Dynamic Characteristics of Multiple-Output Buck-Type Converters with Weighted Voltage Control

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Abstract *- The paper discusses the dynamic characteristics, i.e., audio susceptibilities, output impedances and output transimpedances, of multiple-output buck-type converters with weighted voltage control. The open-loop and closed-loop small-signal characteristics are examined. Based on the analysis, the design issues are addressed*

I. Introduction

On many occasions, weighted voltage control (WVC) and coupled inductors are utilized in multiple-output converters (MOCs), Fig. 1. Most early-published papers about MOCs mainly dealt with stability analysis for MOCs with coupled inductors [1,2]. Some papers published lately discussed MOCs with WVC [3 - 5]. But their focus was on the modeling and analysis. The effects of WVC and coupled inductors on the audio susceptibility and output impedance, which are very important for compensator design, are not well understood. The purpose of this paper is to address the design issues from the standpoint of the small-signal characteristics, *i.e.*, audio susceptibilities, output impedances, and output transimpedances, based on the small-signal model presented in [4].

First, the paper reviews the small-signal model of multiple-output buck-type converters given in [4]. To facilitate discussion, the model is presented in the form of small-signal block diagram. Then the open-loop small-signal characteristics are discussed. The closed-loop small-signal characteristics are examined, which reveal the combined effects of coupled inductors and WVC. Different compensator design schemes are discussed and compared. Based on the analysis, the design issues are discussed.

To provide design insight, the analysis and discussion are performed on a dual-output forward converter.

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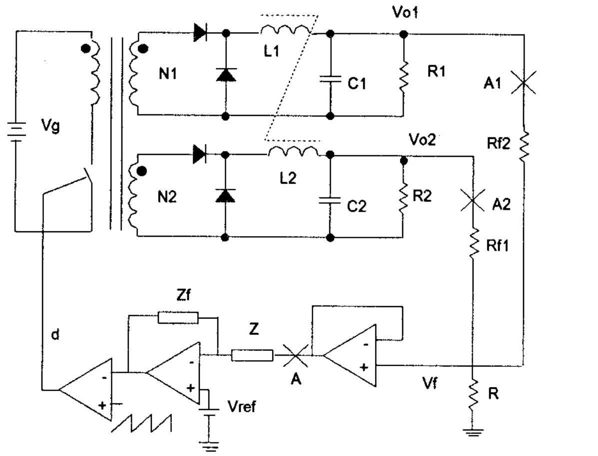

Fig. 1. A multiple-output forward converter with WVC and coupled output filter inductors. *The coupling between the output filter inductors is characterized by the mutual inductance, M.*

II. Small-Signal Model

The small-signal circuit model presented in [4] fully describes the small-signal behavior of the buck-type MOC. By using circuit simulation tools, such as PSpice and Saber, all of the small-signal transfer functions can be obtained. However, when discussing audio susceptibilities and output impedances, it is more convenient to represent the system using a block diagram. The block diagram is more suitable when using system simulation tools, such as Easy 5 and Matlab. Figure 2 shows the small-signal block diagram of a dual-output converter with weighted voltage control and coupled inductors. Each open-loop transfer function is defined and derived as follows.

The duty cycle-to-output transfer functions:

$$G_{vdi} \equiv \frac{v_{oi}}{d} = \frac{Num(G_{vdi})}{Den}, \qquad i = 1,2. \tag{1}$$

The audio susceptibilities, or line-to-output transfer functions:

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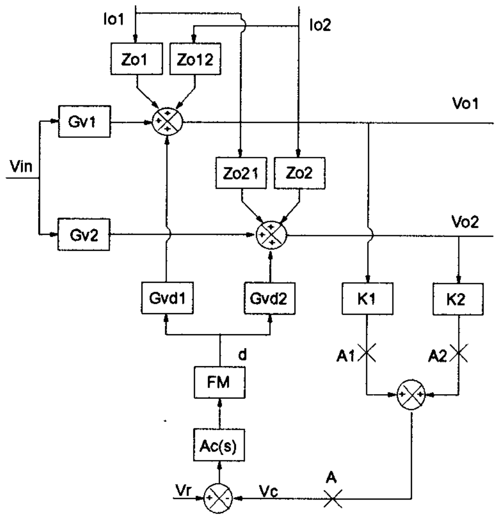

Fig. 2. Small-signal block diagram. *Any perturbations from the line voltage and load currents enter the system through two paths, whose net effects at the outputs tend to cancel each other.*

$$G_{vi} \equiv \frac{v_{oi}}{v_{in}} = \frac{Num(G_{vi})}{Den}, \quad i = 1,2. \tag{2}$$

The output impedance transfer functions:

$$Z_{oi} \equiv \frac{v_{oi}}{i_{oi}} = \frac{Num(Z_{oi})}{Den}, \quad i = 1,2. \tag{3}$$

The output transimpedance transfer functions:

$$Z_{oij} \equiv \frac{v_{oi}}{i_{oj}} = \frac{Num(Z_{oij})}{Den}, \quad i,j = 1,2, \quad i \neq j. \tag{4}$$

The denominator of each transfer function is the characteristic equation of the system, which is therefore invariant. The transfer functions differ from each other by numerators. The analytical expressions of the denominators and the numerators are summarized in Table 1.

III. Open-Loop Characteristics

The open-loop audio susceptibilities and the output impedances are shown in Figs. 3 and 4 respectively. Like a single-output converter, audio susceptibilities describe the rejection of the line disturbance, whereas the output impedances describe the rejection of the load disturbances. One unique feature of multiple-output converters is that each output voltage is dependent not only on its own load current, which is characterized by the output impedance, but also on the other load current, which is characterized by transimpedance as shown in Fig. 5.

It is interesting to notice that the expressions of the output transimpedances $Z_{o12}(s)$ and $Z_{o21}(s)$ are identical, *i.e.*,

Table 1. Analytical expressions of the open-loop transfer functions.

Den	$[1+(R_{C1}C_1+\frac{L_1}{R_1}+R_{L1}C_1)s+L_1C_1s^2][1+(R_{C2}C_2+\frac{L_2}{R_2}+R_{L2}C_2)s+L_2C_2s^2]-[\frac{M_{12}}{R_1}s+C_1M_{12}s^2][\frac{M_{12}}{R_2}s+C_2M_{12}s^2]$
$Num(G_{vd1})$	$V_{in}(1+R_{C1}C_1s)\{N_1[1+(R_{C2}C_2+\frac{L_2}{R_2}+R_{L2}C_2)s+L_2C_2s^2]-N_2[\frac{M_{12}}{R_2}s+C_2M_{12}s^2]\}$
$Num(G_{vd2})$	$V_{in}(1+R_{C2}C_2s)\{N_2[1+(R_{C1}C_1+\frac{L_1}{R_1}+R_{L1}C_1)s+L_1C_1s^2]-N_1[\frac{M_{12}}{R_1}s+C_1M_{12}s^2]\}$
$Num(G_{v1})$	$D(1+R_{C1}C_1s)\{N_1[1+(R_{C2}C_2+\frac{L_2}{R_2}+R_{L2}C_2)s+L_2C_2s^2]-N_2[\frac{M_{12}}{R_2}s+C_2M_{12}s^2]\}$
$Num(G_{v2})$	$D(1+R_{C2}C_2s)\{N_2[1+(R_{C1}C_1+\frac{L_1}{R_1}+R_{L1}C_1)s+L_1C_1s^2]-N_1[\frac{M_{12}}{R_1}s+C_1M_{12}s^2]\}$
$Num(Z_{o1})$	$(1+R_{C1}C_1s)\{[1+(R_{C2}C_2+\frac{L_2}{R_2}+R_{L2}C_2)s+L_2C_2s^2](sL_1+R_{L1})-\frac{s^2M_{12}^2}{R_2}[1+R_2C_2s]\}$
$Num(Z_{o2})$	$(1+R_{C2}C_2s)\{[1+(R_{C1}C_1+\frac{L_1}{R_1}+R_{L1}C_1)s+L_1C_1s^2](sL_2+R_{L2})-\frac{s^2M_{12}^2}{R_1}[1+R_1C_1s]\}$
$Num(Z_{o12})$	$sM_{12}(1+R_{C1}C_1s)(1+R_{C2}C_2s)$
$Num(Z_{o21})$	$sM_{12}(1+R_{C1}C_1s)(1+R_{C2}C_2s)$

$$Z_{oij} \equiv Z_{oji}, \quad i,j = 1,2, \quad i \neq j. \tag{5}$$

This can be explained by examining the relationship between the two outputs from the system configuration as shown in Fig. 2. Since the power stage transformer can be

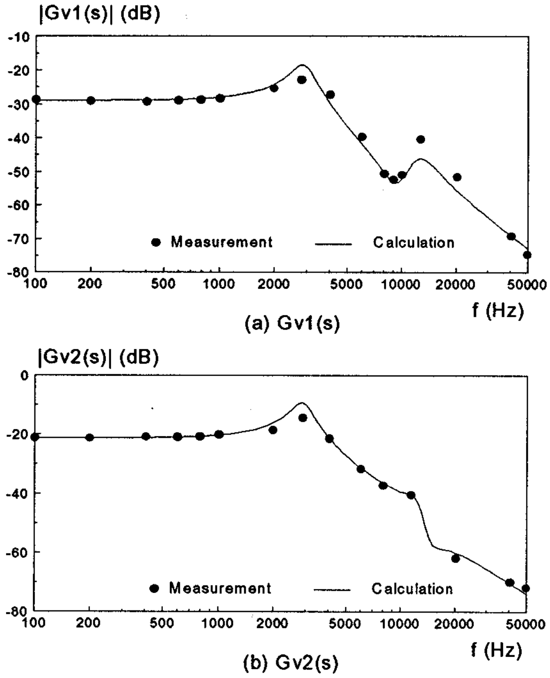

Fig. 3. Open-loop audio susceptibilities. (a) Open-loop audio from the input to output 1. (b) Open-loop audio from the input to output 2.

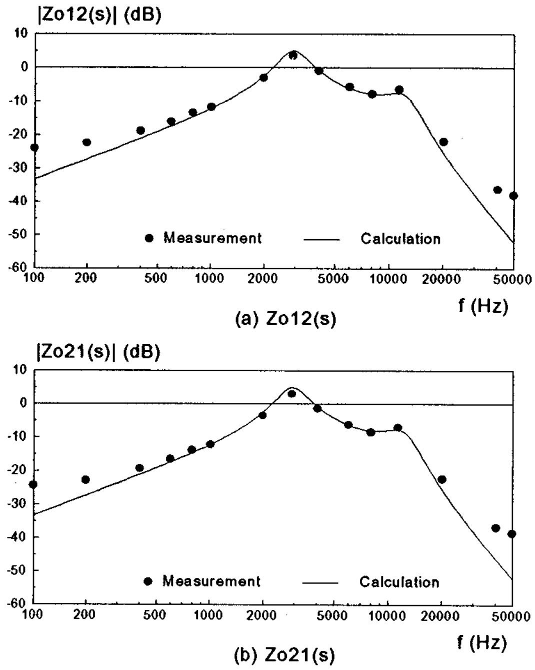

Fig. 5. Open-loop output transimpedances. (a) Output transimpedance $Z_{o12}(s)$. (b) Output transimpedance $Z_{o21}(s)$. The output transimpedances are reciprocal.

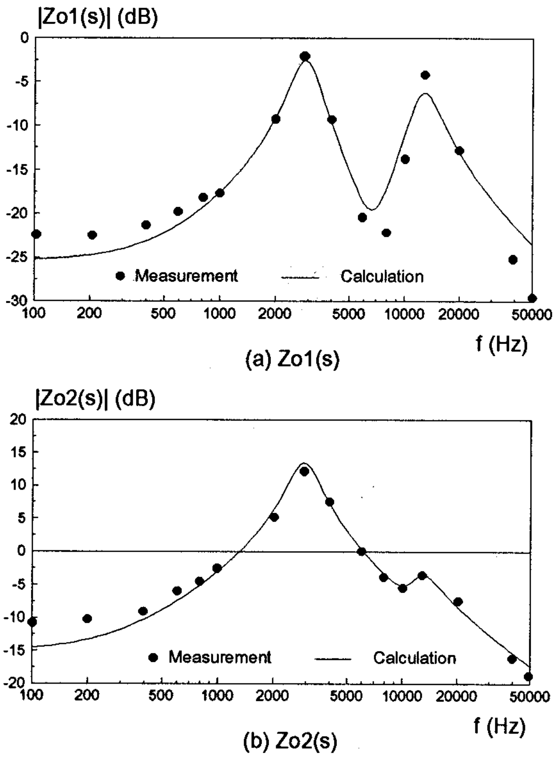

Fig. 4. Open-loop output impedances. (a) Output impedance of output 1. (b) Output impedance of output 2.

approximated as an ideal transformer, all power channels are virtually decoupled from each other. The only physical link between the power channels is through the mutual inductance of the coupled output filter inductors, $M(=M_{12}=M_{21})$, which is the same regardless of the power channel. As a result, the transimpedances are also identical. In other words, the output transimpedances are reciprocal. The experimental results confirm the theoretical derivation.

Due to coupling between the output filter inductors, the converter is no longer a 2nd order system, but rather a 4th order system. As a result, the duty cycle-to-output (feedback) transfer functions have two pairs of complex poles and one pair of complex zeros. The audio susceptibilities and output impedances have two peaks, and they can be comparable in magnitude. The compensator design philosophy for single output converter does not account for this effect, and therefore a new design methodology is needed.

IV. Closed-Loop Audio Susceptibilities and Output Impedances

For a dc-dc converter, more important than the open-loop characteristics is the closed-loop small-signal behavior, which is determined, to a large extent, by the feedback compensator. When discussing closed-loop transfer functions, the loop gain is always used, which is defined by breaking the loop at point A in Fig. 2:

$$T = A_c F_m (K_1 G_{vd1} + K_2 G_{vd2}). \qquad (6)$$

The loop gain as defined can provide stability information and is convenient to use when performing compensator design. Since the product of each weighting factor, the duty cycle-to-output transfer function, the modulator gain, and the compensator transfer function often appear in the expressions of the closed-loop audio susceptibility and output impedance transfer functions, two additional loop gains are defined:

$$T_1 = A_c F_m K_1 G_{vd1}\ , \tag{7}$$

and

$$T_2 = A_c F_m K_2 G_{vd2}. \tag{8}$$

T_1 and T_2 happen to take the same form as the loop gain of the single output converter. It should be noticed that neither loop gain, T_1 or T_2, by itself provides adequate stability information.

The analytical expression for each closed-loop transfer function is derived and summarized in Table 2.

It can be seen that both the denominators and numerators of the analytical expressions of the closed-loop audio susceptibilities and output impedances include the compensator $A_c(s)$. This may mislead one to the conclusion that wide bandwidth, which usually requires high loop gain, does not improve dynamic regulation. The output impedance $Z_{o1}(s)$ is used to demonstrate how high loop gain magnitude improves the dynamic characteristics.

The closed-loop output impedance $Z_{o1(CL)}(s)$ is derived as:

$$Z_{o1(CL)} = \frac{Z_{o1}(1+T_2) - \frac{K_2}{K_1} T_1 Z_{o21}}{1+T}. \tag{9}$$

The following three scenarios are discussed.

(1) Assuming $|T_1|>>1$ and $|T_2|>>1$, $Z_{o1(CL)}(s)$ can be simplified as:

$$Z_{o1(CL)} \approx \frac{Z_{o1}T_2 - \frac{K_2}{K_1} T_1 Z_{o21}}{T} = \frac{K_2(G_{vd2}Z_{o1} - G_{vd1}Z_{o21})}{K_1 G_{vd1} + K_2 G_{vd2}}. \tag{10}$$

The numerator of the closed-loop output impedance is the difference between two terms. The first term $G_{vd2}Z_{o1}$ is caused by the output impedance of the V_{o1} power channel and the second term $G_{vd1}Z_{o21}$ is caused by the output transimpedance between two power channels, which is due to the coupled inductors. The difference between these two terms results in the reduction of the output impedance. Physically, this can be explained by referring to the small-signal block diagram as given in Fig. 2. The current disturbance at $i_{o1}(s)$ is injected into the system through two paths: one through the output impedance $Z_{o1}(s)$, and the other through the output transimpedance $Z_{o21}(s)$. The effect of weighted voltage control tends to cancel out these two disturbances at the output. As a result, the output impedance is decreased by using coupled inductors and weighted-voltage control simultaneously.

Table 2. Analytical Expressions of the Closed-Loop Transfer Functions

Audio Susceptibility 1	$G_{v1(CL)}$	$\frac{G_{v1}T_2 - \frac{K_2}{K_1}T_1 G_{v2}}{1+T}$
Audio Susceptibility 2	$G_{v2(CL)}$	$\frac{G_{v2}T_1 - \frac{K_1}{K_2}T_2 G_{v1}}{1+T}$
Output Impedance 1	$Z_{o1(CL)}$	$\frac{Z_{o1}(1+T_2) - \frac{K_2}{K_1}T_1 Z_{o21}}{1+T}$
Output Impedance 2	$Z_{o1(CL)}$	$\frac{Z_{o2}(1+T_1) - \frac{K_1}{K_2}T_2 Z_{o12}}{1+T}$
Transimpedance from i_{o2} to v_{o1}	$Z_{o12(CL}$	$\frac{Z_{o12}(1+T_2) - \frac{K_2}{K_1}T_1 Z_{o2}}{1+T}$
Transimpedance from i_{o1} to v_{o2}	$Z_{o21(CL}$	$\frac{Z_{o21}(1+T_1) - \frac{K_1}{K_2}T_2 Z_{o1}}{1+T}$
Loop Gain 1	T_1	$A_c F_m K_1 G_{vd1}$
Loop Gain 2	T_2	$A_c F_m K_2 G_{vd2}$
Total Loop Gain	T	$A_c F_m (K_1 G_{vd1} + K_2 G_{vd2})$

It should be pointed out that if coupled inductors are not employed, the second term which is related to the output transimpedance will disappear, and consequently the reduction of the closed-loop output impedance will be far less effective. An interesting result is obtained by assuming there is no coupling between the output filter inductors, *i.e.*, $Z_{o21}=0$. The closed-loop output impedance becomes

$$Z_{o1(CL)} = Z_{o1}\frac{(1+T_2)}{1+T} \approx \frac{Z_{o1}}{1+(K_1 G_{vd1})/(K_2 G_{vd2})}. \tag{11}$$

To facilitate discussion, the closed-loop output impedance for the other power channel is also provided:

$$Z_{o2(CL)} = Z_{o2}\frac{(1+T_1)}{1+T} \approx \frac{Z_{o2}}{1+(K_2 G_{vd2}/K_1 G_{vd1})}. \tag{12}$$

It can be seen that the closed-loop output impedances are mainly determined by the ratio of the weighting factors. Assuming that the two power channels are identical and the outputs are equally weighted, then the closed-loop output impedance will be one half of the open-loop output impedances, or a *6 dB* reduction in the closed-loop output impedances will take place. It can be concluded that reduction of the output impedance by using weighted voltage control is very limited. In addition, it is impossible to reduce the output impedances in both power channels. The reduction of the output impedance in one power channel is at the cost of increasing the output impedance in the other power channel. On the other hand, if coupled inductors are used, the closed-loop output impedance is dependent not only on the open-loop output impedance Z_{o1} but also on the open-loop transimpedance Z_{o21}. These two terms appear in the numerator of the closed-loop output impedance with opposite signs, which results in reduced output impedance. This shows one of the advantages of using coupled inductors.

(2) Assuming $|T_1|<<1$ and $|T_2|<<1$, then $Z_{o1(CL)}(s)$ can be simplified as:

$$Z_{o1(CL)} \approx Z_{o1}. \tag{13}$$

The closed-loop output impedance is approximately the same as the open-loop output impedance. Therefore, under this circumstance, the feedback does not affect load regulation.

Compared with a single-output converter, where the closed-loop output impedance is directly suppressed by the high loop gain, the reduction of the closed-loop output impedance for the dual-output converter is not directly decided by the loop gain, but rather by the difference between $G_{vd2}Z_{o1}$ and $G_{vd1}Z_{o21}$, Eq. (10). This improvement of the output impedance, nevertheless, is under the conditions $|T_1|>>1$ and $|T_2|>>1$, which means that the loop gain should be high over the frequency range of interest. If the magnitude of the loop gain is too low, the closed-loop output impedance will simply take the value of the open-loop output impedance.

(3) When the loop gain is comparable to unity, *i.e.*, $|T_1|\approx 1$ and $|T_2|\approx 1$, which occurs in the vicinity of the cross-over frequency of the loop gain, the closed-loop output impedance has to be determined from Eq. (9). The closed-loop output impedance can be greater or smaller than its open-loop counterpart. The situation is far more complicated than the previous two, and it is difficult to draw any general conclusions. Later, an example will be given to demonstrate a case in which the closed-loop output impedance is greater than the open-loop output impedance.

Figure 6 shows the closed-loop output impedances. The compensator consists of an integrator, two poles and two zeros:

$$A_{c1}(s) = \frac{K_1}{s}\frac{(s+s_{zc1})(s+s_{zc2})}{(s+s_{pc1})(s+s_{pc2})}. \tag{14}$$

The design of the compensator will be elaborated on in the next section. The closed-loop output impedances are reduced compared with their open-loop counterparts. Unlike the case of the single-output counterpart, where the output impedance can be nullified at dc by using an integrator in the loop gain, the output impedances of the dual-output converter always have finite values at dc, even when an integrator is applied in the compensator.

The same argument can be applied to the output transimpedances and audio susceptibilities, and similar results can be obtained. It is interesting to notice that the closed-loop audio susceptibilities, as shown in Fig. 7, can be nullified at dc, since the numerator of each audio susceptibility is the difference of two terms, *i.e.*,

$$G_{v1(CL)} = \frac{G_{v1}T_2 - \frac{K_2}{K_1}T_1G_{v2}}{1+T} = \frac{K_2(DN_1N_2V_{in} - DN_1N_2V_{in})}{(K_1N_1+K_2N_2)V_{in}} = 0, \tag{15}$$

which is consistent with the fact that the dual-output converter is very robust to line voltage disturbances. Again, the drastic reduction of the closed-loop audio susceptibility is not the direct effect of high loop gain, but rather the effect of weighted-voltage control. Referring to the small-signal

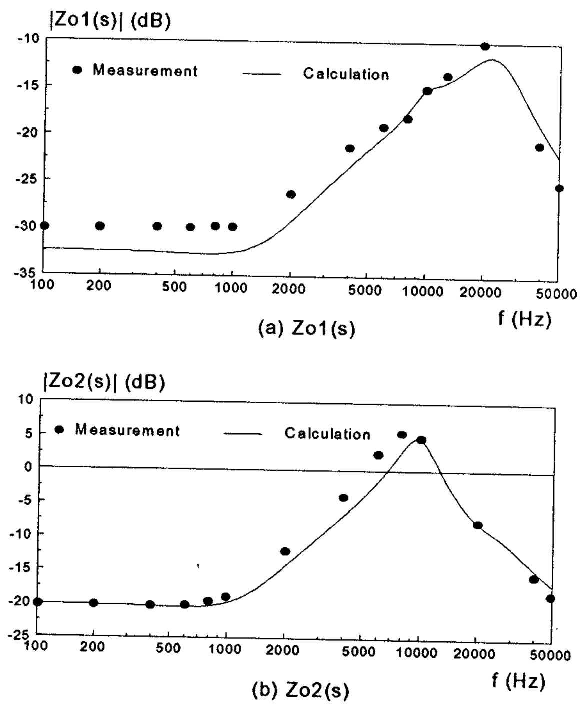

Fig. 6. Closed-loop output impedances. *The closed-loop output impedances are reduced by designing the compensator properly. Unlike the single output converter, the closed-loop output impedances have finite values at dc.*

block diagram as given in Fig. 2, it can be observed that the perturbation from the line voltage, v_{in}, is injected into the system through two paths: one through the open-loop audio susceptibility of power channel one, $G_{v1}(s)$, and the other through the open-loop audio susceptibility of power channel two, $G_{v2}(s)$. The net effects tend to cancel each other. As a result, the closed-loop audio susceptibility is drastically decreased. The conclusion made here is different from that reported in [5], where the reduction of the closed-loop audio susceptibility has been attributed to the high loop gain of the compensator. Actually, only one audio propagation path was accounted for in [5].

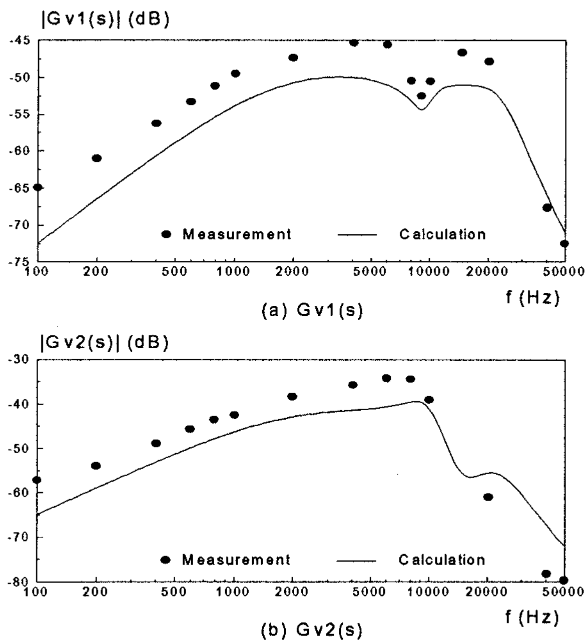

*Fig. 7. **Closed-loop audio susceptibilities.** Reduction of the audio susceptibilities is more significant compared with the closed-loop output impedances. This explains why the dual-output converter is more robust against line disturbance than against load disturbances.*

V. Design Considerations

From the above analyses, it can be concluded that the compensator should be designed such that the cross-over frequency be as high as possible and the loop gain be high over the frequency range of interest to improve the line and load regulations, provided that the compensated system has enough stability margin.

To support the above conclusions, another compensator with an integrator only is applied:

$$A_{c2}(s) = \frac{K_I}{s}. \qquad (16)$$

The loop gains for both cases are plotted in Fig. 8. It can be seen that using an integrator results in a much lower cross-over frequency and a much lower loop gain compared with using the compensator with two zeros and three poles, Eq. (14). Figure 9 shows the open-loop output impedances and closed-loop output impedances employing both types of compensators. The comparison clearly shows that the higher loop gain results in a smaller output impedance. For the lower loop gain, which corresponds to a lower cross-over frequency, the closed-loop output impedances have even higher peaking than the open-loop output impedances. This is because the value of the loop gain around the resonant frequency of the first pair of power stage complex poles is close to unity, but the phase delay is close to -180°. The magnitude of the denominator in Eq. (9) is very small, resulting in higher closed-loop output impedances. Figure 10 shows the large signal simulation of the output responses to a step load current change for two different types of compensators. Again, the comparison is favorable to the compensator which yields higher cross-over frequency and higher loop gain. The compensator design is based on the system duty cycle-to-output (-feedback) transfer functions which can take different forms: (1) those with interlaced complex poles and zeros, and (2) those with non-interlaced poles and zeros. As discussed in [4], for the systems with non-interlaced complex poles and zeros, it is difficult to achieve high gain and wide bandwidth. Therefore, it is preferable to design the duty cycle-to-feedback transfer function such that the complex poles and zeros are interlaced. The pole-zero interlacing condition has been derived as following:

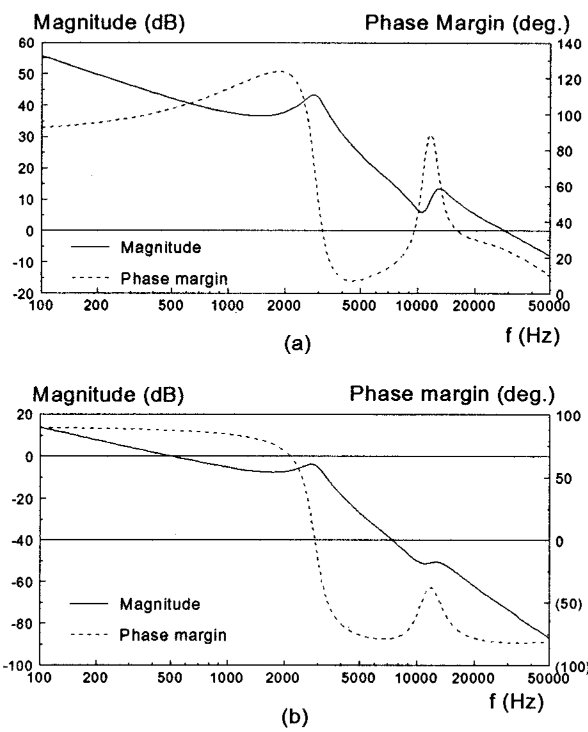

*Fig. 8. **Loop gain using different compensators.** The compensator of an integrator plus 2 poles and 2 zeros (a) results in a higher crossover and a higher loop gain than the compensator of an integrator (b).*

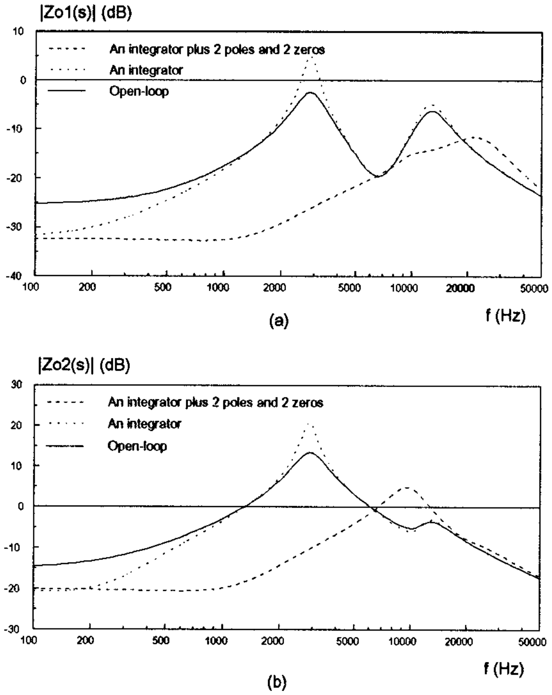

Fig. 9. Comparison of output impedances using different compensators. *Three curves shown here are open-loop output impedances and closed-loop output impedances with different types of compensator. (a) Output 1; (b) Output 2.*

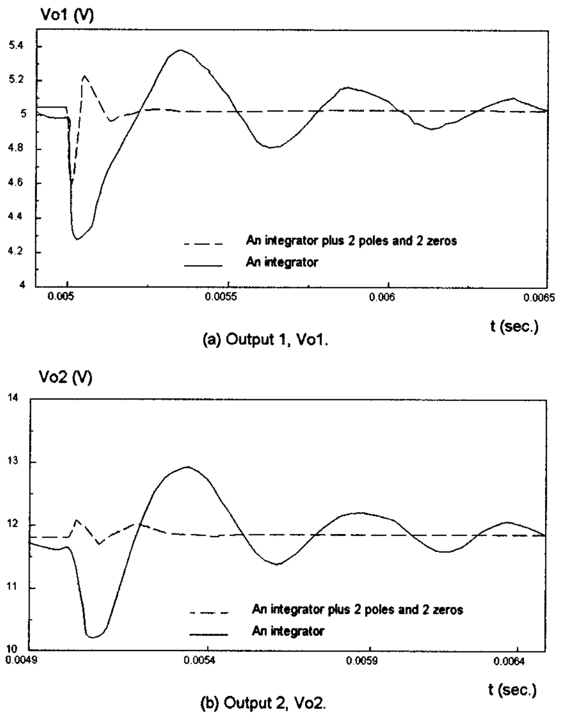

Fig. 10. Transient responses to a step load change at output 1. *The higher loop gain yields smaller output impedances, and consequently faster response with smaller over/under shoot.*

$$k \le \frac{K_1N_1L_2C_2 + K_2N_2L_1C_1}{(K_1N_1 + K_2N_2)L_2\left(\dfrac{C_1C_2}{C_1 + C_2/n_{12}^2}\right)} - 1. \tag{17}$$

Figure 11 shows the right side of Eq. (17) as a function of the ratio of the weighting factors, $u=K_1/K_2$. It can be seen that a large coupling coefficient tends to yield a noninterlaced pole-zero distribution for the given weighting factors, and the system tends to have less stability margin for the same compensator design. It is noticed that once the right side of Eq. (17) exceeds unity, the coupling coefficient, k, no longer affects relative positions of the poles and zeros since k is physically bound to the range (0,1). If the coupling coefficient is chosen outside the shadowed area, the duty cycle-to-output transfer function will always have interlaced complex poles and zeros.

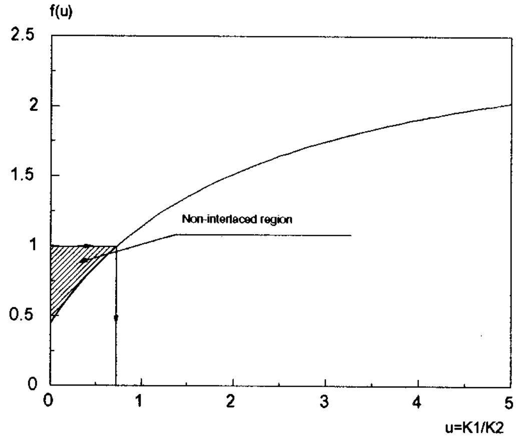

Fig. 11. Design criterion for coupling coefficient, k, to ensure a pole-zero interlaced system. *The vertical axis f(u) is the right side of Eq. (17), which should be greater than the coupling coefficient, k. It is noticed that the coupling coefficient is in the range: $0<k<1$. Once $f(u)>1$, k can take any value between (0, 1) and the system still has interlaced poles and zeros.*

For a converter whose duty cycle-to-feedback transfer function has interlaced complex poles and zeros, the system behavior at both low and high frequencies is similar to that of the single output converter. The commonly-used compensator with an integrator plus two poles and two zeros as shown in Eq. (14) is suggested to be employed. First, the loop gain crossover frequency, f_c, is set according to the resonant frequencies of the duty cycle-to-output (-feedback) transfer function, and the given switching frequency, f_s. Since the peaking of the open-loop audio susceptibilities and output impedances occur at the resonant frequencies of the complex poles, the cross-over frequency, f_c, should be chosen higher than these resonant frequencies. However, the cross-over frequency cannot be raised arbitrarily; it is also limited by the switching frequency. It is suggested that the cross-over frequency be chosen so as not to exceed $f_c=f_s/5$ - $f_s/3$.

The design criterion for the compensator poles and zeros is to make sure that the closed-loop system has high loop gain before the cross-over frequency while maintaining enough stability margin. One compensator pole is placed to cancel the equivalent ESR zero. The other one is placed at approximately one half of the switching frequency to attenuate the switching noise in the modulator. To achieve a high loop gain at the low frequency range, a pole is placed at the origin. The amplitude of the loop gain and stability margin are determined by the placement of the compensator zeros. The first compensator zero is placed slightly below the low frequency complex pole $s_{zc1} \leq \omega_{pt1}$. The position of the second compensator zero determines loop gain over the mid-frequency range and is suggested to be placed as follows:

$$\omega_{pt2} \leq s_{zc2} \leq \omega_c . \tag{18}$$

Furthermore, the second compensator zero, s_{zc2}, should be placed close to the second pair of the complex poles of the duty cycle-to-feedback transfer function to ensure enough phase margin.

A potential problem for this design is that if the first pair of complex poles, ω_{pt1}, and the complex zeros, ω_{zt}, are widely separated, the phase can approach -180°. Under this scenario, the second zero should be chosen at a relatively low frequency to ensure the stability of the system.

Figure 12 shows the measurement and prediction of the loop gain for a pole-zero interlaced system. There is good correlation between the measured and predicted results.

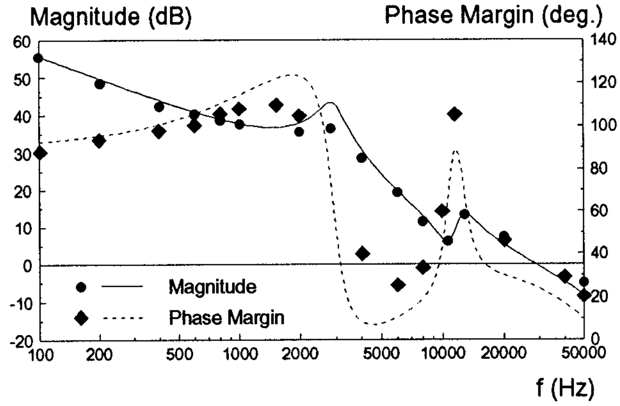

Fig. 12. Loop gain of the dual-output forward converter with WVC. The operation condition is high line. It is necessary to check the stability margin at the lowest line voltage for which the feedback controller starts to regulate the output voltages.

VI. Summary

The dynamic performance of a multiple-output converter are characterized by the audio susceptibilities and output impedances as in the case of its single-output counterpart. The unique quantities for the multiple-output converter are the output transimpedances which describe the output voltage variation caused by the other load current. The analytical expressions for the open- and closed-loop audio susceptibilities, the output impedances, and the output transimpedances contain the compensator transfer function, $A_c(s)$, in both the numerator and denominator, which could mislead one to conclude that the high loop gain does not improve dynamic performance. Actually, the high loop gain does improve the closed-loop performances, as in a single-output converter. The mechanism of decreasing the audio susceptibilities, output impedances are different. Unlike the single-output converter, in which high loop gain knocks down the audio and the output impedance directly, the decreasing of the audio susceptibilities and output impedances is achieved by cancellation of two numerator terms through which the disturbances are injected into the system. Since the numerators of the audio susceptibilities are the differences between two terms, it is possible to nullify the audio susceptibilities at a specific frequency.

The closed-loop dynamic characteristics of a MOC is closely related to the relative positions of the complex poles and zeros of the power stage duty cycle-to-feedback transfer function. The pole-zero interlacing condition is derived, and can be used as a design criterion. A pole-zero interlaced system with a compensator consisting of two zeros and three poles can usually yield a loop gain with a high crossover frequency and high gain, which results in smaller audio susceptibilities and output impedances.

References

[1] K. Harada, T. Nabeshima and K. Hisanaga, "State-Space Analysis of the Cross-Regulation," IEEE Power Electronics Specialists Conf. Rec., 1979, pp. 186-192.

[2] S. S. Kelkar and J. K. Radcliffe, "Dynamic and Static Cross Regulation in Forward Converters," *Proc. of 8th International Telecommunications Energy Conf.*, 1986, pp. 219-227.

[3] C. C. Liu, K. H. Ding, J. R. Young, and J. F. Beutler, "A Systematic Method for the Stability Analysis of Multiple-Output Converters," *IEEE Trans. Power Electronics*, vol. 2, no. 4, pp. 343-353, 1987.

[4] Q. Chen, F. C. Lee, and M. M. Jovanovic, "Small-Signal Analysis and Design of Weighted Voltage Control for a Multiple-Output Forward Converter," *IEEE Power Electronics Specialists Conf. Rec.*, 1993, pp. 749-756.

[5] Y. T. Chen, D. Y. Chen, and Y. P. Wu, "Control-Loop Modeling of Multiple-Output Feedback of Forward Converters," *IEEE Trans. Power Electronics*, vol. 8, no. 3, pp. 320-328, 1993.