Efficiency and Transfer function calculation of the Buck-Boost converter with ideal flow control

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Abstract

The voltage regulator is designed based on a transfer function of the system pre-converter with ideal current control. This transfer function is obtained by first obtaining the system equation for the pre-converter with ideal current control. Subsequently, we look for a possible regime solution for a sinusoidal power supply and then arrive at the transfer function via linearization of the system equation around the solution in regime. The calculations are verified by using Matlab/Simulink on the system with considering two buck boost converters connected in parallel. The control scheme used in simulation is feed forward control scheme.

Introduction

The main goal of all types of the advancement in designing and control of the power electronic converters is to obtain the maximum efficiency. Different types of control techniques are applied for reducing all the types of power losses in a converter circuit. For a DC-DC converter efficiency, the designer should know all the different types of requirements which are responsible for losses minimization in a circuit. For this, a model design is a requirement for the estimation of DC-DC converter efficiency. This design should not only provide the efficiency of the converter for different input values but also it should help designers to improve the control techniques which are helpful in improving the efficiency of power electronic converters. The design of such model should contain a deep study of all different type of the mechanisms that causes power losses inside a system.
In a DC-DC converter, there are different types of power dissipations which are because of the inductor conduction, filter capacitor losses, diode conduction and power dissipated in MOSFET. For calculating the power losses in MOSFET by using the parameters from data sheet is presented in [1] which briefly explains the switching losses as well as conduction losses of a MOSFET. For a fly back converter, the losses due to the output diode is presented in [2] in which the relationship of the losses with respect to input and output voltage is provided. The power losses due to the inductor and the identification of the scenarios which causes the inductor power losses is proposed in [3]. Reference [4] expresses a very basic method for calculating all the different types of losses in a circuit which are effecting the efficiency of the converter.
From the last decade, researchers are working continuously on power factor correction PFC converters [5]–[10]. Amongst these references, [5]–[7] is about single phase analog controller for boost converter. Later on with the advancement of fast digital signal processors DSP along with analog to digital converters and PWM generator, researchers focused on moving from analog to digital control schemes [8]–[10]. The power factor correction for pre regulators and converters are performed by using advanced DSPs with more feasible control options in these references.

Enough literature review has been published for efficiency calculation of DC-DC converters in multiple applications. While for non-idealities to be the part of system, efficiency calculation is performed in [11]. The present work is also focused on providing a general static model of a buck-boost converter and calculating its efficiency by taking into account all the types of non-idealities of the system and performing the power factor correction of the system as well. For initial calculations a single buck boost converter is taken in to account to derive the equations and then these calculations are verified on an already existing model [12]. MATLAB is the software used for providing the results which verifies our derivations.

In this paper, section II is about the single buck boost converter with ideal control method to derive the power balance equation taking all the types of losses in the circuit in to account. Taking the regime conditions in to account the output voltage calculations are performed in section III of this paper for sinusoidal input supply voltage. Considering the small regime deviations, the transfer function for the change in output voltage with respect to the input conductance is done in section IV. Section V is about the elaboration of the previous transfer function for resistive load of the system. The previously conducted case study with simulations and results are presented in section VI of this paper. At the end the work is concluded in section VII.

Buck-Boost converter system comparison

A buck-boost converter system, shown in fig 1, contains two types of elements. The elements which can stores energy and the elements which can't store energy. In fig 1, the inductor L and the output capacitor C_{out} can be the elements that can store energies while the bridge rectifier, switch *S* and the diode *D* with ideal current control don't have this property. Let represent the switch *S*, diode *D* and the ideal current control scheme as a "black box". This "black box" contains the elements which can't store energy so we can represents the power distribution in fig. 1 as;

$$p_{in}(t) = p_1(t) + p_{loss}(t) \tag{1}$$

while

$$p_1(t) = p_{in}(t) - p_{loss}(t) \tag{2}$$

For the value of $p_{loss}(t) = (1-\eta)p_{in}(t)$

$$\eta = \frac{p_1(t)}{p_{in}(t)} \tag{3}$$

From fig. 1, we can express the instantaneous power $p_1(t)$ as a function of input current $i_{in}(t)$ and input voltage $v_{in}(t)$ which is across the input stage of "black box":

$$p_{in}(t) = v_{in}(t)\, i_{in}(t) \tag{4}$$

For an ideal current control scheme, including the input conductance $g_e(t)$ in above equation, so input power can be written as;

$$p_{in}(t) = v_{in}^2\, g_e(t) \tag{5}$$

From the figure 1, it is obvious that,

$$p_1(t) = p_L(t) + p_{out}(t)$$

$$p_1(t) = L\frac{di_1(t)}{dt} i_1(t) + v_{out}(t)\, i_{out}(t) \tag{6}$$

We can represent $i_{out}(t)$ as;

$$i_{out}(t) = i_{out}(t) + i_c(t) \tag{7}$$

$$i_{out}(t) = \frac{p_{out}(t)}{v_{out}(t)} + C_{out}\frac{dv_{out}(t)}{dt} \tag{8}$$

Equation (6) in terms of equation (8) can be represented as;

$$p_1(t) = L\frac{d\big(g_e(t)\, v_1(t)\big)}{dt}\big(g_e(t)\, v_1(t)\big) + v_{out}(t)\left(\frac{p_{out}(t)}{v_{out}(t)} + C_{out}\frac{dv_{out}(t)}{dt}\right) \tag{9}$$

By simplifications we get

$$p_1(t) = \frac{1}{2}L\frac{d}{dt}\big(g_e(t)\, v_1(t)\big)^2 + p_{out}(t) + \frac{1}{2}C_{out}\frac{d}{dt}{v_{out}}^2(t) \tag{10}$$

Solving equation (3) for equation (5) and (9) to obtain power balancing equation for buck boost converter i.e.

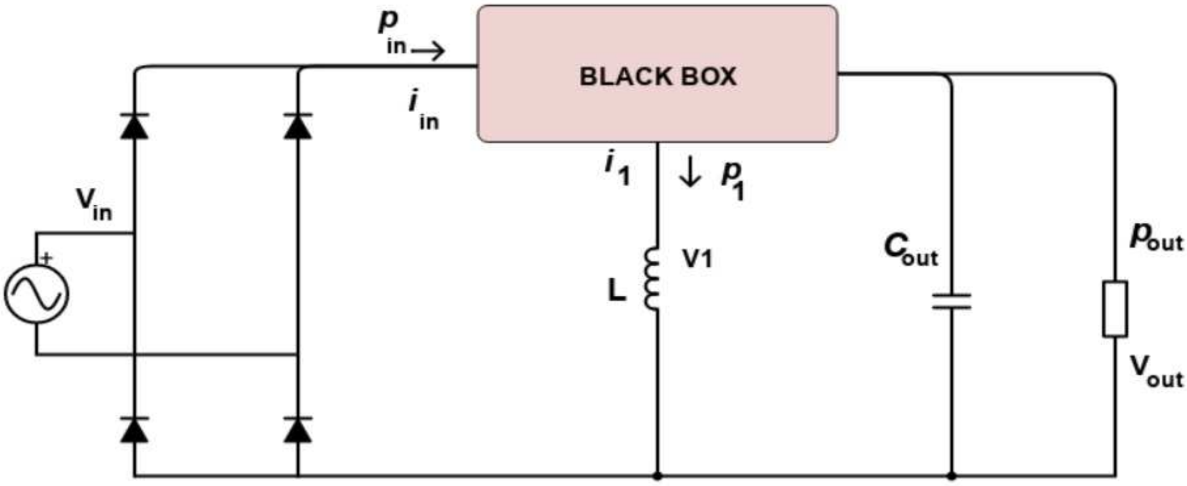

Fig. 1: Buck-boost converter with constant power load.

$$p_{out}(t) = \eta\, v_{in}{}^2(t)\, g_e(t) - \frac{1}{2} L \frac{d}{dt}\big(g_e(t)\, v_1(t)\big)^2 - \frac{1}{2} C_{out} \frac{d}{dt} v_{out}{}^2(t) \qquad (11)$$

Equation (11) is the power balance equation for buck boost converter. From this equation we can see that the power provided by the bridge rectifier is reduced because of the power absorbed by the coil, power losses in the black box and power consumed by the output capacitor.

Regime solution for sinusoidal supply voltage

For regime input supply voltage, we always represents the quantities in regime with the index *r*. We assume that power absorbed by load $p_{out}(t)$ and equivalent input inductance $g_e(t)$ have the fixed regime values represented by $P_{out,r}$ and $G_{e,r}$ respectively. So the output power equation (11) can be written as

$$P_{out,r} = \eta v_{in,r}{}^2(t) G_{e,r} - \frac{1}{2} L\, G_{e,r} \frac{d}{dt}\left(v_{1,r}(t)\right)^2 - \frac{1}{2} C_{out} \frac{d}{dt} v_{out,r}{}^2(t) \qquad (12)$$

For pure sinusoidal supply voltage, we can control the rectified input voltage and also the inductor voltage as;

$$v_{in,r}(t) = |\hat{V}_{in,r} \sin(w_n t)| \qquad (13)$$

The first term on the R.H.S of equation (12) can be written in terms of equation (13) as,

$$\eta v_{in,r}{}^2(t) G_{e,r} = \eta G_{e,r} \hat{V}_{in,r}{}^2 \sin^2(w_n t)$$

$$\eta v_{in,r}{}^2(t) G_{e,r} = \eta G_{e,r} \hat{V}_{in,r}{}^2 \left(\frac{1 - \cos(2w_n t)}{2}\right)$$

$$\eta v_{in,r}{}^2(t) G_{e,r} = \frac{\eta G_{e,r} \hat{V}_{in,r}{}^2}{2} - \frac{\eta G_{e,r} \hat{V}_{in,r}{}^2 \cos(2w_n t)}{2} \qquad (14)$$

In the same way we perform calculations for second term on the R.H.S of equation (12) in terms of equation (13) as,

$$\frac{1}{2} L G_{e,r} \frac{d}{dt}\left(v_{1,r}(t)\right)^2 = \frac{1}{2} L G_{e,r} \frac{d}{dt}\left(\hat{V}_{1,r} \sin(w_n t)\right)^2$$

$$\frac{1}{2} L\, G_{e,r} \frac{d}{dt}\left(v_{1,r}(t)\right)^2 = \frac{1}{2} L G_{e,r} 2\left(\hat{V}_{1,r} \sin(w_n t)\right) \frac{d}{dt}\left(\hat{V}_{1,r} \sin(w_n t)\right)$$

$$\frac{1}{2} L G_{e,r} \frac{d}{dt}\left(v_{1,r}(t)\right)^2 = \frac{L G_{e,r}{}^2 \hat{V}_{1,r}{}^2 w_n \cos(2w_n t)}{2} \qquad (15)$$

While the 3rd term of R.H.S of equation (12) didn't contains any term with input voltage and inductor voltage so we leave it as it is and replacing the 1st and 2nd term of equation (12) by equation (14) and (15) respectively and we get

$$P_{out,r} = \frac{\eta G_{e,r}\hat{V}_{in,r}^{\ 2}}{2} - \frac{\eta G_{e,r}\hat{V}_{in,r}^{\ 2}\cos(2w_n t)}{2} - \frac{LG_{e,r}^{\ 2}\hat{V}_{1,r}^{\ 2} w_n \cos(2w_n t)}{2} - \frac{1}{2}C_{out}\frac{d}{dt}v_{out,r}^{\ 2}(t) \quad (16)$$

Solving the above equation for $V_{out,r}(t)$,

$$\frac{d}{dt}v_{out,r}^{\ 2}(t) = \frac{2}{C_{out}}\left(\frac{\eta G_{e,r}\hat{V}_{in,r}^{\ 2}}{2} - \frac{\eta G_{e,r}\hat{V}_{in,r}^{\ 2}\cos(2w_n t)}{2} - \frac{LG_{e,r}^{\ 2}\hat{V}_{1,r}^{\ 2} w_n \cos(2w_n t)}{2} - P_{out,r}\right)$$

Taking integration on both sides we get;

$$v_{out,r}^{\ 2}(t) = \left(\frac{\eta G_{e,r}\hat{V}_{in,r}^{\ 2} - 2P_{out,r}}{C_{out}}\right)t - \frac{\eta G_{e,r}\hat{V}_{in,r}^{\ 2}\sin(2w_n t)}{2w_n C_{out}} - \frac{LG_{e,r}^{\ 2}\hat{V}_{1,r}^{\ 2}\cos(2w_n t)}{2C_{out}} + b_1^{\ 2} \quad (17)$$

While b_1^2 is an integration constant. The average power supplied by the bridge rectifier stage after losses in "black box" is equal to the constant power absorbed by the load i.e.

$$\frac{\eta G_{e,r}\hat{V}_{in,r}^{\ 2}}{2} = P_{out,r} \quad (18)$$

This term is referred as linearly increasing term so we can neglect it from the equation (17), while the constant b_1^2 is representing the average of output voltage i.e.

$$b_1^{\ 2} = V_{out,r}^{\ 2} \quad (19)$$

so the other terms on R.H.S of the equation (17) pulsating with twice of the line frequency, taking equation (18) and (19) into account, can be written as;

$$v_{out,r}^{\ 2}(t) = V_{out,r}^{\ 2} - \frac{P_{out,r}\sin(2w_n t)}{w_n C_{out}} - \frac{LG_{e,r}^{\ 2}\hat{V}_{1,r}^{\ 2}\cos(2w_n t)}{2C_{out}} \quad (20)$$

In regime, there is an output voltage ripple which is operating at the frequency which is double of the line frequency i.e.

$$v_{out,r}^{\ 2}(t) = V_{out,r}^{\ 2} + V_{out,r}^{\ 2} 2f(2w_n t) \quad (21)$$

So by putting this equation in above equation (20)

$$V_{out,r}^{\ 2} + V_{out,r}^{\ 2} 2f(2w_n t) = V_{out,r}^{\ 2} - \frac{P_{out,r}\sin(2w_n t)}{w_n C_{out}} - \frac{LG_{e,r}^{\ 2}\hat{V}_{1,r}^{\ 2}\cos(2w_n t)}{2C_{out}}$$

$$f(2w_n t) = -\frac{P_{out,r}\sin(2w_n.t)}{2V_{out,r}^{\ 2} w_n C_{out}} - \frac{LG_{e,r}^{\ 2}\hat{V}_{1,r}^{\ 2}\cos(2w_n t)}{4V_{out,r}^{\ 2} C_{out}} \quad (22)$$

For the purpose of calculations, for ideal current control scheme we can take this inductor voltage $V_{1,r}$ as equal to the rectified input voltage $V_{in,r}$ so keeping the equation (18) into account we can write the above equation as,

$$f(2w_n t) = -\frac{P_{out,r}\sin(2w_n t)}{2V_{out,r}^{\ 2} w_n C_{out}} - \frac{LG_{e,r}P_{out,r}\cos(2w_n t)}{2\eta C_{out}} \quad (23)$$

Any function whose value is less than 0.05 for all values of *t* with deviations which are less than 1%, so this implies the following concept i.e.

$$\sqrt{1 + 2g(t)} = 1 + g(t) \quad (24)$$

The efficiency of such converters are always more than 90%. The input inductance value of 0.0152 with adjustable output power values and by putting the other circuit values we can find the maximum amplitude for equation (22),

$$\frac{P_{out,r}}{2V_{out,r}^{\ 2} w_n C_{out}} = 33.* 10^{-3} V^2 \quad (25a)$$

And

$$\frac{LG_{e,r}P_{out,r}\cos(2w_n t)}{2\eta C_{out}} = 173.8 * 10^{-6} V^2 \quad (25b)$$

With these amplitude values we can apply equation (24) on (23) with approximation error to be less than 1% i.e.

$$V_{out,r}(t) = V_{out,r}(1 + f(2w_n.t)) \quad (26)$$

Taking small regime deviations into account

We derives the power balance equations with ideal power regulations and then regime equations for buck boost converter with sinusoidal input supply voltage. Now in this part we will discuss the system behavior in case of small deviations from the regime. So we consider all these small variations in the time variables i.e.

In input voltage;

$$\Delta v_{in}(t) \approx v_{in}(t) - v_{in,r}(t) \tag{27}$$

in output voltage;

$$\Delta v_{out}(t) \approx v_{out}(t) - v_{out,r}(t) \tag{28}$$

in inductor voltage;

$$\Delta v_1(t) \approx v_1(t) - v_{1,r}(t) \tag{29}$$

in equivalent input inductance;

$$\Delta g_e(t) \approx g_e(t) - G_{e,r} \tag{30}$$

in output power;

$$\Delta p_{out}(t) \approx p_{out}(t) - P_{out,r} \tag{31}$$

For observing the effect of these small deviations in regime on the system, we consider all these equations from 27-31 in equation (11),

$$\Delta p_{out}(t) + P_{out,r} = \eta\,(\Delta v_{in}(t) + v_{in,r}(t))^2\left(\Delta g_e(t) + G_{e,r}\right) - \frac{1}{2}L\,\frac{d}{dt}\Big((\Delta g_e(t) + G_{e,r})\,(\Delta v_1(t) + v_{1,r}(t))\Big)^2 - \frac{1}{2}.C_{out}.\frac{d}{dt}\left(\Delta v_{out}(t) + v_{out,r}(t)\right)^2 \tag{32}$$

By rearranging the above equation by neglecting the second and third order derivatives in it we can write the above equation as;

$$\Delta p_{out}(t) = 2\,\eta G_{e,r} v_{in,r}(t)\Delta v_{in}(t) + \eta v_{in,r}{}^2(t)\Delta g_e(t) - G_{e,r}{}^2 L\,\frac{d}{dt}\Big(v_{1,r}(t)\Delta v_1(t)\Big) - G_{e,r}L\frac{d}{dt}\Big(v_{1,r}{}^2(t)\Delta g_e(t)\Big) - C_{out}\,\frac{d}{dt}(v_{out,r}(t)\Delta v_{out}(t)) \tag{33}$$

Putting the values of input voltage, output voltage and inductor voltage in above equation, we get;

$$\Delta p_{out}(t) = 2\,\eta G_{e,r}(|\hat{V}_{in,r}\sin(w_n t)|)\Delta v_{in}(t) + \eta(|\hat{V}_{in,r}\,sin(w_n t)|)^2\Delta g_e(t) - G_{e,r}{}^2 L\,\frac{d}{dt}\Big((|\hat{V}_{in,r}\,sin(w_n t)|)\Delta v_1(t)\Big) - G_{e,r}L\frac{d}{dt}\Big((|\hat{V}_{1,r}\,sin(w_n t)|)^2\Delta g_e(t)\Big) - C_{out}\,\frac{d}{dt}(V_{out,r}(1 + f(2w_n.t))\Delta v_{out}(t)) \tag{34}$$

As we know that periodic functions occurs with a period which is half of the grid period T_n i.e.

$$|\,sin\left(\frac{2\pi}{T_n}t\right), \quad sin^2\left(\frac{2\pi}{T_n}t\right)|\,, \quad \frac{df\left(\frac{4\pi}{T_n}t\right)}{dt}, \quad f\left(\frac{4\pi}{T_n}t\right)$$

Taking the assumption into account that we only consider the small deviations because they vary slowly than the grid frequency so we can simplify the equation (34) on this assumption. By averaging this equation over several network periods we can get a simple equation from it. Averaging means that integrating the equation (34) over a network periods kT_n with $k \in IN$. The product of some integrand from sinusoidal function with period T_n and the time lapse is slower than other sinusoidal function so we can put it outside the integration and taking into account the average from $|sin(w_n t)|$ which is equal to 2π;

$$\frac{1}{nT_n}\int_0^{nT_n}\Delta p_{out}(t)dt = 2\,\eta G_{e,r}\frac{1}{nT_n}\int_0^{nT_n}(\frac{2}{\pi}\hat{V}_{in,r}\Delta v_{in}(t))dt + \eta\frac{1}{nT_n}\int_0^{nT_n}(\frac{1}{2}\hat{V}_{in,r}{}^2\Delta g_e(t))dt - G_{e,r}{}^2 L\,\frac{1}{nT_n}\int_0^{nT_n}\left(\frac{2}{\pi}\hat{V}_{1,r}\frac{d}{dt}\Delta v_1(t)\right)dt - G_{e,r}L\frac{1}{nT_n}\int_0^{nT_n}\left(\frac{1}{2}\hat{V}_{1,r}{}^2\frac{d}{dt}\Delta g_e(t)\right)dt - C_{out}\frac{1}{nT_n}\int_0^{nT_n}(V_{out,r}\frac{d}{dt}\Delta v_{out}(t))dt \tag{35}$$

Due to the constantly assumed time lapse of decisions and their derivatives, we can write the above equation as;

$$\Delta p_{out}(t) = 2\,\eta G_{e,r}\frac{2}{\pi}\hat{V}_{in,r}\Delta v_{in}(t) + \eta\frac{1}{2}\hat{V}_{in,r}{}^{2}\Delta g_e(t) - G_{e,r}{}^{2}L\,\frac{2}{\pi}\hat{V}_{1,r}\frac{d}{dt}\Delta v_1(t) - G_{e,r}L\frac{1}{2}\hat{V}_{1,r}{}^{2}\frac{d}{dt}\Delta g_e(t) - C_{out}V_{out,r}\frac{d}{dt}\Delta v_{out(t)} \quad (36)$$

Transferring this to Laplace transform and keeping the variations in input voltage and output power very slow and taking them equal to zero. Also assuming the inductor voltage to be equal to the rectified input voltage, we can write the following transfer function;

$$\frac{\Delta v_{out}(s)}{\Delta g_e(s)} = \frac{\eta(1-\frac{sLG_{e,r}}{\eta})\hat{V}_{in,r}{}^{2}}{2s\,C_{out}\,V_{out,r}} \quad (37)$$

This above transfer function has a zero point on real axis i.e.

$$s = \frac{\eta}{LG_{e,r}} \quad (38)$$

Solving the above equation by putting the values gives us the value for *s = 61 * 103*. With this value of *s* we can say that influence of this zero point of behavior of the ideal current control buck boost converter is negligible in comparison with pole in the origin. So we can write the above transfer function in simplest form with minimum abnormalities for sinusoidal input voltage supply;

$$\frac{\Delta v_{out}(s)}{\Delta g_e(s)} = \frac{\eta\hat{V}_{in,r}{}^{2}}{2s\,C_{out}\,V_{out,r}} \quad (39)$$

While the disturbance dynamic on output voltage is given by

$$\Delta v_{out}(s) = -\frac{1}{s\,C_{out}\,V_{out,r}}\Delta p_{out}(s) + \frac{4\eta G_{e,r}\hat{V}_{1,r}}{s\pi\,C_{out}\,V_{out,r}}\left(1 - \frac{sLG_{e,r}}{2\eta}\right)\Delta v_1(s) \quad (40)$$

This above equation represents the output voltage of the system with ideal current control.

Transfer Function for resistive load

Let's consider resistive load at the output of the converter in which the output power is dissipated. This dissipated output power as described in [13] is represented by;

$$p_{out}(t) = \frac{v_{out}{}^{2}(t)}{r_{out}(t)} \quad (41)$$

For a fixed value of the resister $R_{out,r}$ we can write the relationship between the changes in output voltage and outgoing power in regime as;

$$\Delta p_{out}(s) = 2\frac{\Delta v_{out}(s)}{R_{out,r}}\,V_{out,r} \quad (42)$$

Now writing equation (39) together with (40);

$$\Delta v_{out}(s) = -\frac{1}{s\,C_{out}\,V_{out,r}}\Delta p_{out}(s) + \frac{\eta\hat{V}_{in,r}{}^{2}}{2s\,C_{out}\,V_{out,r}}\Delta g_e(s) \quad (43)$$

Putting equation (42) in (43) and by solving it gives the following transfer function,

$$\frac{\Delta v_{out}(s)}{\Delta g_e(s)} = \frac{\eta\hat{V}_{in,r}{}^{2}R_{out,r}}{2\,V_{out,r}(2+sR_{out,r}\,C_{out})} \quad (44)$$

As we know that transfer function of any resistive load for first order system is given by

$$\frac{\Delta v_{out}(s)}{\Delta g_e(s)} = \frac{K}{1+\mathrm{T}s} \quad (45)$$

Where *K* and *T* represents amplification factor and time constant respectively. By comparing equation (45) with (44) we get;

$$K = \frac{\eta\hat{V}_{in,r}{}^{2}R_{out,r}}{4\,V_{out,r}} \quad (46)$$

$$\mathrm{T} = \frac{C_{out}\,R_{out,r}}{2} \quad (47)$$

For large values of output resister the system behaves like an integrator. So equation (44) becomes

$$\frac{\Delta v_{out}(s)}{\Delta g_e(s)} = \frac{K}{\mathrm{T}s} = \frac{\eta\hat{V}_{in,r}{}^{2}}{2s\,V_{out,r}\,C_{out})} \quad (48)$$

Case study with simulations and results

The scheme under consideration for the following calculation is presented in [12] in which two buck boost converters are connected in parallel. Both are connected by the same input supply source and current and voltage are controlled by using feed forward control method [12]. The power factor correction is performed with the help of feed forward control scheme with less input current distortion. The output voltage is also controlled to a certain reference value. The input current is purely sinusoidal and in phase with the input supply voltage The simulations are performed for various input voltages and the efficiency calculations are made accordingly.

Fig. 2 show the model under consideration with feed-forward control scheme. Two parallel buck-boost converters are used with inductors values, *L1=5mH* and *L2=0.5mH*. Input voltage of *150V RMS* with line frequency of *50 Hz* is provided to the system with switching frequency of *10kHz*. Reference output voltage is maintained at *250 volts*.

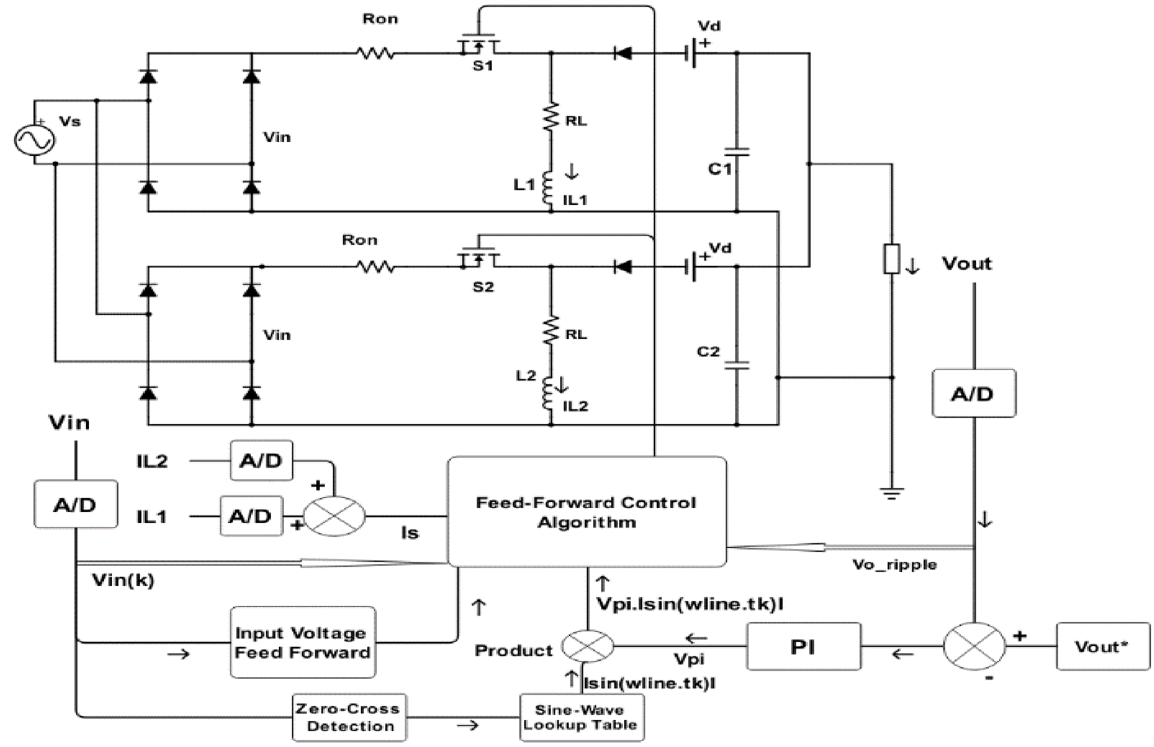

Fig. 2. Two Parallel BBC PFCs with Digital Feed-Forward Algorithm.

Electronic equipment that contains DC-DC converters installed inside the product have comparatively high efficiencies as compared to AC-DC converters. They can achieve up to 95% of efficiency easily. While the efficiency of AC-DC converters is between 80-90% and this can be improved by widespread use of switching methods. The efficiency of above scheme goes almost up to 98% under different calculations. The simulation results can be shown from fig. 3 to 9.

Fig. 3 shows the comparison graph of the efficiency of the above model with respect to the output power and it can be visible the efficiency of the converter is maximum for high output power. The output voltage of the converter can be also seen in fig. 4 which shows that reference voltage is achieved.

The efficiency of the system is also dependent on the output current from the system which is provided to the load. Efficiency has a direct proportionality to the load current. The input current profile is shown in the fig. 5 with respect to the input voltage and can be clearly seen that both are in line and phase to each other. There is no displacement factor between voltage and current and this can lead to the unity power factor of the above considered system.

As it can be seen from fig. 5 that input voltage and current are purely sinusoidal and they are also in phase with each other so it will provide the least THD of the system as shown in the fig. 6. The THD of the system is about 8.66% which will alternately provide the power factor up to 0.99. Changing the input supply voltage value also effect the efficiency as well as the power factor of the system. The relation of the efficiency with input supply voltage is shown in the fig. 7. The power factor of the system also have different graphs for two different input supply voltages which is shown in the fig. 8. The duty cycle values that is provided by the controller at its output towards the switches of the buck boost converter is shown in the fig. 9.

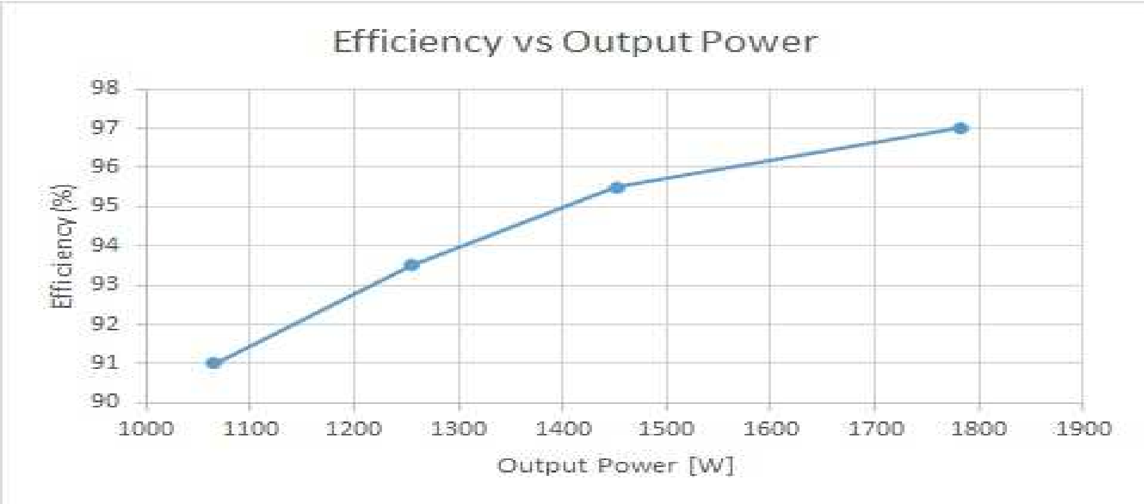

Fig. 3. Efficiency of the system w.r.t increasing output power in [W].

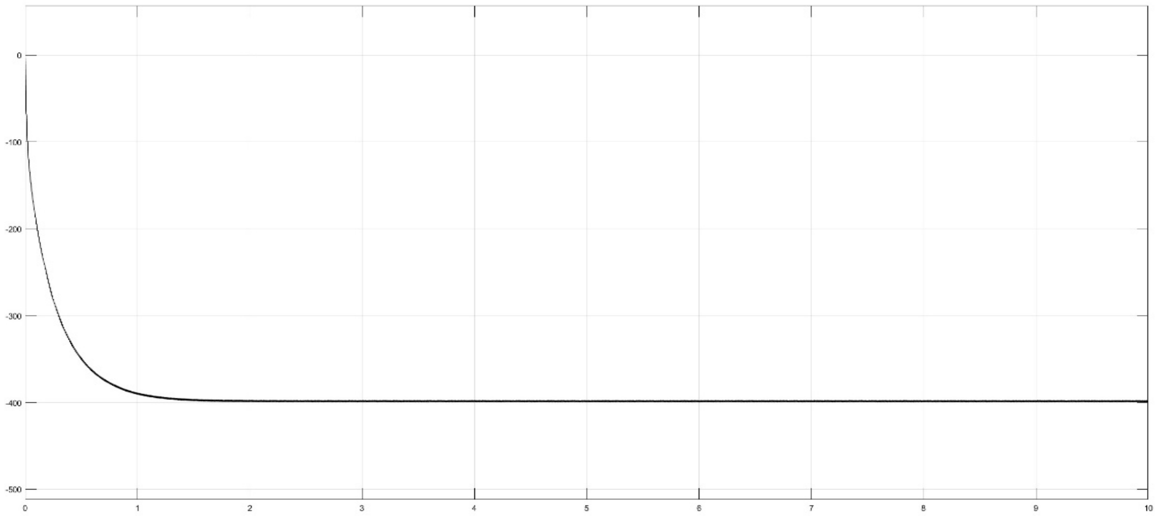

Fig. 4. Output voltage profile.

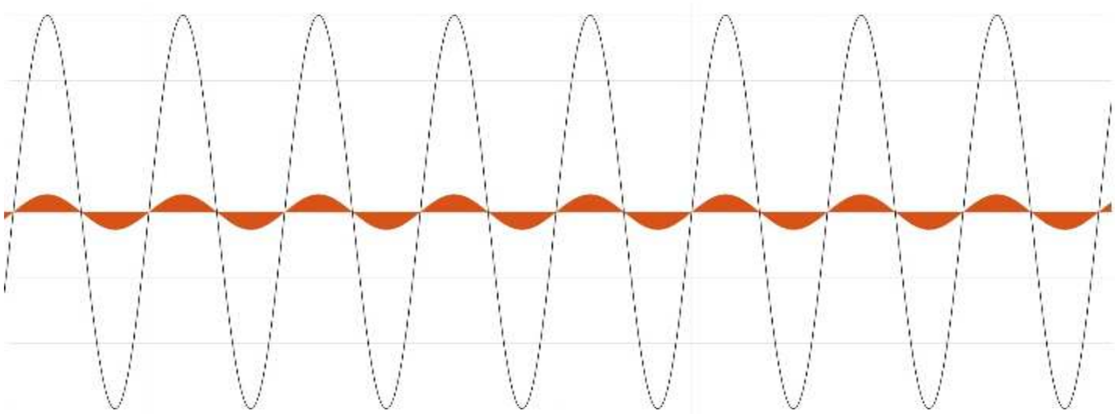

Fig. 5. sinusoidal input voltage and current.

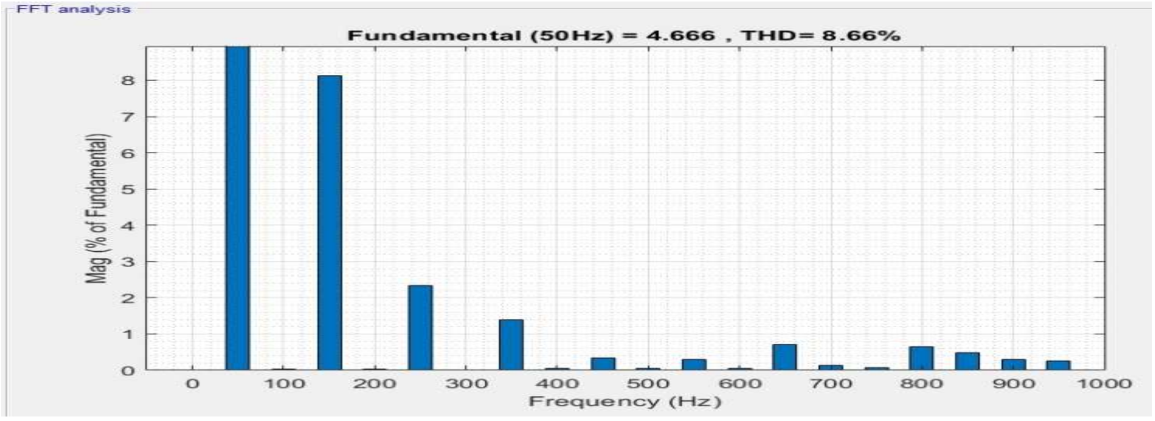

Fig. 6. THD of the case study system.

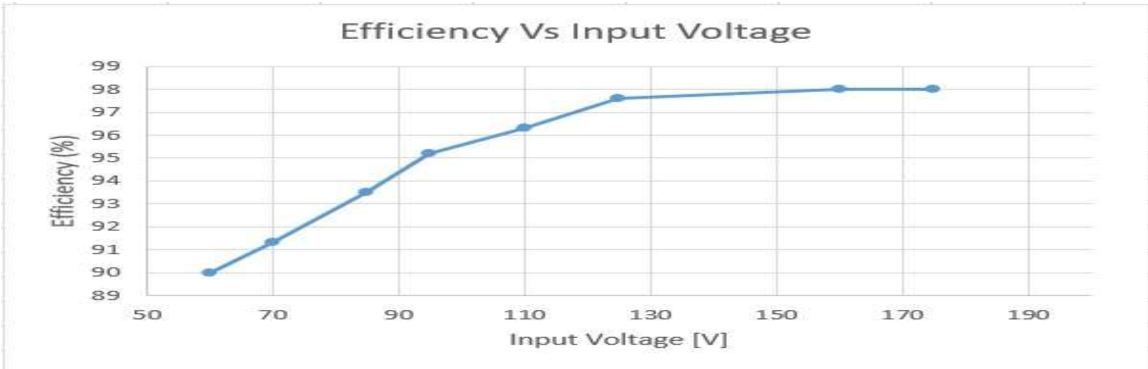

Fig. 7. Efficiency vs input supply voltage of the system.

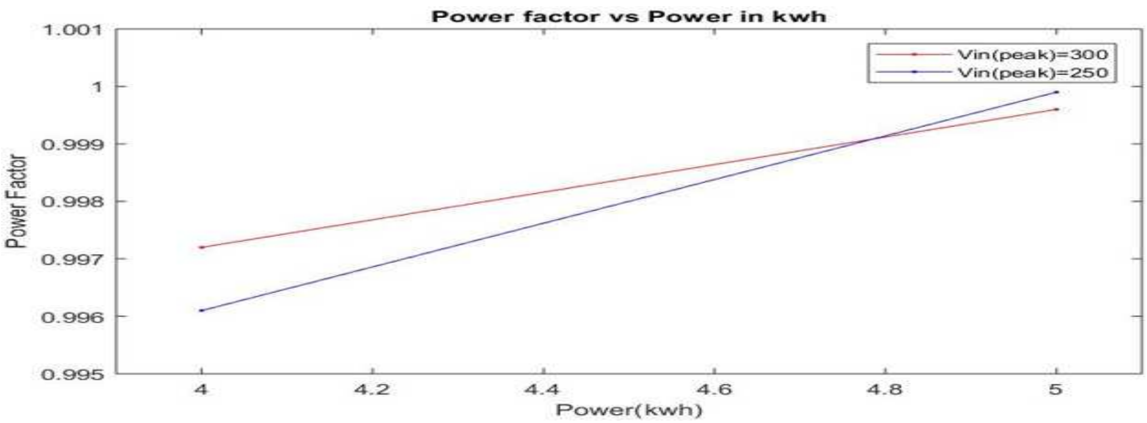

Fig. 8. Power Factor relation with output power for two different input supply voltages.

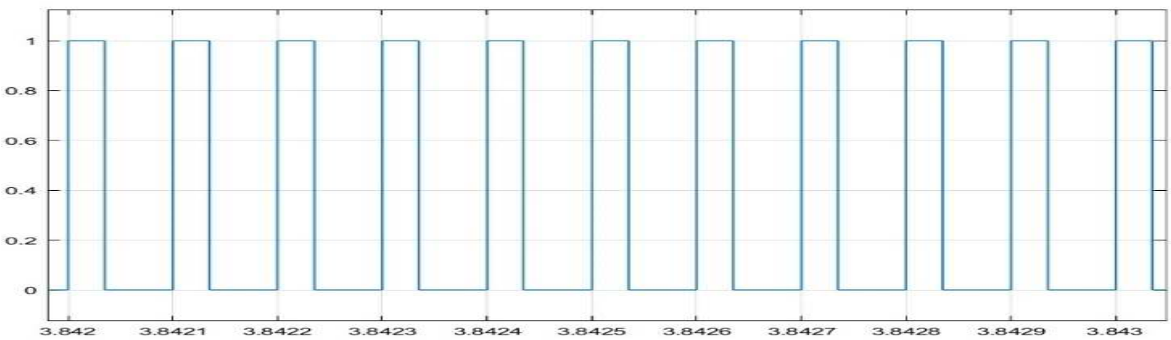

Fig. 9. Duty Cycle values.

Conclusion

In this paper a detailed design approach has been presented to achieve a very high efficiency of the system containing buck boost converter. Initially a single buck boost converter is studied for an ideal current control to obtain power balancing equations including all types of power losses inside the system. A transfer function between equivalent conductance and output voltage is arranged in a dimensionless form. This transfer function is also extended for the resistive output load. These equations are validated against the system in which two buck boost converter connected to each other in parallel for unity power factor and a high efficient system.

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