

Intusoft Newsletter

Personal Computer Circuit & System Design Tools



Copyright © Intusoft, All Rights Reserved

Newsletter Issue #87, June 2017

In This Issue	
page	Story
1	A Generic Power Factor Correction, PFC, model
3	EMI Model
4	Behavioral sample and hold

A Generic Power Factor Correction, PFC, model *(Cont'd)*

Previously, a generic model was developed that relied on equations that described the in/out relationship of an ideal PFC. Now we will dig deeper into the model to see how a control loop can be added and how the generic average model can be transitioned to a switching model and further how to superimpose transient behavior so that EMI filters can be designed.

Controlling the PFC: The small signal gain of a PFC is controlled by the output/input power relationship. And that is independent of the switch mode topology. The basic equation is:

$$V_{in} \cdot I_{in} \cdot G_{in} = I_{out} \cdot V_{out}$$

Using G_{in} as a control variable; that is, G_{in} becomes a controlling voltage that has a much lower bandwidth than twice the AC mains frequency. Then the small signal Gain becomes:

$$A_v = dV_{out}/dG_{in} = V_{in} \cdot V_{in} / I_{out}$$

I_{out} can also be expressed as V_{out}/R_{load} . In any event a multiplier may be required in the control loop to account for variations in I_{out} , R_{out} or V_{out} . A constant value for these variables can be used to study the control loop characteristics, when variations are on the order of 3dB.

The following parameter based equations result in the P-I controller.

$$V_{(con)} = \{2 \cdot \pi \cdot F_c / (V_{rms} \cdot V_{rms})\} \cdot v_{(err)} \cdot 1/s \cdot I_{out} + C_{load} \cdot V_{set}$$

Where

$V_{(con)}$ is the G_{in} control variable

I_{out} is the average load current

V_{set} is the desired output

V_{out} is the average output

$$V_{(err)} = V_{set} - V_{out}$$

V_{rms} is the RMS mains voltage

F_c is the control loop bandwidth in Hz

C_{load} is the output filter capacitance

Control systems use feedback to correct error terms. The error for an output voltage controller is the difference between its desired output, the set point (V_{set}) and the measured output, V_{out} . The steady state error can be reduced to 0 if the error signal is integrated. Small signal control system theory tells us that the “best” loop gain is simply $1/s$, where $I = 2 \cdot \pi \cdot F_c$, and F_c is the closed loop bandwidth. But, the PFC is nonlinear.

The order in which equations are solved plays an important role when compensating nonlinear control systems. The partial transfer function, $I_{out} \cdot 1/s$ vs. $1/s \cdot I_{out}$, can be implemented by multiplying either before or after the integration. The simulation results for a step change in load are completely different as shown in Figure 1.

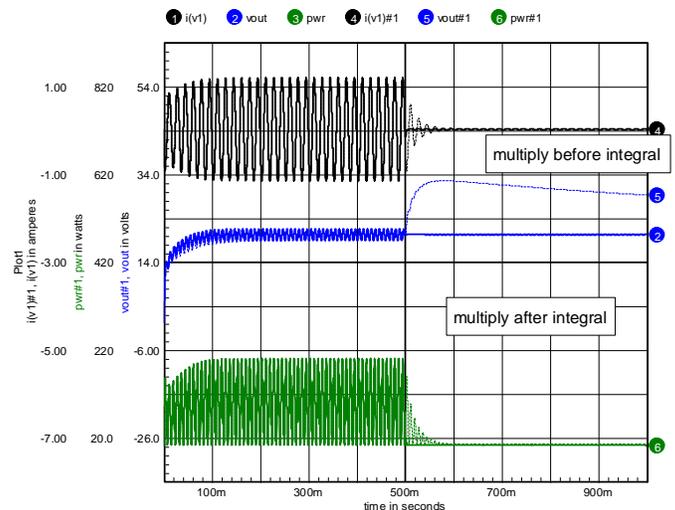


Figure 1. Load switch from 4 ohm to 20 Ohm to at 500 ms. Multiplying the error signal by I_{out} before integration leaves an unwanted error at the integrator output.

Estimating I_{out} : It may be inconvenient to measure I_{out} . The steady state value of I_{out} is given by:

$$I_{out}(frequency \rightarrow 0) = \pi / (2 * \sqrt{2}) * \text{abs}(I_{in}) * V_{rms} / V_{out}.$$

Using this estimate requires the value to be filtered to reduce the 120 Hz ripple, so a 2 stage filter was used and the results are shown in Figure 2. The performance is not quite as good as using the actual value of I_{out} ; but the result is acceptable.

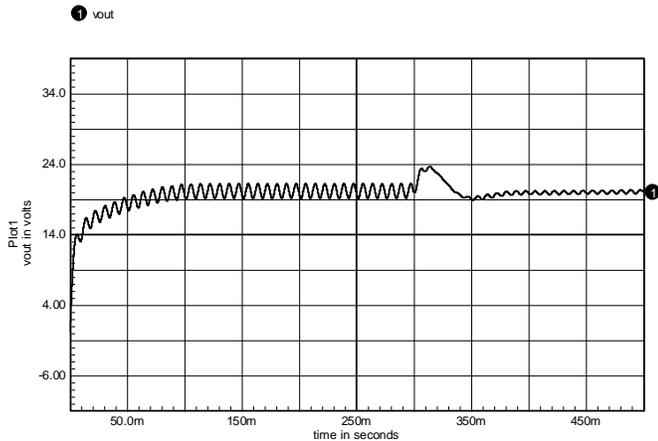


Figure 2. Pout with load switched from 5 Ohm to 20 Ohm at 300 ms some performance is lost using I_{in} to estimate I_{out}

How much filtering of I_{out} is required? I_{out} contains a 120 Hz ripple component. If the ripple is too high, the input current will be distorted with respect to the input voltage and the power factor will begin to fall. Table 1 shows the effect using I_{out} for the gain control

Table 1. AGC2 Filter Properties

Capacitor	Gain@120Hz	Distortion[1]	Power Factor
0.1uf	-1.3dB	37%	.88
0.3uf	-6.5dB	12%	.99
1.0uf	17.9dB	2%	.997

[1] $\text{Mag}(\text{third harmonic}) / \text{Mag}(\text{Fundamental}) * 100\%$

Critical Mode, CM PFC: Switch mode power supplies can be modeled using a 3 terminal equivalent circuit. Many topologies can be expressed with one switch forced on and the other using a freewheeling diode. Using a constant ON time that restarts when the inductor discharging current falls to nearly 0 enforces critical Mode logic. The switch arrangements are shown in Figure CM_topologies. The flyback mode is almost always used with a transformer (coupled inductor) so that the output voltage can be made positive.

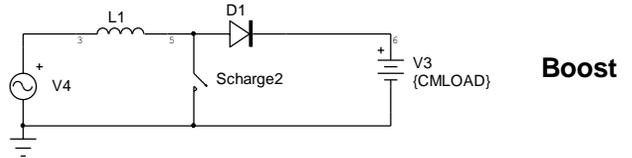
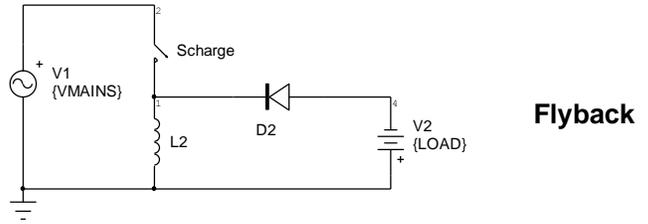


Figure CM_Topologies, The inductor switching logic enforces CM behavior.

Notice the input current in the boost mode contains both charging and discharging currents and the switching frequency changes over the half cycle of the AC mains frequency. This creates a problem for the flyback configuration since power is $i/2L * I^2 * \text{frequency}$ only works as a PFC for constant frequency. Waveforms are shown in Figure CM_Waveforms.

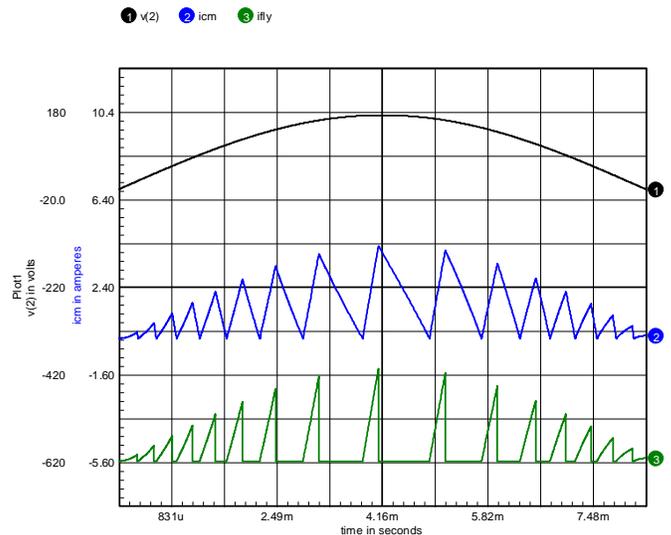


Figure CM_Waveforms, the 3 terminal switching models shown in boost and flyback mode.

A critical mode controller requires a constant Duty ratio and varies frequency with mains voltage. When connected In Boost mode, the inductor charging and discharging current are both seen at the mains terminal, making the average current proportional to the peak current. When the peak current is made proportional to the mains voltage, the desired PFC property is observed because mains current is proportional to mains voltage ($V = I * R$). However, when connected in Flyback mode, only the inductor charging current becomes the mains input current. The variable frequency results in a distorted current waveform,

defeating the desired PFC characteristics. Reference [Flyback PFC] Develops the equations in detail for those interested in perusing this further; however, the net result is that CM controllers aren't appropriate for Flyback converters used in PFC applications.

Input Current Control alleviates many of the problems occurring with output control. If the controller remains in discontinuous mode, runs at constant frequency and its on-time is constant; then, the result is a PFC controller. This can be achieved with most flyback controllers so that it is not necessary to use a "PFC" controller. The result is a constant power controller; that is, if the mains voltage is constant. Otherwise the output power will vary as the square of the mains voltage, and that's less variation than the 4th power behavior of an incandescent lamp. Using this approach and transiting to the previously described voltage controller for the open circuit state provides the control algorithms needed for LED lighting applications. Intusoft newsletter 28, from back in January 1983 describes how to make mode transitions using fuzzy logic.

Adding switching noise: So far the results are independent of the transformer turns ratio, N. That's because the average model is independent of the switch mode topology. The switching waveforms; however, are dependant on the topology. The model can be re-arranged to expose the bridge rectifier output as shown in Figure Bridge. That's where the switching ripple and other transients can be injected. Proper injection allows evaluation of EMI filters. Appendix B. develops the equations for a discontinuous mode, DCM, flyback PFC. A DCM controller running at constant duty ratio and constant frequency is also a PFC controller because the input current varies in proportion to the mains voltage.

functions, dxin and dxout. When divided by $(10 \cdot L \cdot F)$, they become the current charging and discharging waveforms. There are several reasons for using the intermediate dxin and dxout nodes.

The first reason has to do with controlling the simulation transient time step using VSECTOL. VSECTOL is used in analog behavioral models, ABM's, to control the time step when there are no explicit reactive elements. In this case, dxin and dxout reach numerical values that are hundreds of times larger than the charge and discharge currents. That allows the VSECTOL to operate on these nodes within the same-scaled values as the mains voltage.

Secondly, running the simulation using the average models discussed earlier and then switching over to the switched model accomplish the initialization of the transient analysis. The intermediate computation of dxin and dxout during this initialization period is easily eliminated using the behavioral (if, then else) rules as shown in figure "average-transient switch logic".

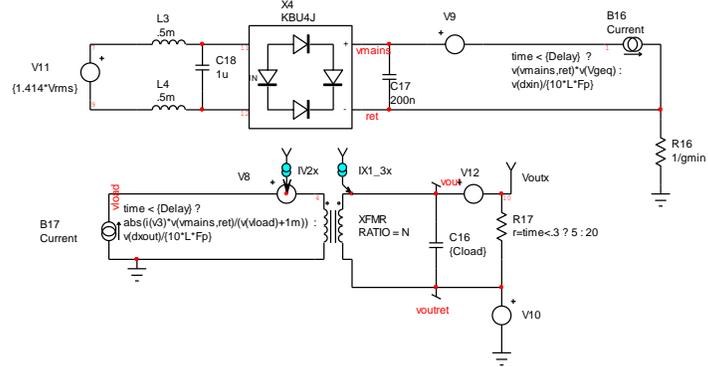


Figure "average-transient switch logic". Average-transient switch logic selects the proper equation for each mode

The complete schematic with equations defining dxout and dxin can be downloaded using the link on our home page. Notice R16 and V10 are removed when constructing a hierarchical or subcircuit model so these nodes can be isolated. Of course there must be a ground somewhere in the higher-level circuit.

Common mode noise: capacitive coupling of the switch voltage to the output circuit causes High frequency EMI. This noise circuit is completed through the system ground. For test purposes, a line impedance stabilizations network, LISN, connects its input ground to the output ground connection. Magnetic coupling occurs through an effective shorted turn in the PCB ground, coupled from the switched inductor primary. Figure EMI shows the equivalent circuits. These transients can be superimposed on the previously developed switching model using SPICE generic voltage and current sources. This test circuit is a fair representation of what happens using a USA 3-wire grounded system. However, the relation between ground and the circuit under test isn't very well defined for European systems

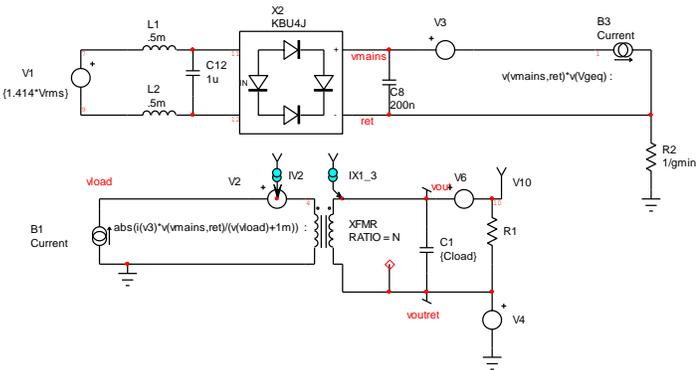


Figure Bridge, The average model with an external bridge rectifier.

Modeling inductor charge and discharge current: The flyback inductor is charged to a current proportional to the input voltage. It is then fully discharged into the output. These 2 currents can be characterized using intermediate

that do not include a mains ground. The LISTEN solves the problem for EMI testing, but does not provide a realistic model coupling the output tin the AC mains input in the actual application.

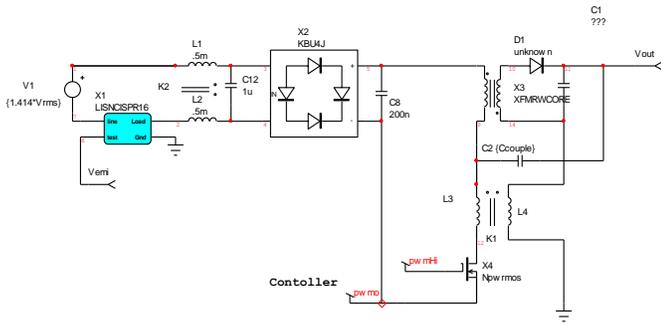


Figure EMI, The EMI equivalent circuit depends on LISN grounding to complete the common mode circuit.

L3 is an estimate of measure of the PWB inductance in the switching loop. Typical values are on the order of 100 nHy. L4 is the estimate of inductance in the PWB load capacitor loop, on the order of 30 nHy. Ccouple is an estimate or measurement of coupling capacitance from The MOSFET drain to the output and is typically on the order of 50 pF. It's a combination of heat sink coupling and transformer coupling. K1 is the coupling coefficient, typically around 0.1. The LISN ground completes the circuit with the output ground. The Listen can be placed in series with L1 or L2 and similar results should be obtained. You will need to adjust these numbers to get agreement with measure data. But, they should result in a ball park estimate for the initial design of your EMI filter

Real world EMI: The EMI standards are concerned with interference at the circuit input. For most systems, the noise affecting different system components powered from a device is left up to the system designer. A classic conundrum occurs when the output of one switch mode device is connected to the input of another. At steady state, the loading device appears to be a constant power device so that its input resistance in negative, illustrated by the following equations

$$P = I*V,$$

$$IdV+VdI = 0,$$

$$dV/dI = R_{in} = -V/I$$

Therefore, the load will have a negative real part as the frequency approaches 0. That could cause a system oscillation if the real part of output impedance of the first device, summed with the real part of the input impedance of the load is negative at any frequency. Cascading SMPS devices occur in many applications, from the power grid to your automobile or the airplane you may be traveling on right now!

Simulation Speed: The average model simulation runs thousands of time faster than switching model. The switching model is needed to evaluate EMI compliance. The switching model only needs to be run at “steady state”, that is, constant duty ratio for DCM controllers.

Therefore several steps are taken to improve simulation speed:

1. Run with the average model to get the initial condition for the switching model
2. Run the switching model with a constant Duty Ratio
3. Generate behavioral models for inductor charge and discharge current
4. Use a SPICE generic voltage source for switching model timing.

Finally, adjust the SPICE simulator options for speed and accuracy.

1. Increase tolerance limits to reflect power electronic levels
 - a. ABSTOL=1u
 - b. VNTOL=1m
2. Select auto ITL4, iteration limit
 - a. ITL4=0 (selects the automatic feature)
3. Select integration method
 - a. METHOD=Trapezoidal
 - b. RELTOL=0.005
4. Set VSECTOL=10u

Generally Gear integration is preferred over Trapezoidal; however, Gear integration tends to attribute some power loss to reactive elements, making circuits more stable; however, power supply efficiency is not accurately calculated. Set METHOD=Gear to check it out.

Behavioral sample and hold: The control loop can be sampled at the end of the steady state computation using R-C networks as shown in Figure SnH.

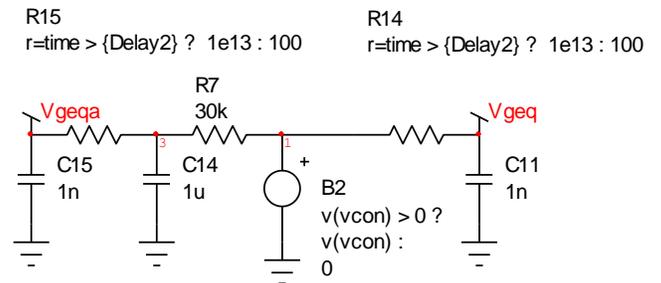


Figure SnH, Sample and hold using a behavioral switched resistor

The average control loop uses Vgeq for duty cycle control. Vgeqa is used to initialize the switched controller. R7-C14 reduces the 120Hz switching ripple in order to make the sampled duty ratio be independent of the 120 Hz ripple.

Sizing the Flyback filter capacitor is a function of the acceptable output ripple. Recently the filter has been removed altogether for off-line LED lighting applications, allowing your eyeball to do the averaging. The result is that

the LED power pulsates at 120 Hz. The same thing happens with Fluorescent lamps, and the only problem is beat frequencies that may be setup between the lamp and other appliances such as a TV etc. When both are at the exact same frequency, you might see an annoying band across the TV screen. If the TV or monitors runs at a slightly different frequency you may see a rolling band. The second problem relates to reduced LED lifetime because of the rapid thermal cycle. This can be investigated using the Norris-Landzberg [NL] equation. This equation calculates the change in failure rate based on a known test. Back of the envelope calculations indicates that there is no significant acceleration in LED failure rate when using this technique.

Schematics that provide working simulations can be downloaded from the link in our home page. You can view these using the current demo software, although some circuits are too complex to run a simulation. To run the simulations you need a recent ICAP/4 package that can read File Type 18.

[Flyback PFC], eq7
http://www.digikey.com/web%20export/supplier%20content/Fairchild_261/mkt/led/High-Power-Factor-Flyback-with-Constant-Current-Output.pdf

[NL]Norris-Landzberg Equation,
www.nxp.com/files/archives/doc/rel_qual_info/FPCR_EPORT.pdf