Bidirectional multi-resonant converter with a wide voltage range

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Abstract
A bidirectional multi-resonant DC/DC converter is proposed in this study. The structure of the converter is symmetrical, and the operating characteristics of the bidirectional operation mode are the same. Adopting synchronous driving, the converter can regulate the output voltage over a wide range. Regardless of a forward mode or reverse mode, the switches of the converter can realize zero-voltage switching over the entire voltage range. The gain of the converter can be zero due to its characteristics similar to the notch filter. The converter can transmit the fundamental wave and the third harmonic at the same time, and the energy utilization rate is high. The converter has a wide voltage range and high efficiency at the same time. A 3.3 kW experimental prototype is designed. In forward mode, the secondary-side voltage varies from 250 to 450 V, and the peak efficiency is 97.7%. In reverse mode, the secondary-side voltage varies from 300 to 450 V, and the peak efficiency is 97.6%.

1 | INTRODUCTION

Electric vehicles are widely used today. The electric vehicle charging system, as a device that provides stable power to the power battery, is supposed to adjust the voltage in a wide range [1, 2]. Adjustable range of power factor correction (PFC) units is limited, thus DC–DC converters are important for meeting the charging demand [3–6]. Through bidirectional DC–DC converters, power batteries can not only be charged but also be used as emergency power sources to provide external equipment with electrical energy. Therefore, bidirectional DC–DC converters have a long-term application prospect.

The power level of the electric vehicle charging system is high. There may be potential electrical safety hazards if a high-frequency transformer is not used. Electric vehicle charging system needs to guarantee the electrical isolation of the primary side and secondary side, so isolated converters should be adopted. The voltage or current in resonant converters can be periodically reduced to zero, thus switches can achieve soft switching, which reduces switching loss. Among resonant converters, LLC resonant converter possesses natural soft-switching characteristics [7–13]. However, when operating in reverse mode, the magnetising inductance is clamped by secondary-side voltage and no longer participates in resonance. The converter operates in the same manner as the series resonant converter which can only work in buck mode, thus the voltage range is narrow. In [14], the CLLC bidirectional resonant converter with a capacitor on the secondary side was proposed. The converter can achieve the buck or boost mode. However, the converter’s resonant networks are different in bidirectional operation modes, so characteristics are not symmetrical, which is not conducive to parameter design. In [15–18], a CLLLC bidirectional resonant converter with symmetrical structure was proposed. When the gain is lower than 1, the decreasing trend of the gain is not visible, which is not conducive for a wide voltage range.

A bidirectional converter with auxiliary inductor was proposed in [19, 20]. The converter has the same operating characteristics as an LLC converter in bidirectional modes. The converter achieves low voltage gain by circulating energy. The wider the driving pulse width of the secondary-side switches, the larger the circulating energy and the smaller the voltage gain. However, the circulating energy also brings additional losses to the converter, which is not conducive to improve efficiency. Wide gain is achieved by switching the operation mode between half-bridge and full-bridge in [21]. The gain increases as phase shift angle increases. But when increases, the gain can only increase from 0.5 to 1, so the converter cannot work in boost mode. To widen the gain range, half-bridge and full-bridge switching can be implemented on both sides in [22]. However,
adoption pulse width and frequency hybrid modulation is complicated. Moreover, the converter uses 12 switches, which are not conducive to the improvement of power density.

To achieve wide gain and high efficiency, a bidirectional converter is proposed. Adopting synchronous driving, the converter can regulate the output voltage over a wide range. This paper is organised as follows. In Section 2, the topology and operation principles of the bidirectional converter are analysed. Section 3 introduces the characteristics of the proposed converter. Experimental results are shown in Section 4. Section 5 gives the conclusions.

2 | PROPOSED CONVERTER AND OPERATION PRINCIPLE

Figure 1 presents the main circuit of the bidirectional converter. The converter uses two full-bridge structures to achieve voltage inversion or rectification. The resonance tank formed by \( L_p \), \( C_r \) and \( C_p \) can realise the function of a notch filter. It is connected in series with the resonance inductor \( L_r \) in the primary-side circuit. Auxiliary inductor \( L_{m2} \) is connected in parallel at the midpoints of the primary-side bridge arms. During forward operation, the notch filter, \( L_r \) and magnetostrictive inductance \( L_{m1} \) form a resonant network together. \( L_{m2} \) is clamped by \( V_p \). It does not participate in resonance but can help switches achieve soft switching. During reverse operation, \( L_{m1} \) is clamped by \( V_s \). The notch filter, \( L_r \) and \( L_{m2} \) form a resonant network together.

Synchronous variable frequency (VF) drive method is adopted in bidirectional power transmission, which includes the equal-width VF method and non-equal-width VF method. The forward mode is explained as an example. \( f_s \) is the switching frequency and \( f_r \) is the resonant frequency. \( S_1, S_4, S_6 \), and \( S_7 \) are turned ON at the same time.

When \( f_s < f_r \), the on-time of \( S_1, S_4, S_2 \), and \( S_3 \) is one-half the switching period, meanwhile the on-time of \( S_6, S_7 \) and \( S_5 \) is one-half the resonance period, that is, when \( i_s = 0 \), \( S_6, S_7, S_5 \), and \( S_8 \) are turned OFF. When \( f_s > f_r \), the driving signals of \( S_1, S_4 \), and \( S_7 \) are completely synchronised. The pulse widths are the same, both of which are 1/2 duty cycle. When \( i_s \) is not zero, \( i_s \) flows through the channel of the metal oxide semiconductor field effect transistor (MOSFET) instead of its body diode. The conduction voltage drop can be significantly reduced, so the conduction loss will be reduced accordingly. Because the converter has characteristics similar to a notch filter, a low gain can

be achieved without using circulating energy, and losses caused by circulating energy can be avoided.

Figures 2 and 3 show the main waveforms and equivalent circuits in different stages when \( f_s < f_r \) in forward mode. Assuming that the ripple of \( V_s \) is ignored, the stages of the converter are analysed below.

Stage I \([t_0, t_1]\): \( S_1, S_4, S_6 \), and \( S_7 \) are turned ON at \( t_0 \). Prior to this time, the parasitic capacitances of \( S_1, S_4, S_6 \), and \( S_7 \) are discharged, and their body diodes are conducting freewheeling. Therefore, \( S_1, S_4, S_6 \), and \( S_7 \) can achieve ZVS. \( L_{m1} \) is clamped by \( n V_s \) where \( n \) denotes windings ratio \( N_p/N_s \) of the transformer. The magnetising current \( i_{m1} \) increases linearly. \( C_r, L_{m1}, C_p \), and \( L_r \) resonate. \( i_s \) rises and is always greater than the magnetising inductance current \( i_{m1} \).

Stage II \([t_1, t_2]\): \( S_6 \) and \( S_7 \) are turned OFF at \( t_1 \). At this time, \( i_s \) is equal to \( i_{m1} \), and \( i_s \) is reduced to zero, so \( S_6 \) and \( S_7 \) achieve ZCS. At this stage, \( C_r, L_{m1}, C_p \), and \( L_r \) resonate together. \( L_{m1} \) is large enough, thus the changing rate of \( i_s \) can be ignored. \( i_s \) and \( i_{m1} \) are equal and remain unchanged.

Stage III \([t_2, t_3]\): \( S_1, S_4 \), and \( S_7 \) are turned OFF at \( t_2 \). The parasitic capacitances of switches are charged or discharged. After this period, the voltage between the midpoints A and B is \(-V_p\). \( C_r, L_{m1}, C_p \), and \( L_r \) resonate. The resonant current \( i_s \) decreases. The voltage between C and D is \(-V_s\). The voltage on \( L_{m1} \) is clamped, and \( i_{m1} \) starts to decrease linearly. The body diodes of \( S_2, S_3, S_5 \), and \( S_8 \) are conducting freewheeling, and the converter enters the second half-cycle stage.

Figures 4 and 5 show the main waveforms and equivalent circuits in different stages when \( f_s > f_r \) in forward mode. The stages of the converter are analysed below.
Stage I \( [t_0, t_1] \): \( S_1, S_4, S_6 \) and \( S_7 \) are turned ON at \( t_0 \). Prior to this time, the parasitic capacitances of \( S_1, S_4, S_6 \) and \( S_7 \) are discharged, and their body diodes are conducting freewheeling. Therefore, \( S_1, S_4, S_6 \) and \( S_7 \) can achieve ZVS. \( L_{m1} \) is clamped by \( nV_S \) where \( n \) denotes windings turns ratio \( N_P/N_S \) of the transformer. The magnetising current \( i_{m1} \) increases linearly. \( C_r, L_p, C_p \) and \( L_r \) resonate. \( i_r \) rises and is always greater than the magnetising inductance current \( i_{m1} \).

Stage II \( [t_1, t_2] \): \( S_1, S_4, S_6 \) and \( S_7 \) are turned OFF at \( t_1 \). The parasitic capacitances of switches are charged or discharged. The resonant current \( i_r \) decreases, and its amplitude is always greater than \( i_{m1} \). When charging and discharging are completed, body diodes of \( S_2, S_3, S_6 \) and \( S_7 \) conduct freewheeling.

Stage III \( [t_2, t_3] \): At \( t_2 \), the amplitudes of \( i_r \) and \( i_{m1} \) are equal. \( i_r \) decreases and its amplitude is smaller than \( i_{m1} \). \( i_r \) changes in polarity, and the voltage between \( C \) and \( D \) is \(-V_S \). The voltage on \( L_{m1} \) is clamped and \( i_{m1} \) starts to decrease linearly. The body diodes of \( S_2, S_3, S_6 \) and \( S_8 \) conduct.

At \( t_3 \), \( S_2, S_3, S_5 \) and \( S_8 \) are turned ON. The converter enters the second half-cycle operation. The reverse mode and the forward mode have symmetrical working waveforms, which will not be repeated here.
3 CHARACTERISTICS ANALYSIS

3.1 Gain analysis

Figure 6 depicts a simplified model of the converter. Using the fundamental wave approximate analysis method, \( R_e \) can be expressed as

\[
R_e = \frac{8n^2R_o}{\pi^2}
\]

\( v_{AB} \) is input square wave voltage. The notch filter is composed of \( C_r, L_p, C_p \). \( Z_{rp} \) is composed of notch filter and \( L_r \), which can be expressed as

\[
Z_{rp} = \left| j\omega L_p + \frac{1}{j\omega C_r} + \frac{1}{j\omega C_p} \right| \left| j\omega L_r + \frac{1}{j\omega C_r} \right| \left| j\omega L_p + \frac{1}{j\omega C_p} \right|.
\]

When \( Z_{rp} = 0 \), the converter has two resonance frequencies, respectively,

\[
f_{rs1} = \sqrt{\frac{1 + k + q - \sqrt{(1 + k + q)^2 - 4kq}}{2kq}} \times f_{s0}
\]

\[
f_{rs2} = \sqrt{\frac{1 + k + q + \sqrt{(1 + k + q)^2 - 4kq}}{2kq}} \times f_{s0}
\]

\[
Q = \sqrt{\frac{I_m C_r}{R_e}}, I_m = \frac{I_o}{I_m}, f_n = \frac{f}{f_{rs1}}
\]

\[
g = \frac{f_{s0}}{f_{s0}} = \sqrt{\frac{1 + k + q - \sqrt{(1 + k + q)^2 - 4kq}}{2kq}}
\]

\[
M(f_n, Q, I_m) = \frac{1}{\left(1 + L_r \times \left(\frac{1}{j\omega L_r f_n} + \frac{1}{j\omega L_p f_n + \frac{1}{j\omega C_r f_n} + \frac{1}{j\omega C_p f_n}}\right)\right)^2 + Q f_{s0} \times \left(1 + \frac{k}{1 + k + q - \sqrt{(1 + k + q)^2 - 4kq}} \right)^2}
\]

Figure 7 depicts the gain curves of the bidirectional converter. Different from LLC converter, the decreasing trend of the gain is evident in \( f_{s0} < f_s \leq f_{fp} \). At \( f_{fp} \), the output voltage
can be zero. Within the operating range, the frequency variation range is narrow, which can improve the efficiency of the converter.

### 3.2 Design procedure

Based on Equations (1), (3), (4) and (7), \( L_r, C_r, L_p, C_p \) and \( L_m \) can be expressed as

\[
L_r = \frac{QR_e}{2\pi f_{rs0}}
\]

\[
C_r = \frac{1}{2\pi f_{rs0}QR_e}
\]

\[
I_m = \frac{L_r}{L_n}
\]

\[
I_p = kL_r
\]

\[
C_p = qC_r.
\]

Once \( f_{rs1}, f_{rs2} \) and \( f_{rp} \) are set, \( f_{rs0} \) can be calculated. Actually, \( L_r, C_r, I_p, C_p \) and \( I_m \) are determined by \( q, k, Q \) and \( L_n \). The influence of \( q, k, Q \) and \( L_n \) on the gain are described below.

#### 3.2.1 Resonant frequencies

Figures 8 and 9 show the effect of \( k \) and \( q \) on \( M \); \( k \) and \( q \) will affect frequency \( f_{rp} \) but not influent the gain when \( M > 1 \). \( r_{AB} \) mainly contains fundamental wave and third harmonic. If \( f_{rs2} = 3f_{rs1} \) and \( f_{rp} = 2f_{rs1} \), third harmonic can be used and the efficiency can be improved [23]. Once \( f_{rs1} = 100 \) kHz, the resonant frequency \( f_{rs2} \) and \( f_{rp} \) can be set as

\[
f_{rs2} = 3f_{rs1} = 300 \text{ kHz}
\]

\[
f_{rp} = 2f_{rs1} = 200 \text{ kHz}.
\]

From Equations (3), (4) and (7), \( k, q, g \) and \( f_{rs0} \) are expressed by

\[
k = \frac{L_p}{L_r} = 2.4
\]

\[
q = \frac{C_p}{C_r} = 0.6
\]

\[
g = \sqrt[3]{1 + k + q - \sqrt{(1 + k + q)^2 - 4kq}} = 0.527
\]

\[
f_{rs0} = \frac{f_{rs1}}{g} = 190 \text{ kHz}.
\]

#### 3.2.2 The design of \( Q \) and \( L_n \)

Figures 10 and 11 depict the effect of \( Q \) and \( L_n \) on the gain. \( Q \) and \( L_n \) do not influent the gain much when \( M \leq 1 \) but affect the peak gain when \( M > 1 \).

The designed voltage range must be wider than the actual required voltage range. In the operating range, gain is inversely proportional to frequency. In the same frequency condition, light-load gain is higher than heavy-load gain. No-load gain is the highest. Therefore, the minimum no-load gain and the maximum full-load gain must satisfy the required gain range.
When $Q = 0$, $M$ can be defined by

$$M(f_n, L_n) = \frac{1}{1 + L_n \left(1 + \frac{1}{1 + \sqrt{Q \cdot k^2 f_n^2}} - 1 \right) \times \frac{1}{1 + \sqrt{Q \cdot k^2 f_n^2}}}.$$  \hspace{1cm} (19)

The voltage gain $M$ at maximum frequency should satisfy

$$M(f_{n_{\text{max}}}) \leq M_{\text{min}}$$  \hspace{1cm} (20)

where $M_{\text{min}}$ is the minimum gain.

Figure 12 shows $M_{\text{min}}$ and the effect of $L_n$ and $f_n$ on $M$. The gain below $M_{\text{min}}$ satisfies the demand. By projecting the gain curve below $M_{\text{min}}$ to $L_n, f_n$ dimension, the value of $L_n$ and $f_n$ that meet the conditions can be obtained, which is shown in the shadow of Figure 12. $L_n$ and $f_n$ are selected at appropriate values. Once $L_n$ is set, $M$ is only determined by $f_n$ and $Q$. The gain $M$ at the minimum frequency should satisfy

$$M(f_{n_{\text{min}}}) \geq M_{\text{max}}$$  \hspace{1cm} (21)

where $M_{\text{max}}$ is the maximum gain.

Figure 13 shows $M_{\text{max}}$ and the effect of $Q$ and $f_n$ on $M$. The gain above $M_{\text{max}}$ satisfies the demand. By projecting the gain curve above $M_{\text{max}}$ to $Q, f_n$ dimension, the value of $Q$ and $f_n$ that meet the conditions can be obtained, which is shown in the shadow of Figure 13. $Q$ is selected at an appropriate value.
3.3 Soft-switching condition

To ensure high efficiency, switches need to achieve ZVS over the entire voltage range. Take S₃ and S₄ as examples for analysis. In forward mode, the parasitic capacitance of S₃ is charged after S₃ is turned OFF. The charging process will continue until \( V_{\text{d3}} = V_p \). The parasitic capacitance of S₃ is discharged until \( V_{\text{d3}} = 0 \). This period takes \( 2C_{\text{oss}}V_p \). During dead time, the current flowing through S₃ remains at \( i_{4,\text{off}} \), which can provide the charge of \( i_{4,\text{off}}r_{\text{dead}} \) during the dead time \( t_{\text{dead}} \). If \( S_3 \) can realize zero-voltage switching, the condition to be satisfied is \( i_{4,\text{off}}r_{\text{dead}} \geq 2C_{\text{oss}}V_p \).

When \( f_i > f_r \), \( i_t \) is equal to \( i_{m1} \) after half a resonance period \( T_{m1}/2 \), that is,

\[
nV_S = I_m \frac{di_{m1}}{dt} \tag{22}
\]

\[
i_{m1}(t) = -\frac{nV_S}{4L_m f_s} + \frac{nV_S}{T_m} t \tag{23}
\]

\[
i_{m1}\left(\frac{T_{m1}}{2}\right) = -\frac{nV_S}{4L_m f_s} + \frac{nV_S}{T_m} \frac{T_{m1}}{2} \tag{24}
\]

The converter enters the intermittent conduction state, and \( i_t \) and \( i_{m1} \) are approximately considered to be equal and remain unchanged. The resonant period can be expressed as \( T_{s1} = f_n / f_r \).

When \( S_1/S_4 \) is turned OFF, \( i_t \) and \( i_{m1} \) are maintained at the value at \( T_{m1}/2 \), which can be defined by

\[
i_{m1-\text{off}} = i_{m1}\left(\frac{T_{m1}}{2}\right) = -\frac{nV_S}{4L_m f_s} + \frac{nV_S f_a}{2T_m f_s} = \frac{(2Mf_n - M)V_p}{4L_m f_s} \tag{25}
\]

After half a switching cycle, \( i_{m2,\text{off}} \) can be expressed as

\[
i_{m2,\text{off}} = \frac{V_p}{4L_m f_s} \tag{26}
\]

The current flowing through \( S_4 \) is the sum of \( i_{m2} \) and \( i_t \), which can be expressed as

\[
i_{4,\text{off}} = i_t + i_{m2,\text{off}} = \frac{(2Mf_n - M + 1)V_p}{4L_m f_s} \tag{27}
\]

The condition for achieving ZVS can be expressed as

\[
I_m \leq \frac{(2Mf_n - M + 1)t_{\text{dead}}}{8C_{\text{oss}} f_s} \tag{28}
\]

When \( f_i > f_r \), \( S_1 \), \( S_4 \) and \( S_6 \), \( S_7 \) are turned OFF at the same time. At the moment when \( S_3 \) is turned OFF, the value of \( i_{m1} \) and \( i_{m2} \) can be expressed as

\[
i_{m1,\text{off}} = \frac{nV_S}{4L_m f_s} \tag{29}
\]

\[
i_{m2,\text{off}} = \frac{V_p}{4L_m f_s} \tag{30}
\]

When \( S_1 \) and \( S_4 \) are turned OFF, the resonance process between \( L_m \), \( C_i \), \( L_p \) and \( C_p \) has not ended or just ended. At this time, \( i_t \geq i_{m1} \).

The current flowing through \( S_4 \) can be expressed as

\[
i_{4,\text{off}} = i_t + i_{m2,\text{off}} \geq \frac{nV_S}{4L_m f_s} + \frac{V_p}{4L_m f_s} = \frac{(M + 1)V_p}{4L_m f_s} \tag{31}
\]

The condition for achieving ZVS is expressed as

\[
I_m \leq \frac{(M + 1)t_{\text{dead}}}{8C_{\text{oss}} f_s}. \tag{32}
\]

4 EXPERIMENTAL VERIFICATION

A 3.3 kW bidirectional converter experimental prototype is built, which is shown in Figure 14. In reverse mode, the battery's stored energy and discharge power decreases, so the converter works under the light-load condition at \( P_o = 1.6 \text{kW} \). The minimum discharge voltage of the battery must be higher than the minimum charge voltage. The operation index is listed in Table 1 and the parameters are listed in Table 2.

In forward mode, the converter operates with a constant current (CC) of 7.3 A and \( V_S \) varies from 250 to 450 V. When \( V_S \) reaches 450 V, the converter operates with constant voltage...
| Item, symbol | Value | Part name |
|-------------|-------|-----------|
| Switching frequencies | 74–141 kHz | Ferroxcube PQ26/25 |
| Resonant inductor $L_r$ | 20 μH | WIMA MKP |
| Resonant capacitor $C_r$ | 34 nF | WIMA MKP |
| Resonant inductor $L_p$ | 50 μH | Ferroxcube PQ32/30 |
| Resonant capacitor $C_p$ | 20 nF | WIMA MKP |
| Magnetising inductance $I_{ref}$ | 200 μH | Ferroxcube PQ50/50 |
| Turns ratio $n = N_P / N_S$ | 1.2 (30:25) | |
| Auxiliary inductor $L_{aux}$ | 200 μH | Ferroxcube PQ32/30 |
| Switches $S_1$–$S_8$ | IPW60R070CFD7 | |

Figure 15 Control diagram of the converter in forward mode

(CV). In reverse operation, $V_S$ varies from 300 to 450 V and $V_P$ maintains at 400 V. Figure 15 shows the control diagram of the converter in forward mode. The control block diagram includes two proportional integral (PI) controllers and two-mode selections. Two PI controllers realise the CC and CV operations of the converter, respectively. Mode selection 1 includes a parallel structure capable of automatically switching between CC and CV modes. After feedback signals are calculated by PI controllers, the smaller value will be output, that is, it can determine whether the converter works in CC or CV mode. The output of PI controllers can reflect the value of $f_s$. In CV mode, non-equal-width VF method is adopted. In CC mode, if $f_s \geq f_t$, mode selection 2 will choose synchronous equal-width VF method, otherwise synchronous non-equal-width VF method is adopted.

Figure 16 shows the main waveforms when $V_S = 335$ V and $f_s = 100$ kHz. It can be seen from the figure that $S_4$ and $S_6$ achieve ZVS, and $S_8$ achieves ZCS. Because the converter can transmit the third harmonic, $i_t$ presents a curve which seems like a saddle. The efficiency can be improved.

Figure 17 shows the main waveforms when $V_S = 450$ V and $f_s = 70$ kHz. It can be seen from the figure that $S_4$ and $S_6$ achieve ZVS. $S_5$–$S_8$ are turned OFF when $i_t = 0$ so that ZCS can be observed. Parasitic capacitances of switches resonant with inductors, so $V_{ds}$ and $i_t$ oscillate, which is normal.

When $f_s > f_t$, $S_1$, $S_4$, $S_6$, and $S_8$ are turned OFF at the same time. Figure 18 shows the main waveforms when $V_S = 250$ V and $f_s = 140$ kHz in forward mode. It can be seen that $S_4$ and $S_6$ achieve ZVS, but $S_8$ cannot achieve ZCS.
In reverse mode, $V_S$ varies from 300 to 450 V and $V_P$ maintains at 400 V. The experimental waveforms are shown in Figures 19–21. The reverse experimental waveforms are symmetrical to the waveforms in forward mode. The charging process and efficiency in forward mode are shown in Figure 22(a), and the efficiency at $P_o = 1.6$ kW in reverse mode is shown in Figure 22(b). The converter has high efficiency near $f_r$ when it is operating in forward or reverse mode. In forward mode, the maximum efficiency is 97.7%. In reverse mode, the maximum efficiency is 97.6%. The converter has higher efficiency over a
wide voltage range. The efficiency of forward mode is higher than 94.5%, and the efficiency of reverse mode is higher than 93.2%.

5 CONCLUSION

A bidirectional converter is proposed in this study. The structure of the converter is symmetrical, and the operating characteristics of the forward and reverse modes are the same. Adopting synchronous driving, the converter can regulate the voltage and power over a wide range. The control method is simple to implement. Regardless of forward or reverse mode, the switches can realise ZVS. Both the fundamental wave and the third harmonic can be used. The converter has obvious advantages in gain and efficiency. A 3.3 kW experimental prototype is designed. In forward mode, the secondary-side voltage varies from 250 to 450 V, and the peak efficiency is 97.7%, while in reverse mode, the secondary-side voltage varies from 300 to 450 V, and the peak efficiency is 97.6%.

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