# *Current-Mode Control of Switching Power Supplies*

by Lloyd H. Dixon, Jr.

Topic 1

# CURRENT MODE CONTROL OF SWITCHING POWER SUPPLIES

Introduction: This paper examines the current mode control method applied to Buck, Boost and Flyback circuits operated in the continuous and discontinuous inductor current modes.

The current mode control method uses two control loops -- an inner, current control loop and an outer loop for voltage control. Figure 1 shows a forward converter (buck family) using current mode control. When the switching transistor is on, current through $R_{sense}$ is proportional to the upward ramping filter inductor current. When the ramp voltage $V_s$ reaches $V_e$ (the amplified output voltage error), the switching transistor turns off. Thus, the outer voltage control loop defines the level at which the inner loop regulates peak current through the switch and through the filter inductor.[1]

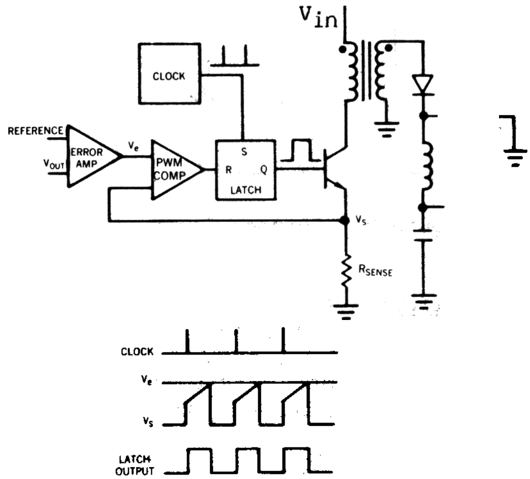


Figure 1.

ADVANTAGES:

- Input voltage feed-forward, resulting in good open-loop line regulation.
- Simplified loop -- inductor pole and 2nd order characteristic eliminated
- Optimum large-signal behavior.
- No conditional loop stability problems.
- Flux balancing (symmetry correction) in push-pull circuits.
- Automatic pulse-by-pulse current limiting.
- Current sharing of paralleled supplies for modular power systems.
- Less complexity/cost (current sense/amp is *not* an added complication).

DISADVANTAGES AND PROBLEMS (continuous mode only):

- Peak/avg. current error and instability -- slope compensation
- Noise immunity is worse because of shallower ramp.
- Half Bridge runaway
- DC open loop load regulation is worse.
- (1-D) current error in Boost or Flyback circuits.
- Loop irregularities with multiple output buck circuits.

## THE BUCK REGULATOR (Continuous Inductor Current mode)

The buck family includes the Forward Converter, Full Wave Center-Tap, Full Bridge and Half Bridge.

Peak vs. Average Current Error: As shown in Figure 2, current mode control regulates the *peak* inductor current. However, in a buck regulator operated in the continuous mode, the inductor drives the output so that load current equals the *average* inductor current. The difference between peak and average inductor current is an error, which is greatest when $V_{in}$ is large.



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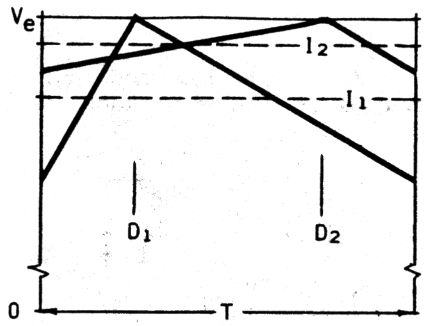


Figure 2. Current Error

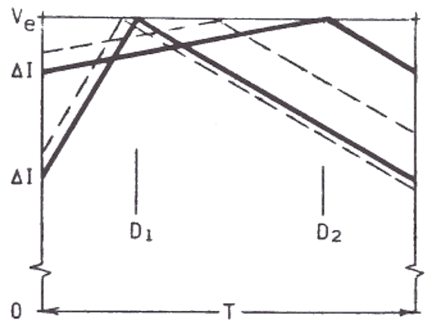


Figure 3. Instability

Instability: With continuous mode operation, controlling the peak inductor current results in circuit instability when the duty ratio, D, is greater than 0.5. Figure 3 shows how a small perturbation in the inductor current is magnified when D is greater than 0.5 (when $V_{in}$ is low) because the downslope is greater than the upslope.

Slope Compensation: The correct amount of slope compensation eliminates both problems simultaneously. The *average* inductor current does not then change regardless of changes in $V_{in}$ and D, and the circuit becomes stable at any duty ratio, as shown in Figure 4. However, in boost and flyback circuits, the large (1-D) error between average inductor current and output diode current is not improved through slope compensation.

Slope compensation is achieved as shown in Figs. 4 and 5 by summing a negative sawtooth ramp voltage with the amplified error voltage, $V_e$, at the control input of the comparator (Ref. Fig. 1). For perfect compensation, the compensation ramp must have a slope equal to exactly 1/2 the downslope of the voltage waveform at the other comparator input, which is the voltage analog of the inductor current downslope as seen across the current sense resistor.

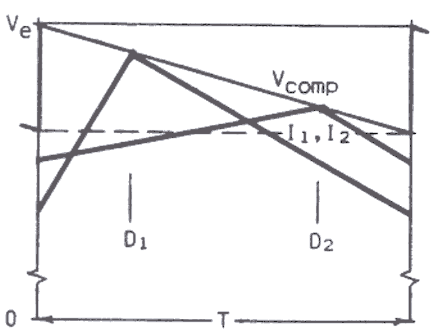


Figure 4. Slope Compensation

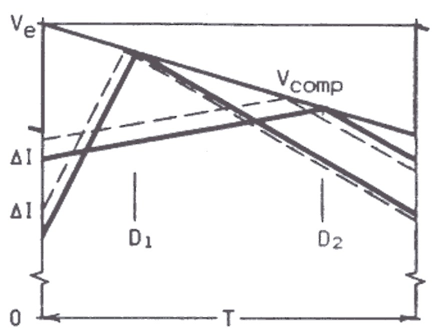


Figure 5. Slope Compensation



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A negative ramp is not usually available within the control IC, but a positive going ramp is available from the IC's clock pulse generator. As shown in Figure 6, it is possible to use this *positive* ramp for slope compensation by applying it through a voltage divider to the *opposite* comparator input, summing it with the voltage analog of the inductor current from the current sense resistor. The voltage divider should set the slope of this positive ramp to be the same magnitude -- 1/2 the downslope of the voltage waveform representing the inductor current.

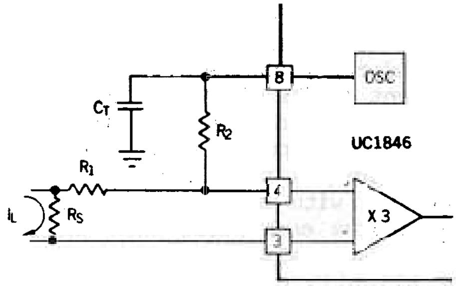

Figure 6. Compensation Circuit

Line Regulation: Current mode control has an inherent input voltage feed-forward characteristic which means the open loop D.C and dynamic line regulation is excellent. This reduces the closed loop gain needed in the outer voltage control loop. The good line regulation is caused by the inner current control loop which maintains constant peak inductor current regardless of changes in $V_{in}$. With slope compensation, this results in constant *average* inductor current. $V_{in}$ changes of any kind -- large signal step changes, spikes, A.C. signals or noise riding on the line are rejected, even with the outer voltage control loop open.

Load Regulation: With voltage mode (duty ratio) control, open loop DC load regulation is inherently excellent, although dynamic load regulation is very bad. With current mode control, and with the outer voltage control loop open, the load regulation is very poor. This because the inner current control loop makes the circuit into a current source. Any current source has poor load regulation. This is not nearly as bad as it seems, because it is so much easier to get high closed loop gain-bandwidth which corrects the problem and gives many other performance advantages.

Paralleling Outputs: With current mode control, it is easy to parallel several supplies for the sake of redundancy or for modular power supply systems. If all the supplies to be paralleled have identical current sense resistors and identical current control loops, a single control voltage common to all supplies will cause them to deliver identical currents. Their outputs may then be paralleled to drive a common load, and they will share this common load current equally. A single voltage reference and error amplifier is used, and this amplified error voltage is the control voltage used to program the current in all the paralleled supplies.

Flux Balancing in Push-Pull Circuits: In any transformer coupled push-pull circuit, small differences in $V_{sat}$ and/or storage time between the switching transistors cause a slight asymmetry in the voltage waveform, resulting in a net small DC voltage across the transformer primary. This DC voltage causes the magnetizing current and core flux to slowly move in one direction until finally, the transformer saturates. The magnetizing current flowing through the DC resistance of the primary circuit causes IR drops that are in a direction to correct this problem, but the resistance is usually nowhere near large enough. This has been a severe problem in push-pull circuits.

Current mode control automatically solves the unbalanced flux problem, by sensing and controlling the emitter currents of the switching transistors. These transistor currents represent the inductor current reflected from the secondary, *plus the transformer magnetizing current.* When the magnetizing current drifts off in one direction, it adds to the inductor current in one transistor and subtracts from it in the other transistor, making the push-pull current pulses unequal in amplitude. With duty ratio control, current pulse inequality will get worse and worse until the transformer saturates. However, with current mode control this cannot happen because the controller terminates each pulse when the same specific current is reached. Thus the pulses must always have the same amplitude, although the pulse widths will be slightly different. This pulse width differential that is created effectively corrects for the original asymmetry.

Runaway in Half-Bridge Circuits: As noted above, when current mode control is applied to any push-pull circuit, small differences in the alternate pulse widths will be created in order to correct any volt-second asymmetry applied to the transformer. Unfortunately, this *causes asymmetry in the current waveforms* drawn through each transistor -- the peak currents are maintained identical, therefore different pulse widths result in small differences in ampere-seconds, or charge, drawn alternately through the switching transistors.

In the half-bridge, one side of the transformer primary is connected to the midpoint of a capacitive voltage divider. When the charge flowing in alternate directions is slightly different because the alternate pulse widths are not the same, the capacitor divider voltage will drift in one direction. The direction is unfortunately such that it tends to reinforce the original volt-second asymmetry. This causes the current mode controller to further correct the volt-second asymmetry, making the charge unbalance worse. Thus, the capacitive divider voltage will run away until it reaches one or the other source voltage extreme.

This small charge imbalance is of no consequence in full bridge or full-wave center-tap circuits which do not involve a "soft" source such as the capacitive divider in the half bridge.

Noise Immunity: In continuous inductor current mode circuits, current mode control suffers from poorer noise immunity than duty ratio control. Both methods acheive control by comparing the amplified output voltage error to a ramp voltage. Referring to Figure 2, with current mode control the ramp represents inductor current. It sits on a pedestal and is quite shallow, especially when $V_i$ is low, making D large. The ramp voltage is never far away from the control voltage, $V_e$. A relatively small noise spike will cause the current pulse to terminate prematurely. This problem is solved by: using care in circuit layout and proper location of ground returns to avoid pulses generated by

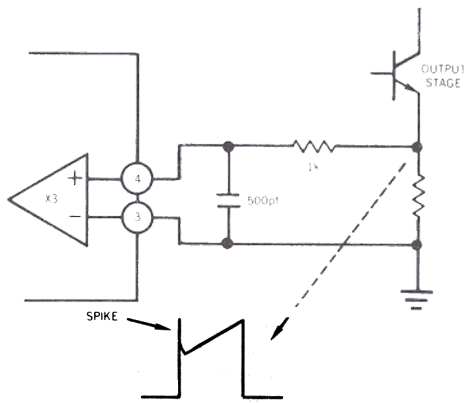

Figure 7



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fast switched high currents through wiring inductance, using differential input current sense amplifiers, using a small filter inductor (consistent with keeping out of the discontinuous mode at minimum load current), and by filtering out any remaining noise spikes with a simple RC filter at the input of the current sense amplifier as shown in Figure 7.

Small-Signal Loop Compensation: When current mode control is used with a continuous mode buck regulator, the inner (current control) loop includes the filter inductor. This eliminates the inductor from the small signal model of the outer (voltage control) loop. The voltage control loop then has only the single pole of the output filter capacitor and load resistance, as shown in Figure 9. Because the 90° phase lag of a single pole is inherently stable without additional compensation, it is easy to get high loop gain and excellent small signal dynamic performance.

With conventional duty ratio control applied to the buck regulator, the LC power filter has a two-pole second order characteristic as shown in Figure 8. There is an abrupt 180° phase lag at filter resonance (often near 100 Hz). This will cause ringing and instability if not compensated. At least one and usually two zeros near filter resonance must be provided in the error amplifier compensation network. This requires much more error amplifier gain-bandwidth than with current mode control and requires large compensation capacitors with time constants in the order of milliseconds.

When properly compensated (although with much greater difficulty), the small signal dynamic behavior of the buck regulator with conventional duty ratio control may be nearly as good as with current mode control. With a 50 kHz switching frequency, small-signal response times in the order of 100 microseconds can be achieved with either control method.

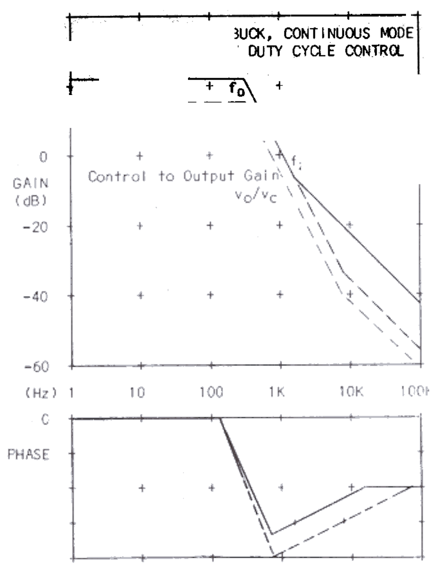


Figure 8.

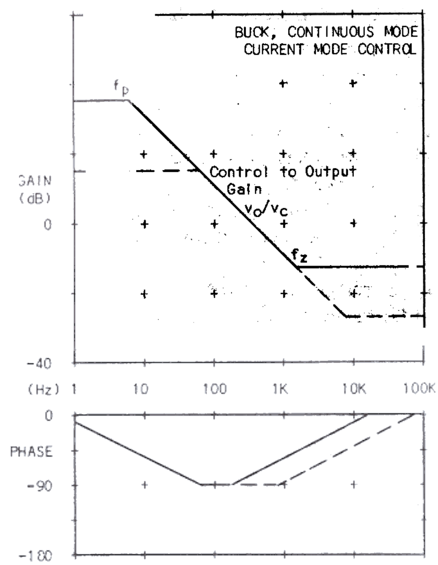


Figure 9

Loop Gain Irregularities with Multiple Outputs: Figure 1 is a current mode controlled circuit of the buck family with one output. It is clear that the current which is sensed and controlled on the primary side is directly controlling the current through the single filter inductor to the output on the secondary side.

Additional outputs can be created by adding additional transformer secondaries, rectifiers, independent LC filters and loads. The current controlled on the primary side now suppies all of the outputs, which are effectively in parallel. Since the DC output of each LC filter must equal the average of the input voltage waveform (above the critical minimum load current), the DC voltages of the different outputs are absolutely related by their secondary turns ratios. The current supplied and controlled from the primary side will automatically apportion itself according to the demands of each load. Otherwise the voltages would diverge, which is not possible at DC and at frequencies below filter resonance. The DC and low frequency AC gain from the control input to the controlled output is proportional to the paralleled combination of all load resistances, adjusted by the square of the turns ratios.

At or above filter resonance, the story is different. With only one output, the single filter inductor is in series with the current source, causing the inductor to disappear. With multiple outputs, there are several LC filters being driven in parallel from one common current source. The inductors now do *not* disappear unless the resonant frequency *and the Q of each filter are exactly the same*, which would make the paralleled filters look like a single common unit. This is not likely, because the Q values are largely determined by load resistances which normally vary considerably.

Only one output is usually sensed and fed back to become part of the closed voltage control loop. The input of the LC filter of this controlled output is driven from the primary current source, but the LC filters of the other outputs attached in parallel to this same driving point. These other LC filters are really series resonant circuits which shunt the common driving point to ground. At the resonant frequency of each of these series resonant filters, they become a very low impedance, especially if its Q is high under lightly loaded conditions. The low shunt impedance at resonance chops a hole in the gain characteristic of the controlled output, with dramatic unexpected phase shifts. This problem is really severe if the controlled output is a lower power output and is shunted by a much higher power output, especially when the high power output is lightly loaded making its Q high.

The best solution to this problem is to couple the filter inductors together by winding them on a common core. The filters are no longer independent and do not have separate resonances. This also dramatically improves the dynamic cross regulation, which is very poor with independent filters.

Large Signal Behavior: Unlike small-signal behavior, when large-signal conditions prevail, such as at startup or with large and rapid load changes, there is a dramatic difference in performance between the two control methods. The large filter inductance values inherent in continuous mode circuits make it impossible for the inductor current to follow rapid changes in load *regardless of the control method*. This limitation in the slew rate of inductor current causes the power supply output voltage to go out of regulation temporarily. The error amplifier is driven into the stops, causing the voltage control loop to become temporarily open until after the inductor current reaches the new load current level.

Under small-signal conditions, error amplifier feedback keeps the inverting input at a DC level within a millivolt of the DC reference voltage applied to the inverting input. But while the inductor current is slew rate limited and the error amplifier is in the stops, the voltage at the inverting input is uncontrolled. Referring to Figure 10, during this large-signal transition the compensation capacitors required with the duty ratio method will charge to voltage levels totally unrelated to normal operation. After the inductor current reaches the new value of load current (in perhaps 100 microseconds) and the error amplifier starts to regulate again, the power supply output voltage will be regulated with a significant error offset (perhaps 6 volts instead of 5) because of the unusual compensation capacitor voltages. The time constants to discharge the capacitors and bring the power supply output voltage back to normal will be 2 to 5 milliseconds.

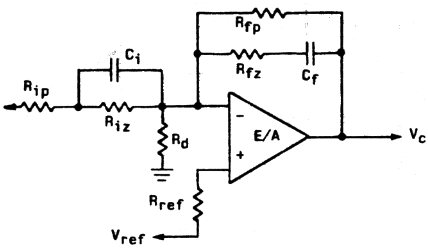


Figure 10.

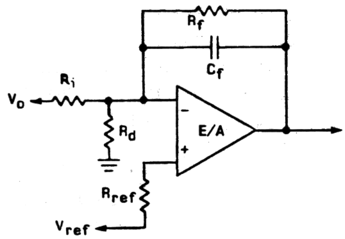


Figure 11.

In other words, the compensation capacitors necessary for good small-signal performance with duty ratio control cause poor large-signal performance -- large output glitches which take a long time to recover. In contrast to the above, current mode control achieves excellent small-signal performance without compensation capacitors other than one small capacitor which cancels the output capacitor's ESR zero. It recovers accurate regulation much more rapidly -- as soon as the inductor current reaches the new value of load current. The circuit of Figure 11 shows the simplicity of the current mode control error amplifier circuit. The single capacitor used is less than one tenth the value and time constant of the 2 capacitors necessary with duty ratio control.

Violation of Conditional Loop Stability: During large-signal episodes such as above, the time averaged loop gain is reduced. The crossover frequency is effectively lowered, which may cause a serious problem with simple duty cycle control if the control loop is undercompensated. If the loop is only conditionally stable (phase shift exceeding 180 degrees at frequencies well below crossover), the reduction in crossover frequency will initiate large-signal oscillation. The circuit will most likely remain in this oscillatory state until it is shut down and restarted. For this reason, conditional loop stability should be avoided in the design of switching power supplies. This is very easy to achieve with current mode control.

(1-D) Current Error: In boost or flyback continuous mode regulators, there is a large error between the average inductor current, $I_L$ (regulated by the current control loop) and the load current, $I_O$. This is because $I_L$ is provided to the output only during periods of diode conduction and *not* continuously as in the buck regulator. The load current $I_O$ equals the average diode current $I_D$ which equals $I_L$ (1-D). For any continuous mode regulator, the duty ratio, D, is a direct function of $V_{in}$. Referring to Figure 12, if $V_{in}$ changes, D must also change, and this changes the relationship between $I_O$ and $I_L$. If $V_{in}$ changes, $I_L$ must then be changed to accomodate a constant load. In other words, the open loop line regulation is poor, unlike the buck regulator where $I_L$ always equals $I_O$, making it independent of $V_{in}$.

The error between peak and average inductor current and the inherent instability of continuous mode circuits can both be eliminated by using slope compensation, just as with the buck regulator. The (1-D) error between $I_L$ and $I_O$ cannot be eliminated dynamically (this is what causes the right-half-plane zero), but the low frequency (1-D) error can be eliminated by appropriate feed-forward of $V_{in}$ into the current control loop.

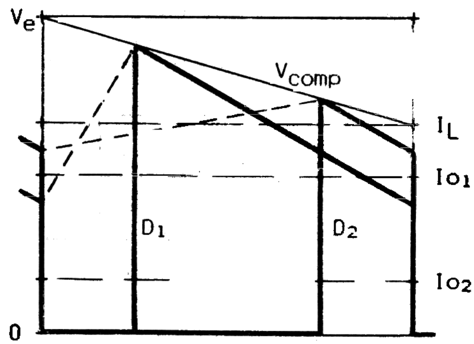


Figure 12

Right-Half-Plane Zero: Current mode control does *not* eliminate the right half-plane zero inherent in boost and flyback continuous mode circuits, although it does eliminate the inductor pole and the 2nd order resonant filter characteristic. The RHP zero gives a 20 dB/decade gain boost with a 90° phase *lag*, which is considered impossible to compensate, and usually forces the designer to roll the loop gain off a decade or more below what could otherwise be achieved. Current mode control does not help in this respect. Refer to the separate paper titled "The Right-Half-Plane Zero -- a Simplified Explanation"

## DISCONTINUOUS INDUCTOR CURRENT MODE

In the discontinuous inductor current mode, the inductor current by definition is at zero during part of each switching cycle, as shown in Figure 13. There are therefore three states during each cycle, rather than two as in the continuous mode. Most of the relevant problems encountered with continuous mode operation are not present in the discontinuous. Likewise, many of the advantages of current mode control in continuous mode operation are irrelevant in the discontinuous mode.

Current mode control does have one important advantage as applied to the discontinuous mode -- good line regulation by virtue of the inherent feed-forward capability. It shares this advantage with feed-forward ICs such as the UC3840. The choice would be mainly on the cost of the IC vs. its functionality. Voltage feed-forward and current mode control are both much better than simple duty ratio control.

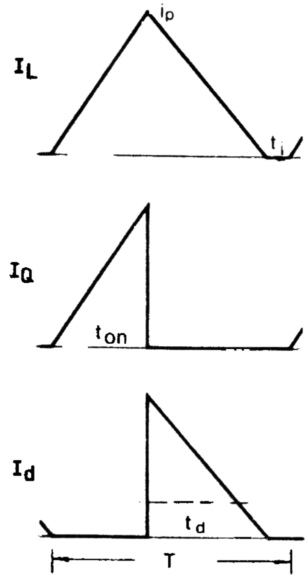

Figure 13.

In the discontinuous mode, all of the energy stored in the inductor is delivered to the load each cycle. Power output equals the energy stored in the inductor each cycle times the frequency. This inductor energy (and the power delivered to the load) is established by the peak inductor current at the time the transistor switch is turned off. With fixed $V_{in}$, current mode control and duty ratio control methods both determine the peak inductor current -- one directly, the other indirectly. With current mode control, the peak current value and the the power output do not change when $V_{in}$ changes. With simple duty ratio control, the peak current varies proportional to $V_{in}$, therefore the open loop line regulation is inherently poor. With voltage feedforward, such as in the UC3840, the open loop line regulation is good because the duty cycle is automatically changed in inverse proportion to $V_{in}$. This results in constant peak inductor current and power output regardless of $V_{in}$ changes.

## REFERENCES:


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