PROPOSAL OF A HYSTERESIS CONTROLLER WITH CONSTANT SWITCHING FREQUENCY

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Abstract – **This paper introduces a new control technique for power factor correction in ac-dc converters. It has the advantages of conventional hysteresis control strategy and the prominent characteristics PWM control, since the limits of the hysteresis band can be adjusted to provide fixed switching frequency. The operating principles, theoretical analysis and results regarding a single-phase boost converter are presented to validate the proposal.**

Keywords – **hysteresis control, low harmonic distortion, power factor correction.**

I. INTRODUCTION

The growing number of nonlinear loads such as diode or thyristor rectifiers, switch-mode power supplies and adjustable speed drives generates harmonic currents causing various problems to other equipment connected to the point of common coupling. Typical problems are overheated machines, transformers and power cables, current flow through the neutral conductor, flicker effects, and malfunctioning of sensitive devices. The reduction of the harmonic content and also high power factor are desirable aspects in ac-dc converters, because they are potential harmonic sources and may affect power quality [1].

Throughout the years, two switching techniques have become popular: PWM and hysteresis modulation [2]. The hysteresis control introduces a minor error in the average input current and provides better dynamic response than the PWM control, but an inherent drawback is the variable switching frequency. Therefore the control circuit is supposed to be designed for a large band, otherwise low frequency harmonics will result [3] [4].

Hysteresis control is essentially an analogic technique. Despite the advantages given by the digital controls, in terms of interfacing, maintenance, flexibility, and integration, their accuracy and response speed are often inadequate for current control in highly demanding applications, such as active filters and high-precision drives [5]-[7]. Indeed, in these applications, current reference waveforms characterized by high harmonic content and fast transient response must be followed by good accuracy. In these cases, the hysteresis method can be a good solution, provided some improvements are introduced to overcome its main limitations, which are the variations of the switching frequency and the sensitivity to phase commutation interferences. To this purpose, a variety of alternatives, both analog and digital, have been proposed by several authors [8]-[15].

However, when high switching frequency is demanded, analog solutions offer the fastest performance with a relatively simple implementation. A fully analog technique, which eliminates the interference and gives constant switching frequency, was presented in [8]. This technique was extensively used and proved to be robust and reliable.

This paper introduces a novel control technique that controls the input current, allowing ac-dc converters to operate with constant switching frequency with the advantages of conventional hysteresis and PWM controllers. The control system can be implemented using analog circuits, and the proposal is discussed as follows.

II. HYSTERESIS CONTROLLER WITH CONSTANT SWITCHING FREQUENCY

A. Theoretical Background

The basic principle of the proposed technique consists in determining the ideal time interval between the switching instant and the crossing instant of the reference and input currents. The input current is then supposed to oscillate around the reference current with fixed frequency.

Fig. 1. shows the behavior of the input current in an incremental time interval, where the reference current is almost constant, since the switching frequency is much greater than the line frequency.

This control strategy can be applied to most of the single-phase ac-dc converters, without any essential change in the original topology [16]. Furthermore, the study is also extended to a three-phase boost converter in [16]. However, the analysis carried out here considers a conventional single-phase boost converter operating in continuous conduction mode due to simplicity, because it has only two operating stages.

In order to analyze the proposed control strategy, the circuit shown in Fig. 2 must be considered, where it can be seen that the converter operation can be defined in two stages. Voltage V_x in the first and second stages is defined as V_{x1} and V_{x2}, respectively, but such values depend on the converter in question. For instance, if the converter is a full-bridge topology, voltages V_{x1} and V_{x2} will be $+V_o$ and $-V_o$, respectively [16]. For a boost converter, voltages V_{x1} and V_{x2}, will be $+V_o$ and null, respectively.

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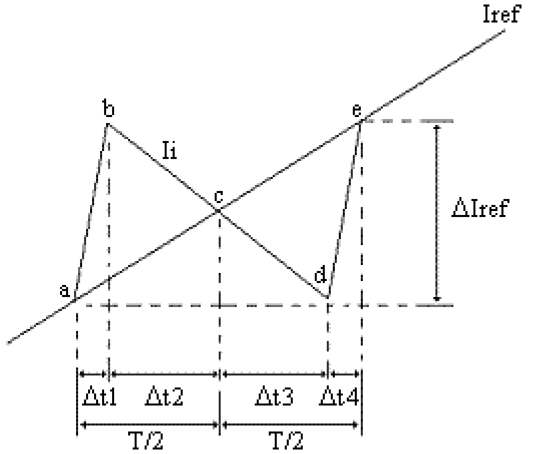


Fig. 1. Input current profile within a switching period.

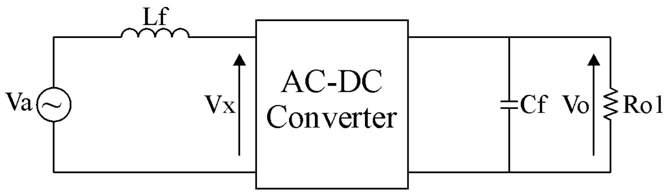


Fig. 2. Generic power converter with two operating stages.

However, independently of the topology, voltage V_{x1} will always be greater than the input voltage, and voltage V_{x2} is less than the input voltage or null. Additionally, if the switching frequency is considered much greater than the line frequency, the behavior of the current through the filter inductor can be described according to Fig. 1, which corresponds to an ideal situation i.e. triangles *abc* and *cde* have the same area. In order to maintain this condition, it is necessary to determine Δt_1 and Δt_3, which correspond to the time intervals during which the switch is turned on and off, respectively. Such intervals are determined as a function of the circuit parameters, in order to assure constant switching frequency.

Expressions (1) and (2) represent the increasing and decreasing rates of the input current, respectively.

$$-V_{x2}+V_a=L_f\cdot\frac{dI_i}{dt} \tag{1}$$

$$-V_{x1}+V_a=L_f\cdot\frac{dI_i}{dt} \tag{2}$$

If the switching frequency is considered much greater than the frequency of the reference current, output voltage and (dI_{ref}/dt) ratio are constant. In addition to this, the increasing and decreasing rates of the current through the filter inductor are constant. Therefore triangles *abc* and *cde* have the same area and the following expressions are valid.

$$\Delta t_2 \cong \Delta t_3 \tag{3}$$

$$\Delta t_1 \cong \Delta t_4 \tag{4}$$

where:

Δt_1 – time interval during which the current derivative is positive, from the crossing point with the reference current to the instant of commutation;

Δt_2 – time interval during which the current derivative is negative, from the instant of commutation to the crossing point with the reference current;

Δt_3 – time interval during which the current derivative is negative, from the crossing point with the reference current to the instant of commutation;

Δt_4 – time interval during which the current derivative is positive, from the instant of commutation to the crossing point with the reference current.

Analyzing Fig. 1, expressions (5) and (6) result.

$$\Delta t_1+\Delta t_2=T/2 \tag{5}$$

$$\int_t^{t+\Delta t_1}\frac{(-V_{x2}+V_a)\cdot dt}{L_f}+\int_{t+\Delta t_1}^{t+T/2}\frac{(-V_{x1}+V_a)\cdot dt}{L_f}=\frac{\Delta I_{ref}}{2} \tag{6}$$

where ΔI_{ref} is the reference current variation, and T is the switching period. Assuming (dI_{ref}/dt) constant in the switching period, ΔI_{ref} can be given as:

$$\Delta I_{ref}=T\cdot\frac{dI_{ref}}{dt} \tag{7}$$

Finally, from expressions (5) to (7), interval Δt_1 is obtained.

$$\Delta t_1=\frac{T}{2\cdot(V_{x1}-V_{x2})}\cdot\left(V_{x1}-\frac{2}{T}\int_t^{t+T/2}V_a\cdot dt+L_f\cdot\frac{dI_{ref}}{dt}\right) \tag{8}$$

If the same procedure is followed, interval Δt_2 can also be calculated.

$$\Delta t_2=\frac{T}{2\cdot(V_{x1}-V_{x2})}\left(-V_{x2}+\frac{2}{T}\int_t^{t+T/2}V_a\cdot dt-L_f\cdot\frac{dI_{ref}}{dt}\right) \tag{9}$$

From time intervals Δt_1 and Δt_2, it is possible to define the upper and lower limits of the hysteresis band, which according to Fig. 3 must be symmetrical to the reference current so that constant switching frequency is achieved.

$$I_{ref(up)}=\frac{1}{L_f}\cdot\int_0^{\Delta t_1}\left(-V_{x2}+V_a-L_f\cdot\frac{dI_{ref}}{dt}\right)\cdot dt+I_{ref} \tag{10}$$

$$I_{ref(low)}=\frac{1}{L_f}\cdot\int_0^{\Delta t_2}\left(V_{x1}-V_a+L_f\cdot\frac{dI_{ref}}{dt}\right)\cdot dt+I_{ref} \tag{11}$$

B. Definition of The Hysteresis Band Considering A Boost Converter

The operation of a conventional single-phase boost converter can be defined in two stages. When switch S is turned off, the voltage across the filter inductor is (V_a-V_o), otherwise it is V_a.

In this case, voltages V_{x1} and V_{x2} are $+V_o$ and null, respectively. Equations (8) and (9) can be simplified as follows:

$$\Delta t_1=\frac{T}{2\cdot V_o}\cdot\left(V_o-V_a+L_f\cdot\frac{dI_{ref}}{dt}\right) \tag{12}$$

$$\Delta t_2=\frac{T}{2\cdot V_o}\cdot\left(V_a-L_f\cdot\frac{dI_{ref}}{dt}\right) \tag{13}$$

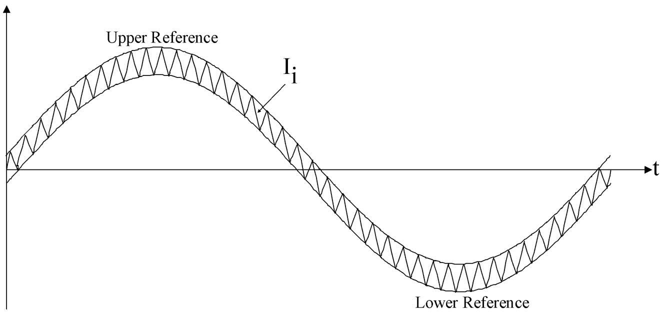


Fig. 3. Hysteresis band.

It can be demonstrated that a maximum relative error due to the conditions given in (3) to (5) results, as follows:

$$\text{Error}(\Delta t_1)=\frac{\omega\cdot T\cdot V_a\cdot\cos(\omega t)}{2\cdot\left(V_o-\int_t^{t+T/2}\left|V_a\right|\cdot dt+L_f\cdot\frac{dI_{ref}}{dt}\right)} \tag{14}$$

$$\text{Error}(\Delta t_2)=\frac{\omega\cdot T\cdot V_a\cdot\cos(\omega t)}{2\cdot\left(\int_t^{t+T/2}\left|V_a\right|\cdot dt-L_f\cdot\frac{dI_{ref}}{dt}\right)} \tag{15}$$

where ω is the line frequency. Since Δt_1 and Δt_2, represent the increasing and decreasing rates of the input current in Fig. 3, respectively, the relative error is consequently related to the switching frequency variation.

From (12) and (13), it is possible to define the upper and lower limits of the hysteresis band as (16) and (17), respectively. According to Fig. 3, they must be symmetrical to the reference current so that constant switching frequency is achieved.

$$I_{ref(up)}=I_{ref}+\frac{T}{2\cdot V_o\cdot L_f}\cdot\left(V_o-V_a+L_f\cdot\frac{dI_{ref}}{dt}\right)\cdot\left(V_a-L_f\cdot\frac{dI_{ref}}{dt}\right) \tag{16}$$

$$I_{ref(low)}=I_{ref}-\frac{T}{2\cdot V_o\cdot L_f}\cdot\left(V_o-V_a+L_f\cdot\frac{dI_{ref}}{dt}\right)\cdot\left(V_a-L_f\cdot\frac{dI_{ref}}{dt}\right) \tag{17}$$

C. Implementation of The Controller

Fig. 4 shows the simplified diagram representing the controller applied to a boost converter. Some aspects must be considered about the implementation of the analogic circuit.

When the input voltage becomes null, and the reference current starts increasing, the increasing rate of the current through the filter inductor is not enough to follow the reference current, until V_a/L_f ratio becomes greater than (dI_{ref}/dt). This can be explained because the increasing rate of I_i depends only on the input voltage. Additionally, this will cause the switching frequency to be reduced at this point, since switch S remains turned on, until the input current equals the reference current.

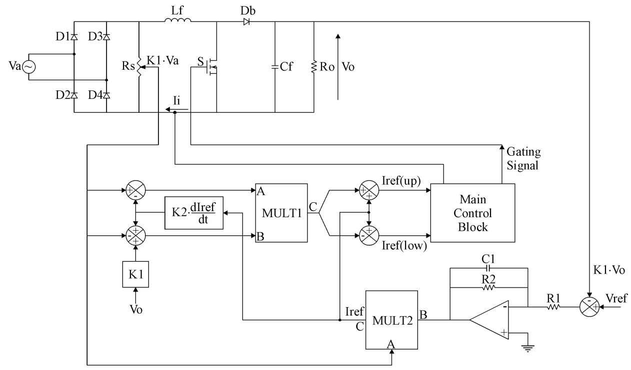


Fig. 4. Proposed controller associated with a boost converter.

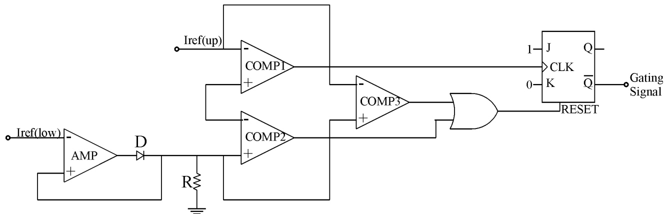


Fig. 5. Representation of the main control block.

Since the current through the filter inductor can not flow in the reverse direction, the lower limit of the hysteresis band is always supposed to be a positive value. If $I_{ref(up)}$ becomes less than $I_{ref(low)}$, switch S must remain turned on.

Reference signals are obtained using sum and multiplier circuits. Parameter K_2 is proportional to inductor L_f and constant K_1, and it must be defined in order to keep voltages $V_{\Delta t1}$ and $V_{\Delta t2}$ within the band limits. The switching frequency is set by constant K_1 (18), which is squared by the multiplier circuit.

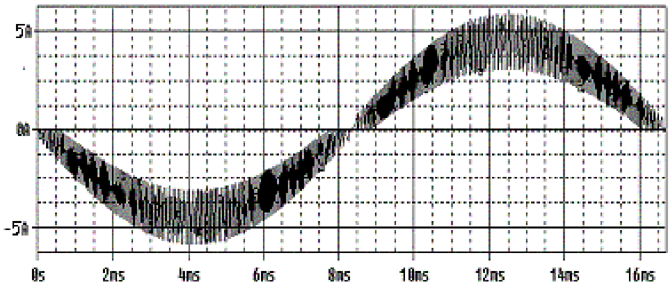
$$K_1=\sqrt{\frac{T}{2\cdot L_f\cdot V_o}} \tag{18}$$

Fig. 5 shows the discrete representation of the main control block, whose operating principle is quite simple. When the input current is less than the reference current, comparator *COMP1* drives the *CLK* input of the JK flip-flop, setting the output to high. When the input current reaches the upper limit of the band, comparator *COMP2* drives the *RESET* input of the *JK* flip-flop again, consequently setting the output to low.

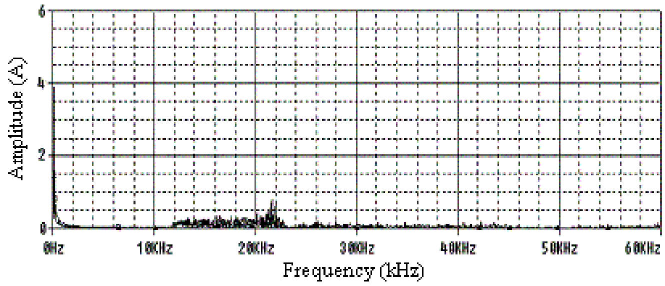
III. SIMULATION RESULTS

Simulation tests were performed on a boost converter to demonstrate the controller performance. Conventional hysteresis control was also implemented to establish an eventual comparison with the proposed technique. The average switching frequency is the same in both cases, as the parameters set in Table I is employed.

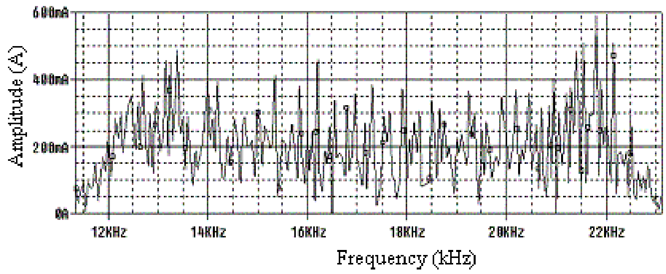
Fig. 6 and Fig. 7 show results concerning the conventional hysteresis controller and the proposed one, respectively. Fig. 6 (a) and Fig. 7 present the input current waveform. Fig. 6 (b) and Fig. 7 (b) correspond to the frequency spectrum of the input current, where switching frequency variation is evidenced. In Fig. 6 (c) and Fig. 7 (c), one can see that such variation is much greater for the conventional hysteresis controller.



(a) Input current

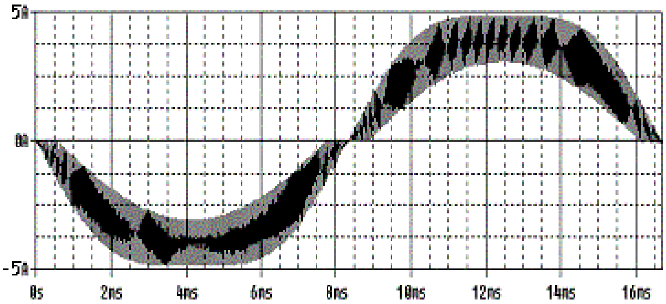


(b) Frequency spectrum of the input current

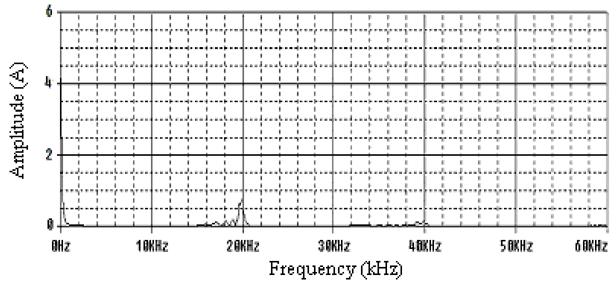


(c) Detailed view indicating the switching frequency variation

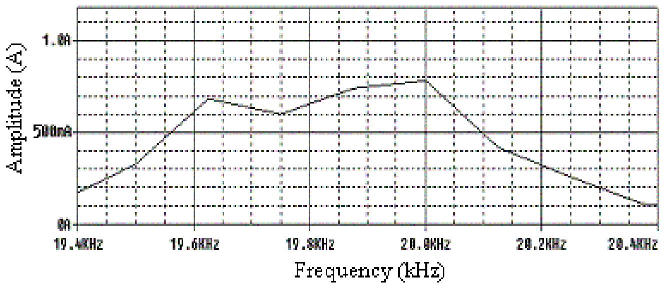
Fig. 6. Simulation results obtained with the conventional hysteresis controller.



(a) Input current



(b) Frequency spectrum of the input current



(c) Detailed view indicating the switching frequency variation

Fig. 7. Simulation results obtained with the proposed hysteresis controller.

Table I
Parameters set used in the tests

Parameter	Value
Input voltage	V_i=127Vrms
Output voltage	V_o=250Vdc
Filter inductor	L_f=2.1mH
Filter capacitor	C_f=1000µF
Load current	I_o=2A
Output Power	P_o=500W
Average switching frequency	f_s=20kHz
Switch *S*	IRFP264
Boost diode D_b	MUR860

Additionally, the current *THD* considering the harmonic content to the fiftieth order is 3.7% for the conventional controller, and 1.5% for the proposed one.

IV. EXPERIMENTAL RESULTS

In order to validate the theoretical assumptions, and also the results in Section III, an experimental prototype of the boost converter associated with the proposed controller was implemented. Results are presented and discussed as follows.

Fig. 8 shows the rectified input current obtained from a current sensor. This waveform represents the hysteresis band itself.

Fig. 9 corresponds to the input voltage and input current waveforms, where power factor correction is evidenced.

Fig. 10 (a) and (b) represent the harmonic content of the input voltage and input current, respectively, as voltage *THD* is 2.45% and current *THD* is 5.74%.

Finally, Fig. 11 shows the frequency spectrum of the input current, where the switching frequency variation can be seen.

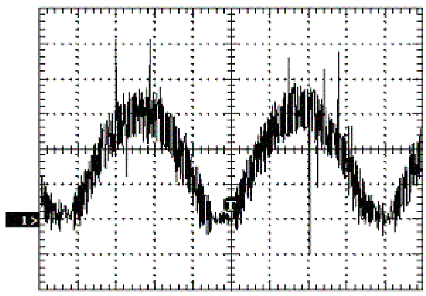


Fig. 8. Rectified input current.
Scales: I_i – 200mA/div.; time – 2ms/div.

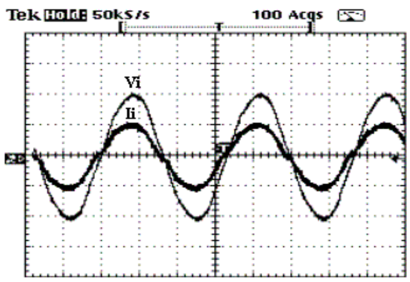
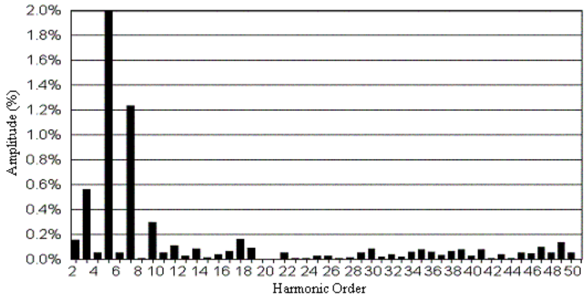
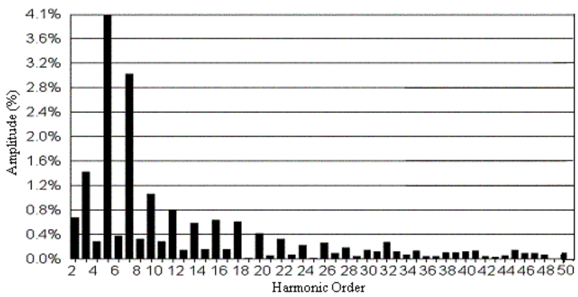


Fig. 9. Input voltage and input current.
Scales: V_i – 100V/div.; I_i – 5A/div.; time – 5ms/div.

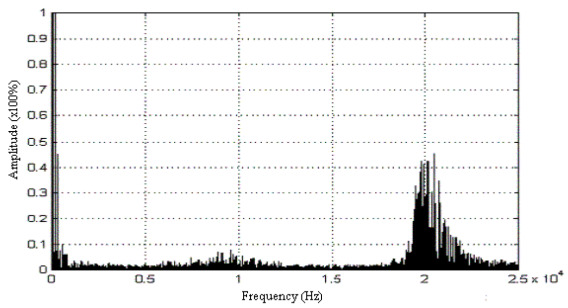


(a) Input voltage

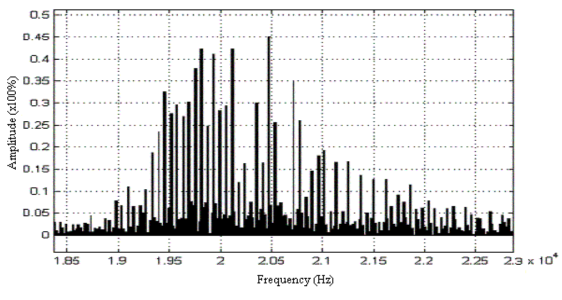


(b) Input current

Fig. 10. Harmonic content.



(a) Frequency spectrum



(b) Detailed view indicating the switching frequency variation

Fig. 11. Frequency spectrum of the input current.

V. CONCLUSION

This paper has presented a novel control technique that has the advantages of conventional hysteresis and PWM control techniques. If the upper and lower limits of the hysteresis band are modified, constant switching frequency can be achieved.

As it can be seen in the tests, the switching frequency variation is reduced, as expected in the mathematical study. It is also possible to say that the relative error decreases when the average switching frequency increases, as it may become negligible.

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BIOGRAPHIES

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