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SERIES RESONANT CONVERTER CONTROL: COMPLETE TUNING OF AN IP REGULATOR

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Abstract - In this paper, the authors deal with the closed-loop behaviour of DC-DC series resonant converters, in which frequency is controlled through an IP regulator. Using a small signal modelling method, linear control strategies have been designed and their closed-loop performances, in particular their robustness, have been determined and compared. This study showed that the IP controller presents good characteristics. So, we propose to expose general rules to tune up a control structure with an IP regulator, yielding to a stable closed-loop behaviour over the whole range of operation. In particular, the large signal operation of a controlled in such a way converter is studied. All these rules are validated on an experimental converter.

INTRODUCTION

In the electrical power conversion domain, the resonance utilization advantages need no longer be demonstrated [1]. However, the presence of a high frequency resonant circuit, which is compulsory in these structures, makes their operation more complex.

In this article, we are mainly concerned by the non reversible DC-DC series resonant converter, operating above its natural frequency. The energy transfer can only be achieved by adjusting the switching instants of the inverter.

If different control methods (frequency control, transistor conduction angle control, ASDTI control, resonant voltage control, optimal trajectory control) are now well known for this particular structures [2], the control implementation by adjusting directly the switching frequency remains the simplest and so, the kind of control has still its place in industrial applications in spite of its well known disadvantages (slow and oscillatory responses, strongly depending on the operating point).

Therefore, the control by using standard linear regulators requires a particular attention.

The choice of the linear control structures derives from the utilization of informations on a small signal model. A theoretical study of the closed-loop behaviour allows evaluating the robustness of this control for this high non-linear system.

A comparison of the performances of each control structures have shown that the IP regulator presents good characteristics. That is why, we propose to give general rules to tune up an IP regulator, yielding to a stable operation over the whole range of frequency.

REVIEW OF THE MODELLING RESULTS

Studied structure

The ideal structure of the non reversible DC-DC series resonant converter is reminded in figure 1. This converter allows generating a DC voltage from another DC voltage source through a high frequency AC link (resonant tank).

![Structure of the non reversible series resonant converter operating above resonance](image)

For any value of $f_s$, the inverter switching frequency, this type of converter always operates in a step-down mode. In our case, we are concerned only with switching frequencies higher than the natural frequency and current reversible bloccable switching devices are required (ZVS mode).

Notations : Normalized Units

The different analysis of the series resonant converter use the following basic units :
- $E$ supply voltage of a complete bridge inverter or half-supply voltage of a half-bridge inverter,
- Characteristic impedance of the resonant circuit $(L/C)^{1/2}$,
- Current $i_f = E (Q/L)^{1/2}$,
- Natural frequency of the resonant circuit $f_0 = \frac{1}{2\pi\sqrt{LC}}$.

The following normalized units can be determined :
- $y$ : normalized current in the resonant circuit $i_f/\beta$,
- $x$ : normalized voltage across capacitor C $V_C/E$,
- $y_{avg}$ : normalized average load current $i_{avg}/\beta$,
- $q$ : normalized load voltage $v_f/E$,
- $u$ : normalized operating frequency $f_u/f_0$,
- $a = C/Q :$ normalized output capacitance,
- $p = (Q/L)^{1/2} :$ normalized load.

These units enable to manipulate normalized equations, characterizing the structure instead of a particular circuit.

Small letters will be used for variable quantities and capital letters will be used for steady quantities.

Modelling : converter behaviour

Considering the output voltage properly smoothed, a study of the state plane diagram, at U constant, allows determining all the steady state quantities. A network of characteristics $Q(U)$ with fixed $p$ can be obtained [1]:

The ideal structure of the non reversible DC-DC series resonant converter is reminded in figure 1. This converter allows generating a DC voltage from another DC voltage source through a high frequency AC link (resonant tank).
A modelling method consists in studying (while modulating the frequency) the evolution of the state variables by writing the state equations and by stating their continuity at switching times [3].

This method allows to obtain a transfer function, that links the output voltage variation to the switching frequency variation, with a second order at the numerator, a third order at the denominator [4]:

\[
\begin{align*}
\hat{\omega}_n(p) &= \frac{G}{(1 + 2p\omega_n + p^2/\omega_n^2)} \\
\hat{\omega}_d(p) &= \frac{1}{(1 + p/\omega_d)(1 + 2p/\omega_d + p^2/\omega_d^2)}
\end{align*}
\]

(2)

These model parameters (steady gain, zeros and poles) significantly evolve with the operating point.

**Steady gain** - G is only depending on the switching frequency and on the connected load. For U higher than 1.2, at a constant load, G is a decreasing function.

**Zeros** - They are conjugate ($\zeta > 1$) or real ($\zeta < 1$).

They can always be neglected since $\omega_z$ is very high in comparison to the poles pulsations. Also, the transfer function can be reduced to:

\[
\hat{\omega}_n(p) = \frac{G}{(1 + p/\omega_n)(1 + 2p/\omega_n + p^2/\omega_n^2)}
\]

(3)

**Poles** - There are always a real pole and two conjugate poles ($\zeta < 1$):

- $\omega_p$ is a decreasing function of U and tends to the $1/R_sC_s$ limit when U increases. It is close to this limit when U > 1.2.
- $\omega_z$ can be approximated, whatever the operating point, by $2\pi f_0(U-1)\,\text{rad/s}$. The damping ratio $\zeta$ essentially depends on the connected load and on the frequency, it decreases when U increases and tends to $p/5$ for high values of U.

Consequently, the real pole is mainly due to the output filter and the conjugate poles to the resonant circuit. However, there is an interaction between the both dynamics and these poles significantly evolve with the operating point. According to their relative placement, the converter behaviour is completely different. This behaviour study consists in considering the ratio between the pulsations $\omega_p$ and $\omega_z$ [4 - 5].

The converter transfer function can be reduced if one dynamics is enough dominant. In particular, if $\omega_p/\omega_n < 0.1$, the output filter dynamics is dominant and the converter can be assimilated to a first order system. We have this phenomenon especially for high frequencies of operation. This operation zone principally depends on the output filter time constant ($\tau = R_sC_s$) in comparison with the resonant pulsation.

Another modelling method consists in linearizing the output characteristics and in considering only the output filter dynamics [1]. It permits to determine a first order transfer function. The obtained steady gain has approximately the same value that before. In the same way, the real pole pulsations are comparable in the area where the output filter is preponderant [5]. For the other operating points, this method can give wrong values of this pole, especially when the switching frequency decreases near the natural frequency. This method presents the advantage of giving analytical results (the first requires heavy computations). It leads to:

\[
\omega_h = \frac{1}{\tau} \left( 1 + \frac{P Q \tan^2 \delta}{\delta (1 + \delta Q / P)} \right)
\]

(4)

\[
Q + P \tan \delta \frac{Q^2 - (1 + \delta Q / P)^2}{(1 + \delta Q / P)}
\]

(5)

If the resonant circuit influence can be neglected, this model gives a good approximation of the converter behaviour.

Normally, even if this first order transfer function is sufficient to represent the converter behaviour, this information is only known when analysing the complete model. However, since $\omega_c$ can be analytically calculated and $\omega_p$ approximated by $2\pi f_0(U - 1)$, the type of behaviour can be evaluated only with the simplest modelling method.

If $\omega_c/(2\pi f_0(U - 1)) < 0.1$, the real pole is dominant and the simplest method gives a good approximation of the transfer function. In this case, only the damping ratio of the resonant circuit dynamics can not be analytically calculated.

**SMALL SIGNAL CLOSED-LOOP PERFORMANCE**

According to the type of behaviour and to the forecasted closed-loop performances, different control structures have been envisaged.

We studied simple structures like PI and IP controllers and more complex structures like a PID regulator associated with a phase lead network.

We chose analogic standard regulators. The utilization of a voltage controlled oscillator (VCO) is necessary to control the switching frequency of the converter. Besides, the switching frequency range must be limited. The lower limit is necessary above the resonant frequency and both limits depend on the used smoothers [1]. The control structure is shown in figure 2.

![Figure 2: Control principle](image)

The knowledge of the model, valid near an operating point, allows to calculate the regulators parameters by different methods (compensation, poles placement, ...). A closed-loop study permits also to determine these control laws performances.

This work showed that the IP controller is a good compromise between the robustness and the possible closed-loop bandwidth [6]. Its structure is reminded on figure 3.

![Figure 3: Principle of the IP regulator](image)

Near operating points with a dominant output filter, the converter behaviour can be represented by a first order transfer
function. The obtained closed-loop transfer function is also:

$$\frac{\dot{V}_0(p)}{V_{rd}(p)} = \frac{1}{(1 + \left(1 + \frac{K d_{mod}}{\omega_d}\right)\frac{\omega_d}{\omega_c^2} + \frac{p^2}{\omega_d^2})^{-1}}$$

(6)

The coefficients can also be calculated by choosing a poles placement (or $\omega_{nd}$ and $\zeta_d$):

$$K = 2\zeta_d \frac{\omega_{nd}}{\omega_c} - 1$$

(7)

$$T_1 = \frac{2\zeta_d \omega_{nd}}{\omega_c} - 1$$

(8)

Such a placement is possible if $T_1 > 0$ that is to say:

$$\omega_{nd} > \omega_c / 2$$

(9)

In most cases, this constraint is not a problem, the designer looks for increasing the bandwidth.

Limitations induced by the resonant circuit dynamics

The performance of a control loop can be represented by its bandwidth which must be as high as possible. However, there is an interaction between the forecasted dynamics, $\omega_{nd}$, and the neglected dynamics (resonant circuit). Consequently, the choice of the closed loop dynamics results from a compromise.

The integration of the resonant circuit dynamics in the control loop leads to a fourth order transfer function:

$$\frac{\dot{V}_0(p)}{V_{rd}(p)} = \frac{1}{(1 + \frac{2\zeta_{mod}}{\omega_{mod}} + \frac{p^2}{\omega_{mod}^2})(1 + \frac{2\zeta_{mod}}{\omega_{mod}} + \frac{p^2}{\omega_{mod}^2})^{-1}}$$

(10)

with $\omega_{nd} > \omega_{nd}, \zeta_{mod} < \zeta_d, \omega_{mod} < \omega_n, \zeta_{mod} < \zeta_c$

The "mod" index means that the pulsations (or the damping ratio) are modified by the interaction between the dynamics. $\omega_{nd}, \zeta_{mod}, \omega_{mod}$ and $\zeta_{mod}$ are not very different respectively of $\omega_{nd}, \zeta_d, \omega_n$ and $\zeta_c$ if the interaction is not really efficient. $\omega_{nd}$ and $\omega_n$ must be sufficiently apart in the frequency plane. $\omega_{nd} < \omega_n / 4$ is the appropriate order of magnitude. Near an operating point characterized by $U$, it leads to:

$$\omega_{nd} < \pi f_0 (U-1) / 2$$

(11)

Otherwise, the resonant circuit dynamics is too modified by the closed loop and can generate instabilities since its damping ratio $\zeta_{mod}$ decreases below 0.

These results are valid for all the operating points whether on the real pole is dominant or not. However, the converter and regulator parameters can be analytically known owing to analytical knowledge of $G$, $\omega_c$ and $\omega_n$, except near the natural frequency (for $U < 1.3$).

Closed-loop behaviour for a varying operating point

The great variation of the open-loop static and dynamic characteristics has a strong influence on the closed-loop performance. The closed-loop behaviour significantly evolves and can degrade itself [6].

With a constant load, when the switching frequency decreases near the natural frequency, the static gain increases if U is greater than 1.2. The closed-loop dynamics increases proportionally to the product $(G\omega_c)^{1/2}$ (6). Moreover, the resonant circuit dynamics slows down. The influence of the resonant circuit, with the oscillations and often the instabilities which it implies, appears quickly because of the combination of these two phenomena [6].

At the opposite, when the frequency increases, the closed-loop system slows down. In this case, the second order assumption stays always valid. However, the evolution of the damping ratio $\zeta_{mod}$ must be considered. These variations are more difficult to analyse but the numerical applications show that they are low.

For example, with a regulator calculated near $U = 1.6$ for $\rho=0.65$ and $C_d/C=55$, with $\zeta_d=0.7$ and $\omega_{nd} = \omega_n/10$, the damping ratio $\zeta_{mod}$ stays, for $1.35 < U < 2.5$, in the range [0.6, 0.8]. Moreover, the damping ratio stays higher than 0.6, then the response overshoot is always low. This closed-loop evolution is shown in Nyquist plane on figure 4.

![Figure 4: Poles and zeros variations, in the Nyquist plane, with an IP calculated around $U = 1.6$, $\rho = 0.65$ with $\omega_{nd} = \omega_n/10$ and $\zeta_d = 0.7$](image)

Experimental validation

The precedent results have been validated [6] on an experimental half-bridge converter (1.5 kVA), using MOS switches (300V, 40A) and operating above natural frequency (20 kHz), shown on figure 5.

The behaviour of this converter is equivalent to that of a complete bridge converter which would have an input voltage of 150 V and a resonant capacitor of 540 nF. We use this analogy...
to model the converter.

The IP regulator has been analogically realized. Time constant $T_1$ and proportional gain $K$ can be independently adjusted. The frequency control uses a VCO with a gain of 1700 Hz/V.

$$E = 300 \text{ V}$$
$$8.8 \text{ nF}$$
$$120 \mu \text{H}$$
$$30 \mu \text{F}$$
$$8.8 \text{ nF}$$
$$270 \text{ nF}$$
$$30 \mu \text{F}$$
$$270 \text{ nF}$$

Figure 5: Structure of the experimental converter

The comparison between theory and experience shows that the model knowledge permits to have a good representation of the closed-loop behaviour. These tests have been realized for small variations of the reference voltage.

Figures 6 show the comparison between theoretical and experimental Bode diagrams, with a closed loop using an IP regulator, calculated around the nominal point (load $R_s = 10 \Omega$, switching frequency $F_s = 30$ kHz). The difference between theory and experience is mainly a higher damping for the experimental results because of the loss existing in a converter.

Figure 6: Comparison between the theory (1) and experience (2) around the point ($R_s=10\Omega$, $F_s=30$ kHz) with an IP regulator: $\omega_{nd} = 2 \omega_0$ (or $\omega_{nd} = 0.95 \omega_0$); $\delta_l = 0.7$

On figure 7, the evolution of the experimental closed loop when the operating point varies is studied in the same regulation conditions as previously. The theoretical results concerning the influence on the operating point variation can be validated.

The bandwidth always decreases when the switching frequency increases and the damping ratio is not very sensible to the operating point (Figure 7).

In fact, the resonant circuit dynamics is the most variable parameter. By its structure, the IP regulator leads to a closed-loop behaviour where this dynamics is well filtered and its influence can be neglected. We can also impose a large closed-loop bandwidth and obtain a relatively robust operation.

These results are derived from the analysis of a small signal model. They have been validated for small variations of the reference voltage. In the next paragraph, we are rather concerned by the large signal behaviour of this converter.

Figure 7: Experimental closed-loop Bode diagrams near different operating points with the same IP regulator: (1): $F_s = 28$ kHz, (2): $30$ kHz, (3): $33$ kHz

LARGE SIGNAL TUNING OF AN IP REGULATOR

A linear control law can ensure the forecasted dynamics only around the point which the regulator has been designed for. Then, we propose to experimentally study the performance of such a control for large variations and to define a way of tuning up the controller in this case.

At first, large variations of the switching frequency are not ever accepted by the real converter equipped with snubbers (figure 5). The control law must be correctly designed in order to reduce the resonant circuit perturbations when the reference voltage or the load varies significantly.

The design essentially depends on the specifications:
- operation in tracking control or in disturbance regulation,
- fixed performances near a special point or operation range to satisfy.

If performances are fixed for an operating point, the controller is automatically designed and we can only study the control robustness. In the other case, a regulator which leads to a stable operation on the whole operating range can be determined. We are concerned by this second possibility and we must define rules to design such a control law. At first, the design point must be determined and then, the closed-loop performances around this point must be chosen in order to determine the regulator parameters $K$ and $T_1$.

Choice of the regulator calculation point

Because of the snubbers, the frequency is generally limited in [1.3 $f_0$, 2.5 $f_0$]. Then, for a constant load, the output voltage (or for a constant reference voltage, the connected load range) can take values in a fixed interval.

For an operating point, $\omega_{nd}$ must be choosen less than $\omega_{nd}$ in order to dilter the resonant circuit oscillations. This choice is correct for each reference voltage lower than the calculation reference voltage or for each load higher than the calculation load.

In the other case, the resonant circuit dynamics is not sufficiently apart from the principal dynamics and oscillations can appear. Figure 8 presents the experimental responses,
obtained, with a load of 10 \( \Omega \), for the same reference step from 70 V (\( U = 1.55 \)) to 90 V (\( U = 1.4 \)), with two different regulators. Both are designed with oscillations. The oscillations obtained, with the corrector designed around 70 V, the response presents oscillations and the current and voltage overshoots cannot be controlled. With the other, the response is slower but there is no oscillations.

Moreover, the current overshoot during the transient period is in the middle of the operating range. From this way, the closed-loop behaviour stays correct on both sides.

**Choice of the dynamic performances**

Two parameters have to be determined around the design point: the pulsation \( \omega_{nd} \) and the damping ratio \( \zeta_d \).

- The more the operation range is important, the more the forecasted bandwidth must be unpretentious. This choice essentially depends on the output filter time constant.
- Since the calculation point is in the middle of the operating range, \( \omega_{nd} \) must necessary be chosen less than \( \omega_n/4 \). \( \omega_{nd} = \omega_n/10 \) is a good order of magnitude.
- Moreover, \( \zeta_d > \omega_n/(2 \omega_{nd}) \) is a necessary condition to a poles placement. Also, if we choose \( \omega_{nd} = \omega_n/10 \), the damping ratio \( \zeta_d \) must be higher than 0.35.
- If it is possible, \( \zeta_d \) is often equal to 0.7 which is the optimal damping ratio concerning the 5% response time [8-9]. It can be chosen higher than 0.7, in particular to respect the precedent condition (for the operating points area where \( \omega_n/\omega > 0.14 \)).

These design rules have been validated for different cases whether on the output filter time constant is important or not (different values of the load and of the output capacitor). In the next paragraph, two examples are detailed.

**Examples of tuning and experimental validation**

**Example 1:** point with an important output filter. Given a load of 10 \( \Omega \). With the experimental control structure, the switching frequency range is \([1.3 f_0, 2.3 f_0]\) or \([25500 \text{ Hz, } 50000 \text{ Hz}]\). According to (1), the output voltage can take values in the interval \([35 \text{ V, } 105 \text{ V}]\). The operating point, at the frequency 31000 Hz (71 V), is also in the middle of the operating range.

Using (4) and (5), we find, for this point, the characteristics:

\[
G = -0.004 \text{ V/Hz, } \omega_n = 4840 \text{ rd/s}
\]

Moreover, \( \omega_n \) can be approximated by \( 2 \pi (f_s - f_0) = 70550 \text{ rd/s} \). We have also \( \omega_n/\omega_n = 0.07 \).

The real pole is dominant and (4) and (5) allow to have a good approximation of \( G \) and \( \omega_n \).

For this point, we choose \( \omega_{nd} = \omega_n / 10 = 7055 \text{ rd/s} \). The damping ratio \( \zeta_d \) must be higher than 0.35. We can choose 0.7 and we determine \( K \) and \( T_1 \): \( K = -0.16 \), \( T_1 = 100 \mu s \).

This corrector has been experimentally tested. The closed-loop response, obtained for a reference voltage step from 50 V to 90 V is shown on the figure 9.a. It can be compared to the converter open-loop response, obtained for a frequency step from \( (40 \text{ kHz, } 50 \text{ V}) \) to \( (27 \text{ kHz, } 90 \text{ V}) \). It can be verified that this such designed regulator leads to satisfying results. The response time has been reduced. Moreover, the current overshoot during the transient state is less important when the converter operates in closed-loop.

Also, the IP controller allows to attenuate the resonant circuit perturbations, especially for large steps.

**Example 2:** point with a low output filter. Given a load of 10 \( \Omega \). With the experimental control structure, the switching frequency range is \([1.3 f_0, 2.3 f_0]\) or \([25500 \text{ Hz, } 50000 \text{ Hz}]\). According to (1), the output voltage can take values in the interval \([35 \text{ V, } 105 \text{ V}]\). The operating point, at the frequency 31000 Hz (71 V), is also in the middle of the operating range.

Using (4) and (5), we find, for this point, the characteristics:

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For this point, we choose \( \omega_{nd} = \omega_n / 10 = 7055 \text{ rd/s} \). The damping ratio \( \zeta_d \) must be higher than 0.35. We can choose 0.7 and we determine \( K \) and \( T_1 \): \( K = -0.16 \), \( T_1 = 100 \mu s \).

This corrector has been experimentally tested. The closed-loop response, obtained for a reference voltage step from 50 V to 90 V is shown on the figure 9.a. It can be compared to the converter open-loop response, obtained for a frequency step from \( (40 \text{ kHz, } 50 \text{ V}) \) to \( (27 \text{ kHz, } 90 \text{ V}) \). It can be verified that this such designed regulator leads to satisfying results. The response time has been reduced. Moreover, the current overshoot during the transient state is less important when the converter operates in closed-loop.
Example 2: point with a small output filter time constant - For a smaller load of 3 \( \Omega \), the same frequency range leads to an output voltage included in [10 V, 50 V]. The middle output voltage is obtained for \( f_s = 27000 \) Hz.

By using expressions (4) and (5), we find:
\[
G = -0.0027 \text{ V/Hz}, \ \omega_n = 11790 \text{ rad/s}
\]

The approximation of \( \omega_n \) by \( 2\pi(f_s-f_0) \) leads to 57990 rad/s and we have: \( \omega_n/\omega_n = 0.2 \)

In this case, the real pole is not dominant. The use of the complete model is necessary. This method gives:
\[
G = -0.0028 \text{ V/Hz}, \ \omega_n = 11390 \text{ rad/s}, \ \omega_m = 58860 \text{ rad/s}, \ \zeta = 0.09
\]

We note that the error induced by the simplest method is negligible (2.4 % for \( G \), 3.5 % for \( \omega_n \) and 1.5 % for \( \omega_m \)). However, a verification of the converter parameters is necessary by using the complete model. The simplest method can give wrong results.

For this point, we choose \( \omega_m = \omega_n/10 = 5886 \text{ rad/s} \). The damping ratio \( \zeta_d \) must be chosen higher than 0.95. We cannot take a value of \( \zeta_d \) too close of this limit: a little error on the parameters of the model (or of the controller) can lead to an unstable behaviour. We choose 1.3 and determine: \( K = -0.07, T_1 = 97 \mu s \)

As for example 1, the open- and closed-loops experimental results have been compared (figures 10.a and 10.b). Both corresponds to the passage from 18 V to 42 V.

![Image 10.a: closed-loop response](image1)

![Image 10.b: open-loop response](image2)

**Figure 10**: Experimental responses with \( R_s = 3 \Omega \)

In this case, the converter dynamics (output filter and resonant circuit) impose to choose, around the middle point, a closed-loop bandwidth smaller than the open-loop one with an important damping ratio. Then, the response time has been significantly decreased in order to filter the resonant circuit oscillations, which are very visible in open loop. Such an open-loop frequency step can not be increased any more because a switching default appears for a larger step (43 V to 17 V).

A correctly designed IP regulator permits always to obtain a satisfying closed-loop behaviour. For converters with a small output filter, the resonant circuit dynamics can be filtered but to the detriment of the response time.

**CONCLUSION**

The exploitation of a small signal model allows designing linear control strategies for the series resonant converter and determining the closed-loop performances. Experimental results permit to verify all this study.

From this way, all the usually observed phenomena, in particular the apparition of oscillations and sometimes of instabilities when the switching frequency decreases near the natural frequency, can be explained. These oscillations are induced by the bad control of the resonant circuit behaviour. This allows determining the limit of these different structures use, essentially depending on the resonant circuit dynamics.

A preliminary study showed that the IP regulator is a good compromise to control such a converter. The obtained closed-loop performances are satisfying when respecting the validity assumptions.

All the main rules to design such an IP regulator (from the converter modelling to the controller tuning) have been exposed in this paper and experimentally validated.

The obtained results confirm that an IP controller, correctly designed, allows well filtering the few damped oscillations of the resonant circuit (whatever the output filter and for small or large variations of the operating point).

On the other hand, when the output filter time constant has a low value, the regulator design imposes to slow down the converter dynamics. In this case, a control, using informations on the resonant circuit state, can be preferable if the user desires higher dynamics performances.

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