

Control Topology Options for Single-Phase UPS Inverters

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Abstract—Four control topologies for single-phase uninterruptible power system (UPS) inverters are presented and compared, with the common objective of providing a dynamically stiff, low total harmonic distortion (THD), sinusoidal output voltage. Full-state feedback, full-state command controllers are shown, utilizing both filter inductor current and filter capacitor current feedback to augment output voltage control. All controllers presented include output voltage decoupling in a manner analogous to “back-electromotive force (EMF)” decoupling in dc motor drives. Disturbance input decoupling of the load current and its derivative is presented. An observer-based controller is additionally considered and is shown to be a technically viable, economically attractive option. The accuracy transfer function of the observer estimate is used to evaluate its measurement performance. Comparative disturbance rejection is evaluated by overlaying the dynamic stiffness (inverse of output impedance) frequency response of each controller on a single plot. Experimental results for one controller are presented.

Index Terms— Control systems, dc-ac power conversion, impedance, inverters, observers, state space methods, state estimation, transient analysis, voltage control.

I. INTRODUCTION

IN all uninterruptible power system (UPS)-style inverters the goal is to maintain the desired output voltage waveform over all loading conditions and transients. In the past, sine-wave inverters relied on open-loop feedforward control to produce the shape of the waveform, while a relatively slow output voltage rms feedback loop regulated the magnitude. While these types of controllers could maintain a desired steady-state rms output voltage, their response to step changes in load were noticeably slow (several cycles of the output waveform), and nonlinear loads could greatly distort their output voltage waveform. Today, various modern feedback control techniques are available to control the output voltage waveform continuously, rather than on an rms basis. These so-called “instantaneous” controllers offer many performance advantages including faster (sub-cycle) transient response, better total harmonic distortion (THD), and improved disturbance rejection via lower output impedance.

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Many instantaneous controllers have been presented in the literature that actively control the inverter’s output over the entire waveform. Digital controllers incorporating various forms of state feedback have shown good performance, but at the cost of a relatively fast μ -processor which must compute inverter duty cycle on a pulse-by-pulse basis [3], [6], [10]. Several hysteresis-type controllers have been presented in [9] and [12]. These controllers can suffer from relatively high and variable switching frequencies. Analog-based controllers utilizing inductor current feedback are found in [4], [7], and [11], while capacitor current feedback topologies are found in [1], [2], [5], [8], and [10]. The technique of dc bus voltage decoupling has been shown in [1] and [8], and the advantages of output voltage, or “back-electromotive force (EMF)” decoupling have been presented in [1]. The importance of the inverter’s closed-loop output impedance characteristic has been recognized in [1], [4], and [6].

This paper will present several state-space control topology options for a single-phase PWM inverter with an LC output filter. Two basic feedback topologies are explored: 1) filter inductor and load current sensing and 2) filter capacitor current sensing, where both approaches use a full-state command structure. For the case of inductor current feedback, two methods of load current decoupling will be considered. In the case of capacitor current feedback, a Luenberger-style observer for capacitor current will also be considered in lieu of a current sensor. All controllers presented employ active decoupling of both the dc bus and the “back-EMF” of the output voltage. The output dynamic stiffness (inverse of output impedance) of each controller is evaluated and compared on a single plot. Experimental results are presented for the capacitor current feedback controller topology.

II. SINEWAVE INVERTER COMPONENTS AND STATE-SPACE MODEL

Sine-wave inverters are typified by the components shown in Fig. 1. At the heart of the system is a full-bridge inverter which modulates a dc bus into a cycle-by-cycle average output voltage, e_i .

The amplitude of e_i is directly proportional to the commanded duty cycle of the inverter, which varies over $\pm 100\%$, and the amplitude of the dc bus voltage, V_{dc} . Thus, e_i can range from $+V_{dc}$ to $-V_{dc}$.

The output of the inverter is in turn passed through a second-order LC filter to block all but the desired fundamental

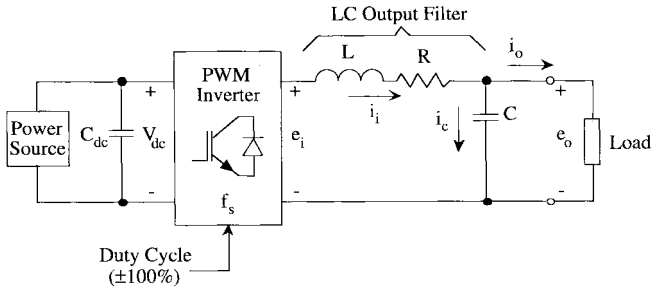


Fig. 1. Single-phase UPS inverter system.

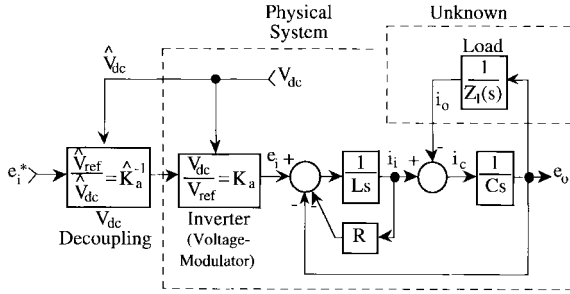


Fig. 2. State-space model with dc bus voltage decoupling.

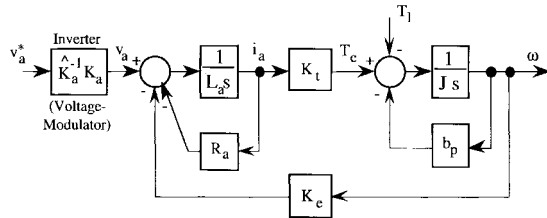


Fig. 3. Analogous state-space dc motor system diagram.

frequency (50/60/400 Hz). The resistance of the output filter inductor is represented by R . Filter capacitor ESR is ignored since the break frequency (typically above 200 kHz) appears far above the range of concern. The load shown in Fig. 1 can be any type of ac load: resistive, inductive, capacitive, or nonlinear.

The power source shown in Fig. 1 can be of variable voltage (such as a battery) which will tend to “droop” under heavy loads. Fig. 2 shows a state-space model of the system with V_{dc} decoupling to compensate for changes in the bus voltage; permitting e_i to be commanded directly. The carets, $\hat{\cdot}$, denote estimated values such as \hat{e}_o .

Because the switching frequency, f_s , of the inverter is usually several orders of magnitude above the fundamental, the dynamics of the inverter are usually ignored. The V_{dc} compensated inverter is thus depicted as a simple unity gain block in all subsequent diagrams.

In studying the physical system depicted in Fig. 2, it was noted that the LC filter was directly analogous to a dc motor model, as shown in Fig. 3, where $K_e = K_t = 1$, and $b_p = 0$. The filter capacitance is analogous to rotor inertia. Thus, all of the advanced dc motor control techniques previously developed can be applied directly to the control of the inverter and LC filter.

III. FILTER INDUCTOR CURRENT REGULATOR

If inductor current is controlled, it becomes possible to implement various forms of disturbance input decoupling (which is also known as disturbance feedforward control).

A. Load Current Decoupling

If it is economically feasible to measure the load current, then it can readily be decoupled. Fig. 4 shows a full-state feedback controller with load current (disturbance input) decoupling, where time derivatives are indicated by “ \bullet ” and commanded values are indicated by “ $*$,” such as \hat{e}_o^* . The controller is shown in a cascade format, with an inner inductor current loop and an outer output voltage loop. The controller additionally uses full-state command feedforward and state feedback decoupling of the output voltage and filter inductor IR drop. Note that with the state feedback decoupling terms, the remaining control inputs command voltage across the filter inductor, e_L^* , directly. Nulling the state coupling in this manner allows a simple proportional gain, R_a , to be used in forming the inductor current loop.

The disturbance input decoupling is implemented by sensing the load current, i_o , and using it as an additional current loop command to produce the needed load current without waiting for errors in voltage to occur. This leaves only the capacitor current to be commanded, $i_c^* = \hat{e}_o^* \hat{C}$, which is independent of load current. Thus, load transients can be effectively rejected up to the bandwidth of the inductor current loop. This bandwidth (set by R_a) can be as high as $1/5$ – $1/4$ of f_s . If the bandwidth were infinite, the disturbance input decoupling would be perfect, and the dynamic stiffness (defined in Section V) would be infinite.

As shown in Fig. 4, \hat{e}_o^* is used to give a full command vector, though in practice (for a 50- or 60-Hz command) the relative magnitudes of L and C will render e_{i-cff} to approximately $1/1000$ th of the total e_i^* command. Thus, the majority of e_i^* is actually determined by the e_o decoupling state feedback. This may be explained by recognizing that the fundamental voltage drop across L is quite small when compared to e_o . From Fig. 4, the command response transfer function is found as

$$\frac{e_o}{e_o^*} = \frac{\hat{C}\hat{L}s^3 + \hat{C}R_a s^2 + R_a K_v s + R_a K_i}{CLs^3 + CR_a s^2 + R_a K_v s + R_a K_i} \quad (1)$$

If the estimated parameters \hat{L} , \hat{R} , and \hat{C} are accurate, the controller will exhibit perfect command tracking up to the bandwidth limit of the voltage modulator. Since f_s is usually several orders of magnitude above the fundamental, this does not pose a limit. From Fig. 4, the output dynamic stiffness is found as

$$\frac{i_o}{e_o} = \frac{CLs^3 + CR_a s^2 + R_a K_v s + R_a K_i}{Ls^2} \quad (2)$$

This is plotted in Fig. 8, along with the dynamic stiffness of the other controllers for comparison. Table II lists the controller gains and the eigenvalues used for all the controller alternatives.

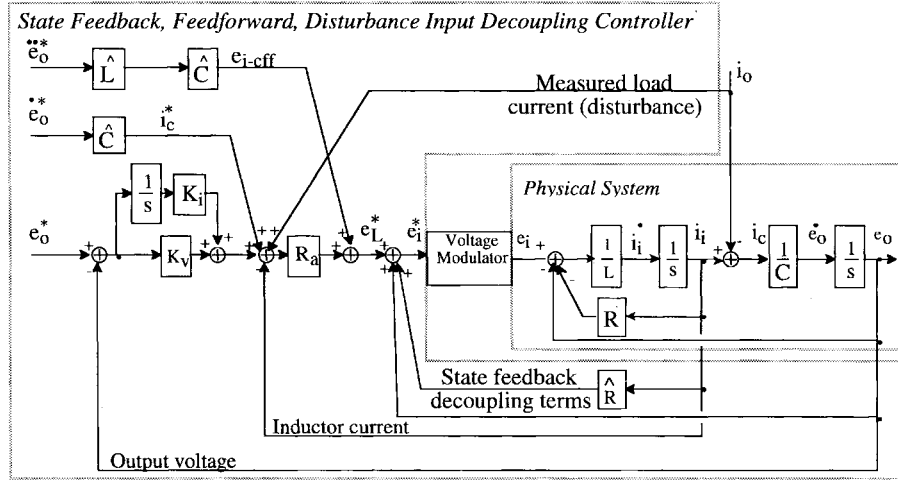


Fig. 4. State feedback controller with state command feedforward, decoupling state feedback, and disturbance input decoupling via load current sensing.

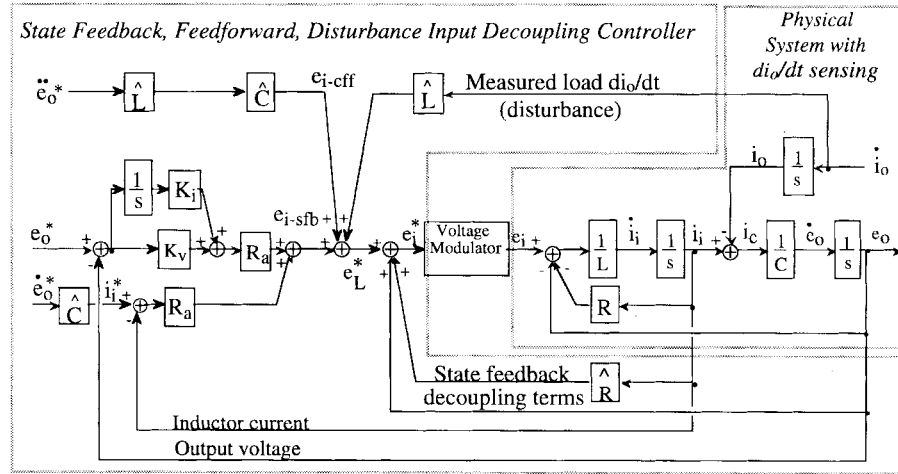


Fig. 5. State feedback controller with state command feedforward, decoupling state feedback, and disturbance input decoupling via load current di_o/dt sensing.

B. Load Current Decoupling With di_o/dt Feedback

An alternative approach to disturbance input decoupling which further improves the dynamic stiffness uses the derivative of the output current, $di_o/dt = \dot{i}_o$, which can be sensed with a small choke and fed back, as shown in Fig. 5. In conjunction with the output voltage and inductor resistance decoupling, the di_o/dt term can now be used to fully decouple load-induced voltage transients across the output filter inductance. Thus, the system will now exhibit infinite dynamic stiffness up to the bandwidth of the voltage modulator, provided that the estimate of inductance \hat{L} is accurate; it should be noted that \hat{L} may be mapped as a nonlinear function of i_o to improve the parameter estimate.

This increased dynamic stiffness, and the likelihood that sensing di_o/dt with a small choke will be less expensive than a full current sensor for i_o , makes this an attractive control topology. As such, the topology in Fig. 5 is considered the “upper bound” of the controllers represented in this paper.

IV. FILTER CAPACITOR CURRENT REGULATOR

If capacitor current is controlled, dynamic stiffness can be improved substantially. The key issue for capacitor current is how the sensing is performed, i.e., either via direct measurement or via an observer.

A. Capacitor Current Sensed

As an alternative to sensing inductor current and load current (or load di_o/dt), the filter capacitor current, i_c , may be measured and used in a state feedback controller as shown in Fig. 6. It is especially relevant because the derivative of the output voltage, \dot{e}_o , is proportional to i_c . Because i_c is small and ac in nature, it may be sensed with a small inexpensive current transformer. From a disturbance rejection point of view, capacitor current feedback directly senses changes in load current, as the capacitor current is the sum of inductor and load currents. Thus, without some form of disturbance input decoupling, as shown in Figs. 4 and 5, a capacitor current feedback topology will exhibit better dynamic stiffness than that of a controller with inductor current feedback alone.

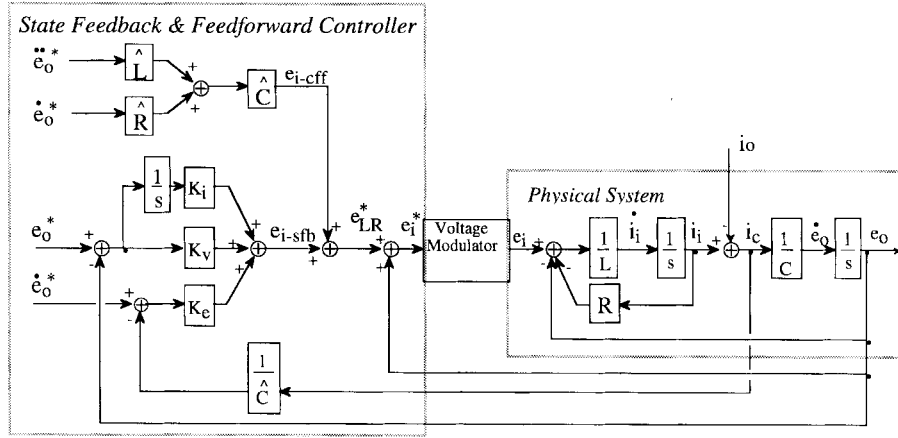


Fig. 6. State feedback controller based on capacitor current sensing with state command feedforward and output voltage decoupling state feedback.

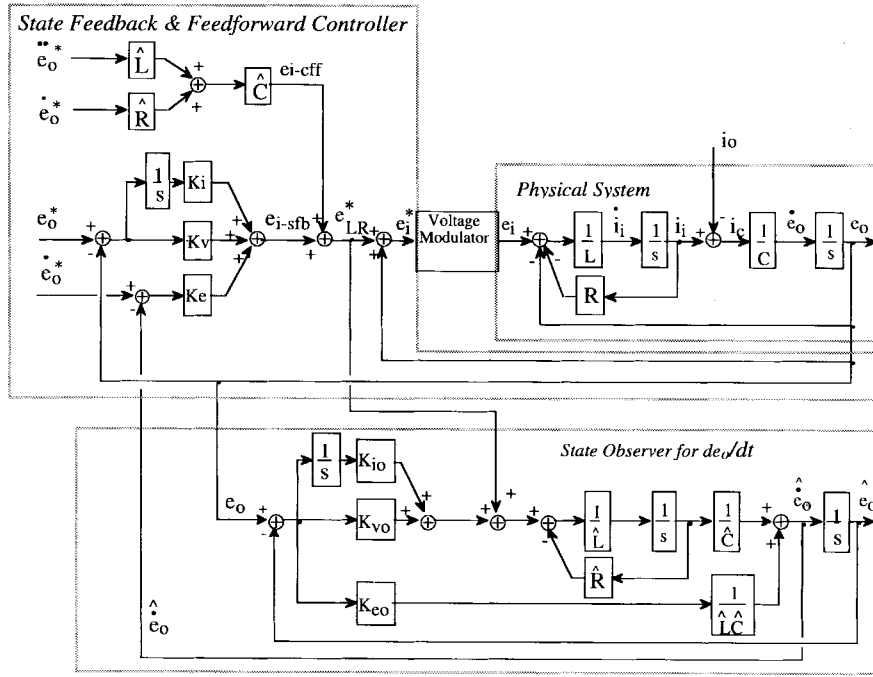


Fig. 7. Observer-based capacitor current feedback state-space controller.

The command response and dynamic stiffness transfer functions can be found as

$$\frac{e_o}{e_o^*} = \frac{\hat{C}\hat{L}s^3 + (\hat{C}\hat{R} + \hat{K}_e)s^2 + K_v s + K_i}{CLs^3 + (CR + K_e)s^2 + K_v s + K_i} \quad (3)$$

and

$$\frac{i_o}{e_o} = \frac{CLs^3 + (CR + K_e)s^2 + K_v s + K_i}{Ls^2 + Rs} \quad (4)$$

respectively.

Because i_i is not measured, the inductor resistance is not decoupled through state feedback, but rather through the feedforward control path. Because of this, the resistance affects the dynamic stiffness, in the form of a low break frequency, R/L . Again, the controller has perfect command tracking for all frequencies if the parameter estimates are correct. While the inductor resistance and nominal inductance may change significantly over temperature and loading, respectively, the

capacitor value is usually quite stable. Fig. 8 depicts the dynamic stiffness frequency response and Table II lists the controller gains and eigenvalues used for all of the controller alternatives.

B. Capacitor Current Feedback via Observer

If only output voltage e_o is measured, the derivative term, \dot{e}_o , may be estimated by an enhanced Luenberger observer. The Luenberger observer, introduced in [13] with design examples in [14]–[16], makes use of all available manipulated inputs as command feedforward information to allow the observer to track commanded inputs with the same response as the physical system. It also actively controls the error in its estimate of the measured physical state e_o and so forces convergence of the estimated state, \hat{e}_o , to the actual state value. An integration state is added to the observer controller to force

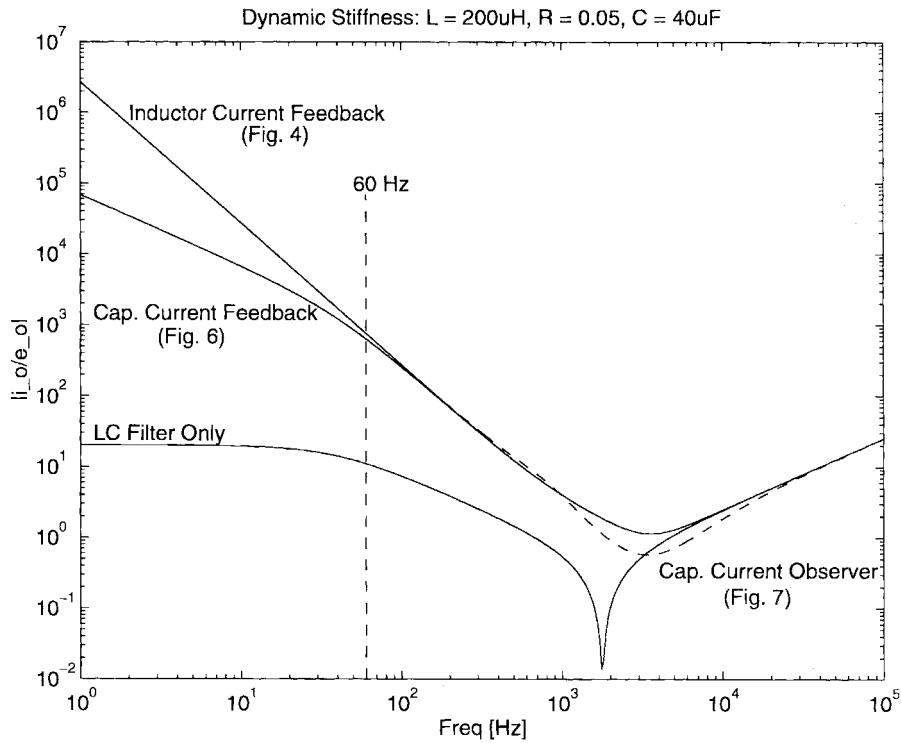


Fig. 8. Dynamic stiffness of controllers.

 TABLE I
 IGBT INVERTER PARAMETERS

Parameter	Value	Units
Switching Frequency, f_s	20	kHz
Nominal DC Bus Voltage	275	V
Rated Output Voltage	120	V _{rms}
Rated Output Frequency	60	Hz
Rated Output Power	8	kVA
Rated Output Current	67	A _{rms}
Filter Inductor, L (Nom.)	200	μ H
Inductor Resistance, R	0.050	Ω
Filter Capacitor, C	40	μ F

zero steady-state errors. Fig. 7 shows the complete controller topology, including the observer. With the exception of the observer, the controller is identical to that of Fig. 6, and the same controller gains K_i , K_v , and K_e are used.

The dynamic performance and accuracy of the observer estimate is best evaluated by viewing the observer as a transducer alternative and evaluating its frequency response characteristic. This is given by the following equation, and will be unity at all frequencies only if the parameter estimates, \hat{L} , \hat{R} , and \hat{C} , are exactly correct:

$$\frac{\hat{e}_o}{e_o} = \frac{\hat{e}_o}{e_{LR}^*} \frac{e_{LR}^*}{e_o} \quad (5)$$

or as shown in (6), at the bottom of the page.

The observer characteristic polynomial is the denominator in this equation.

Since robustness is not a concern with observers, the observer gains are chosen based on nominal parameter values to make all eigenvalues equal, corresponding to an observer bandwidth of 2.0 kHz, or $0.1f_s$; this bandwidth limits the observer response to switching noise, while still providing adequate overall dynamic stiffness. The resulting gains are summarized in Table II. The effects of parameter variation may be minimized by setting the observer bandwidth as high as possible, although switching frequency components will be corresponding larger in the estimated output voltage derivative.

The observer-based controller dynamic stiffness response function is

$$\frac{i_o}{e_o} = \frac{CLs^3 + (CR + G_{obs}K_e)s^2 + K_v s + K_i}{Ls^2 + Rs} \quad (7)$$

where

$$G_{obs} = \frac{\hat{e}_o}{i_o} \frac{i_o}{e_o} \quad (8)$$

The resulting dynamic stiffness transfer function has a sixth-order numerator and a fifth-order denominator.

Fig. 10 depicts the system response with an 8 kW step load. Note that the output voltage sinewave recovers almost instantly; the transient cannot be noticed on an ordinary 60 W lightbulb on the same circuit.

$$\frac{\hat{e}_o}{e_o} = \frac{CL\hat{L}s^3 + (C\hat{L}R + \hat{L}K_{e_o})s^2 + (\hat{R}K_{e_o} + \hat{L}K_{v_o})s + \hat{L}K_{i_o}}{\hat{C}\hat{L}^2s^3 + (\hat{C}\hat{L}\hat{R} + \hat{L}K_{e_o})s^2 + (\hat{R}K_{e_o} + \hat{L}K_{v_o})s + \hat{L}K_{i_o}} \quad (6)$$

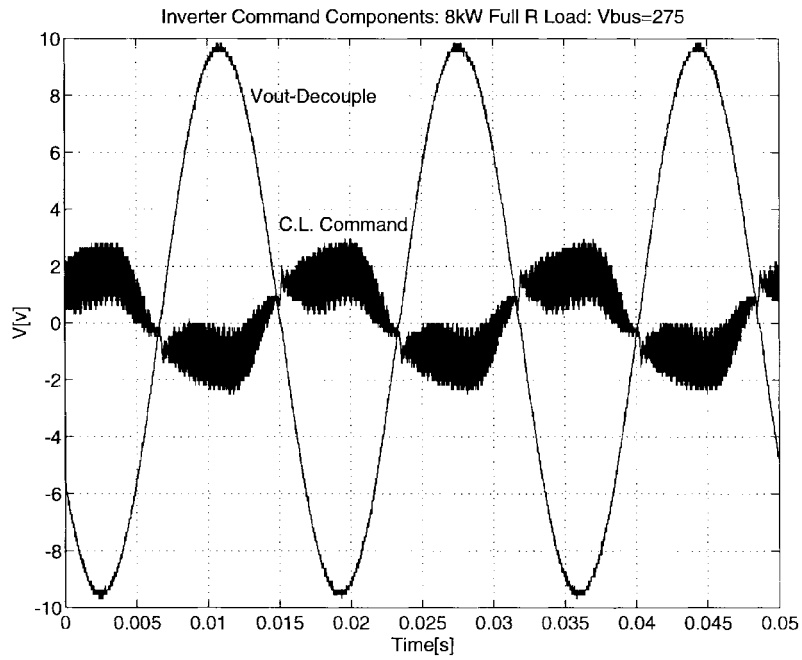
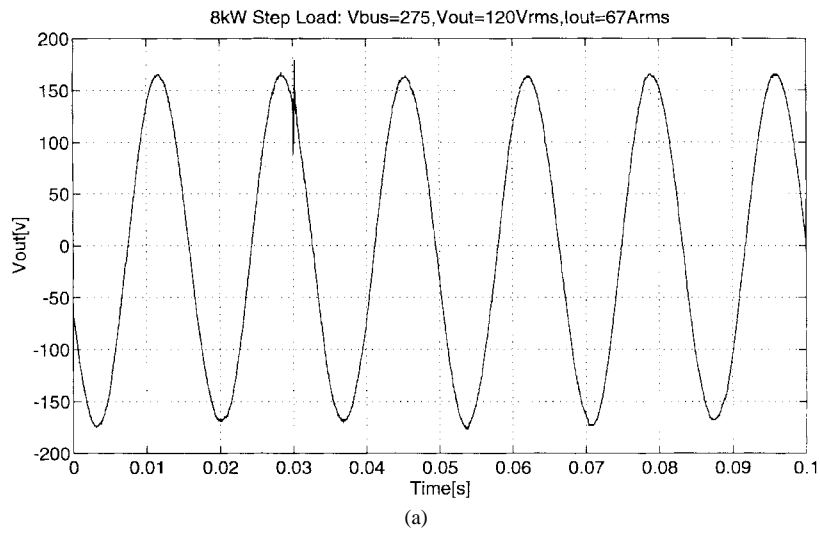
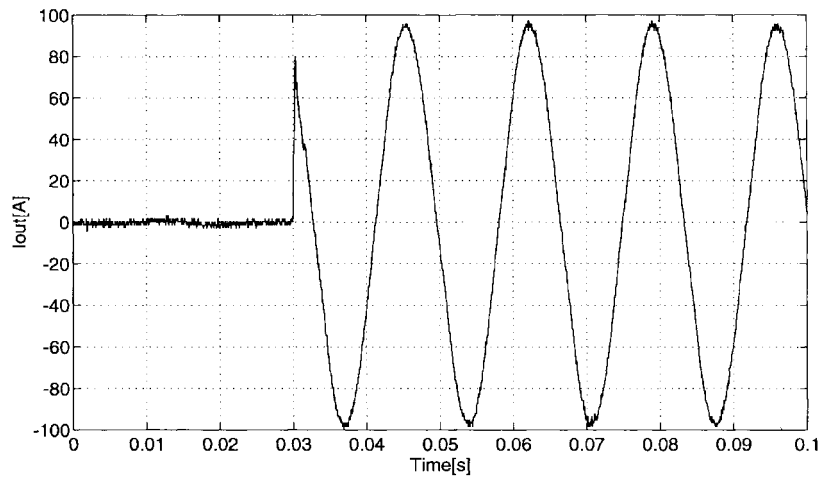


Fig. 9. Inverter command components under full load (8 kW).



(a)



(b)

Fig. 10. 8-kW step load response. (a) Output voltage. (b) Output current.

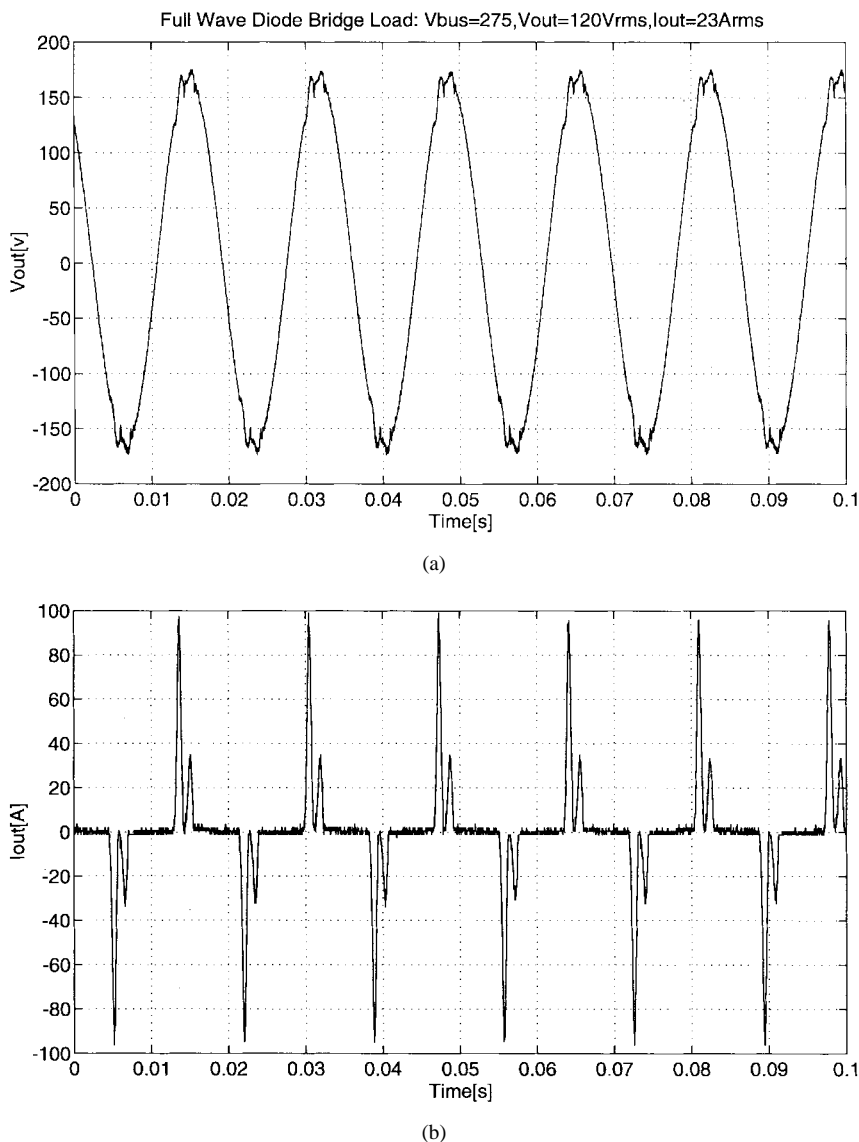


Fig. 11. FW diode bridge load. (a) Output voltage. (b) Output current.

V. DISTURBANCE REJECTION COMPARISON VIA DYNAMIC STIFFNESS

The dynamic stiffness of a UPS system is defined as the magnitude of output load current that causes a unit deviation in output voltage magnitude: $|I_o(s)/E_o(s)|_{s=j\omega}$. Fig. 8 shows the dynamic stiffness frequency response of the controllers considered earlier. For comparison purposes, all the controllers were evaluated with eigenvalues that corresponded to those implemented in the laboratory. The controller with di_o/dt feedback, Fig. 5, is not included in Fig. 8, as it is infinitely stiff if parameter estimates are correct. With errors in the parameter estimates, the controller stiffness maintains the overall shape of the inductor current feedback curve, but is shifted up by several orders of magnitude.

The disturbance input decoupled controller with i_o sensing (Fig. 4) provides superior disturbance rejection compared to the capacitor current controller (Fig. 6) over the low frequency range, but provides negligible improvement for

load-current frequency components above 60 Hz, which predominate in most UPS applications. Note that the equivalent 60-Hz output impedance of the controllers is $\approx 2 \text{ m}\Omega$, as compared to the LC filter impedance of $\approx 100 \text{ m}\Omega$.

Over most of the frequency range, the observer-based controller (Fig. 7) provides the same stiffness as the controller with measured capacitor current, but the former suffers decreased performance above the bandwidth of the observer, especially if the L and C parameters vary.

VI. EXPERIMENTAL RESULTS

The standard capacitor current feedback controller depicted in Fig. 6 (less e_{i-cff}) was implemented in the laboratory. Table I lists the specifications of the insulated gate bipolar transistor (IGBT) full-bridge inverter used in the tests. Table II lists the eigenvalues and controller gains for all controllers.

TABLE II
CONTROLLER GAINS AND BANDWIDTHS

Parameter	Value	Units
Eigenvalues (All Controllers)	890, 2200 ± j2710	Hz
Inductor Current FB: (Fig. 4)		
R_a	6.65	Ω
K_v	0.765	Ω^{-1}
K_i	3.23e3	Ω^{-1}/s
Capacitor Current FB: (Figs. 6, 7)		
K_e	2.64e-4	s
K_v	5.09	-
K_i	2150	s^{-1}
Observer Eigenvalues	2000	Hz
Cap. Current Observer: (Fig. 7)		
K_{e0}	3.10e-4	s
K_{v0}	3.72	-
K_{i0}	1.59e4	s^{-1}

Fig. 9 depicts the relative magnitude of the components making up the inverter command e_i^* (refer to Fig. 6). Note that the output voltage decoupling makes up 80%–90% of the total inverter command; since the filter inductor impedance is relatively small ($X_L \approx 75 \text{ m}\Omega @ 60 \text{ Hz}$), the fundamental voltage drop across it is also small. With most of the inverter command being created from the output voltage (i.e., “back-EMF”) decoupling, the closed-loop controller gains can be lower and more robust, while still providing excellent disturbance rejection.

To simulate a worst-case loading, a full-wave diode bridge rectifier load was tested. The output of the diode bridge was connected directly to a 1000- μF capacitor, with resistive load of $\approx 18 \Omega$. Fig. 11 depicts the results. Note that although the current is drawn in 100-A spikes, the voltage waveform is only slightly distorted. Note additionally that with relatively stiff voltage sources on either side of the diodes, small oscillations in the voltage waveform produce significant current transients. Table III lists the THD of the output voltage under various load conditions.

VII. CONCLUSIONS

Several UPS inverter control topologies have been presented and evaluated by comparing their dynamic stiffness characteristics. Both filter inductor and filter capacitor current feedback control topologies with full-state command structures have been examined. Various options for physical state feedback decoupling have been presented, each including a “back-EMF” decoupling of the output voltage. It has been shown that a controller utilizing load-current derivative feedback can exhibit infinite dynamic stiffness up the dynamic limits of the voltage modulator.

As a low-cost alternative, the filter capacitor current feedback controller exhibits outstanding performance, as can be seen in the experimental results, less than 0.5% THD with a single-phase full load of 8 kW. Voltage distortion following

TABLE III
OUTPUT VOLTAGE THD PERFORMANCE

Load Condition	i_o [A _{rms}]	i_o Peak [A]	i_o THD [%]	v_o THD [%]
No Load	0	0	-	0.35
Full Rated Resistive-Load	67	95	-	0.44
Full-Wave Diode Bridge with C and R-load	23	100	160	2.77
FWDB Load in parallel with R-Load	59	150	32	1.75

full power load transients cannot be observed in the intensity output of an average incandescent bulb sharing the same circuit. An observer-based version of this controller is predicted to perform nearly as well, with a further reduction of sensor cost.

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