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for Reduction
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Loop
Interactions
In Magamp
Supplies

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Techniques for Reduction of Control-loop Interactions In Magamp Supplies

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Abstract

Everyone that routinely designs magamp post regulators soon becomes familiar with the tendency of the magamp to interact with the main control loop. Yet, there has been almost nothing published on the subject. Modeling is not easy and the math is difficult both to generate and to generalize. This paper ignores the math but does explain the more common sources of interaction and provides practical design techniques for minimizing these control-loop interaction effects.

Introduction

Interaction between magamp post regulators and the main PWM control loop can occur from a number of sources. Perhaps the most common interaction derives from use of primary current sensing in a current-mode supply. Another source is use of an output other than the magamp for reset of the saturable reactor in for example the type B magamp of ref. 2. Magamps can also interact with each other. Other interactions can and do occur during large scale transients (start-up, current limiting and during severe load changes).

The authors do not apologize for the lack of math in this article. The basic math for various magamp control loops is given in references 1, 2, 5 and 6. However, in each of the references the effects of control loop interactions have been ignored. Lab data, where it does appear, was obtained from supplies in which techniques were used to minimize most of the control-loop interactions. To the best of the authors' knowledge, very little if any work has been performed in the academic community to describe these interactions. One doesn't have to look very far to discover the reasons. Few

people in universities have had practical experience with magamps. Those that do have other pressing priorities. In the field, engineers are busy doing projects and trying to meet schedules. Regrettably, the SMPS companies that can still afford some basic research have other priorities. Also, there is probably some reluctance by companies with a good understanding of these interactions to "share their trade secrets". Spice models that include the interaction effects are theoretically possible. (The authors are not aware of anyone that has done accurate spice models that include the interaction effects.)

However, we, when doing magamp designs, cannot wait for someone else to do the math. That wait is likely to be a long one. Plus the math will be complex, difficult to understand and probably even more difficult to generalize. Thankfully, it is possible with a little lab work for one to develop an intuitive feel for the causes and effects of these control loop interactions. Many an Armed Forces radar technician has been taught the fundamentals of radar with very little math in the course of instruction. Also most baseball players reach the apex of their careers without having to

write the differential equation to describe the trajectory of a pop fly.

The apology then is not for the lack of math, but that this control loop interaction information was not presented sooner. Thus many novices in the magamp field may have seen results that differ from those described in previous papers - differences which likely can be attributed to control-loop interactions.

Sources of Interaction

Source #1 -- Sensing the primary current of the power transformer

The primary current consists of four basic components. They are:

- 1) The magnetizing current of the power transformer,
- 2) The DC load current from the main PWM controlled output reflected to the primary.
- 3) The DC load currents from auxiliary outputs reflected to the primary, and
- 4) The ramp up of currents in the output chokes of both the main and auxiliary outputs reflected into the primary.

In addition to these four basic components is a current spike at the leading edge of the pulse due to a number of factors, including the input/output capacitance of the transformer plus the snubbers across output rectifiers. There is another current spike when the saturable reactor saturates.

The pulse in figure 1 illustrates these four basic components as seen in a typical forward converter with a single magamp. The initial ramp up of current is due to the magnetizing current of the transformer to which has been added the ripple current in the output choke in the main output reflected into the primary. The ramp up of current after the saturable reactor saturates contains both of these components plus the choke current in the magamp reflected into the primary.

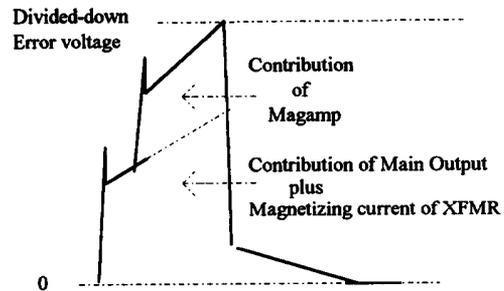


Figure 1. Pulse Components in the Primary Current of a Forward Converter with Magamp

Assume current-mode control with peak current detection. From the waveform in figure 1, it is easy to see that if the load current of the magamp is suddenly increased without a change in peak current allowed then there is an effect that tends to shorten the pulse width temporarily. Obviously what happens in the magamp can effect the main control loop because the current modulator in the main control loop is detecting the peak current.

Also obvious is that if we had a scheme for subtracting out the contribution of the magamp to the voltage analog of the primary XFMR/switch current prior to it being routed to the PWM in the main control loop, the interaction just described would disappear.

Another way of looking at the inherent interaction due to use of current-mode via sensing the primary current is in the frequency domain. Assume that the main loop of a current-mode supply has a high unity-gain crossover frequency, e.g., 10 kHz and a minus 1 slope for the Bode plot. Also assume that the magamp has a crossover frequency of 2 kHz. Then for perturbations in the main loop of less than 2 kilohertz the magamp will draw nearly constant average current from the transformer. In other words, the magamp rejects the changes in average voltage seen from the power transformer. However, for frequencies much above 2 kHz the magamp's control loop is not fast enough to reject these perturbations in the main PWM control loop and some of the

changing current will be diverted into the magamp.

The net effect as shown in Figure 2 is that the gain of the main loop drops in the region of the crossover frequency of the magamp. Remember that the gain of an ideal current-mode converter is directly proportional to the impedance of its output caps. At frequencies above the crossover frequency of the main output, the current-mode control begins to see the impedance of the output cap of the magamp in parallel with the output caps in the main 5-volt output. When there is a drop in gain there is also a phase shift.

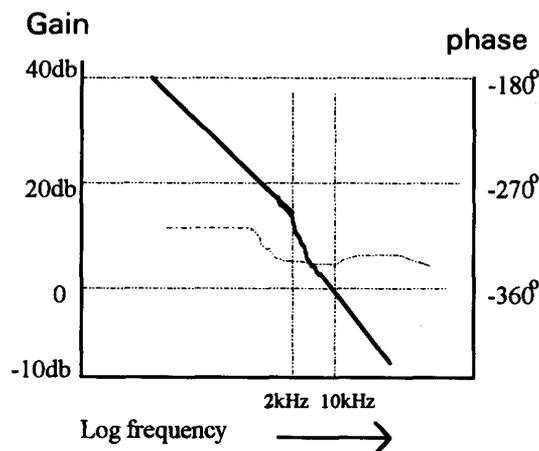


Figure 2. Typical Bode plot for fast 5-volt control loop with slower 12-volt magamp.

Some observations are in order.

1. To reduce the effect one can choose output caps for the magamp that have higher impedance (less capacitance and higher ESR)

but that requires a larger choke for the same ripple voltage on the output.

2. One should separate the crossover frequencies by one or more octaves to get good phase margin.

3. Normally it is easier to make the PWM loop faster than the Magamp, partly because of the inherent phase shift in the magamp due to the time between reset and saturation of the saturable reactor.

Two common ways to avoid the interaction in a current-mode supply are to 1) subtract the contribution of the magamp from the voltage analog of the primary current prior to that voltage being presented to the current-mode PWM or 2) sense the 5-volt current in the secondary via a current sense transformer. Either scheme removes the wiggles from the 5-volt Bode plot seen in figure 2. This is because the current in the 5-volt choke will now rise to its required value as determined by the error voltage (at least for small scale signals).

Another obvious way to avoid the interaction is to not use current mode. Use voltage-mode control for the main output.

Figures 3a) and 3c) show circuits that can be used to subtract the magamp's contribution to the voltage analog of the primary switch current if a current-sense resistor is used to sense the primary switch current. Figure 3b) shows a circuit that can be used when the primary switch current is sensed by a current-sense transformer. There are other circuits but these three examples convey the basic idea.

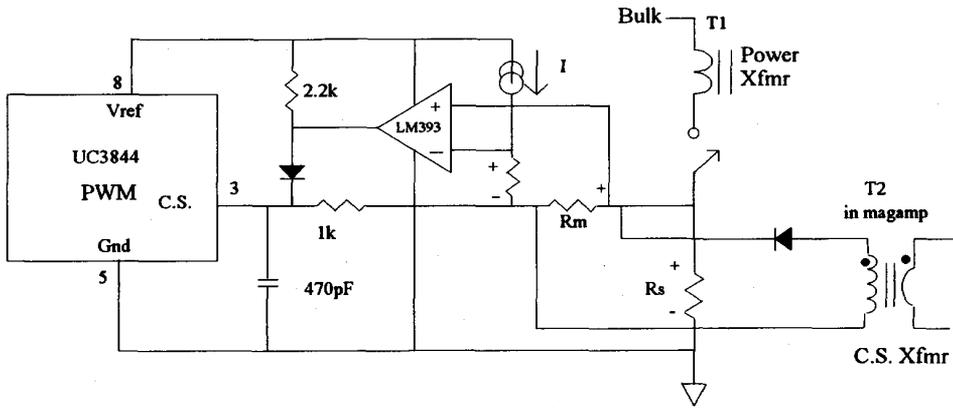


Figure 3 a) Subtraction Circuit for Primary C. S. Resistor

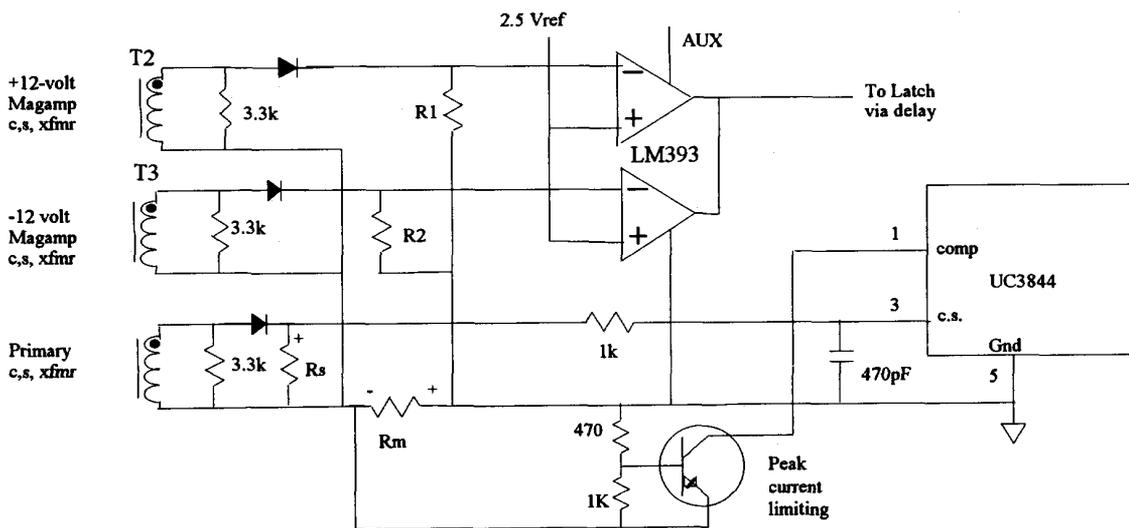


Figure 3b) Subtraction circuit for c.s. transformer in primary

An additional improvement of the circuit in figure 3b) is obtained by tying the return of T3 and T4 to the top of R_s instead of to the left side of R_m . Then value of R_m is reduced. New R_m is old $R_m - R_s$. This reduces the power in R_s and R_m plus makes the c.s. transformer more accurate since less voltage is required in its secondary, thus there is less magnetizing current in the c.s. transformer.

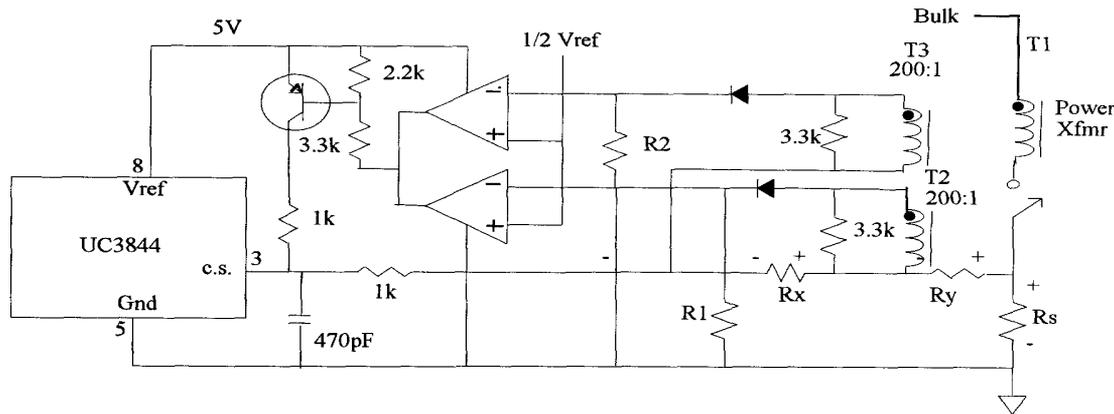


Figure 3c) - Improved subtraction scheme for 2 magamps
(T2 and T3 are c.s. transformers in the magamps)

A word of caution. When the power to the magamp is much greater than the power to the 5-volt output, the subtraction may not be accurate because of tolerances in component values. Also, even though the scheme works well for a forward converter, be leery of using it in a half or full-bridge supply. In the latter two, the magnetizing current of the power transformer jumps over to the secondary between pulses and unbalances the freewheel currents. (See ref. 5.) Some squirrely waveforms can appear in the subtraction circuit because of transients in the magnetizing current. One sees very few half-bridge magamp supplies. The interactions seen between the magamps and the main control loop is one of major reasons for this scarcity.

Obtaining the current for current-mode control via a current-sense transformer in the secondary of the power transformer for the main output also works well. Disadvantages are:

- 1) More leakage inductance energy is given to the primary inverter at the end of the pulse.
- 2) Above about 30 amps the current-sense transformer becomes larger and more expensive.

- 3) The magnetizing current of the magamp is lost thus the need for slope compensation is increased.

Source #2 - Interaction via Reset of Magamp

In the Type B magamp of ref. 2, the reset voltage is derived from a voltage other than the output of the magamp. The type B magamp shown in figure 4a) is normally used when the magamp has to provide short circuit protection via reset of the saturable reactor. As seen from the circuit model in fig. 4b), perturbations in the other output will effect the reset and thus can be a source of crosstalk. Normally, this feedback is not a factor in the lower-frequency range for which the error amplifier in the magamp has a gain much larger than 1. Also, if the reset voltage is from the main output there is some negative feedback at the lower frequencies since more reset will occur if there is an increase in the voltage of the main PWM controlled output. This will at the lower frequencies fight the tendency of the magamp's voltage to rise because of a higher average voltage from the power transformer. However, the crosstalk via this path can be significant at the higher frequencies.

Typical ways to address this problem are to:

- 1) Use a secondary auxiliary voltage for reset of the saturable reactor, or
- 2) In the case of a floating magamp, the voltage ahead of the saturable reactor can be regulated and used for reset.

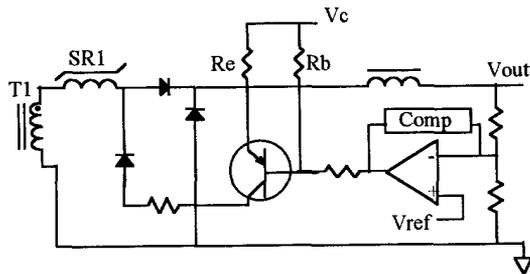


Figure 4a) Type B magamp

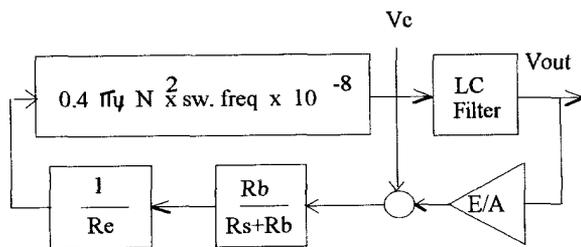


Figure 4b) Circuit model of type B magamp

Feed Forward - One advantage in having a magamp mated to a off-line current-mode supply is that current-mode control has inherent feed forward for rejection of the 120 Hertz ripple voltage on the bulk caps. The main PWM control loop removes most of the 120 Hertz ripple that would appear if both the main and magamp control loops were not fast enough to reject it. The author has seen magamp control loops with crossover frequencies as low as 50 Hertz that had less than 10 millivolts (peak-to-peak) of 120-Hertz ripple. If voltage-mode control is used then either the control loops must be much faster or a feed-forward feature must be added to the voltage-mode control scheme in order to reject the 120 Hertz..

Steering of ripple voltage -- The following is an example of a typical interaction that one

might see with magamps. One of the authors once had a supply with a 5-volt main output, a 12-volt magamp, and a low-current, minus 12-volt semi-regulated output. A large fan on the 12-volt magamp pulled current in 800-Hertz chunks. The unity-gain crossover frequency of the magamp was about 3 kilohertz, which did a good job of reducing the 800-Hertz ripple on the 12-volt output.

The main 5-volt loop had a crossover frequency of about 6 kHz which in turn did a good job of rejecting the perturbations of the magamp control loop. Imagine the author's surprise when he found a large 800-Hertz ripple voltage on the output of the minus 12 volt output, which had no post-regulator. In other words, the fan ripple current in the plus 12-volt magamp caused an 800-Hz ripple voltage on the minus 12-volt output.

To fix the problem, two turns were added to the transformer winding for the -12 volts and a linear post regulator was used to remove the 800 hertz ripple voltage from the -12 volt output. Then all of the 800 Hertz ripple was confined to the headroom voltage for the -12 volt. When the magamp control loop was disabled in an experiment, the 800 hertz ripple voltage disappeared from the -12 volts section and went back to the plus 12-volt output. The ripple voltage ahead of the -12 volt linear regulator was much larger than that seen on the plus 12-volt output when the magamp control loop was disabled (because of smaller caps for the -12 volt headroom voltage).

The above example gives insight into one of the most important rules to be followed in a magamp supply. **All of the control loops need to be fast enough to reject a dynamic load even if only one output is dynamically loaded.** How fast the control loops must be is a function of the di/dt of the load, the caps on the outputs, the inductance of the output chokes and how much attenuation is required.

Source #3 Interaction due to Hunting

There is a way in which magamps can interact with each other in a fashion that one early researcher has called 'hunting'. This phenomenon occurs when the saturable reactors in two magamps both saturate at almost the same instant. When a saturable reactor saturates, there is a short temporary drop in the power delivered to the main output as the current through the leakage inductance of the power transformer is diverted into the magamp. When there are two saturable reactors saturating at the same time, for example if a 12 volt and -12 volt windings have the same number of turns and near equal coupling to the main output winding, one sees a jitter in the leading edge of the pulses downstream of the saturable reactor in each magamp as each magamp control is trying to ignore the effects of the other magamp. This type of oscillation is low amplitude, is a normal phenomenon and in most cases can be ignored. The biggest problem is that "It doesn't look right on the oscilloscope". Also, sometimes it produces a low amplitude audible hiss (which is easily blanketed by the fan noise).

One of the authors has on several occasions added a turn to the -12 volt winding (the lower-current magamp) "to clean up the waveforms" so as not to confuse technicians on the production floor. The technicians had been trained to be suspicious of any leading edge jitter on a magamp's current or voltage pulse because in most magamps that is an indication of instability. The added turn separates the times at which the saturable reactors in the ± 12 volt magamps saturate.

One phenomenon that is sometimes associated with "hunting" is that 10 or even 15 degrees of phase margin may be taken away from one magamp's control loop and added to the other magamp. The mechanism behind the cause is not completely understood. However, the authors have seen it repeatedly in Bode plots. This is another reason in a 5v, ± 12 volt supply why one may wish to add a turn to one of the

transformer's secondary windings and prevent the 'hunting' phenomenon.

Source #4 - Interactions in a Power-Factor-Corrected Supply

PFC supplies using inherently noisy continuous-current boost circuits are notorious for the PFC stage talking to the downstream converter(s). Much of the crosstalk is via common-mode noise. Typically, the major crosstalk symptom is an audible low-frequency buzz with short glitches of about 1/2 milliseconds duration seen in the low-voltage outputs at a 240-Hertz rate. This is caused by the pulses of the downstream converter terminating a bit prematurely whenever the trailing edge of the PFC pulses and the downstream converter pulses match up. Then the output of the error amp in the downstream converter tends to slew high and often causes a slight overshoot as the trailing edges of the pulses are later separated. Typical counters to reduce this effect are:

- 1) Synchronize the PWMs (works for 90-to-264 vac range in a boost to 400 volts).
- 2) Put Faraday shield in power transformer.
- 3) Put balun in 400-volt bus.
- 4) Use voltage mode for downstream converter
- 5) Put some filtering in sense line ahead of error amplifier.
- 6) Put PWMs on different P.C. boards or put PWM of downstream converter in the secondary.

Normally, if the control loop of the main PWM output is made immune to the crosstalk from the PFC boost stage, then the magamp should be OK. Sometimes, a little filtering ahead of the magamp's error amp will be needed in the output voltage sense line. Also, layout can be critical.

Source #5 - Interaction due to 2nd inverter.

The magamp is wonderfully immune to most noise spikes - much more immune than either a synchronous-switch post regulator or a PWM down switcher. Once the saturable reactor has

been reset, no noise spike is going to cause it to set prematurely. Once the saturable reactor is set, no noise spike can cause it to revert back to its high impedance state. This immunity to noise is one reason why in the olden days, magamps were used as PWMs in military supplies for high radiation environments. However, the magamp post regulator is a slave to its parent main output which is PWM controlled. If another inverter or a downstream switcher is put into the supply, it generates noise spikes that can confuse the magamp's parent PWM. The inherent dependency of the magamp upon its parent can cause the magamp to produce erratic behavior. Also, the noise spikes associated with the magamp can have an effect upon the 2nd inverter.

The way to stop this nonsense is to use a PWM IC in the MC34065 family (basically two UC3844s in the same package). With it the two inverters can be made to operate 180 degrees out of phase so that there is never any crosstalk. The author once used a MC34065 to mate a four-output flyback with a four-output forward converter with 2 magamps. There was no crosstalk between the inverters or the magamps.

Source #6 - Parallel Operation

Yet another source of interaction can occur when two supplies are paralleled. Normally, since the magamp is time slaved to its own master PWM output, there is no crosstalk due to the noise spikes at the leading and trailing edges of the pulses. However, when two supplies are paralleled, there can be some crosstalk between the magamp of one supply and the PWM of the other supply or between the two PWMs. Sometimes, the magamp is unable to reject the quick changes in pulse width and also becomes part of the act. The symptom is usually a low-amplitude buzzing or faint frying sound. A similar phenomenon occurs when the paralleled supplies are separated by a foot or so because the user wants to nest his logic boards in between the two paralleled supplies. Here the two supplies

seem to be fighting each other for control of the output bus voltage. In general this is not a major problem. The author has found that a 0.001 μ F cap from the output of the error amp in the PWM to return reduces the hissing.

Oscillations -- Quite often when one first brings up a magamp prototype supply on the bench he sees oscillations. The first step is to try to determine if the oscillations are due to the main PWM control loop or to the magamp. Often that is not as easy as it sounds since the main loop may be stable by itself and only misbehaves when the magamp is connected. In general, the magamp is the chief culprit if the step up in the primary current waveform that occurs as the saturable reactor saturates has considerable jitter on its leading edge. Some small jitter in the trailing edge of the pulse is normal since the main control loop must reject 120 hertz. However, if the control loops are stable and there is no pulse loading then there should be almost no jitter on the leading edge of the step due to the magamp.

If the subtraction of the magamp's contribution to the voltage analog of the current sensed by the PWM is done - or if voltage-mode control is employed then the task of determining which control loop is the culprit is simplified. In a current-mode supply without one of these techniques being used, an oscillation in one control loop is practically guaranteed to cause a sympathetic oscillation in the other. (Oscillations can occur even when one of these techniques is used!) Often some trial and error iterations are needed to get the supply to behave, particularly if there is a poor layout and noise is getting into the control circuits. One can save a lot of time in the lab if he first takes steps to minimize the common sources of control loop interactions. (The amount of time saved increases roughly with the square of the number of magamp post regulators.)

One effect that is not a control loop interaction but often a factor in misbehavior of the magamp is the following: The reverse recovery current in the reset diode when the saturable

reactor saturates sees a forward biased collector-base junction in the reset transistor. Part of the reverse recovery current can find its way to the output of the error amp, travel around to the input via the feedback path around the error amp and thus confuse the reset circuit. Typical counters are to put a diode from anode of the reset diode to return (which also prevents saturation of the reset transistor during the pulse) or to put a small cap from the output of the error amplifier to return as shown in fig. 5.

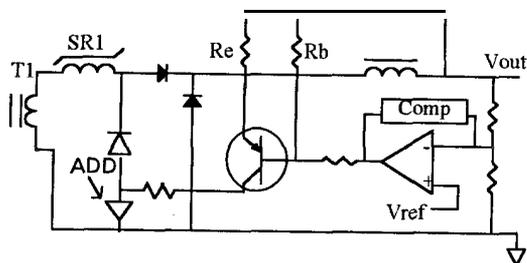


Figure 5 - Added diode prevents problem due to reverse current in reset diode.

Large scale transients - During large step loads, there is another source of interaction that has not been addressed in previous magamp papers. For example, voltage overshoots are common after a large step load - even with phase margins of 60% or more. Why? Typically, in a large scale transient, the output of the error amp is driven to either its upper or lower rail. There is a delay after the return of voltage in the output before the error amp can slew to its steady state value. What also limits the response to a large increase in output load is the headroom voltage to the choke and the value of the choke. (The headroom voltage is defined as that voltage that appears on the output of the magamp with no reset to the saturable reactor.) Often another turn on the transformer winding will do far more toward speeding up the response of the magamp to a step load than will a change to the compensation of the error amplifier.

A tactic that can reduce the overshoot is to put a clamp on the output of the error amplifier so

that the output of the error amplifier has less voltage to slew as it comes out of the stops.

Race problems during startup or during an abnormal are also common with use of the type B magamp (fig. 4.) in which the reset voltage is obtained from another output -often the main 5-volt loop. For example after a momentary short circuit on the 5 volt output, if there is no latch off, the main loop may go through some gyrations that tend to confuse the control circuit of the magamp. This effect can be reduced but not eliminated by use of a secondary aux voltage for reset of the saturable reactor in the magamp. In any case, the magamp is inseparably interlocked power wise to the main output. Anytime the error amplifier in the magamp is not satisfied with its output voltage it tends to slew toward a rail. A similar effect, although usually less severe, can occur with the type A magamp. There is no perfect world.

Transients can shock an otherwise stable magamp supply into oscillation. Control loops tend to oscillate at their unity-gain crossover frequencies if the oscillations are small in amplitude (control circuit remains linear). The audible noise emitted is a pure tone, usually in the kilohertz frequency range. In large scale oscillations, the control loop goes non-linear because the output of the error amp begins rail bumping and also, at low line the PWM is probably bumping up against its maximum pulse width limit every few cycles. The audible noise generated is usually a loud buzzing. Prevention of this type of oscillation can involve changing the control loop so that there is less phase shift at frequencies well below the small-signal crossover frequency. Also a change in the quiescent voltage at the output of the error amplifier or a change in the upper or lower rail voltage at the output of the error amplifier is sometimes needed to keep our magamp supply from being shocked into oscillation as a result of a large-scale load transient on one of the outputs. For large scale transients, the author recommends that one forget Bode analysis and let the magamp tell you what it is doing.

(It is always a good idea with any large-scale oscillation to look at the outputs of the error amplifiers to see what they are doing prior to going back to one's desk to do more control loop analysis.)

Regulating the Primary Aux Voltage

When the specification allows the load on each of the outputs to be zero while the other outputs are fully loaded, the temptation is to PWM regulate the primary aux and magamp each of the outputs. This approach can be implemented with voltage-mode control with some bleed on the aux if the aux has a LC filter. Enough bleed is needed to keep the current in the aux inductor continuous. The author advises not to try this scheme with current-mode control unless a large amount of slope compensation is used (which turns the control into voltage mode). Otherwise the control loop interactions can be severe.

If the main PWM current-mode control loop is closed around the aux (or any other output at light current), the subtraction schemes of fig. 3 may not be an option because the magamp's power will be many times the aux power. Sensing only the aux current for the current-mode control instead of the switch current may still be an option.

Another variation is to let the PWM run wide open (fixed duty cycle - no pulse-width modulation) and use a magamp to regulate each output. This removes the control loop interaction since there is no control loop for the inverter. The major disadvantages are that 1) the saturable reactors have to block a lot more voltage at high input line voltage and thus need more $NAe \Delta B$ product and 2) the big change in flux swing causes a fairly large change in gain of the saturable reactor. In general, this approach should be reserved for supplies that have a relatively narrow input voltage range, for example in a DC-to-DC supply and not in an off-line switcher.

Interaction in current limiting

Some of the most severe interaction that one is ever likely to observe in a magamp supply can occur if the designer chooses to do current limiting of the main output via cutting back the pulse width of the PWM for the main control loop. Normally, wild oscillations in current limit are the result. When the magamp senses an over-current and the pulse width of the main PWM is squeezed then the voltage control loop in the magamp shortens the delay time due to the saturable reactor. In other words the voltage first falls in the main output prior to falling in the magamp! The greater the foldback in current limit, the more difficult it is to implement this scheme.

With foldback, it usually takes a lot of trial and error to tame "the beast" that one creates in use of this approach. The authors' recommendation is to do current limiting via the magamp - or to do a latch-off of the PWM - or to put the supply into a burp mode.

The Flyback Magamp Supply

One of the most frequently asked questions about magamps is "How does one implement a magamp in a flyback supply?" There are two basic approaches for the discontinuous-current flyback - both of which come with their own set of control-loop interactions.

The first is the approach in reference 8. In this scheme, the flyback current is diverted away from the main output and into the magamp when the saturable reactor saturates. The shape of the primary and secondary currents are shown in figure 6. Obviously, if the main loop demands more current, the magamp will (unless its control loop is faster) receive most of the initial increase in energy. Ref. 8 does not satisfactorily explain how to address this problem.

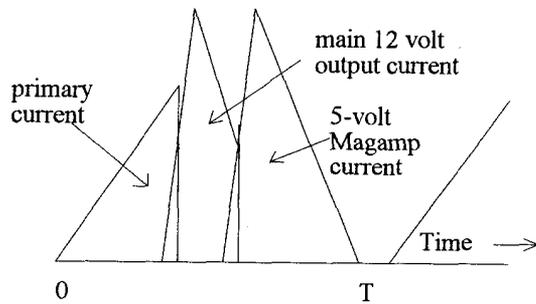


Figure 6. Currents in Flyback Magamp Scheme #1.

The second approach transfers power to the magamp during the pulse just as in a forward converter. The magamp has a choke and a freewheel diode. It looks very much like the magamp in a forward converter. The PWM is controlled by a flyback output. Now the currents as shown in figure 7 look much different. During the pulse there is a sharp increase in primary current at the instant that the saturable reactor saturates.

The authors have no lab experience with this 2nd approach. However, some obvious conclusions are:

1. With current-mode control in which the primary current is sensed, variations in line voltage should have little effect on either the magamp or main loop. There is automatic feed forward for both outputs.
2. Variations in load are more difficult to address. They are expected to produce huge control-loop variations.
3. Unlike the typical flyback, considerable minimum load for the main output is required unless the saturable reactor is designed to handle a large variation in the volt-microseconds that it must block.

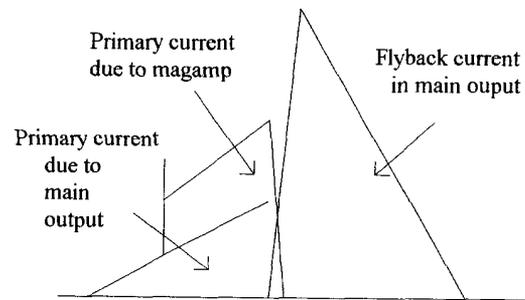


Figure 7. Primary and Secondary Currents for Flyback Magamp Scheme #2

Neither author has not done the math for this scheme nor developed the design rules for the saturable reactor. To derive the equations for static operation does not appear difficult. The math for transient behavior is far more complex.

Use of 2nd Inverter

Some of the techniques used to allow the main loop and the magamp post regulator to both have excellent transient response under extreme operating conditions can become expensive and in a development project for the first time can be time consuming. **Often, in a quick turnaround project, one of the best ways to avoid problems with a magamp is to not use a magamp.** Instead, consider use of a 2nd inverter to make the two outputs independent of each other - a goal which can never be perfectly achieved with a magamp since the magamp is inherently slaved to another output. The MC34065 family of PWMs or the CS3865 PWMs are good choices for avoiding the crosstalk so commonly associated with the use of multiple converters.

Summary

Several of the common types of interaction between the control loop of a magamp and other parts of a switching supply have been exposed. Practical design techniques to minimize these effects have been suggested. The two most important basic rules to remember are:

1. In a current-mode supply with magamp that does not sense the main output current either by c.s. transformer in the secondary or via a subtraction method then the unity-gain crossover frequencies of the control loop should be separated by one or more octaves.

2. All of the control loops need to be fast enough to reject a dynamic load even if only one output is dynamically loaded.

Other steps may be required to minimize interaction in PFC supplies, and to prevent oscillations as a result of large transients.

Acknowledgment

The authors wish to recognize Art Renkes, who in 1984 developed the first known circuit for subtracting out the contribution of three magamp post regulators from the voltage analog of the primary current in a current-mode power supply. Art was one of the first pioneers to study some of current-mode control-loop interactions described in this paper and to develop counter-measures to minimize their effects. One of Art's supplies was used by Unitrode as a test vehicle in development of the UC3838 magamp control IC.

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Appendix I

Subtraction Method - Additional Remarks and Sample Calculations.

The circuits shown in three figures 3a, 3b, 3c depict only 3 ways of implementing the basic subtraction scheme for avoiding most of the control loop interactions. There are a number of other circuits, some of which are in mass production. However, the same general design procedure for subtracting out the contribution of a magamp from the voltage analog of the primary current can be used for most of these primary current sensing schemes.

The basic philosophy is to sense the primary switch current. Then, from its voltage analog to subtract out the contribution of the magamp current prior to routing the remaining voltage to the current sense pin of the current-mode PWM that controls the main output. This remainder can also be used for current limiting of the main output.

A scheme that has been used very successfully is to have two stages of current limiting for the main output. The higher threshold is for a very fast pulse-by-pulse current limit using the 1-volt threshold of the current-mode PWM. The second stage is done either via a comparator with time delay with a somewhat lower threshold, e.g. , 0.7 volt, or by sensing a prolonged undervoltage on either the main output (or a slave output whose voltage is proportional to that of the main output). This 2nd current limit circuit with delay is used to set either a latch or to trigger a burp time-off circuit. The main reason for either a latch or burp circuit is to reduce stress on power components so that they will survive when there is an overload on an output that causes the fan to stop turning.

The voltage analog of the magamp current needed for current limiting of the magamp can be derived from the same current-sense transformer that is used in the subtraction circuit.

If one desires to protect the switching FETs in the event of a shorted main output rectifier, then a comparator can be used that senses the voltage analog of the primary current prior to its use in the subtraction circuit. The options available with various subtraction schemes are many.

Basic choices for sensing the primary switch current in an off-line current-mode supply are:

- 1) Use of a resistor between the source of the primary MOSFET and the return of the primary bulk cap or
- 2) A current-sense transformer with its one-turn primary in series with the primary of the power transformer.

A suggested basic design procedure for the design of a subtraction circuit is as follows:

Step #1. Assume that there is no contribution to the primary current from the magamp. Choose the current sense resistor in series with the switch (or the current sense resistor in the secondary of the c.s. transformer) so that the voltage seen across this resistor is the nominal one volt needed by the current-sense pin of a UC384X PWM for the following condition - The main output current is at its nominal knee current (at the current limit threshold). To this current add 1/2 of the peak-to-peak ripple current in the output choke. Reflect the sum of these two currents to the primary of the power transformer via the turns ratio of the transformer. For example, a 50-amp output at the knee added to 2-amps (due to ripple current in output choke) for a 5-volt output is only 4 amps in the primary if the primary-to-secondary turns ratio is 13. To this reflected current in the primary add the magnetizing current of the power transformer (perhaps another 1/2 amp).

Step #2 Calculate the voltage drop across the resistor due to the magamp current when reflected to the primary. For example, the turns ratio, primary-to-secondary, for a 12-volt magamp may be 26 to 6. **Assume any current in the magamp.** Reflect that current to the primary. Calculate the voltage drop across the resistor of step #1 that current will produce.

Step #3 Design a subtraction circuit that uses a c.s. transformer in series with the secondary winding for the magamp that will subtract the voltage drop in step #2 from the total voltage across the resistor in step #1 for the same assumed magamp current in step #2. Use one of the 3 circuits in figure 3 or any other circuit that achieves this goal.

No matter how accurate the calculations for the subtraction circuit, switching noise will cause results in the lab to be somewhat different. Some trial and error experimentation with a prototype unit is normally required for selection of final component values. For a single magamp, the following lab procedure is useful if the current limit for the main output is via the 1-volt threshold at the current sense pin of a current-mode UC384X PWM.

1. With the magamp load set for minimum, hold the line voltage constant and increase the load on the main output until current limit for the main output is reached. Record the main output current at the knee of foldback or at latch off.
2. Increase the magamp load to maximum. Again increase the load on the main output to its current limit knee.
3. If the knee current in 2) is larger than the knee current in 1), then there is too much subtraction. If the latter value is substantially smaller than the first, then there is too little subtraction. Change circuit values accordingly in the subtraction circuit to make the two values closer. In supplies with the possibility of a large magamp current and a very light main output current, it is best to error on the side of too little subtraction (to avoid erratic behavior when the main output is light loaded).

As always, there are side effects. For example, as the PWM pulse width is squeezed as a

result of an overcurrent in the main output, the output current in current limit tends to rise due to two effects.

- 1) The peak magnetizing current in the power transformer is less because there are fewer volt-microseconds in the pulse.
- 2) Noise filtering on the c.s. pin of the PWM prevents the full voltage from the subtraction circuit from being seen by the c.s. pin of the PWM. This effect becomes more pronounced as the pulse width decreases.

The use of most of the standard slope compensation circuits will also cause a rise in output current as the pulse width decreases (also a function of line voltage). Thus avoid placing tight limits on the current due to a short circuit on the output. The tendency of short circuit currents to be considerably higher than the knee current is a powerful argument for using a delayed peep off (burp) circuit for short circuit protection. (Another argument for a latch off is that the fan in most supplies stops when an output is shorted).

Example calculations for the Subtraction scheme of Figures 3a and 3b.

Circuit of figure 3a.

1. Choose R_s so that the sum of the main output current at its nominal current-limit point plus 1/2 of the ripple current in the output choke when reflected to the primary and added to the magnetizing current of the transformer, all times R_s is 1 volt. The formula for R_s is:

$$R_s = \frac{1 \text{ volt}}{\left[I_{out} + \Delta I_L \right] \frac{N_{pri}}{N_{sec}} + I_{mag} \text{ for XFMR}} \quad (\text{Eq.}\#1)$$

2. The value of R_m is chosen so that $\frac{I_{mag} \text{ amp} \times R_m}{\text{Turns ratio of c.s. xfmr}} = \frac{I_{mag} \text{ amp} \times N_{mag} \text{ amp} \times R_s}{N_{pri}}$

$$\text{or } R_m = \frac{N_{mag} \text{ amp} \times \text{Turns ratio of c.s. XFMR} \times R_s}{N_{pri}} \quad (\text{Eq.}\#2)$$

Examples:

1) Given: Nominal current limit for main 5V output is 30 amps. The switch-frequency ripple current for the 5-volt output choke is 4 amps peak-to-peak. The magnetizing current for the power transformer is 1/2 amp. Transformer primary turns are 26.5V turns are 2. The secondary turns for the 12 volt magamp are 6. The c.s. xfmr is 200: 1.

Calculations of R_s and R_m - circuit of fig. 3a.

$$R_s = \frac{1 \text{ volt}}{\left[30\text{A} + \frac{4\text{A}}{2} \right] \times \frac{2\text{T}}{26} + \frac{1}{2}\text{A}} = \frac{1 \text{ volt}}{2.96\text{A}} = 0.34\text{ohm.} \quad R_m = \frac{6\text{T}}{26\text{T}} \times 200 \times 0.34 = 15.7\text{ohms}$$

Calculation of R_s and R_m - circuit of figure 3b.

Use same data as for example above. Assume perfect subtraction and that all c.s. transformers have 200: 1 turns ratio.

$$\left[\left[30A + \frac{4A}{2} \right] \times \frac{2T}{26T} + \frac{1}{2} A \right] \times \frac{1T}{200T} \times R_s = 1 \text{ volt.} \quad \text{Solve for } R_s = \frac{1 \text{ volt}}{2.96A} \times \frac{200T}{1T} = 68 \text{ ohms.}$$

The equation for calculating Rm in circuit 3b is to recognize that the voltage developed across Rm from the 200: 1 c.s. xfmr in the magamp by a change in the 12-volt magamp current must equal the voltage developed across Rs due to the same change in magamp current reflected to the primary and sensed by the 200: 1 c.s. xfmr in the primary and Rs.

$$\frac{\Delta I_{12v \text{ magamp}} \times R_m}{200T} = \Delta I_{12v \text{ magamp}} \times \frac{6T}{26T} \times \frac{1}{200T} \times R_s$$

$$\text{Solve for } R_m = \frac{6T}{26T} \times R_s = 0.231 \times 68 \text{ ohmS} = 15.7 \text{ ohms.}$$

Circuit of figure 3c.

Here the math gets very complicated and is left as an exercise for the reader. Although the circuit has some advantages over those of figures 3a and 3b, the authors do not recommend it for the novice magamp designer. The circuit avoids the control loop interaction problem but has the disadvantage of a huge effect on the main output when either magamp is put into current limit. The safer approach is to let the output of the comparator set a latch.



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