Improved Modeling and Analysis of a High-Performance ac/dc Converter for High-Frequency Power Distribution System

This paper presents an improved modeling and detailed analysis for determining the steady-state and transient performance of a high-performance ac/dc converter topology that can be utilized in high-frequency power distribution systems. The output voltage is regulated by controlling the turn on time of the anti-parallel semiconductor control switches. Both the steady-state and transient performance is explored in details. It is shown that the converter has almost unity power factor of close to 99%, input current is almost sinusoidal with a total harmonic distortion of less than 2%, a conversion efficiency of close to 96% and very low inrush current of around 110% can be achieved, when appropriate circuit components are used. Based on predetermined value of the input current total harmonic distortion, consideration is given to the design of important circuit components.

Keywords: Unity power factor, total harmonic distortion, efficiency, Power electronics

NOMENCLATURE

\[ F \] : Supply frequency (Hz) \\
\[ \omega \] : Supply angular frequency (rad/s) \\
\[ v_{in} \] : Input voltage (instantaneous value) (V) \\
\[ V \] : rms value of \[ v_{in} \] (V) \\
\[ i_{in} \] : Input current (instantaneous value) (A) \\
\[ R_s \] : Resistance of the inductor (\( \Omega \)) \\
\[ L_s \] : Series inductor (H) \\
\[ C_s \] : Series capacitor (F) \\
\[ R_D \] : On-state resistance of a diode (\( \Omega \)) \\
\[ THD \] : Input current total harmonic distortion \\
\[ V_o \] : Output voltage (V) \\
\[ PF \] : Input power factor \\
\[ C_o \] : Output voltage filter capacitor (F) \\
\[ \Phi \] : Current zero crossing angle referred to input voltage zero crossing (\( \text{o} \)) \\
\[ R_o \] : On-state resistance of semiconductor switch (\( \Omega \)) \\
\[ R_o \] : Load resistor (\( \Omega \)) \\
\[ \alpha \] : Control angel referred to input voltage zero crossing (\( \text{o} \))

1. INTRODUCTION

Extensive publications that focus on single-stage power factor corrected ac/dc converters are available in the literature [1-5]. Garcia et al. [1] listed 40 of such publications while Singh et al. [2] listed 463 of them. A vast majority of these publications are of pulse width modulated type or switch mode power supply type, boost, buck, buck-boost and multilevel with unidirectional or bidirectional power flow. As such they have complex control circuits, ending up with rather bulky circuits. However in some advanced high-frequency power distribution systems, such as the ones used in Space Stations[10], the following stringent requirements are placed on the design of an ac/dc converters: 1) high efficiency; 2) \( PF \) of close-to-unity; 3) near-sinusoidal input current (very low \( THD \)); 4) light weight and high power density; 5) controlled output voltage; 6) low input inrush current; 7) low EMI; 8) high reliability (with minimum circuit components), and 9) Compact design. Built on earlier topologies [6-9], Tanju et al. [10] presented a simple, easy to build, compact and low
cost scheme, that meets all the aforementioned design objectives, as shown in Fig. 1. The circuit was built and tested on a bread-board version.

![Schematic of the high-performance ac/dc converter circuit](image)

**Fig. 1. Schematic of the high-performance ac/dc converter circuit [10]**

Although the experiments conducted on the circuit showed good results, however, modeling and analysis presented in [10] suffered from the following major drawbacks:

1) Except for the standard well-known equations, all the equations that were derived for the analysis of the circuit are incorrect (out of 19 major equations 16 of them are wrong). Therefore the analysis given in [10] can’t predict the performance of the given topology. The main reason for this is the wrong Fourier series expression of the bridge rectifier input voltage, $V_{ab}$ (see Fig. 1), which is used in all of the subsequent equations. As correctly stated in [10], with a sufficiently high value of the output filter capacitor, $C_o$, $V_{ab}$ will have a quasi-square wave shape, as shown in Fig. 2. The Fourier series expression of $V_{ab}$ derived in [10] is given in Appendix A, and a plot of it is shown in Fig. 3, for given values of $V_o$, $\alpha$ and $\Phi$, and is compared with the plot of the correct Fourier series expression proposed by this paper. It is clearly evident from this figure that Fourier series expression given in [10] has no relation with the correct shape of $V_{ab}$. Since all the subsequent derivations involve equation of $V_{ab}$, naturally all those equations will be incorrect and accordingly the whole analysis will be wrong.

2) In [10] the upper and lower limits of the control angle $\alpha$ are taken as 90° and 180° for all cases, without considering the values of the circuit components. Of course this is not true, because the upper limit of $\alpha$ is limited by the value of $\Phi$, as there is no control at and beyond $\alpha=\Phi$. On the other hand, the value of $\Phi$ is obtained from the solution of equation (20), which is controlled by the values of the circuit elements. It was found that in any case the lower limit of $\alpha$ is above 90° and upper limit of it is always less than 180°. Details of this will be discussed in a later section.

3) In [10] nothing is mentioned with regards to the value of $\Phi$, without which the performance of the given topology can’t be analyzed. This is most probably due to the fact that the equations derived in [10] did not converge to any logical value for $\Phi$.

4) For transient analysis no modeling for simulation is presented in [10]. Only results obtained from the simulator ATOSEC 5 [11] are presented, which most probably, as in SPICE, requires only the schematic representation of the circuit, with component models and values.
5) In the abstract and conclusion of [10] it is claimed that a conversion efficiency of more than 96% is achieved, but there is no analysis or any kind of results in the body of the presented work to prove that.

For the aforementioned five very important reasons, it became vital to give a completely new modeling and analysis for the steady-state and transient performance of the given topology. Therefore, the purpose of this paper is to present a complete new modeling and analysis for both the steady-state and transient performance, as well as discussing the results in details. Moreover, a special consideration is given to the design of resonant network components (resonant network), $L_s$ and $C_s$.

2. OPERATING PRINCIPLES

Operating principles of the topology given in Fig. 1 is explained in details in [10]. However for the sake of completeness, brief operating principles will be given here, using the operating diagram shown in Fig. 2. The control part of the circuit is composed of a series resonant network ($L_s$ and $C_s$) and a back-to-back connected pair of semiconductor switches $Q_A$ and $Q_B$. The series resonant network is closely tuned to the supply frequency, showing a small impedance to the fundamental component of the input current and a very large impedance to the harmonic currents. From $\omega t=0$ to $\omega t=\alpha$ none of the semiconductor switches is conducting, hence the bridge rectifier is powered from the supply through the resonant network and thus $V_{ab}=V_o$ (assuming sufficiently large output filter capacitor, $C_o$).

When switch $Q_A$ is turned on at $\omega t=\alpha$ and it is allowed to conduct until current crosses zero at $\omega t=\Phi$. During the period $\Delta \omega t=\Phi-\alpha$ the bridge rectifier input is shorted by $Q_A$, energy is stored in resonant network ($L_s$ and $C_s$), thus $V_{ab}=0$ and $C_o$ discharges through the output resistor $R_o$, keeping the voltage across it almost constant and the current through it continuous. From $\omega t=\Phi$ to $\omega t=\pi+\alpha$ again none of the switches is conducting and thus the bridge rectifier is powered from the supply and the energy stored in the resonant network is boosted to the output circuit, but this time, and due to the negative polarity of the supply voltage, $V_{ab}=-V_o$. At $\omega t=\pi+\alpha$, switch $Q_B$ is turned on and it is allowed to conduct until current zero crossing at $\omega t=\pi+\Phi$. During this period, $Q_B$ shorts the bridge rectifier input and thus $V_{ab}=0$, while the energy is stored in resonant network and $C_o$ discharges through $R_o$, keeping the voltage across it almost constant and the current through it continuous. At $\omega t=\pi+\Phi$, $Q_B$ is turned off at current zero crossing and due to positive polarity of the supply voltage, $V_{ab}=V_o$ and then the cycle starts repeating itself again. By this mean, the rectifier bridge input voltage $V_{ab}$ becomes a quasi-square wave as can be seen in Fig. 2. The outputs power and voltage are regulated by controlling the semiconductor switches’ ($Q_A$ and $Q_B$) turn on time (control angle $\alpha$). Hence, with a $C_o$ of about 10 $\mu$F or more, a continues and regulated output voltage with very low ripple content is achieved.
3. STEADY-STATE ANALYSIS

In steady-state analysis it is assumed that the supply voltage is an ideal sine-wave signal and all the circuit elements are ideal, i.e., on-state resistance of the semiconductor switches, QA and QB (assuming them to be IGBTs) and diodes in addition to the inductor’s resistance in the control circuit are negligible.

The input voltage is:

\[ v_{in} = \sqrt{2}V \sin(\omega t) \quad (1) \]

As was explained earlier, the rectifier bridge input voltage, \( V_{ab} \), ignoring the ripple content of the output voltage, is a quasi-square wave, as shown in Fig. 2. The Fourier series expression of \( V_{ab} \) can be expressed as:

\[ V_{ab} = \frac{2V}{\pi} \sum_{n=1}^{\infty} \frac{\sqrt{a_n^2 + b_n^2}}{n} \sin(n\omega t - \beta_n) \quad (2) \]

Where \( n=1 \) gives fundamental component, \( n=3,5,7, \ldots \) give harmonics and;

\[ a_n = \sin(n\alpha) + \sin(n\Phi), \quad b_n = \cos(n\alpha) + \cos(n\Phi) \quad (3) \]

\[ \beta_n = \tan^{-1}(a_n/b_n) \quad (4) \]

Using equation (2) a plot of \( V_{ab} \) is given in Figure 3, which confirms the correctness of the equation (2). Figure 3 clearly shows that in reference [10] \( V_{ab} \) is expressed in an incorrect way, which was the caused all of the related analysis to be incorrect.
This paper by Tanju at. al. [10] introduces a Fourier series expression for the rectifier input voltage, $V_{ab}$.

For this reason in reference [10] a total of 13 major equations, that are involving the Fourier series expression of $V_{ab}$, are incorrect, which completely destroys the analysis of the given circuit. The Fourier series expression of $V_{ab}$ given in reference [10] is given in Appendix and its plot is included in Figure 3. Observation of Fig. 3 proves that a new analysis should be given for the power factor correction circuit under investigation.

Representing the part of the circuit including the rectifier bridge and beyond with a voltage source, which is the Fourier series expression of $V_{ab}$, the complete circuit can be represented with an equivalent circuit given in Fig. 4. The voltage equation governing the equivalent circuit given in Fig. 4 is:

$$\sqrt{2}V \sin(\omega t) = -\frac{2V_m}{\pi} \sum_{n=1,3,5,...}^\infty \frac{\sqrt{a_n^2 + b_n^2}}{nZ_n} \sin(n\omega t - \beta_n) + L_s \frac{d}{dt} \int i_{in}dt$$

(5)

Solution of (5) gives:

$$i_{in} = -\sqrt{2}V \frac{\sin(\omega t)}{Z_L} - \frac{2V_m}{\pi} \sum_{n=1,3,5,...}^\infty \frac{\sqrt{a_n^2 + b_n^2}}{nZ_n} \cos(n\omega t - \beta_n)$$

(6)

Where:

$$Z_L = \text{fundamental impedance} = \frac{X_L - X_C}{\omega L_s - \frac{1}{\omega C_s}}$$

(7)

$$Z_n = \text{nth harmonic impedance} = nX_L - \frac{X_C}{n} = n\omega L_s - \frac{1}{n\omega C_s}$$

(8)

Since $V_o$ and $\Phi$ (current zero-crossing angle) are not known, the input current and hence the performance of the circuit can’t be computed from the equations given above. Therefore two equations are needed to solve for $V_o$ and $\Phi$, and then determine the performance of the circuit. The input current is = 0 when $\omega t = \Phi$, and for this condition equation (6) gives:

$$\cos(\Phi) + \frac{V_m \sqrt{2}Z_L}{V} \sum_{n=1,3,5,...}^\infty \frac{\sqrt{a_n^2 + b_n^2}}{nZ_n} \cos(n\Phi - \beta_n) = 0$$

(9)
For a given control angle $\alpha$, the value of $\Phi$ would have been calculated using equation (5), if the value of $V_o/V$ are known. The value of $V_o/V$ is obtained by setting input power equal to output power.

Since the input voltage contains no harmonics, the fundamental average input power can be expressed as;

$$P_{in} = V I_f \cos(\delta) \tag{10}$$

Where, $I_f$ is the rms value of the fundamental component of the input current and $\delta$ is the angle between the fundamental current and supply voltage. From equation (6), the fundamental current can be expressed as;

$$i_f = \sqrt{A^2 + B^2} \sin(\omega t - \delta) \tag{11}$$

where;

$$A = -\frac{\pi \omega}{\alpha_2} \sqrt{\alpha_1^2 + \beta_1^2} \sin(\beta_1), \quad B = \frac{\pi V_2}{\beta_2} + \frac{\pi \omega}{\beta_2} \cos(\beta_2) \tag{12}$$

$$\delta = \cos^{-1} \left( \frac{A}{\sqrt{A^2 + B^2}} \right) \tag{13}$$

in which $\delta$ is the angle between fundamental current and the supply voltage and $\cos(\delta)$ is the fundamental power factor, then the average fundamental power can be expressed as;

$$P_{inf} = V \frac{\sqrt{A^2 + B^2}}{\sqrt{2}} \frac{A}{\sqrt{A^2 + B^2}} = \frac{VA}{\sqrt{2}} \tag{14}$$

Substituting (12) into (15) yields;

$$P_{inf} = -\frac{\pi \omega}{\alpha_2} \sqrt{\alpha_1^2 + \beta_1^2} \sin(\beta_1) \tag{15}$$

Then substituting (3) and (4) (for fundamental terms) into (15) the fundamental input power becomes;

$$P_{inf} = -\frac{\pi V}{\pi \beta_2} \left( \sin(\alpha) + \sin(\Phi) \right) \tag{16}$$

The output power is;

$$P_2 = V_2^2 / R_2 \tag{17}$$

Ignoring losses and setting the input power equals to the output power yields;

$$V_0 = -\frac{\pi V R_2}{\pi \beta_2} \left( \sin(\alpha) + \sin(\Phi) \right) \tag{18}$$

from which we obtain;

$$\frac{V_0}{V} = -\frac{\pi R_2}{\alpha_2} \left( \sin(\alpha) + \sin(\Phi) \right) \tag{19}$$
Substituting (19) into (9) yields;

\[ \cos(\Phi) - \frac{2R_o}{a_x^2} (\sin(\alpha) + \sin(\phi)) \sum_{n=1,3,5}^{\infty} \frac{\sqrt{a_x^2 + b_z^2}}{n-z_n} \cos(n\Phi - \alpha_n) = 0 \] \hspace{1cm} (20)

For a given control angle \( \alpha \), load resistor \( R_o \) and impedance \( Z_n \) the value of \( \Phi \) is obtained from (20) by using secant method [11].

3.1. Calculation of Input Current Total Harmonic Distortion (THD)

\( THD \) is an important measure of the quality of the input current, and as is well known lowering the \( THD \) will result in higher power factor which is the main required objective to be achieved using the circuit under investigation. The \( THD \) is defined as [13]:

\[ THD = \frac{\sqrt{\sum_{n=1}^{\infty} I_n^2}}{I_1} \] \hspace{1cm} (21)

where \( I_n \) is the rms value of the \( n \)th harmonic current. Substituting (6) and (11) into (21), the \( THD \) becomes:

\[ THD = \frac{2k}{a_x} \sqrt{\sum_{n=1}^{\infty} \left( \frac{a_x^2 + b_z^2}{A^2 + b^2} \text{Re} \left( a_x^2 + b_z^2 \right) \right)} \] \hspace{1cm} (22)

3.2. Calculation of the Input Power Factor (PF)

The power factor in a circuit containing a non-sinusoidal voltage or current is given by [13]

\[ PF = \text{displacement factor} \times \text{distortion factor} \] \hspace{1cm} (23)

Using (11), (13) and (21) in (23) the power factor \( PF \) becomes:

\[ PF = \frac{A}{\sqrt{(A^2 + B^2)(1 - THD^2)}} \] \hspace{1cm} (24)

4. DESIGN OF RESONANT NETWORK COMPONENTS

The values of the components in the resonant part of the circuit shown in Fig. 1, i.e. values of \( Z_1 \) and \( L_s \) are closely related to the value of the load resistor \( R_o \). Moreover, the value of \( C_s \) depends on the values of \( Z_1 \) and \( L_s \). Therefore the value of the \( R_o \) will be taken as the base impedance, while \( Z_1 \) and \( X_L \) will be expressed as per-unit (p.u.) of \( R_o \). As mentioned in [10], to achieve near unity PF and very low \( THD \) the value of \( Z_1 \) should be a small value leading impedance. It was found that optimum value of \( Z_1 \) lies between -0.2 p.u. to -0.3 p.u. When absolute value of \( Z_1 \) is less than 0.2 p.u., the equation (20) does not converge to valid values of \( \Phi \) for values of the control angle \( \alpha \) between 90° to 110°, which limits the maximum value of the output voltage and output power. It was found that the optimum value of \( Z_1 \) is around -0.25 p.u. Therefore is proposed that \( Z_1 \) to be fixed at -0.25 p.u., i.e.

\[ Z_1 = -0.25R_o \text{ or } Z_1 = -0.25 \text{ p.u.} \] \hspace{1cm} (25)

The value of \( L_s \) or \( X_L = \omega L_s \) is very much related the amplitude of the third harmonic current, and hence to the value of \( THD \). To achieve a \( THD \) of around 2%, the third harmonic impedance should be around 50 times the amplitude of \( Z_1 \). This argument yields the following:

\[ X_L = 5X_1 = 50(0.25R_o) = 3X_1 = (X_1 + 0.25R_o)/3 \] \hspace{1cm} (26)

\[ Z_1 = 50 |Z_1| = 50(0.25R_o) = 3X_1 = (X_1 + 0.25R_o)/3 \] \hspace{1cm} (27)

Equation (27) gives \( X_L = 5R_o \) or \( X_L = 5 \text{ p.u.} \). Therefore, for near optimal performance, the resonant part components \( L_s \) and \( C_s \) are selected as:

\[ L_s = 5R_o/\omega \text{ and } C_s = 1/(5.25\omega R_o) \] \hspace{1cm} (28)
For \( R_o = 12 \, \Omega \), \( f = 20 \, \text{kHz} \) and \( Z_l = -0.25 \, \text{p.u.} \), the values of \( L_s \) and \( C_s \) are obtained as 0.48 mH and 12.6 \( \mu \text{F} \), respectively.

5. **STEADY-STATE PERFORMANCE**

In the following analysis, all results are given for \( V = 40 \, \text{V} \) and \( f = 20 \, \text{kHz} \).

Since the performance of the given topology is very much dependent on the relation between \( \alpha (\text{ALFA}) \) and \( \Phi (\text{FI}) \), in the first attempt the variation of \( \Phi \) versus \( \alpha \) has been computed from equation (20) for three values of \( Z_l \) (-0.25 p.u, -0.5 p.u, and -0.75 p.u) with \( X_L = 5 \, \text{p.u.} \) The results are shown in Fig. 5. Observations of Fig. 5 show that the upper limit of \( \alpha \) is limited by the value of \( \Phi \), because when \( \alpha = \Phi \) the rectifier input voltage \( V_{ab} \) becomes a square wave rather than a quasi-square wave. Beyond that, equation (20) converges to value of \( \Phi \) less than the value of \( \alpha \), which, which although mathematically correct, but means nothing from physical point of view. Because for \( \alpha > \Phi \) the current zero-crossing is already passed, the conducting semiconductor switch will continue conducting until the next zero-crossing of the current and the control over the second half cycle will be lost. Also since \( Z_l \) has a small value of impedance, during the second half cycle current may be large enough to damage some of the circuit elements in the control part of the circuit. Observations of the curves shown in Fig. 5 also show that the upper and lower limits of \( \alpha \) is controlled by the value of \( Z_l \). The smaller the absolute values of \( Z_l \) the wider is the controlling range (difference between upper and lower limit of \( \alpha \)). However when the absolute value of \( Z_l \) falls below a certain value, for values of \( \alpha \) between 90° to 110°, equation (20) does not converge to valid values for \( \Phi \). In this investigation it was found that for overall near optimal performance, \( Z_l \) should be around -0.25 p.u., as it was stated earlier.

Beside the incorrect theoretical analysis, all the above vital concepts were missed in [10].

In [10], \( \Phi \) is computed using the following equation;

\[
\cos(\Phi) + \frac{1}{2} R_o (\sin(\alpha) - \sin(\Phi)) \sum_{n=2}^{\infty} \left( 1 - \cos(n(\alpha - \Phi)) \right) = 0
\]  

(29)

And the output voltage is computed using;

\[
V_o = -\frac{2\sqrt{2}}{\pi Z_L} R_o \sin\left(\frac{\alpha - \Phi}{2}\right) \cos\left(\frac{\alpha + \Phi}{2}\right)
\]  

(30)
Using equations (29) and (30), $\Phi$ and $V_o$ are computed and plotted in Figure 6.a and 6.b respectively. Observation of Figure 6.a shows that according to the analysis given in [10], $\Phi$ is always smaller than $\alpha$, which is not possible for the reasons explained in section 2. Another proof of this is that the output voltage values shown in Figure 6.b, is always negative. This trend is also obvious from equation (30), since; $Z_1$ is a negative number, ($\alpha-\Phi$) is $>0$ (when $\Phi<\alpha$), and ($\alpha+\Phi$)$>\pi$, then $\sin((\alpha-\Phi)/2)$ always $>0$, and $\cos((\alpha+\Phi)/2)$ always$<0$, and finally the output voltage obtained from equation (30) always will be a negative number. On the other hand it is evident from equation (30) and Figure 6.b, that according to the analysis given in [10], the output voltage $=0$ when $\alpha=\Phi$, which is not realistic. All these results clearly show that analysis given in [10] is not correct.

![Fig. 6. Variation of $\Phi$ (FI) and $V_o$ with $\alpha$ (ALFA) computed using analysis given in [10]](image)

The effect of $Z_1$ (for $X_L=5$ p.u.) on the power factor and $THD$ is illustrated in Fig. 7. Observations of this figure reveal that the power factor deteriorates very significantly as the absolute value of $Z_1$ gets larger. Also within the valid values of $\alpha$, the $THD$ increases (deteriorates) as the amplitude of $Z_1$ increases. Therefore fixing the value of $Z_1$ as low as at -0.25 p.u. provides the best advantages.
Fig. 8. Harmonic currents and THD versus $\alpha$ (ALFA) for three different values of $X_L$, for $R_o=12$ $\Omega$.

Fig. 9. Harmonic currents and THD versus $\alpha$ (ALFA) for three different values of $X_L$, for $R_o=24$ $\Omega$. 
For a fixed value of $Z_1 (Z_1 = -0.25 \text{ p.u.})$, the effect of $X_L$ on the harmonic currents and THD is shown in Fig. 8 for three different values of $X_L$ (5 p.u., 10 p.u. and 15 p.u. respectively). Examination of this figure reveals that as the value of $X_L$ increases, both the amplitude of harmonic currents and THD decreases. This might appear as an advantage, but larger size means more expensive inductor, which is the most expensive circuit component of the given topology. Therefore since with an inductor of 0.48 mH a THD of 2\% and a power very close to unity, and, as will be shown in the following section, a very low inrush current is achieved, and there is no need to increase the value of $X_L$ more than 5 p.u.

As long as the relations given in equations (25), (26) and (28) are observed, the p.u. or \% harmonic current content and the value of THD are independent of $R_o$. To prove this argument, the harmonic current content and THD calculations are repeated for $R_o=24 \ \Omega$, i.e., for 100\% increase in $R_o$ and results are demonstrated in Figure 9. Comparison of figures 8 and 9 clearly show that the two figures are identical and the p.u. harmonic current content and THD are independent of the value of load resistor.

### 6. TRANSIENT ANALYSIS

Observations of Fig. 2 show that during one complete cycle there are five different operating modes for the given topology. Therefore to analyze the transient performance, the differential equations governing each mode must be solved. In transient analysis, the power dissipated in various circuit elements is included in calculations and these are $R_f=55 \ \text{m} \Omega$, $R_D=6 \ \text{m} \Omega$ and $R_s=145 \ \text{m} \Omega$. Value of $R_s$ is obtained from theoretical design of a 10 A and 0.48 mH inductor and calculating its resistance. On-state resistance of semiconductor switches ($R_f$) (assuming them to be IGBTs) and diodes used in the rectifier are obtained from some previous measurements performed by the authors.

*Model 1:* $0 < \omega t \leq \alpha$ (Non of the semiconductor switches is conducting (Fig. 2))

The differential equations governing this mode are:

1. $v_{\text{in}} = R_s i_{\text{in}} + L_s \frac{d^2 i_{\text{in}}}{dt^2}$
2. $\frac{dQ_1}{dt} = i_{\text{in}}$
3. $\frac{\text{abs}(i_{\text{in}})}{R_s} = \frac{Q_1}{R_s C_1}$

where $Q_1$ is the electrical charge across $C_s$, while $Q_2$ is the charge across $C_o$. ‘abs’ is used for absolute value.

*Model 2:* $\alpha < \omega t \leq \Phi$ (Semiconductor switch $Q_A$ is conducting (Fig. 2))

1. $v_{\text{in}} = (R_s + R_T) i_{\text{in}} + L_s \frac{d^2 i_{\text{in}}}{dt^2} + \frac{Q_2}{C_2}$
2. $\frac{dQ_2}{dt} = i_{\text{in}}$
3. $\frac{\text{abs}(i_{\text{in}})}{R_s} = 0$

*Model 3:* $\Phi < \omega t \leq \pi + \alpha$ (None of the semiconductor switches is conducting (Fig. 2))
Mode 4: $\pi + \alpha < \omega t \leq \pi + \Phi$ (Semiconductor switch $Q_B$ is conducting (Fig. 2))
Differential equations given for Mode 2 are valid for this mode as well.

Mode 5: $\Phi + \pi < \omega t \leq 2\pi$ (none of the semiconductor switches is conducting (Fig. 2))
Differential equations given for Mode 1 are valid for this mode as well.

The above given differential equations are solved using fourth-order Runge-Kutta method [12]. In order to meet, as close as possible, the sharp changing boundaries of $V_{ab}$, a very small computation step size of 0.125 $\mu$s (400 solutions per cycle) was selected.

7. TRANSIENT PERFORMANCE

By solving the differential equations given in the previous section, the transient performance of the given topology is computed for $\alpha=135^\circ$, $R_o=12$ $\Omega$, $Z_L=-0.25$ p.u., $X_L=5$ p.u. ($L_s=0.48$ mH), $f=20$ kHz, $C_s=12.6$ $\mu$F, $C_o=10$ $\mu$F, and $V=40$ V.

Fig. 10 shows the transient input current and transient output voltage for $\alpha=135^\circ$. Examinations of this figure show that the inrush current is very small during the starting of the circuit, which is one of the expected behavior from the topology under investigation. The input current and output voltage settles to their steady-state value in about 3.8 ms. This figure also shows that the output voltage builds very smoothly to its final value with experiencing a small amount of overshoot (around 10%). The average of the final value of the output voltage is around 53 V. On the other hand, Fig. 5.b. shows that for $\alpha=135^\circ$ the output voltage is around 52 V. Therefore there is a good correlation between the steady-state and transient analysis presented in this paper. These results are a clear verification of the correctness of the analysis presented in this work.
Fig. 11. Demonstration of phases of the input current and input voltage (obtained from the last two cycles of the transient solution, when the waves are settled to their steady-state regime.)

One of the main objectives of the given topology is to provide near unity \( PF \). Fig. 6.a confirms that the given topology realizes this objective. To further confirm this claim, the instantaneous input voltage and input current, which were obtained from transient analysis, are compared in Fig. 11 for observing the time phase difference between them. From this figure it is evident that the both signals are almost in phase, which confirms near unity \( PF \) and the input current is almost pure sinusoidal (THD of only 2%).

8. CALCULATION OF CONVERSION EFFICIENCY

The average input power is:

\[
P_{\text{inav}} = \frac{1}{T} \int_{0}^{T} v_{\text{in}} i_{\text{in}} \, dt
\]

(40)

where \( T \) is the supply voltage period (\( T=1/f \)). Similarly the average output power is given by (rectified signal has twice frequency as the supply frequency):

\[
P_{\text{out}} = \frac{2}{T} \int_{0}^{T} \left( \frac{1}{R_0}\right) \left( \frac{Q_2}{C_0} \right)^2 \, dt
\]

(41)

As stated earlier, \( Q_2 \) is the electrical charge across \( C_0 \). The conversion efficiency is given by:

\[
\eta = \frac{P_{\text{out}}}{P_{\text{inav}}}
\]

(42)

As stated earlier, the transient performance is computed using a very small computation step size of 0.125 \( \mu \)s, i.e., 400 solutions per cycle. Therefore the integrations given in (40) and (41) can be obtained numerically on cycle by cycle bases and hence the average input and output powers and conversion efficiency can be calculated. Fig. 12 illustrates the results obtained by this means. Observation of Fig. 12.a shows that the difference between the average input and output powers is few watts. From close observation of Fig. 12.b it can be seen that the final (steady-state) value of power conversion efficiency is 95.96% \( \pm 96 \% \).
9. CONCLUSION

The specific conclusions of this paper are summarized as follows:

1. Improved modeling and analysis are given for both the steady-state and transient performance of the high-performance the ac/dc converter given in [10] to eliminate the incorrect analysis given therein.

2. A more reliable design guideline is given for the design of the resonant network component values.

3. Both the steady-state and transient performance are presented and results are discussed in details. It is shown that the input current is almost sinusoidal with a total harmonic distortion of less than 2%, and a power factor of close to 99%, conversion efficiency of around 96% and very low inrush current of around 110% is achieved.

4. Wherever possible, the results obtained from transient analysis are compared with their counterparts obtained from steady-state analysis and excellent agreement is obtained between them. Those comparisons confirm the correctness of the modeling and analysis presented in this paper.

5. The output voltage can be regulated by controlling the turn-on time of the semiconductor switches QA and QB (or control angle \( \alpha \))

REFERENCES


APPENDIX

In [10] Fourier expression of the rectifier bridge voltage $V_{ab}$ is given as:

$$V_{ab} = \frac{4V_o}{\pi} \sum_{n=1,3,5,...}^{\infty} \frac{1}{n} \sin \left( \frac{n(\alpha - \Phi)}{2} \right) \cos \left( n\omega t - n \left( \alpha + \Phi \right) \right)$$

(A1)