Series-Resonant Single-Phase AC-DC Power Supply with Control of Reactive Power

J. Ben Klaassens, Marinus P. N. van Wesenbeeck, and Hian K. Lauw

Abstract—The series-resonant converter is one of the major classes of electronic power converters that are extensively applied in dc-dc conversion. This technology transfers bulk energy efficiently through a series-resonant circuit in the direct path of the energy transfer. This resonant converter forms a high-frequency ac link between the voltage sources connected to the input and output terminals. The high-frequency link is exploited in the application of a modulation process, thus avoiding the use of bulky and expensive low-order harmonic filters. The achievable high-pulse repetition frequency is rooted in the efficient turn-off mechanism of the semiconductor switches and does not compromise their reliability.

The classical objective of an ac-dc converter is to control the (average value of the) output waveform (often the output voltage). The energy emanated from the source is adapted by passive filters, often a compromise between volume and effective use. A rectifier-filter network presents a major disadvantage: a power factor less than one caused by distortion of the source current.

The resolution of the high-frequency input current of the series-resonant converter introduces the possibility of accurately controlling the waveform of the source current. In combination with the accurate positioning of the high-frequency current pulses, it is possible to influence the exchange of power with the source by the introduced method of active filtering. A power factor equal to one would lead to optimal operation.

However the zero crossing of the ac source voltage introduces the necessity to store energy. Bulky low-frequency storage elements increase the specific volume and weight of the converter, while the exchange of energy with storage elements decrease the overall efficiency. The energy stored in the resonant circuit is fundamentally too low to solve this problem. The stored energy in the output capacitor necessary to decrease the high-frequency ripple voltage is available for these purposes. To meet the optimal conditions for the power factor at the source, a bipolar flow of energy is required.

I. POWER CIRCUIT CONFIGURATION

EACH input and output line is connected to the switching elements. The configuration in Fig. 1 applies to a single-phase input and output converter. As shown in this figure the sinusoidal source voltage e_s is connected to the switching matrix SM1, whereas the load with a dc voltage u_o is connected to the switching matrix SM2. Both switching matrices are of the half-bridge configuration. The switching matrices are interconnected by a series-resonant circuit constructed with the passive components L_r and

C_r . It is further to be noted that each switching element is realized by two semiconductors in antiparallel in order to be able to switch the bipolar resonant current i_r for each polarity of the voltages connected to the board. This topology facilitates dc and ac input and output and provides the capability of bidirectional power flow, i.e., four-quadrant operation.

The resonant circuit elements of the proposed converter in Fig. 1 are formed by the inductor L_r and the parallel combination of the capacitors $C_{r1} = \frac{1}{2}C_r$ and $C_{r2} = \frac{1}{2}C_r$. Capacitors C_s and the combination of $C_{o1} = \frac{1}{2}C_o$ and $C_{o2} = \frac{1}{2}C_o$ are the input and output filter capacitors that are inserted to form a low-impedance path for the high-frequency components of the source and output currents, respectively. Capacitors C_s and C_o should be large compared to the resonant capacitor C_r .

This power circuit does not need additional components in order to ensure

- 1) Natural commutation
- 2) Unity power factor at the input (no power factor boost circuit is required)
- 3) An internal frequency as high as necessary by the desire to meet the requirements with respect to the output performance specification, size, and power density without the need to include bulky and expensive low-order harmonic filters
- 4) Operation of the converter with ac and dc input
- 5) Bidirectional power flow and four-quadrant operation

The use of thyristors makes it possible to build a series-resonant converter with an internal converter frequency of up to a hundred kilohertz limited by the turn-off time of commercially available thyristors. The use of power FET's makes it possible to increase this frequency to beyond 500 kHz. The use of staggered power modules increases the pulse repetition frequency at the output port of the converter without increasing the internal frequency and, consequently, improves the indicated mode of modulation [10]. It is noted that size advantage with increasing frequency is not present in this class of power converters, which has to apply a dc link. The thyristor is significantly cheaper and more robust than gate turn-off switches (GTO's) in addition to facilitating convenient interfacing between the control circuit and the power circuit due to the low gate current required.

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Fig. 1. Diagram power circuit of the series-resonant single-phase ac-dc converter.

Fig. 2. Process of pulse positioning and modulation. (a) Equivalent circuit diagram. (b) Characteristic waveforms.

Fig. 2 presents the process of pulse positioning and modulation of the source-sided current pulses i_{SM1} on the switching matrix SM1. The high-frequency carrier i_r as generated by the series-resonant converter is modulated in frequency and in combination with a low-frequency, bipolar programmed process of rectification; a sinusoidal source current i_{SM1} is generated at the input of the switching matrix SM1. The fundamental component of this current (as depicted in Fig. 2(b) by the dotted line), in combination with the capacitor current i_{Cs} , has to be positioned in phase with the sinusoidal source voltage e_s . The harmonic components of the pulse modulation process have only a minor influence on the power factor. The current i_{SM1} is generated independently of the source voltage e_s . It is necessary to operate the source-sided switching matrix SM1 as an inverter-type rectifier energized by an en-

ergy source different from the voltage source e_s . Active compensation of the power factor of a system is only feasible for *four-quadrant operation*. Two-quadrant operation (exchange of power with the source in one direction) will result in a degradation of the power factor, especially at low-load conditions.

II. CONVERSION PROCESS

The exchange of energy is activated by the combination of the two voltage sources e_s and u_o by selecting a proper combination of switches. Switching matrices SM1 and SM2 excite the series-resonant circuit by the voltage u_{LC} as a combination of the AC voltages e_{SM1} and e_{SM2} :

$$u_{LC} = e_{SM1} - e_{SM2} \quad (1)$$

where $e_{SM1} = \frac{1}{2}e_s$ and $e_{SM2} = \frac{1}{2}u_o$ as usual for a half-bridge switching matrix. It is sufficient to introduce two voltage sources $|e_{SM1}| = \frac{1}{2}e_s$ and $|e_{SM2}| = \frac{1}{2}u_o$. The converter network is now replaced by the equivalent network shown in Fig. 3. Both voltage sources represent the selected sources at the outside of the converter for a single resonant current pulse. The operation of the controlled series-resonant circuit is based on the excitation by the resulting voltage u_{LC} resulting in a modulated high-frequency carrier i_r .

The resonant current i_r will pass a selection of both voltage sources as connected to the switches. Depending on the configurations of the active switches in the matrix SMp an ac voltage e_{SMp} is supplied at the terminals of the matrix or by *selecting* the voltage source or by *bypassing* the voltage source. The ac voltage e_{SMp} ($p = 1, 2$) is selected from the *two* possible values $+e_s$ or $-e_s$ and $+u_o$ or $-u_o$. For the half-bridge configuration as presented in Fig. 1, the resulting excitation voltage u_{LC} (1) is limited to *four* combinations of $\frac{1}{2}e_s$ and $\frac{1}{2}u_o$ [9]:

$$\pm \frac{1}{2}(e_s + u_o), \pm \frac{1}{2}(e_s - u_o). \quad (2)$$

The result is a sinusoidal resonant current i_r and resonant capacitor voltage u_{Cr} with low damping. The multiplication of the resonant current i_r and the excitation voltage u_{LC} describes the exchange of energy with the resonant circuit.

The control of the semiconductor switches in both matrices has to guarantee two conditions:

- 1) A net amount of energy has to be distributed in the desired direction for each individual current pulse.

Fig. 3. Double excited resonant circuit.

- 2) A recurrent change of stored energy in the network has to be avoided.

Electric energy is distributed from a source e_s to a source u_o by a proper selection of sources: $(e_s - u_o)i_r > 0$. The difference of the energy emanated from the source and the energy distributed to the load will change the energy level of the resonant circuit. For $e_s > u_o$ the converter system can be regarded as a *step-down converter* for which the stored energy in the resonant capacitor will recurrently increase. For $e_s < u_o$ the converter system can be regarded as a *step-up converter* for which the stored energy in the resonant capacitor will recurrently decrease. To stabilize the amplitude of the resonant capacitor voltage u_{Crmax} at the end of each individual current pulse in order to obtain a cyclic stable mode of operation, the energy distribution to the resonant circuit has to be controlled during each current pulse. A resonant current pulse is hence build up from *two segments* (i.e., two levels for the excitation voltage). One current segment is applied to distribute power. The other current segment will stabilize the stored energy by a process of active damping in contrast to passive damping, which will increase the circuit losses. The energy level of the LC circuit has to be *increased* by selecting an additional excitation voltage u_{LC} : $u_{LC}i_r > 0$ or *decreased* by selecting an additional excitation voltage u_{LC} : $u_{LC}i_r < 0$.

The characteristic waveforms of the resonant circuit are shown in Fig. 4(a) for the *step-down mode*. The energy is exchanged during the time interval indicated as ψ_{fk} where the excitation voltage u_{LC} is the difference of both voltage sources e_s and u_o . The excess of energy during the time interval ψ_{fk} is returned to both external voltage sources during the time interval ψ_{rk} . To satisfy the turn-off conditions of the thyristor switches time intervals ψ_{rk} are preceding ψ_{fk} . A variation of the phase angle ψ_{rk} provides a change in the maximal capacitor voltage u_{Crmax} . The characteristic waveforms of the resonant circuit are shown in Fig. 4(b) for the *step-up mode*. The energy is exchanged during the time interval indicated as ψ_{rk} where the excitation voltage u_{LC} is the difference between both voltage sources e_s and u_o . For this mode it is necessary to supply additional energy from both voltage sources e_s and u_o to the resonant network to prevent the interrupt of the distribution of power from the source to the load. Therefore, a boost phase with a time length ψ_{fk} is introduced with an excitation voltage u_{LC} equal to the sum of both

Fig. 4. Resonant inductor current i_r , resonant capacitor voltage u_{Cr} , and excitation voltage u_{LC} for (a) step-down mode, (b) step-up mode.

external voltages. The indicated phases ψ_{rk} and ψ_{fk} are repeated sequentially with alternating polarity. The control of the time intervals ψ_{rk} and ψ_{fk} of each current segment is accomplished for the exchange of energy with the resonant circuit and the distribution of power between the source and the load or vice versa over one half cycle. The direction for the exchange of energy between the source e_s , the load with a voltage u_o , and the resonant circuit is selected by the appropriate switches in both matrices. The equal rank of all matrices in the conversion network allows the direction of the power flow to be mutually exchanged. It is possible to exchange electric energy between both sources in two directions.

Each conversion process is implemented by the selection of at least one switching matrix connected to the voltage source with the lowest voltage level e_l operating as a common rectifier *synchronized* at the zero crossings of the resonant current and one switching matrix operating as a rectifier connected to the voltage source with the highest voltage level $e_h > e_l$ *synchronized* to the resonant current but *shifted* in phase with respect to the zero crossings of the resonant current to guarantee the introduced process of *active damping* by the phase control of the switching instants. A bipolar switching matrix current i_h is exchanged with this voltage source e_h in order to stabilize the stored energy (and consequently the amplitudes of the critical waveforms) of the resonant circuit.

Selecting the correct combination of the excitation voltages opens the way to the exchange of electric energy between two voltage sources interconnected by a resonant circuit for any ratio of the voltages e_s and u_o (even for $e_s \approx u_o$). In reality, a converter shows electric losses that are rooted in the voltage drop of the semiconductor switches and the conduction losses of the passive components and wiring. The transfer of energy will be interrupted for a series-resonant conversion system if its conversion ratio $q \approx 1$. For this mode of operation, electric

Fig. 5. Waveforms generated by the resonant circuit for different values of the source voltage e_s and output voltage u_o . i_r : resonant current. i_{SM2} : unfiltered output current. i_{SM1} : unfiltered source current.

energy is distributed for an effective excitation voltage $u_{LC} = 0$ indicating that no energy is added to the resonant circuit. Ohmic losses will finally decrease the amplitude of the resonant capacitor voltage until the oscillation is interrupted. For each current pulse the losses have to be compensated to maintain the energy level of the switched resonant circuit. It is required to transfer energy to the resonant circuit. The configuration of the switches is modified in order to generate the necessary excitation voltage u_{LC} for which $u_{LC}i_r > 0$ in order to *boost* the stored energy in the resonant circuit as discussed with reference to Fig. 4(b).

Fig. 5 illustrates the resonant current i_r and the unfiltered source and output current i_{SM1} and i_{SM2} characterizing the described processes in the *double excited series-resonant circuit*. The source voltage e_s is a time-varying (sinusoidal) waveform, whereas the output voltage u_o is a constant dc voltage. The current segments indicating the return of energy from or the supply of energy to the resonant circuit are well indicated.

III. ENERGY STORAGE

A sinusoidal voltage source $e_s = \hat{e}_s \sin(\omega_s t)$ is connected to the switching matrix SM1. The phase shift between the source voltage e_s and the fundamental source current i_s is presented by the phase angle φ_s . The source current i_s is constructed from two components i_{sq} and i_{sd} equal to the orthogonal functions.

$$i_s = i_{sq} + i_{sd} = \hat{i}_s \sin(\omega_s t + \varphi_s) \quad (3)$$

where

$$i_{sd} = \hat{i}_{sd} \sin(\omega_s t) \quad (4)$$

$$i_{sq} = \hat{i}_{sq} \cos(\omega_s t). \quad (5)$$

The momentary value of the power as emanated from the source by the converter is

$$S_s = P_s + Q_s = e_s i_s \quad (6)$$

$$= \frac{1}{2} \hat{e}_s \hat{i}_s [\cos(\varphi_s) - \cos(2\omega_s t - \varphi_s)] \quad (7)$$

where P_s is constant. To obtain a time-independent flow of power, the time-varying reactive power Q_s has to be generated by a separate energy buffer. To avoid additional circuitry the output capacitor C_o is available to fulfill this function next to its function as ripple suppressor. Energy can now be returned to the source voltage with very low values. This explains the four-quadrant operation.

A constant flow of power at the output terminals is achieved if an energy buffer is added to the conversion system. The amount of power absorbed by the dc load is nearly constant in magnitude and equals the real input power P_s . The output capacitor absorbs the total amount of

$$Q_{Co}(t) = \frac{1}{2} \hat{e}_s \hat{i}_s \cos(2\omega_s t - \varphi_s^u) \quad (8)$$

where the phase angle φ_s^u is introduced as the phase angle for an unregulated source current, i.e., the fundamental component of the matrix current i_{SM1} (see Fig. 2(b)) is in phase with its source voltage e_s . The optimal capacitor size C_{omin} is given by the relation

$$C_{omin} = \frac{P_o}{\omega_s U_o \Delta u_o \cos(\varphi_s^u)}. \quad (9)$$

For this condition the power factor is equal to one. Equation (9) shows that the output voltage U_o has to be unequal to zero as is expected. The output capacitor C_o as found in resonant dc-dc converters without reactive power control is about seven times smaller in value for the same peak-to-peak output ripple voltage Δu_o .

IV. CONTROL

The shape of the source current i_s is controlled by a pulse modulation process as presented in Fig. 2 and has to be sinusoidal. The control has to minimize the quadrature component i_{sq} of the source current resulting in a power factor equal to one. The amount of distributed power is controlled by the direct component i_{sd} of the source current. The block diagram of the control system is shown in Fig. 6. In the control system two control loops are distinguished:

- 1) The low-frequency *outer loop* governs the (average) dc output voltage and produces a reference current

i_{ref} for the inner-loop ASDTIC controller (ASDTIC: analog signal to discrete time interval converter [21]).

- 2) The high-frequency *inner loop* controls the pulse modulation process of the converter. An ASDTIC-pulse modulation process enables the generation of the low-frequency input current i_s with low distortion

Four control functions are identified and implemented in both control loops: 1) control source current, 2) control output voltage, 3) predictor control, and 4) pulse modulator. The outer loop independently controls the d and q component of the source current i_s . The reference signal i_{dref} and i_{qref} are directly derived from the sinusoidal source voltage e_s . The amplitudes of both current components i_{sd} and i_{sq} can be controlled independently by the dc reference signals U_{oref} and U_{oref} .

The amplitude of i_{dref} is controlled by an error signal $\epsilon_{uo} = U_{oref} - U_o$. The amplitude of this reference signal is controlled by a multiplier MP1. This value is the error signal of the control loop for the output voltage U_o with its reference signal U_{oref} . The inevitable internal converter losses are compensated automatically. The amplitude of i_{qref} is under the control of the error signal $\epsilon_{Qs} = i_s - u_{Qref}$ by multiplier MP2. For a power factor of one, reference voltage u_{Qref} has to be set to zero. Reference signal i_{ref} is composed of i_{refq} and i_{refd} (following (3)). The current i_{SM1} is multiplied by the polarity pol (i_{ref}) for which $pol(i_{ref}) = +1$ or -1 for $e_s \geq 0$ or $e_s < 0$, respectively. The pulse-position control is realized by an ASDTIC controller with its input signals i_{ref} and the rectified switching current $|i_{SM1}|$ where

$$\int_{t_k}^{t_{k+1}} [|i_{SM1}| - |i_{ref}|] dt = 0. \quad (10)$$

The charge of each individual current pulse over each time interval $[t_k, t_{k+1}]$ is equal to the average value of the reference signal $|i_{\text{ref}}|$ over the same interval. The thyristor switches are fired by the output signal of the comparator

at the time t_k ($k = 1, 2, 3, \dots$). An uninterrupted flow of power can be guaranteed by using a *dual-predictor control* algorithm [9]. The predictor circuit will maintain a constant level of the maximal energy in the resonant capacitor (i.e., amplitudes of the resonant capacitor voltage) for each individual current pulse in a predictive way.

V. RESULTS OF THE SIMULATION

The results of a computer simulation of a series-resonant converter including losses confirm the indicated phenomena. The operation of the introduced method is illustrated in Fig. 7. The controlled firing process of the semiconductor switches in matrix SM1 will generate a current i_{SM1} as indicated in this figure. The individual pulses are derived by a modulated process of rectification from the high-frequency carrier i_r . The source current i_s is the result of the current through the input filter capacitor C_s and matrix current i_{SM1} . In Fig. 7(a) the source current i_s is controlled for a source-sided phase shift $\varphi_s = 0$ whereas in (b) this current is presented for $\varphi_s = \pi/4$ (leading current). In Fig. 7(a) the value for the reference signal is $i_{qref} \neq 0$, whereas in Fig. 7(b) it is obvious that the same reference signal is now zero (unregulated mode of operation).

VI. RESULTS OF EXPERIMENTS

An experimental series-resonant converter will be tested with respect to functional integrity under adverse conditions of operation. The conversion system applying thyristor switches, generating a constant dc output voltage at the output terminals, has been designed and constructed to exchange 1 kW of power with an sinusoidal ac source with a maximum voltage $\hat{e}_s = 311$ V ($e_{s\text{rms}} = 220$ V) and a constant frequency $f_s = 50$ Hz. The dc output voltage U_o has a value of 300 V. The output power is 1 kW and is the result of a pulsating flow of energy but with varying pulse frequency. The energy stored in the output capacitor

Fig. 7. Results of simulation for the following waveforms. (a) Capacitor voltage u_{Cr} . (b) Input current switching matrix i_{SM1} . (c) Reference signal i_{ref} . (d) Reference signals i_{dref} . (e) Reference signals i_{qref} . (f) Source voltage e_s . (g) Source current i_s . (h) DC output voltage u_o . For two values of the source-sided phase shift φ_s . (a) $\varphi_s = 0$. (b) $\varphi_s = \pi/4$ (leading current).

is also used to compensate the reactive power. Therefore the resonant converter is designed for a peak power of about twice the average output power. The peak device voltage is 2.2 times $\hat{e}_s$, whereas the peak device current is 6.5 times the dc output current. The maximal stored energy in the passive resonant components are equal to $\epsilon_{Cr} = 0.07$ J and $\epsilon_{Lr} = 0.07$ J. The characteristic waveforms of the source voltage and current confirm the char-

acteristics of the power conversion process to have a linear resistive input impedance and are presented in Fig. 8 for a power factor of one or one-half. These waveforms are derived from a similar converter system and confirm the presented method and can be compared with the results of simulation in Fig. 7. For all conditions of operation the amplitude of the resonant capacitor voltage is kept constant by the dual-predictor controller.

Fig. 8. Experimental waveforms for (a) phase angle $\varphi_s \approx 0$, (b) phase angle $\varphi_s = \pi/3$ (lagging).

VII. CONCLUSIONS

The classical rectifier-filter network, as usually applied to ac-dc converters, limits the available grid power due to the degradation of the power factor [1]. An harmonic free interface achieves a power factor of one and permits a substantial increase in the power.

Different schemes for switching networks with active power factor compensation including dc operation, were proposed [3]. The presented method of active filtering is implemented by a conversion system with thyristor switches generating a high-frequency carrier for the distribution of energy.

To avoid additional low-frequency energy storage elements, a four-quadrant operation is applied (i.e., bidirectional power flow) to maintain the waveform of the source current. The high value of the pulse frequency makes it possible to operate as an accurate and fast-reacting current generator capable of shaping the desired current waveform by the injection of a proper compensating current.

The system performs the function of an active filter optimizing the distortion of the source current. A power factor of one minimizes the distortion of the source current to $\approx 1\%$ without the application of additional filters. Various values of the power factor can be obtained within the limits of the design criteria.

Compensation of the internal losses is executed by control logic integrated into the cascaded control loops.

The four-quadrant operation makes it possible to apply the discussed proposal to dc and ac motor drives.

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