ONE-CYCLE CONTROL OF SWITCHING CONVERTERS

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Abstract

A new nonlinear control technique is conceived to control the duty-ratio d of the switch in real time such that in *each cycle* the average value of the chopped waveform at the switch rectifier output diode is *exactly* equal to the control reference. Experiments demonstrate that switching converters with this new One-Cycle Control reject input-voltage perturbations in only one switching cycle and follow the control reference very quickly. This new control method is very general and directly applicable to all dc-to-dc switching converters in either pulse-width-modulated or quasi-resonant modes.

1 Introduction

Switching converters are pulsed and nonlinear dynamic systems. There has been no standard way to model and control nonlinear systems. At present, most control schemes are approached by first linearizing the governing equations and then applying a linear feedback technique. This approach greatly restricts the capability of switching nonlinear systems.

The objective of this work is to achieve large-signal nonlinear control of switching converters. The motivation is that pulsed nonlinear systems under pulsed nonlinear control should be more robust, have faster dynamic response, and better input-perturbation rejection than the same system under linear control. A new pulsed nonlinear control technique, One-Cycle Control, is introduced in this paper. This technique takes advantage of the pulsed and nonlinear nature of switching converters and achieves instantaneous control of the average value of the chopped voltage or current. This technique provides fast dynamic response and good input-perturbation rejection. It is suitable for the control of pulse-width-modulated (PWM) converters and quasi-resonant (QR) converters.

In Section 2, the One-Cycle Control theory is developed based on the analysis of the basic buck converter with conventional feedback control and current-mode control. In Section 3, experiments were conducted to verify the feasibility of One-Cycle Control of the buck converter. In Section 4, the dynamic behavior of One-Cycle Controlled switching converters is studied. The Ćuk converter is used as an example for the analysis and experiments. In Section 5, the One-Cycle Control theory is generalized to control all types of switching converters: constant frequency, constant turn-on time, constant turn-off time, and variable switches. Conclusions and some further discussions are given in Section 6 .

2 Development of New Control Technique

A new control technique is developed based on the fundamental study of the basic buck converter with conventional feedback control and current-mode control. The duty-ratio of the conventional control is a linear function of the control reference, the duty ratio of the current-mode control contains some nonlinear state feedback, while the duty ratio of the new control technique is a completely nonlinear function of the control reference and some state variables.

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2.1 The Basic Buck Converter

The simplest configuration buck converter, shown in Fig. 1, is used as an example to study the features of control techniques mentioned above. The DC line input voltage is v_g and the switch S is operated

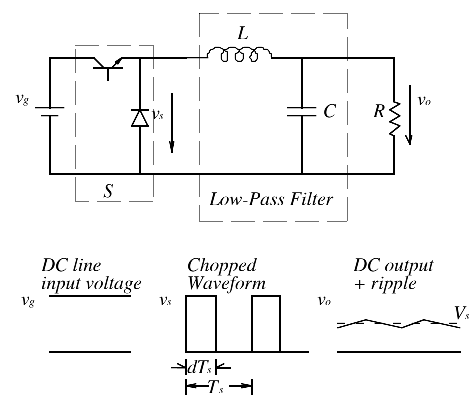

Figure 1: The Buck Converter.

with a constant frequency $f_s = \frac{1}{T_s}$. When the transistor is ON, the diode is OFF, and the diode-voltage v_s is equal to the input voltage v_g. When the transistor is OFF, the diode is ON, and the diode-voltage v_s is zero. The DC line-input voltage is chopped by the switch resulting in a chopped waveform v_s. The average, or DC, of this waveform is V_s.

$$V_s = \frac{1}{T_s}\int_0^{T_s} v_s dt = dv_g \tag{1}$$

The LC low-pass filter transmits this value to the output while rejecting most of the undesired switch frequency f_s. Therefore, the output voltage contains the desired DC value dv_g and a small residual switch ripple. The buck converter has a conversion rate equal to its duty-ratio d. By controlling the duty-ratio d, the output DC voltage is controlled.

2.2 Conventional Feedback Control

In conventional control, the duty ratio pulses are produced by comparing the control reference signal with a saw-tooth signal. As a result, the control reference signal is linearly modulated into the duty ratio signal:

$$d = \alpha v_{ref}, \tag{2}$$

where α is a constant. With feedback, the above equation becomes

$$d = \alpha(v_{ref} - v_o) \tag{3}$$

A buck converter with conventional feedback is shown in Fig. 2. The duty-ratio is modulated in the direction to reduce the error.

Suppose the input voltage is perturbed, for example by a large step up, the duty-ratio control does not see the change instantaneously since the error signal must change first. Therefore, the output voltage jumps up and the typical output voltage transient overshoot is observed as illustrated in Fig. 3. The feedback signal is compared with the reference, and the error is amplified to control the duty-ratio. The

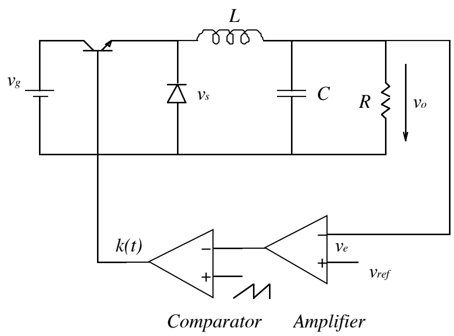

Figure 2: The Conventional Feedback Buck Converter.

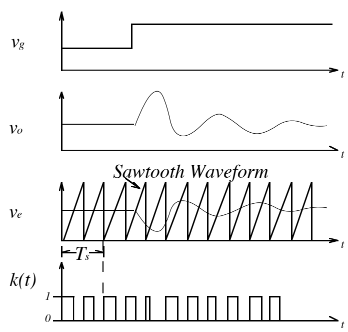

Figure 3: The Feedback Transient.

duration of the transient is dictated by the loop-gain bandwidth and a large number of switching cycles is required before the steady-state limit is reached. The output is always influenced by the input-voltage perturbation. Furthermore, the addition of an input filter might cause oscillations due to interactions with the originally stable closed-loop buck regulator.

2.3 Current-Mode Control

Current-mode control, shown in Fig. 4, utilizes some of the pulsed and nonlinear nature of the switching converter. The switch current is sensed and compared with the control reference. A constant frequency clock pulse turns the transistor on. When the switch current reaches the control reference the comparator changes its state and turns the transistor off. The duty ratio signal is determined by the following relation, which contains some nonlinear state feedback.

$$i_L + m_1 \frac{dT_s}{2} = i_{ref}, \tag{4}$$

where i_L is the inductor current and m_1 is the rising slope of the inductor current, for the buck converter $m_1 = \frac{v_g - v_o}{L}$.

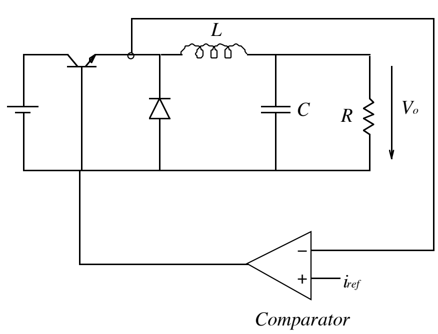

Figure 4: Current-Mode Control of the Buck Converter.

When the input voltage is perturbed, for example by a step up, the current ramp immediately increases to control the duty-ratio, as

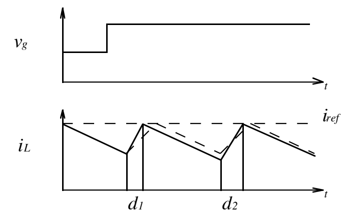

Figure 5: Current-Mode Control of Buck Converter with Artificial Ramp.

shown in Fig. 5.

$$d_{n+1} - d_n \approx -\frac{v_o}{v_g - v_o}(d_n - d_{n-1}) \tag{5}$$

$$d_n\,|_{n=\infty} = \begin{cases} d_{new} & \frac{v_o}{v_g} < 0.5 \\ oscillates & \frac{v_o}{v_g} > 0.5 \end{cases} . \tag{6}$$

When $\frac{v_o}{v_g} < 0.5$, the transient process converges, however, it takes several cycles for the system to reach the new steady state. When $\frac{v_o}{v_g} > 0.5$, the transient process does not converge; an artificial ramp can be employed to stabilize the system. If the artificial ramp is chosen to be exactly equal to the falling slope s_f of the switch current, the system fully rejects the input voltage perturbations. The falling slope of the switch current of the buck converter is determined by the output voltage and the output filter inductance L.

$$s_f = \frac{v_o}{L} \tag{7}$$

When the output voltage changes, the artificial ramp must change accordingly; therefore, only the buck converter operating at a constant output voltage satisfies this condition. For converters other than the buck converter the falling slope of the switch current may be a function of the input voltage, the voltage across the energy-transfer capacitor, and/or the output voltage. Therefore, the artificial ramp can no longer match the falling slope of the switch current. Due to this mismatch, current-mode control is unable to reject input-voltage perturbations.

2.4 One-Cycle Control

Let's go back to the original buck converter shown in Fig. 1. Close observation of the diode voltage leads to an interesting discovery. The output voltage of the buck converter is the average value of the diode voltage, which is equal to the area under each diode-voltage pulse divided by the switching period.

$$V_s = \frac{1}{T_s}\int_0^{T_s} v_s dt = \frac{1}{T_s}\int_0^{dT_s} v_g dt \tag{8}$$

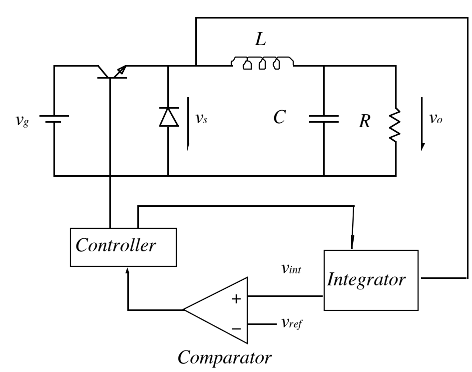

Figure 6: One-Cycle Control of Buck Converter.

A new control scheme, shown in Fig. 6, is conceived for constant switching frequency.

The controller uses constant frequency pulses to simutaneously turn ON the transistor and activate the integrator. The diode voltage is integrated and compared with a control reference. As soon as the integrated diode voltage reaches the control reference, the transistor is turned OFF and the integrator is reset to zero.

If the control reference is constant, then the average of the diode voltage is constant; therefore, the output voltage is constant, as shown in Fig. 7. The slope of the integration is directly proportional to the input voltage. The integration value is continuously compared with the constant control reference. If where the input voltage is higher, the slope of the integration is steeper; therefore, the integration value reaches the control reference faster. As a result, the duty ratio is smaller. If the input voltage is lower, the duty ratio is larger.

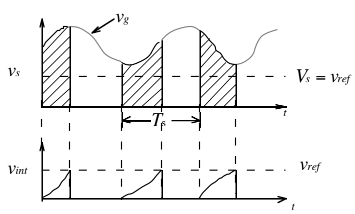

Figure 7: Constant Control Reference.

If the control reference is changing, then the average of the diode voltage is equal to the changing control reference in each cycle; therefore, the output voltage equals the control reference. Fig. 8 shows the case where the control reference changes its value in a single step up. The integration value of the diode voltage keeps up with the control reference immediately.

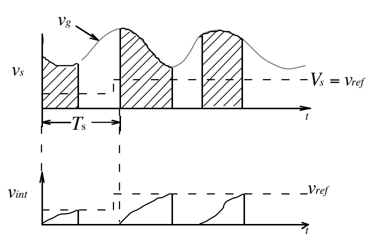

Figure 8: Variable Control Reference.

With this control scheme, the duty ratio d is determined by

$$\frac{1}{T_s}\int_0^{dT_s} v_g dt = v_{ref} \tag{9}$$

which is a nonlinear function of the input voltage and the control reference. If this control concept is practically realizable, the transient of the average value of the diode voltage would be completed within one switching cycle. This control scheme is defined as One-Cycle Control.

3 Feasibility of One-Cycle Control

Experiments were conducted to verify the feasibility of One-Cycle Control. According to the analysis of last section, the diode-voltage of the One-Cycle Controlled switching converter is exactly equal to the control reference; therefore, the average value of the diode voltage should completely reject the input voltage perturbations and instantly follow the control reference. The circuit used for the experiments is shown in Fig. 9. The operating condition for the experiments is $V_g = 15V$, $f_s = 30kHz$, $L_2 = 0.48mH$, $C_2 = 30\mu F$, $R = 25\Omega$. A, B, C, D, and E are the test points used in the experiments.

The diode-voltage is fed back to the real-time integrator. The integration value is compared with the control reference in real time. When the output voltage of the integrator reaches the control reference, the transistor is turned OFF and the integration is immediately reset to zero to prepare for the next cycle.

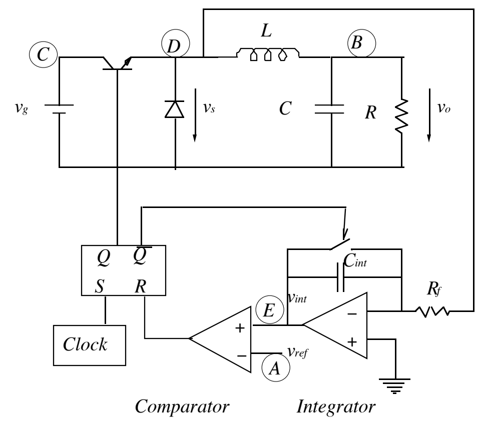

Figure 9: One-Cycle Control Buck Converter.

In each cycle, the diode-voltage waveform may be different; however, as long as the area under the diode-voltage waveform in each cycle is the same as the control reference signal, instantaneous control of the diode-voltage v_s is achieved.

3.1 Input-Voltage Perturbation Rejection

Suppose the control reference and the load are constant while the input voltage v_g is perturbed by an arbitrary pattern. The changing diode-voltage is integrated in real time and the slope of the integrated diode-voltage changes exactly and immediately when changes occur in the diode-voltage. Therefore, the input voltage directly and instantly affects the duty-ratio d such that the integration of the diode-voltage is constant in each cycle. In Fig. 10 the input voltage is stepped

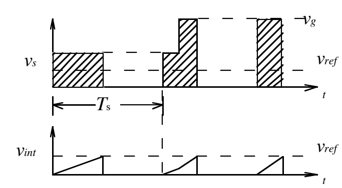

Figure 10: Rejection of Input-Voltage Perturbations.

up while the transistor is ON. The slope of the integration changes immediately; therefore, the speed to reach the control reference is adjusted instantaneously in order to keep the integrated value of the diode-voltage the same as the control reference. Theoretically, this control technique completely rejects input-voltage perturbations.

Experiment 1 The response of the diode-voltage to a step-up perturbation of the input voltage was measured. A step-up function from $10V$ to $20V$ was injected into the input voltage v_g at Point C, while the load and the control reference were held constant. The output response of the integrator v_{int} was measured at Point E. Note that the input voltage has been reduced by a factor of two in order to fit it into the plot of the experimental results shown in Fig. 11. The spikes on the input voltage are caused by the non-zero output impedance of the power source. These spikes did not influence the average value of the diode-voltage, because the spikes are included in the real-time integration that is compared to the reference voltage. The input voltage stepped up while the transistor was on and the slope of the integration of that cycle changed immediately; therefore, the duty-ratio was adjusted instantaneously.

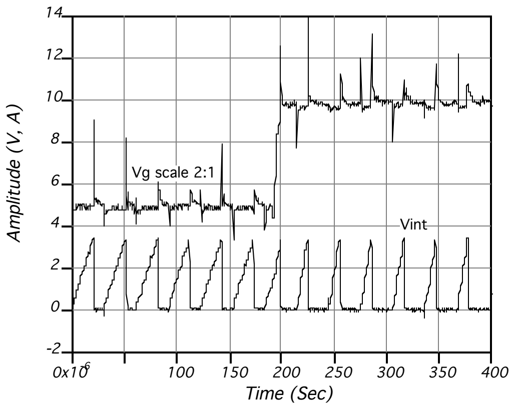

Figure 11: Buck Converter Rejects a Step-Up in the Input Voltage.

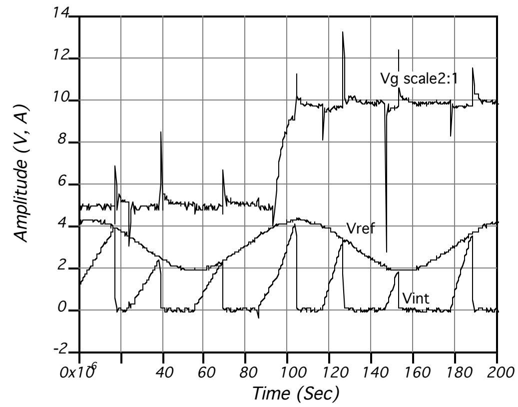

Figure 13: Buck Converter Response to a Step-Up in the Input Voltage and a Sinusoid Change in the Control Reference.

3.2 Load-Disturbance Rejection

Suppose the control reference and the input voltage are constant, whereas the load current is perturbed. If the input voltage source has some output impedance, the amplitude of the diode-voltage will be perturbed because the disturbing current generates a voltage disturbance across the input impedance. This disturbance is equivalent to the case when the input voltage is perturbed. One-Cycle control completely rejects load disturbances at the diode-voltage, and keeps the average of the diode-voltage constant. However, the output voltage is disturbed because of the dynamics of the output filter.

3.3 Following the Control Reference

Suppose the input voltage and the control reference are changing at the same time. For example, the input voltage has a step up perturbation while the control reference changes sinusoidaly. The slope

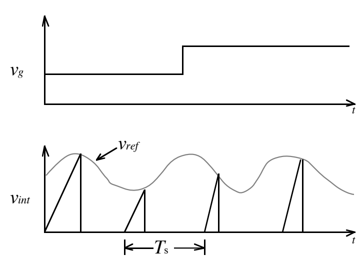

Figure 12: Following the Control Signal and Rejecting the Input Voltage Perturbation.

of the integration becomes steeper when the amplitude of the input voltage steps up. No matter how the integration slope changes, the integration value still keeps up with the sinusoid control reference in each cycle. Therefore, the average value of the diode-voltage does not see the input perturbation and it follows the control reference in one cycle, as shown in Fig. 12.

Experiment 2 The capability of the diode voltage to reject a step-up input-voltage perturbation while following a sinusoidally varying control reference was measured. A step-up function from $10V$ to $20V$ was injected into the input voltage at Point C, while the control reference was varied with a sinusoid wave $v_{ref} = 3.1 + 1.2 sin\omega t$, $f = 10kHz$, at Point A. The output response of the integrator was measured at Point E. Note that the input voltage has been reduced by a factor of two in order to fit it into the plot of the experimental results shown in Fig. 13. The slope of the integration changed immediately when the input voltage stepped up. The envelope of the integration waveform kept up with the control reference exactly. Therefore, the average of the diode-voltage was not influenced by the input disturbance and was fully controlled by the control reference.

4 Stability of One-Cycle Control

Experiments proved the feasibility of One-Cycle Control of the buck converter. If converter is more complicated, such as the Ćuk converter, will One-Cycle Control still work? Is the system globally stable?

Fig. 14 shows the One-Cycle Controlled Ćuk converter. The clock triggers the RS flip-flop to turn ON the transistor with a constant frequency. The diode-voltage is integrated and compared with the reference voltage v_{ref}. When the integrated value of the diode-voltage

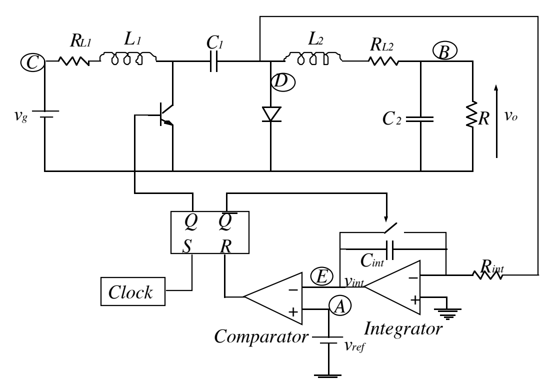

Figure 14: The Experimental Ćuk Converter with One-Cycle Control.

reaches the control reference, the comparator changes its state, which resets the RS flip-flop and consequently turns OFF the transistor. A,B,C,D, and E are the test points. The circuit operating condition is $V_g = 20V$, $f_s = 50kHz$, $L_1 = 2.39mH$, $L_2 = 2.34mH$, $C_1 = 100\mu F$, $C_2 = 1000\mu F$, $R_{L1} = 1\Omega$, $R_{L2} = 1\Omega$, $R = 10\Omega$.

4.1 Global Stability of the One-Cycle Controlled Ćuk Converter

With One-Cycle Control, the average value of the diode voltage of the Ćuk converter is exactly equal to the control reference. Therefore, the dynamics of the system is isolated by the diode voltage. The output voltage is not influenced by the input filter dynamics or by the input voltage perturbations. The Switching Flow-Graphshown in Fig. 15 reveals that the system is seprated into two subloops. The output loop, which is a second-order linear system, is always stable. The input loop is a non-linear second-order system[1]. The state-space equations for input loop are obtained from Fig. 15.

$$L_1 \frac{di_{L1}}{dt} = v_g - R_{L1} i_{L1} - (1-d) v_{C1} \quad (10)$$

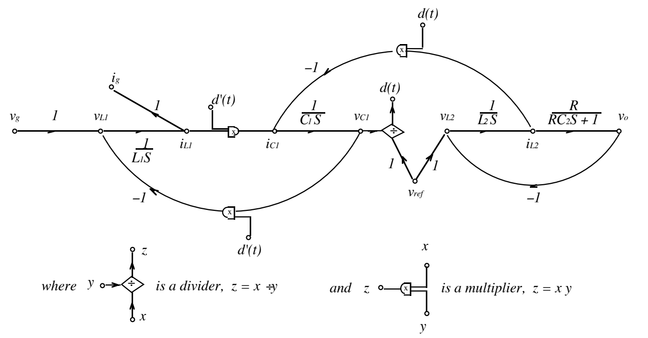

Figure 15: The Large-Signal Model of the One-Cycle Controlled Ćuk converter.

$$C_1 \frac{dv_{C1}}{dt} = (1-d)i_{L1} - d i_{L2} \quad (11)$$

$$d = \frac{v_{ref}}{v_{C1}} \quad (12)$$

Two singular points, P_1 and P_2, and a singular line, $v_{C1} = 0$, are easily found by setting the derivatives in Equations (10) and (11) equal to zero.

The global dynamic behavior, simulated by the TUTSIM program, is shown in Fig. 16. The x-axis represents the voltage across the input

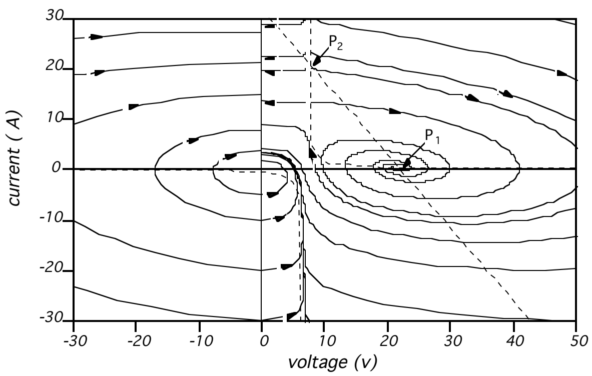

Figure 16: The Global Dynamic Behavior of the One-Cycle Controlled Ćuk Converter.

capacitor v_{C1} and the y-axis represents the input inductor current i_{L1}. The system is not globally stable. P_1 is a stable spiral point and P_2 is an unstable saddle point. The region around P_1 is the desired working region. The lower part of the y-axis is an unstable region, and the upper part of the y-axis is stable.

In practice, there is a physical restriction on the duty-ratio, $D_{min} \leq d \leq D_{max}$. When $v_{C1} \leq \frac{v_{ref}}{D_{max}}$, the system operates at the maximum duty-ratio D_{max}; therefore, the state space Equations (10) and (11) become linear with $d = D_{max}$.

When $v_{C1} \geq \frac{v_{ref}}{D_{min}}$, the system operates at the minimum duty-ratio D_{min} and the system Equations (10) and (11) becomes linear again with $d = D_{min}$.

If the maximum duty-ratio is artificially restricted such that $D_{max} < \frac{v_{ref}}{V_2}$, then the unstable saddle point $P_2 = (V_2, I_2)$ is avoided and the system is globally stable. The global dynamics are shown in Fig. 17. The definition of the axes is the same as that in Fig. 16.

4.2 Local Dynamic Behavior

In a linear feedback system, an infinite loop-gain is required in order to have instantaneous control over some variables. However, all physical systems have limited bandwidth. Consequently, when the loop-gain is higher than a certain value, the loop becomes unstable. Therefore, it is impossible to achieve instantaneous control in a linear feedback control system.

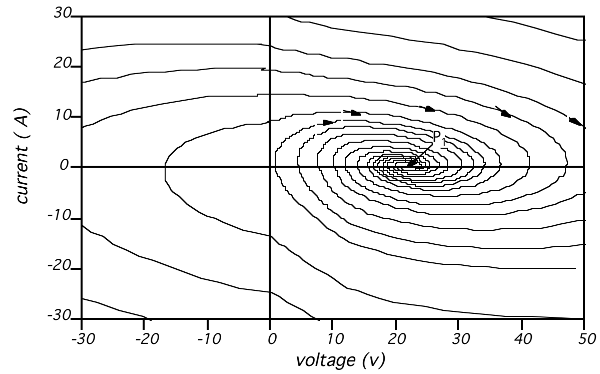

Figure 17: The Global Dynamic Behavior with Duty-Ratio Limitation.

However, instantaneous control is possible in One-Cycle Controlled converters. For the One-Cycle Controlled Ćuk converter, the average value of the diode-voltage actually has an instantaneous response to the control reference. To further understand One-Cycle Control, a study of the linearized local dynamic behavior and the loop-gain is necessary.

The output loop does not contain any switching branches, therefore, it is a stable linear second-order system. The input loop is nonlinear. Suppose the One-Cycle Controlled switch operates around the steady-state point, V_{ref}, V_{C1}, and D with small-signal perturbations, $\hat{v}_{ref}$, $\hat{v}_{C1}$, and $\hat{d}$. The linearized small-signal transfer function of the One-

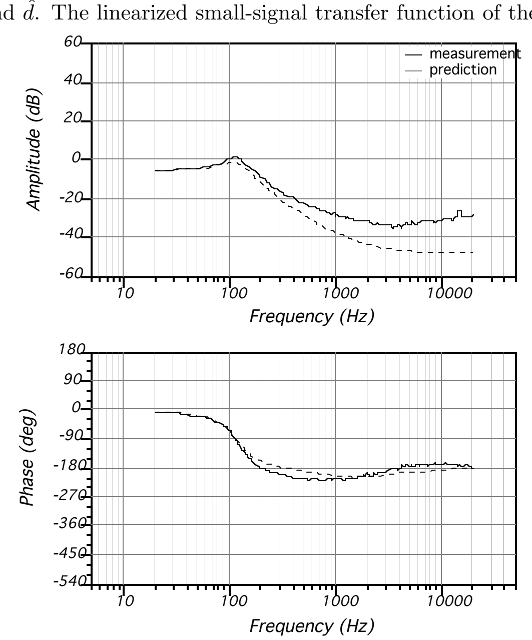

Figure 18: The Loop-Gain of the One-Cycle Controlled Ćuk Converter.

Cycle Controlled Ćuk converter with parasitic resistance is

$$G = \frac{D(RD'^2 - R_{L1}D)}{RD'^3} \frac{(1 - \frac{DL_1S}{RD'^2 - R_{L1}D})(1 + C_1R_{L2}S)}{1 + \frac{C_1(R_{L1}+D'^2R_{L2})}{D'^2}S + \frac{L_1C_1}{D'^2}S^2}. \quad (13)$$

A digital injector [3] was built to measure the loop-gain. The predicted and the measured loop-gains are plotted in Fig. 18. The loop-gain of the One-Cycle Controlled Ćuk converter is not infinite; it is actually lower than $0db$!

One-Cycle Control instantaneously controls the average value of the diode-voltage. Nevertheless, the loop-gain is not infinite. All the other state variables inside the loop obey the physical laws. The variables

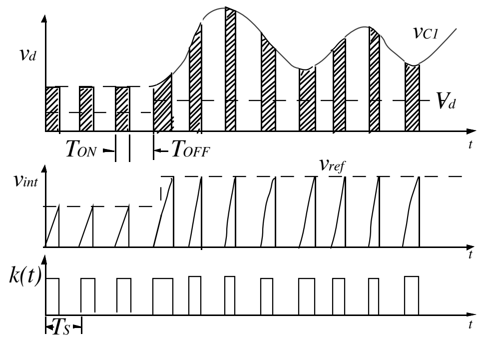

Figure 19: Take Advantage of the Pulsed and Nonlinear Nature.

actually move along the state-space trajectory shown in Fig. 17. As a matter of fact, the voltage across the diode has a finite transient. One-Cycle Control takes advantage of the pulsed and nonlinear nature of the switching converter, and adjusts the average value of the diode-voltage instantaneously. For example, when the control reference steps up, the voltage across the input capacitor undergoes an attenuating oscillation. The input signal of the One-Cycle Controlled switch is the capacitor voltage v_{C1}. The output signal of the One-Cycle Controlled switch is the diode-voltage v_d, which has an envelope equal to the capacitor voltage v_{C1}. The real-time integrated value is compared with the control reference in each cycle. Therefore, the duty-ratio is precisely adjusted such that the average of the diode-voltage is exactly equal to the control reference. The real transient of the diode-voltage is not instantaneously controlled, as shown in Fig. 19.

4.3 Frequency Response Measurements

Experiment 3 The control-to-diode-voltage frequency response was measured. A sweeping frequency signal was injected into the control

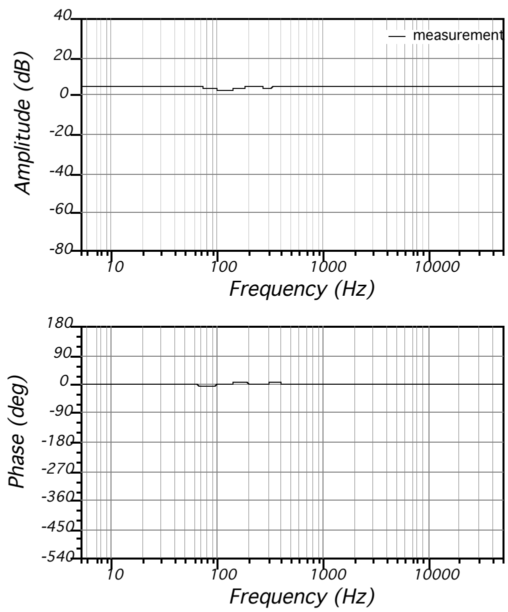

Figure 20: The Control-to-Diode-Voltage Frequency Response of the One-Cycle Controlled Ćuk Converter.

reference at Point A, while the diode-voltage response was measured at Point D. The experimental result is plotted in Fig. 20. Since the average value of the diode-voltage was fully controlled by the control reference, it was predicted that the frequency response of the diode-voltage to the control reference should be flat. The detected frequency response has a very flat amplitude response and phase lag over the frequency range 5Hz to 50 kHz.

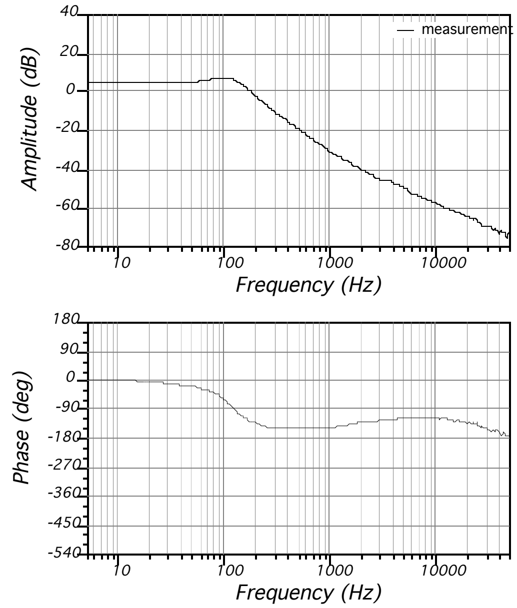

Figure 21: The Control-to-Output Frequency Response of the One-Cycle Controlled Ćuk Converter.

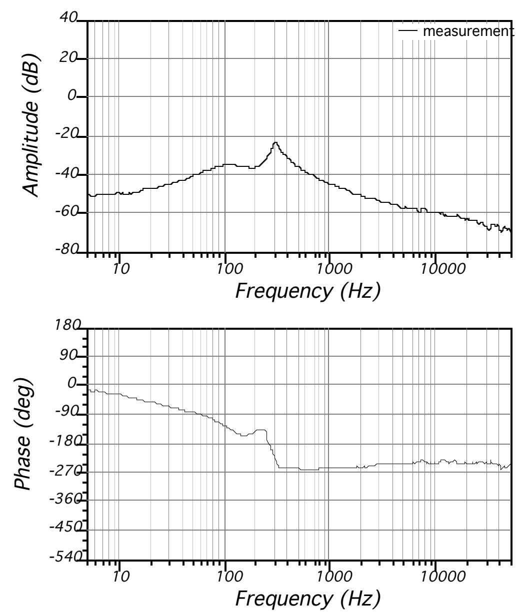

Figure 22: The Input-to-Output Frequency Response of the One-Cycle Controlled Ćuk Converter.

Experiment 4 The control-to-output frequency response was measured. A sweeping frequency signal was injected into the control reference at Point A, and the output-voltage response was measured at Point B. The experimental result is plotted in Fig. 21. The frequency response of the One-Cycle Controlled Ćuk converter is equivalent to a second-order system as expected.

Experiment 5 The input-to-output frequency response was measured. A sweeping frequency signal was injected into the input voltage at Point C, while the output-voltage response was measured at Point B. The experimental result is plotted in Fig. 22. Theoretically, the system should completely reject the input-voltage perturbation. The experimental data show that the input perturbation is attenuated by more than 20 db over the frequency range 5Hz to 50 kHz. There is a peak near the corner frequency, $300Hz$, of the input filter. That is due to the fact that the real diode has a non zero conducting resistance and the wire wrap circuit has some AC coupling.

5 Extension of One-Cycle Control

The One-Cycle Control technique found for the constant frequency switching converter is extended to general theory. The implementation circuits are found for any type of switch, constant frequency, constant ON-time, constant OFF-time, and variable.

5.1 General Theory

A switch operates according to the switch function $k(t)$ at a frequency $f_s = \frac{1}{T_s}$.

$$k(t) = \begin{cases} 1 & 0 < t < T_{ON} \\ 0 & T_{ON} < t < T_s \end{cases} \tag{14}$$

In each cycle, the switch is ON for a time duration T_{ON} and is OFF for a time duration T_{OFF}, where $T_{ON} + T_{OFF} = T_s$. The duty-ratio $d = \frac{T_{ON}}{T_s}$ is modulated by an analog control signal $v_{ref}(t)$. The switch input signal $x(t)$ is chopped by the switch. The frequency and the pulse width of the switch output $y(t)$ is the same as that of the switch function $k(t)$, while the envelope of $y(t)$ is $x(t)$, as shown in Fig. 23.

$$y(t) = k(t)x(t) \tag{15}$$

Four types of switches are considered here:

switch type	T_{ON}	T_{OFF}	T_s
constant frequency	variable	variable	constant
constant ON-time	constant	variable	variable
constant OFF-time	variable	constant	variable
variable	variable	variable	variable

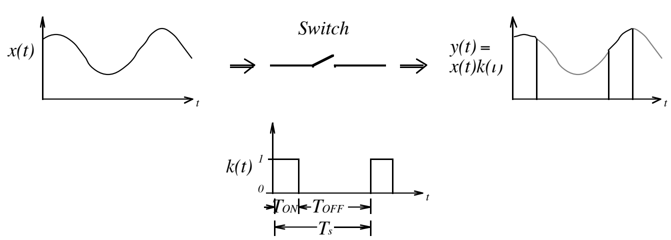

Figure 23: The Switch Function

Suppose the switch frequency f_s is much higher than the frequency bandwidth of either the input signal $x(t)$ or the control signal $v_{ref}(t)$; then the effective signal carried in the switch output is

$$y(t) = \frac{1}{T_s}\int_0^{T_{ON}} x(t)dt \tag{16}$$

$$\approx x(t)\frac{1}{T_s}\int_0^{T_{ON}} dt \tag{17}$$

$$= x(t)d(t) \tag{18}$$

$$= x(t)v_{ref}(t) \tag{19}$$

The output signal $y(t)$ of the switch is the product of the input signal $x(t)$ and the control signal $v_{ref}(t)$; therefore, the switch is nonlinear. If the control signal $v_{ref}(t)$ is constant, for example $v_{ref}(t) = D$, the output signal of the switch is $Dx(t)$, which is the case when the switch is used for digital signal processing. In power processing applications, for example a power amplifier, the input $x(t)$ usually represents the power, while the control signal $v_{ref}(t)$ represents the signal to be amplified. In the ideal case the input power $x(t)$ is constant X; therefore, the output signal $y(t) = Xv_{ref}(t)$. However, in reality perturbations always exist in the input power $x(t)$; hence, the output signal $y(t)$ contains the power disturbance as well.

If the duty-ratio of the switch is modulated such that the integration of the chopped waveform at the switch output is exactly equal to the integration of the control signal in each cycle, ie.

$$\int_0^{T_{ON}} x(t)dt = \int_0^{T_s} v_{ref}(t)dt, \tag{20}$$

then the average value of the chopped waveform at the switch output is exactly equal to the average value of the control signal in each cycle, ie.

$$\frac{1}{T_s}\int_0^{T_{ON}} x(t)dt = \frac{1}{T_s}\int_0^{T_s} v_{ref}(t)dt. \tag{21}$$

Therefore, the output signal is instantaneously controlled within one cycle, ie.

$$y(t) = \frac{1}{T_s}\int_0^{T_{ON}} x(t)dt = \frac{1}{T_s}\int_0^{T_s} v_{ref}(t)dt = v_{ref}(t) \tag{22}$$

The technique to control switches according to this concept is defined as the One-Cycle Control technique. With One-Cycle Control, the effective output signal of the switch is

$$y(t) = v_{ref}(t). \tag{23}$$

The switch fully rejects the input signal and linearly all-passes the control signal v_{ref}; therefore, the One-Cycle Control technique turns a non-linear switch into a linear switch.

5.2 One-Cycle Control of Constant Frequency Switches

For a constant frequency switch, T_s is constant. The object of One-Cycle Control is to adjust the switch ON-time T_{ON} in each cycle, such that the integrated value of the chopped waveform is exactly equal to the control reference.

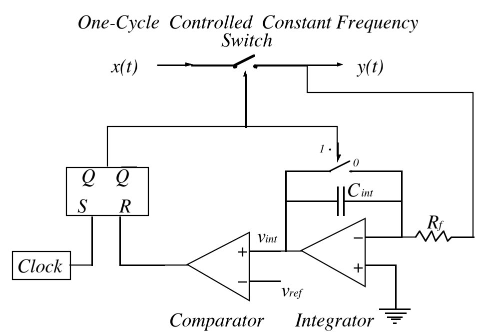

Figure 24: The One-Cycle Controlled Constant Frequency Switch.

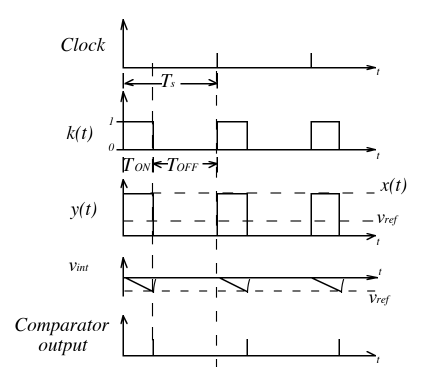

Figure 25: The Waveforms of the One-Cycle Controlled Constant Frequency Switch.

The implementation circuit for One-Cycle Control of a constant frequency switch is shown in Fig. 24. The key component of the One-Cycle Control technique is the real-time integrator. The real-time integration is started the moment the switch is turned ON by the fixed frequency clock pulse. The integration value,

$$v_{int} = \frac{1}{T_s}\int_0^t x(t)dt, \quad (24)$$

is compared with the control signal $v_{ref}(t)$ in real time. At the instant when the integration value v_{int} reaches the control signal $v_{ref}(t)$, the controller sends a command to the switch to change it from the ON state to the OFF state. At the same time, the controller resets the real-time integrator to zero to prepare for the next cycle. The duty-ratio d of the present cycle is determined by the following equation:

$$\frac{1}{T_s}\int_0^{T_{ON}} x(t)dt = v_{ref}(t) \quad (25)$$

Since the switch period T_s is constant and the duty-ratio is controlled, the average value of the waveform at the switch output $y(t)$ is guaranteed to be

$$y(t) = \frac{1}{T_s}\int_0^{T_{ON}} x(t)dt = v_{ref}(t) \quad (26)$$

in each cycle. Fig. 25 shows the operating waveforms of the circuit.

5.3 One-Cycle Control of Constant ON-Time Switches

For a constant ON-time switch, T_{ON} is constant. The object of One-Cycle Control is to adjust the OFF-time T_{OFF} in each cycle, such that the average value of the chopped waveform is exactly equal to the control reference.

The implementation circuit for One-Cycle Control of a constant ON-time switch is shown in Fig. 26. The monostable multivibrator generates a constant pulse width.

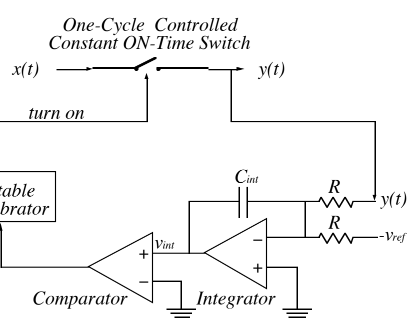

Figure 26: The One-Cycle Controlled Constant ON-Time Switch.

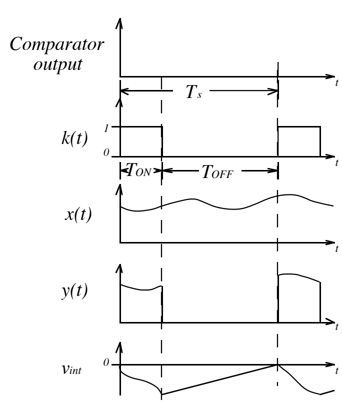

Figure 27: The Waveforms of the One-Cycle Controlled Constant ON-Time Switch.

The real-time integration is started at the moment the switch is turned ON. From $t = 0$ to $t = T_{ON}$, v_{int} decreases. When the monostable multivibrator changes its state from high to low, the switch is turned OFF. From $t = T_{ON}$ to $t = T_s$, v_{int} increases.

$$v_{int} = \begin{cases} \int_0^t v_{ref}(t)dt - \int_0^t x(t)dt & 0 < t < T_{ON} \\ \int_0^t v_{ref}(t)dt - \int_0^{T_{ON}} x(t)dt & T_{ON} < t < T_s \end{cases}. \quad (27)$$

At the instant when v_{int} reaches zero, the comparator changes its state from low to high, which triggers the monostable multivibrator to high and turns the switch back ON. The present switching cycle is completed and the switch starts the next cycle.

The OFF-time T_{OFF} of the present cycle is determined by the following equation:

$$\int_0^{T_{ON}} x(t)dt = (T_{ON} + T_{OFF})v_{ref}(t) \quad (28)$$

The waveform at the switch output $y(t)$ is guaranteed to be

$$y(t) = \frac{1}{T_s}\int_0^{T_{ON}} x(t)dt = v_{ref}(t). \quad (29)$$

where T_s is time dependent and T_{ON} is constant. Fig. 27 shows the operating waveforms of the circuit.

5.4 One-Cycle Control of Constant OFF-Time Switches

For a constant OFF-time switch, T_{OFF} is constant. The object of One-Cycle Control is to adjust the ON-time T_{ON} in each cycle, such that the average value of the chopped waveform is exactly equal to the control reference.

The implementation circuit for One-Cycle Control of a constant OFF-time switch is shown in Fig. 28. The monostable multivibrator generates a constant pulse width. The real-time integration is started

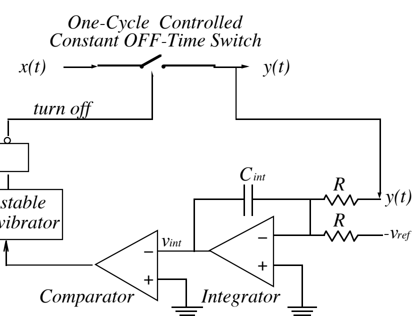

Figure 28: The One-Cycle Controlled Constant OFF-Time Switch.

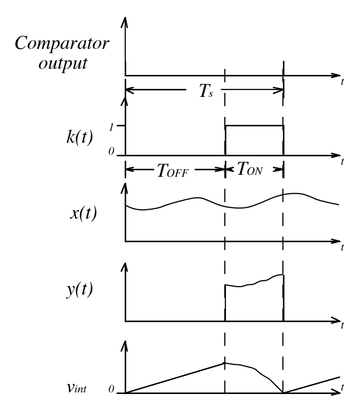

Figure 29: The Waveforms of the One-Cycle Controlled Constant OFF-Time Switches.

the moment the switch is turned OFF. From $t = 0$ to $t = T_{OFF}$, v_{int} increases. When the monostable multivibrator changes its state from high to low, the switch is turned ON. From $t = T_{OFF}$ to $t = T_s$, v_{int} decreases.

$$v_{int} = \begin{cases} \int_0^t v_{ref}(t)dt & 0 < t < T_{OFF} \\ \int_0^t v_{ref}(t)dt - \int_{T_{OFF}}^t x(t)dt & T_{OFF} < t < T_s \end{cases}. \quad (30)$$

At the instant when v_{int} reaches zero, the comparator changes its state from low to high, which triggers the monostable multivabrator to high and turns the switch OFF. The present switching cycle is completed and the switch starts the next cycle.

The ON-time T_{ON} of the present cycle is determined by the following equation:

$$\int_{T_{OFF}}^{T_{OFF}+T_{ON}} x(t)dt = (T_{ON} + T_{OFF})v_{ref}(t) \tag{31}$$

Since the OFF-time T_{OFF} of the switch is constant and the ON-time T_{ON} is controlled, the average value of the waveform at the switch output $y(t)$ is guaranteed to be

$$y(t) = \frac{1}{T_s}\int_{T_{OFF}}^{T_{OFF}+T_{ON}} x(t)dt = v_{ref}(t) \tag{32}$$

in each cycle. Fig. 29 shows the operating waveforms of the circuit.

5.5 One-Cycle Control of Variable Switches

For a variable switch, there are two adjustable control parameters, T_{ON} and T_{OFF}. Usually, one parameter is governed by the particular application. If a particular application requires the ON-time vary in a particular pattern, then the One-Cycle Control can be implemented in an approach similar to the one described for the constant ON-time switches. If a particular application restricts the OFF-time by some function, then the One-Cycle Control can be implemented in an approach similar to the one described for the constant OFF-time switches.

6 Conclusion

The One-Cycle Control technique is designed to control the duty-ratio d of the switch in real time, such that in *each cycle* the average of the chopped waveform at the switch output is *exactly* equal to the control reference. Experiments show that a switching converter with One-Cycle Control, rejects input-voltage perturbations, and follows the control reference quickly. Implementation circuits are found for any type of switch, constant frequency, constant ON-time, constant OFF-time, and variable. Therefore, the One-Cycle Control technique is suitable for large-signal robust control of PWM switching converters and quasi-resonant converters, inverters, and rectifiers. This technique may also useful for signal processing and other applications.

Theoretically, converters with One-Cycle Control are capable of rejecting the input-voltage perturbations, and the diode-voltage is able to follow the control signal instantaneously, within one cycle. Therefore, the One-Cycle Controlled converter is equivalent to a controllable voltage source with an output filter. However, in practice, the switches, the transistors, and the diodes are not ideal switches and the integration is not instantaneous. Therefore, the accuracy of One-Cycle Control is greatly dependent upon the circuit design. The experimental circuits of a buck converter and a Ćuk converter in this work show a very close match between the measurements and the theoretical predictions. The dynamic behavior of the Ćuk converter with One-Cycle Control, for both the large-signal and the small-signal case, is analyzed. The Switching Flow-Graph model shows that the One-Cycle Control Ćuk converter is not globally stable. However, imposing a limitation on the duty-ratio $D_{min} \leq d \leq D_{max}$ prevents the converter from becoming unstable while operating in the previously unstable regions. As a result, the system is globally stable and behaves like a second-order linear system. However, the system transient takes longer then one cycle if the it has to pass the D_{max} or D_{min} limited regions.

The One-Cycle Control concept is straightforward and its implementation circuits are simple; yet it provides excellent control of switching converters.

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