respective switch in each switching cycle.

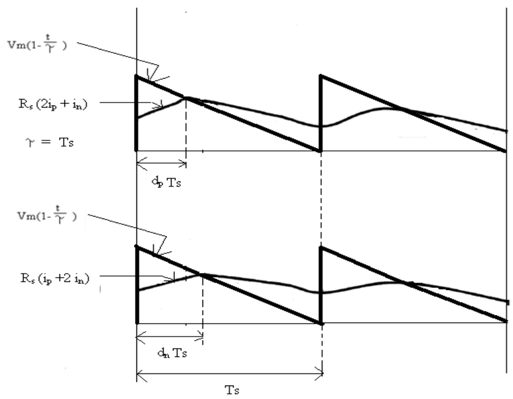


Fig.6 Operational waveforms of the One-cycle controlled APF controller.

The One-Cycle Control Approach has below Features

- Three-phase unity-power-factor and low total harmonic distortion (THD) can be obtained by one integrator with reset ansd also several logic and linear components. It is reliable and simple.
- Only ac main current and the voltage zero-crossing points are sensed. No sensors for load current and the APF inductor currents are required.
- No requirement to calculate the reference for APF inductor current so that complicated digital computation is removed.
- No need of the multipliers.
- Constant switching frequency, which is more desirable for industrial applications, is obtained.
- For the three-phase converter, only the two switches are operated at higher frequency, and the switching losses are reduced compared to the PWM converter.

THE DESIGN CONSIDERATIONS

Dc-Link Capacitor Design values

The output dc-link capacitor of the voltage source converter is obtained by its output voltage ripple. The equation is as below

$$c \geq \frac{P_o}{2 * f_{line} * (V^2_{omax} - V^2_{omin})} \quad \text{-- (10)}$$

Suppose power is 7000 W for example; with 2% ripple content, APF and the output voltage is 400 V. The line frequency is around 60 Hz. The capacitance is measured as 4800µF.

Selection of the APF Inductance

The concept of this recommended control is using one-cycle control to apply the above control key equation as follows:

$$R_s.i_{eq} = V_{m.}(1\text{-}d) \quad \text{-------------- (11)}$$

Where as

$$i_{eq} = (2i_p + i_n) \quad \text{(OR)} \quad (i_p + 2\ i_n)$$

Similar to peak current model control, there is a convergence condition.

The stability condition is given as

$$m_c \geq \frac{(m_2 - m_1)}{2} \quad \text{------------- (12)}$$

where m_2 is a OFF slope of the input current and m_1 is a ON slope of the input current; m_c is the equivalent slope of the carrier signal, which is implemented by an integrator with reset. Taking these into considerations that the load current is at low frequency and influence of load current can be ignored, we only consider that the inductor current in the stability analysis, we have further

Where as

$$m_1 = Rs \cdot \frac{V_g}{L}$$

$$m_2 = Rs \cdot \frac{V_o - V_g}{L} \quad \text{------------ (13)}$$

$$m_c = \frac{V_m}{\gamma} = \frac{V_m}{Ts}$$

$$\gamma = Ts$$

Substitution of (13) into (12) yields a convergence condition

$$V_m \geq \frac{Rs \cdot Ts}{2L} \cdot (V_o - 2|V_g|) \quad \text{---------- (14)}$$

$$\geq \frac{Rs \cdot Ts}{2L} \cdot (V_o - 2 \cdot V_{grms}|\sin(wt)|)$$

this condition of convergence is dependent on a angular angle of input voltage and the V_m and wt , which is more related to the output power and input voltage. When this convergence condition is satisfied partly, the system will still be more stable.

According to (14), the convergence condition for the region 0^o~360^o is given as

$$V_m \geq \frac{Rs \cdot Ts}{2L} \cdot V_o \quad \text{----------- (15)}$$

However

$$Vm = \frac{V_o Rs}{Re} \quad \text{------------(16)}$$

V_m is related to the input voltage and output power through the equation (16) . It can further be rewritten as

$$V_m = \frac{P_o \cdot Rs \cdot V_o}{\eta \cdot V_{grms}^2} \quad \text{-------------(17)}$$

where η is estimated efficiency.

Merging of the above mentioned equations yields

$$L \geq \frac{1}{2} \cdot \eta \cdot Ts \cdot \frac{V_{grms}^2}{P_o} \quad \text{--------------(18)}$$

The above equation used to conclude the size of inductor. At the full load condition and maximum input voltage condition, the system must completely be stable and then the inductor can be selected as

$$L \geq \frac{1}{2} \cdot \eta \cdot Ts \cdot \frac{\max(V_{grms}^2)}{\max(P_o)} \quad \text{--------(19)}$$

For η=90% , Ts=20μs, max(P_o)=7000W, max(V_{grms})=170 v then the minimum inductance is calculated as L = 250μH.

VI SIMULATION RESULTS

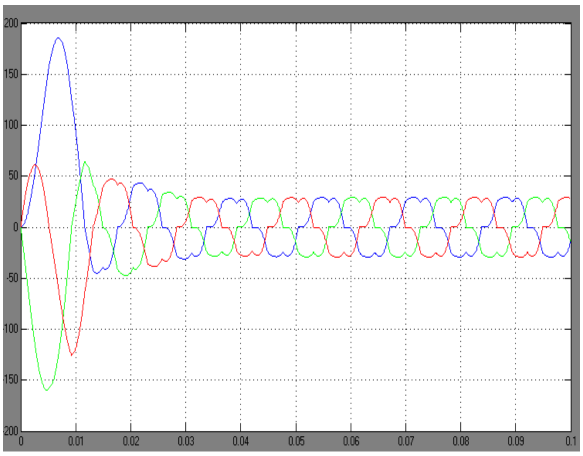


Fig.7 Source current without shunt filter

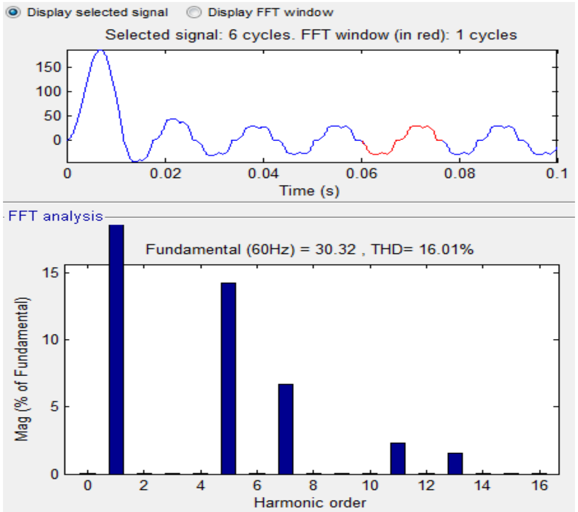


Fig.8 Total Harmonic Distortion without shunt filter

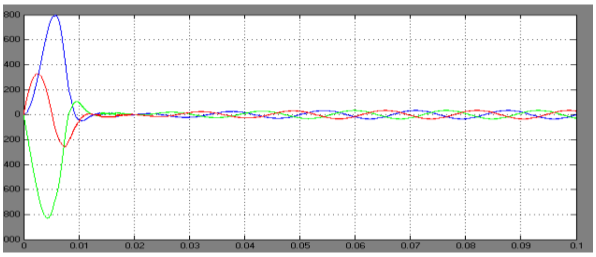


Fig.9 Source current with pi one cycle control

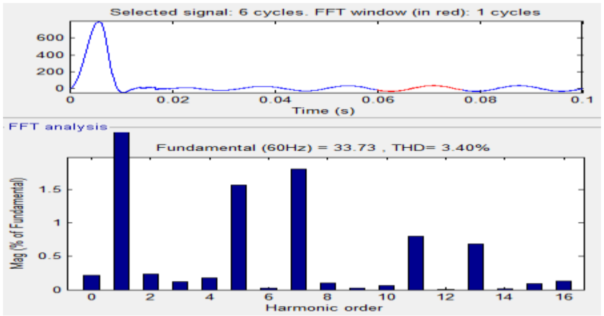


Fig.10 Total Harmonic Distortion with pi one cycle controller

CONCLUSION

A three phase shunt harmonic filter with one cycle PI controller has been designed in this paper. This new controlling method is determined with more reliability and fast harmonic compensating capacity. The main advantage of this method is it senses only source current and voltage but not the load current and voltage. The intensive computation is removed as there is no need of calculating the reference filter inductor current. When a nonlinear load is connected to a three phase source some harmonics are introduced. In order to overcome these harmonics in the source, a pi one cycle control based filter is connected in parallel with the load. The total harmonic distortion without shunt filter is observed to be 16.1%, with PI one cycle control is 3.40%,

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