A New Soft-Switched Resonant Buck-Type Rectifier
With Constant Switching Frequency

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Abstract - A new three-phase soft-switched AC-to-DC buck converter operating at constant switching frequency and drawing current at unity power-factor from the source is proposed. Unlike other three-phase resonant switch buck-type rectifiers reported in the literature, the switching frequency of this converter is held constant. The proposed converter uses two switches, which are soft-switched. This results in reduced switching losses and electromagnetic interference (EMI). The principle of operation and detailed analysis of the converter are given and the design procedure of the rectifier is presented. In order to predict the performance of the proposed rectifier, detailed simulation studies using SABER are carried out. A laboratory prototype of the rectifier operating from three-phase 110-Vrms 50 Hz supply and delivering 250 W at 50 V is designed and developed. The source current is sinusoidal and power-factor is unity. The total harmonic distortion (THD) of the line current is less than 5% at full load.

I. INTRODUCTION

The low power-factor and high harmonic content in the input line current of controlled and uncontrolled rectifiers can lead to voltage distortion, overheating and saturation of transformers, transmission and distribution losses, potential resonant conditions with capacitors in the utility and electromagnetic interference (EMI). The IEEE has recommended that modern power electronic equipment should confirm to IEEE-519 standard to control power-factor and harmonic currents. Hence, there is a recognized need for high quality rectifiers, which present high power factor loads to the ac power system. Over and above it is desirable to have high efficiency, lower input and output filter requirements, simple control circuit, reduced number of components and active switches, and low EMI. Many publications have treated the power-factor-correction of three-phase ac-to-dc power supplies [1-6]. Resonant techniques have been successfully used in three phase high power-factor buck-type rectifiers [1-3]. However, the switching frequency of three-phase buck [1-4] and boost [3-6] resonant switch rectifiers reported in the literature is a function of load. The resonant buck-type rectifiers described in [1-2] require a high value of input filter inductor to maintain linear (resistive) input characteristic and THD within the required limit. Also, the ripple in the input current increases at low switching frequency. Hence the input current does not remain constant in a switching cycle. Thus, requirement of large input filter inductors makes these rectifiers bulky and costly. Since the switching frequency is a function of load, the output filter requirement increases at light load conditions. In these rectifiers, the switch is turned off at the end of positive half resonant cycle. During transient condition the current in the resonant branch may not come to zero at the end of the positive half cycle. Hence, soft switching of the switch cannot be ensured under these conditions. The zero-current (ZC) switched buck-type rectifiers reported in [3] have discontinuous input line current and lower value of input inductors compared to those in [1,2]. In addition, these rectifiers require a filter inductor at the output. The THD in the line current of these rectifiers is higher compared to that in [1,2]. Since the input line current is discontinuous (pulsating), an EMI filter is essential at the input. The design of this filter is difficult because the switching frequency is a function of load. This is also true in the case of ZC switched three-phase multi-resonant boost rectifier reported in [5].

In all the above-mentioned rectifiers there is no electrical isolation between the active rectifier stage and the load. In order to eliminate the above-mentioned drawbacks, a new rectifier circuit shown in Fig. 1 is proposed. The switching frequency of the converter is held constant and the design of the control circuit is independent of the values of resonant components. The load is completely isolated from the active rectifier stage. The proposed converter uses two switches, which are soft-switched. The input line current is continuous for output power varying from 40%-100% of the rated power, and the size of input filter inductors is small. Moreover during transient condition, the sudden change in dc link voltage does not affect the functioning of the rectifier and soft switching is ensured. Also, with resonant capacitors at the input side (for voltage-fed buck-type converters), it is possible to vary the amplitude of the source current and hence the power while maintaining the switching frequency constant. The rectifiers reported in the literature do not have this feature.

The proposed rectifier is derived from zero voltage transition pulse-width modulated (ZVT-PWM) single-phase boost type power-factor-correction rectifier [7] by analogy. The transformation from single-phase to three-phase is possible because the input capacitors of the buck-type (voltage-fed) three-phase rectifiers [2] can be modelled as a single equivalent capacitor. This equivalent capacitor is analogous to the resonant capacitor connected across the main boost switch of the single-phase ZVT-PWM rectifier [7]. The principle of operation of the rectifier remains similar.
to that of the ZVT-auxiliary circuit in ZVT-PWM rectifier except for the first interval (interval 1).

![Fig. 1. Proposed constant switching frequency controlled resonant buck-type rectifier.](image)

### II. CIRCUIT DESCRIPTION AND PRINCIPLE OF OPERATION

The proposed converter consists of a diode bridge rectifier, main buck switch $S_1$, switch $S_2$, three input inductors ($L_a$, $L_b$, and $L_c$), input resonant capacitors ($C_r1$, $C_r2$, and $C_r3$), resonant inductor $L_r$, capacitor $C_d$, isolation transformer $T_r$, rectifier diodes $D_1x$ and $D_2x$, and output filter capacitor $C_f$. Inductor $L_r$ and input resonant capacitors ($C_r1$, $C_r2$, and $C_r3$) form a resonant tank circuit. Switch $S_1$ is turned ON under zero-current (ZC) condition and turned OFF under zero voltage (ZV) condition, while $S_2$ is turned ON and OFF under ZV condition. The voltage waveform across resonant capacitors ($C_r1$, $C_r2$, and $C_r3$) is pulsating sinusoidal (quasi-sinusoidal) in nature, whose peak value can be varied by varying the ON time of $S_2$. Varying this ON time and keeping switching frequency of $S_1$ constant can vary the output power. When current in the resonant inductor $L_r$ is flowing in the positive direction, a voltage is impressed across the primary of $T_r$, and the secondary voltage makes diode $D_1x$ conduct. Likewise, diode $D_2x$ conducts when current in the resonant inductor $L_r$ is flowing in the negative direction. In both cases the primary voltage is clamped to a magnitude of $V_p = V_a + N_1$, where $N_1 = N_a/N_2$ is the primary to secondary turn's ratio of $T_r$ and $V_a$ is the output voltage of the rectifier. The use of transformer not only provides the isolation but also reduces the current stress on $S_1$ and $S_2$.

To analyse the proposed constant switching frequency controlled three-phase resonant rectifier shown in Fig. 1, it is sufficient to consider an operating point at time $t = \pi/2$ of three-phase input voltages. At this instant, phase voltage $V_a$ is at its peak value and $V_b$ and $V_c$ are both negative and equal in magnitude, which is one half of $V_a$. Under this condition $C_r2$ and $C_r3$ (shown in Fig. 1) charge and discharge in a similar manner. Hence, the input side resonant capacitors $C_r1$, $C_r2$, and $C_r3$ can be replaced by an effective capacitor $C_x$, which is equal to the series connection of $C_r1$ and equivalent capacitance of parallel-connected $C_r2$ and $C_r3$. The input voltage sources and the input inductors $L_a$, $L_b$, and $L_c$ can be replaced by a current source. One switching cycle is divided into seven intervals. The equivalent circuit for these intervals and the various waveforms are shown in Figs. 2 and 3 respectively. These equivalent circuits are drawn by assuming negligible ripple in the source current. While referring to the circuit equations in any interval it should be noted that $V_{C_dX}$, $V_{C_X}$, and $I_{L_r}$ are the values of $V_{C_d}$ (voltage across $C_d$), $V_{C_x}$ (voltage across $C_x$) and $I_{L_r}$ (resonant inductor current) at the end of interval $X$. The operation of the circuit is as follows:

![Fig. 2. Equivalent circuits for various operating intervals of the converter.](image)

![Fig. 3. Gating signals and circuit waveforms of the rectifier during one switching cycle.](image)
Interval 0: \( (t_0 \leq t \leq t_1) \): In this mode, all switches are OFF. Initial conditions for this interval are \( I_{in}=0 \) and \( V_{Cd} = V_{C3} \). In this interval, tank capacitors \( C_1-C_3 \) charge linearly at a rate proportional to their respective line currents. The equation for capacitor voltage is given by:

\[
V_{Ct0} = \frac{1}{C_t} \int I_{in} dt
\]

(1)

This mode will continue until switch SI is turned ON under ZC condition. Since the input line-to-line voltages \( V_{ab} \) and \( V_{ac} \) are equal, turning ON of SI makes diodes D1, D5 and D6 to conduct. This initiates the next interval.

Interval 1: \( (t_1 \leq t \leq t_2) \): This mode starts when SI is turned ON under ZC condition. Initial conditions for this interval are \( I_{L1} = 0 \), \( V_{Cd} = V_{C3} \) and \( V_{Cc} = V_{Cd} \). In this interval we see that D1, D5, D6, D1x and SI are ON. Capacitors C1-C3 ring with resonant tank inductor \( L_r \). Current through \( L_r \) starts increasing from zero. This mode ends when current in \( L_r \) becomes equal to \( I_{in} \). The equations for this mode are:

\[
I_{L1} = \left( \frac{V_{Ct0} + V_{Cd5} - V_s}{R_0} \right) \sin \omega_0 t + I_{in} \left( 1 - \cos \omega_0 t \right)
\]

(2)

\[
V_{Ct1} = \left( V_{Ct0} - V_{Cd5} - V_s \right) \cos \omega_0 t + \frac{I_{in}}{C} \left( 1 - \frac{\sin \omega_0 t}{\omega} \right)
\]

(3)

\[
V_{Ct1} = \left( V_{Ct0} - V_{Cd5} - V_s \right) \cos \omega_0 t + \frac{I_{in}}{C} \sin \omega_0 t + \left( V_{Cd5} + V_s \right)
\]

(4)

Where,

\[
\omega_0 = \sqrt{\frac{2}{L_r C}}, \quad R_0 = \sqrt{\frac{2L_r}{C}} \quad \text{and} \quad C = C_x = C_d.
\]

Interval 2: \( (t_2 \leq t \leq t_3) \): This mode starts when \( I_{in} \) is greater than the input current \( I_{in} \). In this mode D1, D5, D6, D1x, and SI are ON. Initial conditions for this interval are \( I_{L2} = I_{L1} \), \( V_{Cd} = V_{C3} \) and \( V_{Cc} = V_{C1} \). This mode ends when all the three resonant tank capacitors C1-C3 are fully discharged. At this instant all the three-phase bridge rectifier diodes and anti-parallel diode of S2 get forward biased.

\[
I_{L2} = \left( A \cos \omega_0 t + B \sin \omega_0 t \right) + \frac{C_p}{C_x}
\]

(5)

\[
V_{Cd2} = \left( \frac{E}{C_d} \cos \omega_d t + \frac{F}{C_d} \sin \omega_d t \right) + \frac{C_p}{C_d} t + V_{Cd1} + \frac{E}{C_d}
\]

(6)

\[
V_{Cd2} = \left( \frac{E}{C_d} \cos \omega_d t + \frac{F}{C_d} \sin \omega_d t \right) + \frac{C_p}{C_d} t + V_{Cd1} + \frac{E}{C_d}
\]

Where,

\[
C_p = \frac{C_d C_x}{C_d + C_x}, \quad A = \frac{C_p}{C_d}, \quad B = \left( \frac{V_{Cd1} - V_s - V_{C3}}{\omega_0} \right)
\]

\[
E = \frac{B}{\omega_0}, \quad F = -\frac{A}{\omega_0} \quad \text{and} \quad \omega_0 = \sqrt{\frac{2L_r}{C_d L_r}}
\]

Interval 3: \( (t_3 \leq t \leq t_4) \): In this mode all the bridge rectifier diodes, anti-parallel diode of S2, and \( D_1, D_5, D_6, D_1x \) are ON. Switch S2 is turned on under ZV condition. Initial conditions for this interval are \( I_{L3} = I_{L2} \) and \( V_{Cd} = V_{C3} \). Current flowing through \( L_r \) at the end of this interval is the same as the input current \( I_{in} \). The circuit equations for this mode are:

\[
I_{L3,4} = \left( I_{L2} \cos \omega_0 t - \frac{G}{\omega_0 \omega_0} \sin \omega_0 t \right)
\]

(7)

\[
V_{Cd3,4} = \left( G \cos \omega_0 t + \frac{I_{L2}}{\omega_0 \omega_0} \sin \omega_0 t \right) - V_s
\]

(8)

\[
G = V_{Cd2} + V_s \quad \text{and} \quad \omega_0 = \sqrt{\frac{1}{L_r C_d}}
\]

This mode ends when the current through anti-parallel diode of S2 becomes zero.

Interval 4: \( (t_4 \leq t \leq t_5) \): In this interval \( D_1, S1 \) and \( S2 \) are ON. Initial conditions for this interval are \( I_{L4} = I_{L3} \) and \( V_{Cd} = V_{C3} \). Current begins to flow through S2 at \( t = t_4 \), and current flowing through the tank inductor \( L_r \) is still positive but less than the input current \( I_{in} \). This mode ends when \( I_{L4} \) becomes zero. The circuit equations for this mode are the same as those given in interval-3. However, the initial conditions are different.

Interval 5: \( (t_5 \leq t \leq t_6) \): In this interval \( D_2, D_1, S1 \) and \( S2 \) are ON. Initial conditions for this interval are \( I_{L5} = 0 \) and \( V_{Cd} = V_{C3} \). Current flowing through S2 is the sum of input current \( I_{in} \) and current flowing through \( L_r \) is still positive but less than the input current \( I_{in} \). The circuit equations for this mode are:

\[
I_{L5} = \frac{V_{Cd4} - V_s}{\omega_0 \omega_0} \sin \omega_0 t
\]

(9)

\[
V_{Cd5} = \left( V_{Cd4} - V_s \right) \cos \omega_0 t + V_s
\]

(10)

\[
V_{Cd5} = \left( V_{Cd4} - V_s \right) \cos \omega_0 t + V_s
\]

(11)

Interval 6: \( (t_6 \leq t \leq t_7) \): In this interval, switch S2 is ON. Input current \( I_{in} \) flows through switch S2. Since the voltage
across Cr1-Cr3 is zero, S2 can be turned OFF under ZV condition. The ON time of S2 depends on load. However, the minimum ON time of switch S2 is the sum of the time durations of intervals-4 and 5 so that the resonance cycle gets completed. It should be noted that the range of load that the converter can supply is determined based on this condition.

III. ANALYSIS

The analysis of the proposed converter can be simplified by using the ideal simplified single-phase model shown in Fig. 4. As shown in figure the entire converter with switches can be modelled as a single bi-directional switch S connected across the capacitor C. This C and L represent the input filter and Vin is the single-phase input voltage source. When S is turned ON, C discharges instantaneously and voltage across L is the supply voltage. When S is turned OFF, C charges linearly and voltage across L is the difference between source voltage and capacitor voltage. Fig. 5 shows the ac source voltage at t=π/2 and voltage across L & C during one switching cycle. The duty cycle of S is maintained at 50%. From simulation studies it is found that if duty cycle is above 50% and switching frequency of S is twice as that of resonance frequency of the circuit, there is a phase displacement between input (inductor) current and ac voltage source. However, the current remains sinusoidal.

![Ideal simplified single-phase model of the rectifier.](image)

Fig. 4. Ideal simplified single-phase model of the rectifier.

Assuming that the ac voltage source (Vin) remains constant over a switching cycle, for average voltage across the inductor to be zero, area A1 should be equal to area A2. For Fig.5, the following equations can be written:

\[ \frac{1}{2} T_{OFF} \cdot V_{C_{peak}} = T_S \cdot V_{in_{peak}} \]  

(12)

\[ V_C = \frac{1}{C} \int I_L \, dt \]  

(13)

Where,

- \( V_C \) = Capacitor voltage
- \( T_S \) = Switching period
- \( I_L \) = Inductor current

Fig. 6 shows the typical steady state voltage waveform of \( V_{Ct} \) and gating pulses for S1 and S2 of the proposed converter. The figure also shows the approximate shape of the input resonant capacitor voltage waveform, which is a triangle of area \( \frac{1}{2} T_{off} \cdot V_{C_{peak}} \) where \( T_{off} \) is the OFF time of S2. Hence the use of simplified model for the analysis is justified.

From the above equations it can be inferred that in order to maintain zero average voltage across the inductor over a switching cycle, the peak value of capacitor voltage should increase with ON time of S.

![Ideal circuit waveforms.](image)

Fig. 5. Ideal circuit waveforms.

![Gate pulses for S1 and S2, Simulated voltage waveform of VCt and approximate waveform used for analysis.](image)

Fig. 6. Gate pulses for S1 and S2, Simulated voltage waveform of VCt and approximate waveform used for analysis.

![Input voltage (Vin) and inductor current waveforms for L=1.643mH and C=155nF with switch duty cycle of 50% and 25% respectively. Switching frequency is held constant at 20 kHz.](image)

Fig. 7. Input voltage (Vin) and inductor current waveforms for L=1.643mH and C=155nF with switch duty cycle of 50% and 25% respectively. Switching frequency is held constant at 20 kHz.
It should be noted that as duty cycle increases, the available time for the capacitor to charge decreases. Now that the capacitor should charge to a higher value than the previous at a lesser time, this could be achieved by forcing a higher current in the inductor. From the simulation study on the simplified model, it is found that if the value of switching frequency of $S$ is twice that of resonance frequency of the circuit, the average value of current over half a cycle of ac source voltage increases by approximately 135% when duty cycle is increased from 25% to 50%, while maintaining current in the inductor continuous and in phase with input voltage. This variation in duty cycle is required to vary the output power of the converter from 40 to 100%. Fig. 7 shows the simulated inductor current waveform during half a cycle of the input for 50% and 25% duty cycle of switch S. As seen from the figure, the current through inductor is sinusoidal with finite ripple and in phase with the input voltage. The average value of current drawn from the input over half a cycle is also indicated. In order to account for the ripple in the input current at the design stage the following exercise is carried out.

Figs. 8(a) shows the plot of peak value of ripple free input current versus its average value over half a cycle of ac source for switching frequency of 20 kHz. For a particular value of the peak current and 50% duty cycle, the value of L and C are determined from equations (12) and (13). The simplified model is then simulated using these values, and the average value of current drawn from the source over half a cycle is determined. This variation is shown in Fig. 8(a). Figure also shows this variation at 25% duty cycle. A similar exercise is carried out for a switching frequency of 30 kHz and the results are shown in Fig. 8(b). It can be noted that the ratio of average input current without ripple to average input current with ripple at a switching frequency of 20kHz and 50% duty cycle, is approximately 1.2. Moreover, the ratio of average input current with ripple at 25% duty cycle and that at 50% duty cycle is approximately 0.4. So, it can be inferred that by varying the duty cycle of S2 from 50% to 25%, the power drawn by the rectifier varies from 100% to 40%.

Thus, it can be shown that the switching timings of S1 and S2 and values of circuit elements (Cr1-Cr3), Lr and Cd can be chosen in such a manner that a required amount of average ac input current over half a cycle of ac source and hence the input power is drawn by the rectifier. Under ideal conditions, if the losses taking place in the converter are neglected, then the input power is equal to the output power. The following guidelines are made use of when designing the rectifier:

IV. DESIGN

The design of the single ended converter is carried out to achieve the following specification.

1) Output power: $P_o = 250$W (max) at 50 V for a switch duty cycle of 50%.
2) Output power: $P_o = 100$W (min) at 50 V with a switch duty cycle of 25%.
3) Switching frequency $F_s = 20$ kHz (constant).
4) Supply voltage: 3 phase 110Vrms (L.L) 50 Hz.

Using the ideal model of the rectifier shown in Fig. 4 and equation (12), the peak value of input resonant capacitor voltage ($V_{Cr1}$) of phase ‘a’ is given by,

$$\frac{1}{2} T_{OFF} * V_{Cr1_{peak}} = T_s * V_{in_{peak}}$$

$$\therefore V_{Cr1_{peak}} = 360 \text{ volts.}$$

The required peak value of input phase current assuming negligible ripple and unity power factor is given by,
The transformer turns ratio $N_x$ affects the circuit performance as follows: As the turns ratio increases, the current stress on $S1$ decreases while time available for turning ON of $S2$ under ZV condition, decreases. Hence a judicious selection of $N_x$ is essential. The following iterative procedure is used to arrive at this value:

When $S1$ is turned ON, the governing equation is:

$$-\sqrt{3} \times V_{Cr1_{\text{res}}} + V_X - V_{Cd} = 0 \quad (17)$$

Since $V_{Cr1_{\text{res}}}$ and $I_m$ are known, initially $V_{Cd}$ can be determined by assuming a low value of $V_X$. In order to turn ON $S2$ under ZV condition, $V_{Cd}$ should become zero and remain at this value for some time. This can be verified by solving Equations (4) and (6). If this condition is satisfied, the value of $V_X$ is further increased and same procedure is repeated. The maximum value of $V_X$ for which this condition is satisfied is chosen to determine the turns ratio. This turns ratio is given by $V_X / V_D$.

V. SIMULATION RESULTS

In order to validate the analysis and to predict the performance of the rectifier, detailed simulation study is carried out on SABER. The input voltage to the rectifier is 50 Hz, 110Vrms (L-L), and circuit parameters are: $Cr1 = Cr2 = Cr3 = 102.88nF$; $Lr=68.46\mu H$, $Cd=102.88nF$, and $La=Lb=Lc=1.643mH$ respectively. Transformer ($T_1$) primary to secondary inductance ratio is chosen to be five. The value of load resistance is set equal to 108Ω. For these circuit parameters the converter can supply a load of 250 W at 50 V for 25µs ON time of $S2$. The switching frequency is maintained constant at 20 kHz. Fig. 9(a) shows phase ‘a’ current and its corresponding Fourier spectrum while supplying a load of 250 W at 50 V.

These results for an ON time of 12.5µs are shown in Fig. 9(b). At this value of ON time the converter is supplying 100 W at 50 V. It can be observed that the source current is sinusoidal and THD at full and 40% load are 3% and 4.2% respectively. Fig. 10 shows the other important voltage and current waveforms. There is a close resemblance between these waveforms and those shown in Fig. 3.
VI. EXPERIMENTAL RESULTS

In order to validate the simulated results a laboratory prototype of the proposed converter is designed and fabricated. The converter is designed to operate from three-phase 110Vrms source and can supply a load of 250W at 50 V. The switching frequency of the converter is maintained at 21kHz and the component values are: L_a=L_b=L_c=1.62mH, L_r=64μH, C_{r1}=C_{r2}=C_{r3}=160nF, and C_d=110nF. A Eupec manufactured BSM 50GB 120DN2 IGBTs are used for S1 and S2, an International rectifier 60EPF06 for three-phase diode bridge and transformer T_x secondary winding rectifier diodes. The various waveforms obtained from the prototype are shown in Fig. 11. From Fig. 11 (a, b) it can be observed that the source voltage waveform is distorted. This has affected the source current waveform and its Fourier spectrum. It can be observed that waveforms obtained from the prototype closely resemble the corresponding simulated waveforms.
VII. CONCLUSION

This paper proposes a constant switching frequency controlled three-phase buck-type soft-switched rectifier that draws sinusoidal current from the source. The main features of this converter are constant switching frequency operation under variable load condition, reduced size of input filter inductors and output filter capacitors, and inherent isolation between source and load. Due to these features there is a significant reduction in the size of the rectifier. Also, the performance of the control circuit is independent on the resonant inductor and capacitor values. The operation of the converter is explained and detailed analysis is carried out. In order to predict the performance of the converter detailed simulation studies on SABER are performed and these results are experimentally validated.

VIII. REFERENCES