Mitigation of Stability Problem in a Boost Converter having an Input Filter.

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ABSTRACT

This paper presents a theoretical analysis of the stability status of a boost converter having an input filter using a small-signal equivalent circuit model. The addition of an L-C input filter distorts the dynamic responses of the basic converter model, owing to the presence in the resulting transfer function of double-pair of complex conjugate poles, and a pair of left half-plane complex conjugate zeros. The glitch consequence in the vicinity of filter resonant frequency noticed in the Bode plot and the attendant reduction in phase margin is mitigated by the damping of the input filter. An improved R_f-L_b parallel damping approach resulted in appreciable increase in the system phase margin.

(Keywords: input filter, stability, damping, phase margin)

INTRODUCTION

Usually, a switching converter injects a pulsating current $i_g(t)$ into the power source $v_g(t)$ of the converter. It has been observed that the Fourier series of the pulsating current contains harmonics at multiples of the switching frequency of the converter [1]. The magnitudes of the higher-order harmonics can also be significantly affected by the current spike caused by diode reverse recovery, and also by the finite slopes of the switching transitions.

The large high-frequency current harmonics i(t) can interfere with television and radio reception, and can disrupt the operation of nearby electronic equipment. To meet limits on conducted electromagnetic interference (EMI) and to improve the reliability of the system, it is necessary to add an input filter to the converter. The roles of the input filter are to: (1) attenuate the switching harmonics that are present in the converter input current waveform, (2) protect the converter and its load from transients that appear in the input voltage $v_g(t)$, thereby improving the system reliability, and (3) provide compliance with regulations that limit conducted EMI.

Several schemes involving soft-switched converters and quasi-resonant converters are documented in the literature [2] – [4]. Other approaches aimed at causing the converter input current $i_g(t)$ to be proportional to the applied input voltage $v_g(t)$ have also been documented [5]. An example of such an approach employs a full-wave diode rectifier network, cascaded by a dc-dc converter. The controller must cause the conversion ratio given in Equation (1) to vary between infinity (at the ac line voltage zero crossings) and some minimum value (at the peaks of the ac line voltage waveform):

$$M(d(t)) = \frac{v(t)}{v_r(t)} = \frac{V}{V_m\left|\sin(\omega t)\right|} \qquad \textbf{(1)}$$

The capability of producing the M(d(t)) of Equation (1) is limited to boost, buck-boost, SEPIC, and $\bar{c}$uk converters, but still require minimal input EMI filtering. These schemes all fall short of addressing the problems outlined in (2) and (3), hence the need for a low-pass filter having attenuation sufficient to meet conducted EMI specification at the converter input.

The addition of input filter, nonetheless, changes the dynamics of the converter, often in a manner that degrades regulator performance. The transient response is modified, and control system may even become unstable. The output impedance may become large over some frequency range, possibly exhibiting resonances [1]. The line-to-output, or audio-susceptibility, may be degraded.

The stability problem arising from the addition of an input filter is mitigated by introducing damping into the input filter and by designing the input filter such that its output impedance is sufficiently small [6] – [7].

Input filters are, in most cases, required to operate normally when transients or periodic disturbances are applied to the power input. Such conducted susceptibility specifications require that a design for the damping of the input filter resonances be inculcated, so that input disturbances do not excite currents or voltages within the filter or the converter.

ANALYSIS OF THE BASIC CONVERTER MODEL

When, for example, a single L-C input filter is added to a boost converter as in Figure 1, the small-signal equivalent circuit model [8] is modified as shown in Figure 2(b).

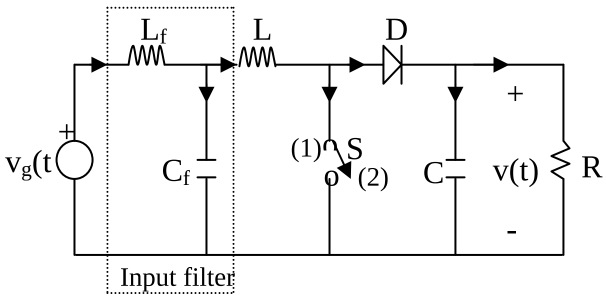


Figure 1: Addition of a Single L-C Low-Pass Filter to the Input Terminals of the Boost Converter.

The line-to-output transfer function $G_{vg}(s)$ of the small-signal model of the boost converter, Figure 2 (a), is obtained by setting all the $\hat{d}$ sources to zero.

The source $\hat{v}_g(s)$ and the inductor are pushed through the transformer to arrive at Figure 3. The line-to-output transfer function $G_{vg}(s)$ is given by:

$$G_{vg}(s) = \left.\frac{\hat{v}(s)}{\hat{v}_g(s)}\right|_{\hat{d}(s)=0} = \frac{1}{D'} \frac{1}{1 + s\frac{L}{D'^2 R} + s^2 \frac{LC}{D'^2}} \quad \textbf{(2)}$$

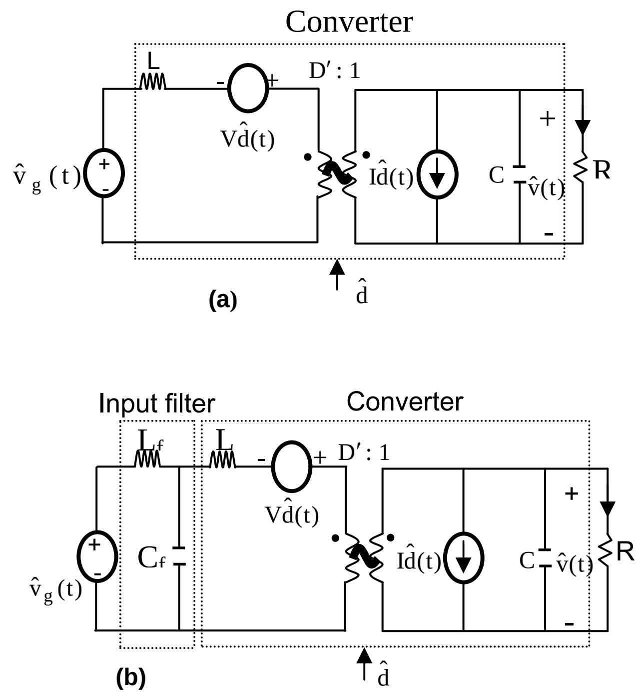


Figure 2: Small-Signal Equivalent Models of the Boost Converter (a) Basic Converter Model, (b) With Addition of Input Filter.

Comparing Equation (2) with the standard normalized transfer function given by:

$$G_{vg}(s) = G_{go} \frac{1}{1 + \frac{s}{Q\omega_o} + \left(\frac{s}{\omega_o}\right)^2} \quad \textbf{(3)}$$

where G_{go} is known as DC gain, the salient features of the line-to-output transfer function are:

$$G_{go} = \frac{1}{D'} \quad \textbf{(4)}$$

$$\omega_o = \frac{D'}{\sqrt{LC}}, \quad f_o = \frac{D'}{2\pi\sqrt{LC}} \quad \textbf{(5)}$$

$$Q = D'R\sqrt{\frac{C}{L}} \quad \textbf{(6)}$$

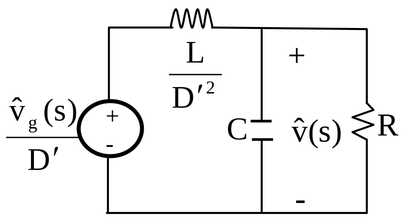


Figure 3: The Source $\hat{v}_g(s)$ and the inductor pushed through the transformer.

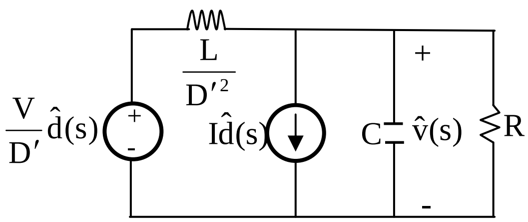


Figure 4: The $\hat{d}$ voltage source and the inductor pushed through the transformer.

The analysis of the control-to-output transfer function $G_{vd}(s)$ is achieved by applying the principle of superposition owing to the presence in the small-signal model of more than one generator that depend on $\hat{d}(s)$. The control-to-output transfer function $G_{vd}(s)$ is obtained by setting $\hat{v}_g$ source equal to zero. The $\hat{d}$ voltage source and the inductor are pushed through the $D':1$ transformer to obtain the circuit of Figure 4.

Further analysis which involve individual setting of the current and voltage sources to zero are made to arrive at the transfer function given by:

$$G_{vd}(s) = \left.\frac{\hat{v}(s)}{\hat{d}(s)}\right|_{\hat{v}_g(s)=0} = \frac{V}{D'}\,\frac{\left(1 - s\frac{LI}{D'V}\right)}{1 + s\frac{L}{D'^2R} + s^2\frac{LC}{D'^2}} \qquad \textbf{(7)}$$

Comparing Equation (7) with the standard normalized transfer function given by:

$$G_{vd}(s) = G_{do}\,\frac{\left(1 - \frac{s}{\omega_z}\right)}{\left(1 + \frac{s}{Q\omega_o} + \left(\frac{s}{\omega_o}\right)^2\right)} \qquad \textbf{(8)}$$

The salient features of the transfer function of Equation (7) which are derived by equating coefficients in Equations (7) and (8) are:

$$G_{do} = \frac{V}{D'} \qquad \textbf{(9)}$$

$$\omega_o = \frac{D'}{\sqrt{LC}}, \quad f_o = \frac{D'}{2\pi\sqrt{LC}} \qquad \textbf{(10)}$$

$$Q = D'R\sqrt{\frac{L}{C}} \qquad \textbf{(11)}$$

$$\omega_z = \frac{D'^2R}{L}, \quad f_z = \frac{D'^2R}{2\pi L} \qquad \textbf{(12)}$$

The denominators of Equations (2) and (7) are identical; hence $G_{vd}(s)$ and $G_{vg}(s)$ share the same ω_o and Q. The approximate Bode plot of the magnitude and phase of the control-to-output transfer function G_{vd} and line-to-output transfer function G_{vg} are constructed in Figure 5 and Figure 6, respectively.

EFFECT OF UN-DAMPED INPUT FILTER ON THE BOOST CONVERTER TRANSFER FUNCTIONS

In this paper, we limited our study of the effect of the input filter on the stability of the boost converter model, to the variations in the line-to-output transfer function. To determine the line-to-output transfer function $G_{vg}(s)$ in the presence of the input filter, we set all the $\hat{d}$ sources to zero. The source $\hat{v}_g(s)$, the inductor L, and the filter inductor and capacitor, L_f and C_f, respectively, are pushed through the transformer to arrive at Figure 7.

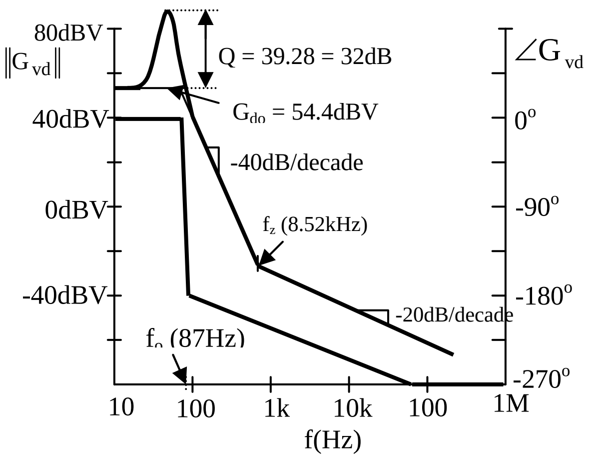


Figure 5: Bode Plot of the Control-to-Output Transfer Function G_{vd} for the Boost Converter Model.

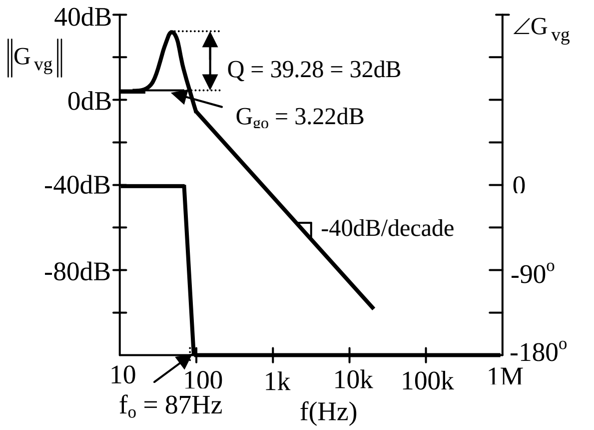


Figure 6: Bode Plot of the Line-to-Output Transfer Function G_{vg} for the Boost Converter Model.

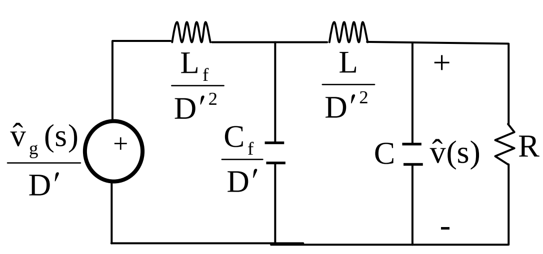


Figure 7: The Source $\hat{v}_g(s)$, the Inductor L, and the Filter Inductor L_f and Capacitor C_f pushed through the Transformer.

The line-to-output transfer function $G_{vg}(s)$ is a fourth degree polynomial given in Equation (13).

DAMPING OF THE INPUT FILTER

Several schemes are available for the effective damping of the input filter. The choice of any scheme depends among other factors on (a) the desire to prevent transients and disturbances in $v_g(t)$ from exciting filter resonances, (b) the attainment of desired filter attenuation, (c) the desire to minimize the size of reactive elements [9], etc. In this paper, a damping resistor R_f, is connected across the filter inductor L_f. The choice of R_f across L_f is to avoid power dissipation in R_f since the dc voltage across the inductor L_f is zero. However, the resulting transfer function derived from Figure 8 after the usual manipulations in the small signal model contains a high frequency zero.

The result of connecting R_f across L_f is that much of the filter attenuation is provided by C_f. The line-to-output transfer function $G_{vg}(s)$ of the converter model with damped input filter is also a fourth degree polynomial Equation (14).

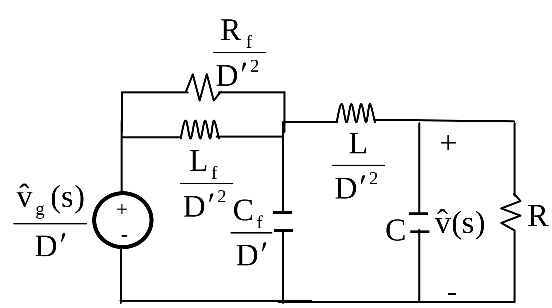


Figure 8: Addition of R_f to Figure 7.

A more practical and standard approach, however, is to connect a low value, high-frequency blocking inductor L_b (usually, $L_b < L_f$) in series with the damping resistor R_f. The high-frequency blocking inductor L_b also contributes to the effective attenuation provided by the input filter. The addition of the inductor L_b effectively boosts the system phase margin, Pm. A low value damping resistor, R_f results in a very high value phase margin at an angular frequency much lower than the pre-input filter model value. To boost the frequency and still maintain reasonable phase margin requires an appreciable increase in the value of the damping resistor R_f.

$$G_{vg}(s) = \frac{1}{D'} \frac{1 + s^2 \frac{L_f C_f}{D'^3}}{1 + s\frac{(L + L_f)}{RD'^2} + s^2 \frac{(LCD' + L_f CD' + L_f C_f)}{D'^3} + s^3 \frac{LL_f C_f}{RD'^5} + s^4 \frac{LL_f C_f C}{D'^5}}$$

(13)

$$G_{vg}(s) = \frac{\frac{1}{D'} \quad 1 + s\frac{L_f}{R_f} + s^2 \frac{L_f C_f}{D'^3}}{1 + s\frac{(L + L_f + \frac{L_f}{R_f} D'^2)}{RD'^2} + s^2 \frac{(LCD' + L_f CD' + L_f C_f + L\frac{L_f}{RR_f} D')}{D'^3} + s^3 \frac{(L\frac{L_f}{R} C_f + LC\frac{L_f}{R_f} D'^3)}{D'^5} + s^4 \frac{LL_f C_f C}{D'^5}}$$

(14)

The benefit of increasing the R_f is that there is no power dissipation in it since the dc voltage across the inductor L_f is zero. The resulting line-to-output transfer function is a fifth order polynomial and contains a high frequency zero and is given by:

$$G_{vg}(s) = \frac{1}{D'} \frac{s^3 A + s^2 B + sE + 1}{s^5 F + s^4 G + s^3 H + s^2 J + sK + 1} \quad \textbf{(15)}$$

where,

$$A = \frac{L_b L_f C_f}{R_f D'^3} \quad \textbf{(15a)}$$

$$B = \frac{L_f C_f}{D'^3} \quad \textbf{(15b)}$$

$$E = \frac{L_b + L_f}{R_f} \quad \textbf{(15c)}$$

$$F = \frac{LL_b C_f C}{R_f D'^5} \quad \textbf{(15d)}$$

$$G = \frac{LC_f}{D'^5}\left(L_f C + \frac{L_b}{RR_f}\right) \quad \textbf{(15e)}$$

$$H = \frac{1}{D'^2}\left(\frac{L_b L_f C}{R_f} + \frac{LL_b C}{R_f} + \frac{LL_f C}{R_f} + \frac{LL_f C_f}{RD'^3} + \frac{L_b L_f C_f}{R_f D'}\right) \quad \textbf{(15f)}$$

$$J = \frac{1}{D'^2}\left(L_f C + LC + \frac{L_b L_f}{RR_f} + \frac{LL_b}{RR_f} + \frac{LL_f}{RR_f} + \frac{L_f C_f}{D'}\right) \quad \textbf{(15g)}$$

$$K = \frac{L_f}{RD'^2} + \frac{L}{RD'^2} + \frac{L_b}{R_f} + \frac{L_f}{R_f} \quad \textbf{(15h)}$$

DISCUSSION OF SIMULATION RESULTS

Some simulations were run to verify the stability conditions of the three model states of the converter: pre-input filter model, addition of input filter model, and damped input filter model. Test techniques employed in the simulations include Nyquist plots, Bode plots, and step response plots. Data for the basic converter model used both in the analysis and simulations are shown in Table 1.

Filter inductor L_f, and capacitor C_f, values were selected to meet specifications while damping resistor R_f was iteratively got through series of simulation runs. The Nyquist plot, Bode diagram, and step response for the pre-input filter converter model are shown in Figure 9. The Bode diagram, Figure 9 (b), indicates a stable system with a positive phase margin and infinite gain margin. The transient response of the system,

Figure 9 (c), to a step input shows that a steady state was achieved in less than two seconds after the transient excitation.

Table 1: Data for the Basic Converter Model.

Input voltage, V_g	248.3V
Duty cycle, D	0.31
Inductor, L	800uH
Capacitor, C	2000uF
Load resistor, R	90Ω
Switching period, T	100us
Filter inductor, L_f	500uH
Filter capacitor, C_f	470uF
Damping resistor, R_f	50Ω

Source: [10]

Figure 10 shows the stability plots for the basic converter model with the input filter added. In the vicinity of the filter resonant frequency, Figure 10 (b), the transfer function contains a double-pair of complex poles, and also a pair of left half-plane complex conjugate zeros. These cause a glitch in the magnitude plot, with a contribution of 180^o of lead to the phase of the original plot. When the 180^o contributed by the addition of input filter is added to -180^o contributed at high frequency, $(\omega/\omega_o) > 1$, by the two poles of the original $G_{vd}(s)$, a high frequency phase asymptote of 0^o is obtained. The introduction of the input filter left-shifted the basic converter angular corner frequency from ω_o = 545 rad/s to ω_o = 402 rad/s, with a resulting decrease in quality factor, Q, from 98 to 82. The transient response of the system to a step input, Figure 10 (c), shows an extra 10 seconds delay before steady state was attained.

In Figure 11 (a), it is explicitly evident that the -1+j0 point was not enclosed by the Nyquist plot. The glitch in the magnitude plot in the vicinity of the filter resonant frequency which dipped to about -150dB was reduced to below -50dB by damping the input filter, Figure 11 (b). At an optimum value of $R_{f,optimum}$ = 50Ω, the high-frequency spike-like quality factor, Q, experienced an appreciable damping, however, to the detriment of an increase in the low frequency Q. At this $R_{f,optimum}$, the phase margin is about 0.22^o at an angular frequency 599.41 rad/sec. When a 50μH high-frequency blocking inductor was connected in series with $R_{f,optimum}$, the phase margin rose to 63.69^o at an angular frequency 527.64 rad/sec.

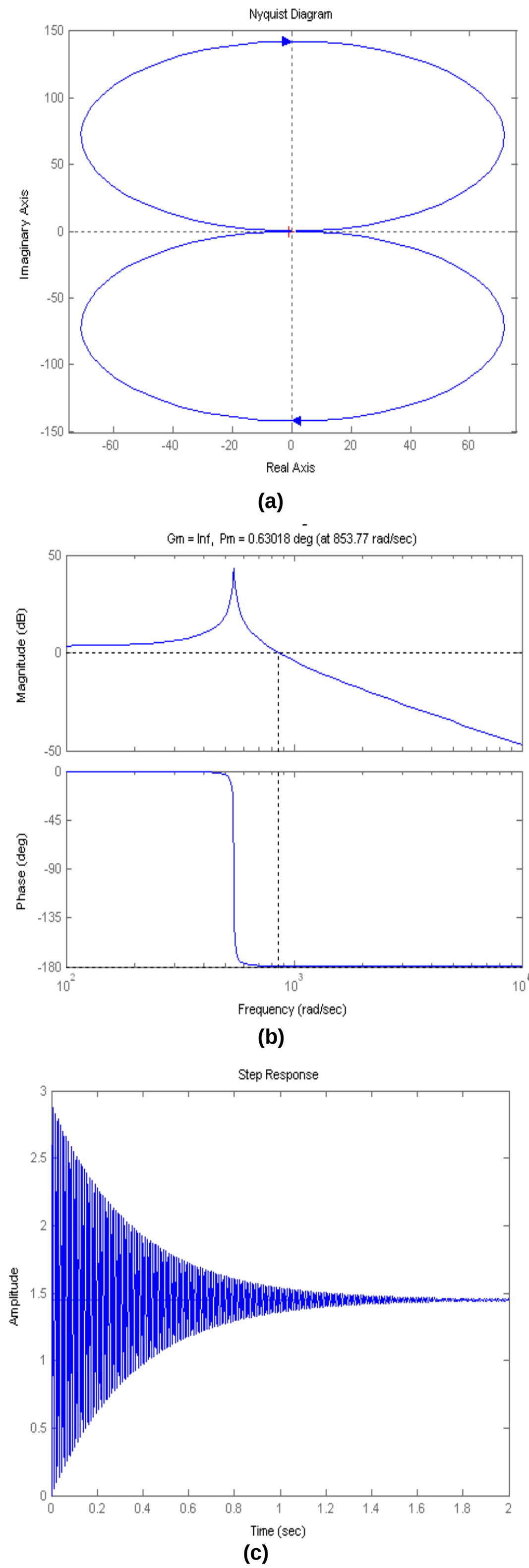
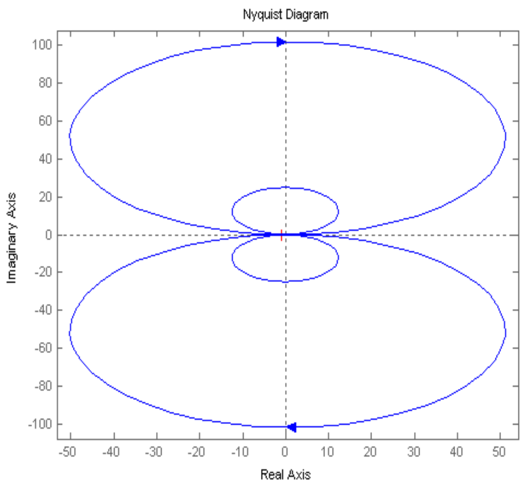
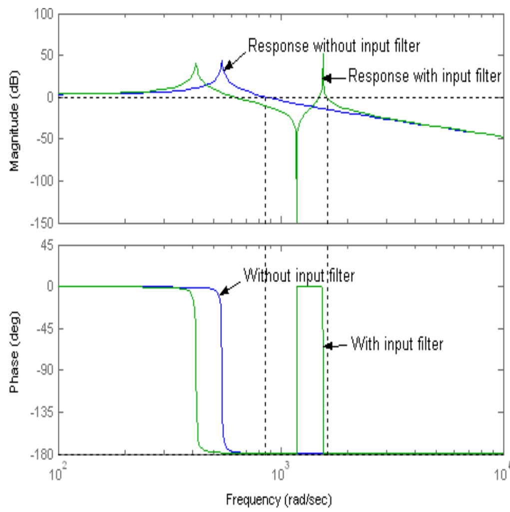


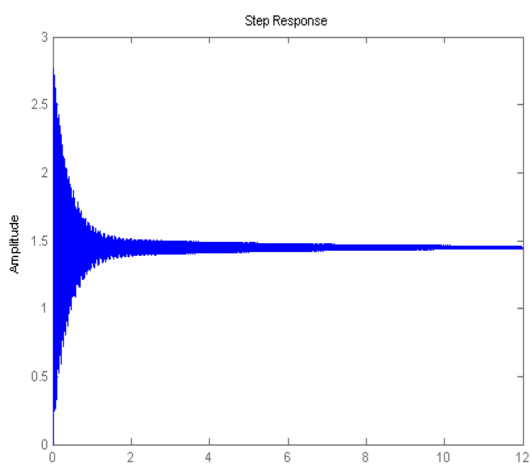
Figure 9: Stability Plots for the Basic Converter Model (a) Nyquist Plot, (b) Bode Plot, (c) Step Response Plot.



(a)

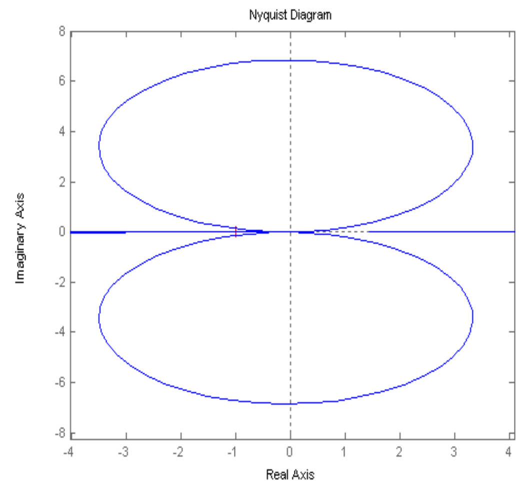


(b)

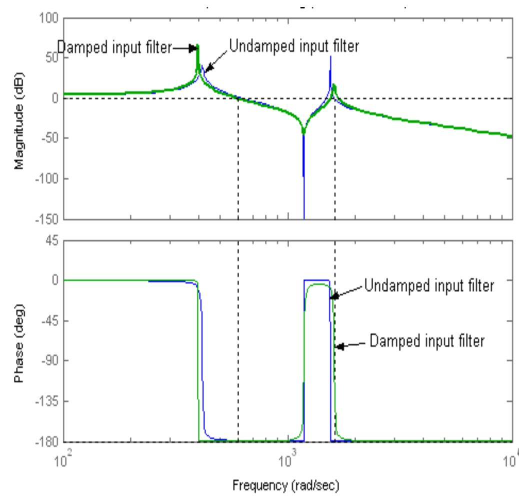


(c)

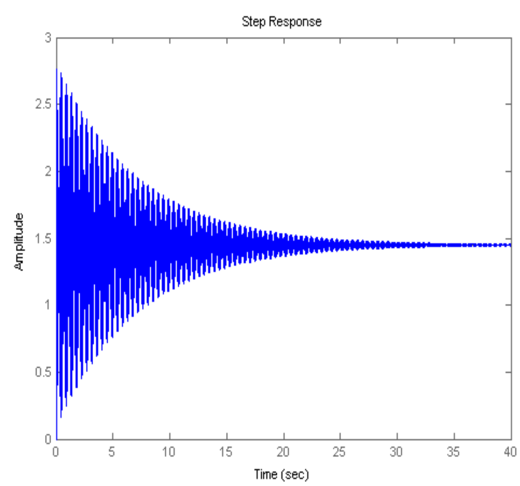
Figure 10: Stability Plots for the Basic Converter Model with Undamped Input Filter (a) Nyquist Plot, (b) Bode Plot, (c) Step Response Plot



(a)

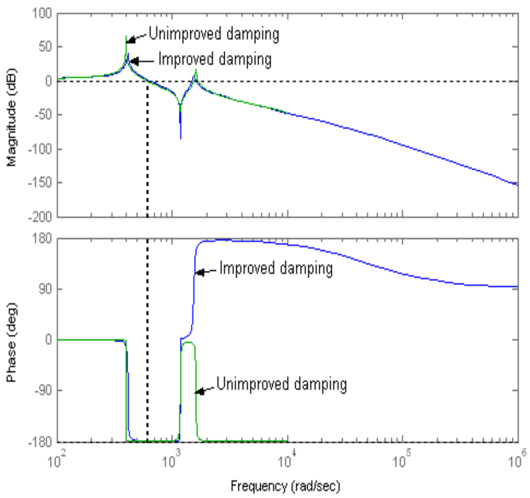


(b)



(c)

Figure 11: (Continued on Next Page).



(d)

Figure 11: Stability Plots for the Basic Converter Model with Damped Input Filter (a) Nyquist Plot, (b) Bode Plot, (c) Step Response Plot, (d) Improved Damped Response.

Through an iterative procedure, a final optimum value, $R_{foptimum}$ = 5KΩ, was selected. Figure 11 (d) displays the improved Bode response of the system. In addition to a double-pair of complex poles and a pair of left half-plane complex zeros, the transfer function also contains one each of negative real pole and zero. They both combine to contribute a 360^o of lead to the phase of the original plot.

When the 360^o contributed by the damped input filter is added to -180^o contributed at high frequency, $(\omega/\omega_o) > 1$, by the two poles of the original $G_{vd}(s)$, a high frequency phase asymptote of 180^o is obtained. The low-frequency Q was effectively damped. The phase margin of 1.11^o at an angular frequency of 624.34 rad/sec was achieved.

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