SMPS Buck Regulator This is background information for those not versed in SMPS design and/or control system design.

Overview, Figure 1 shows the block diagram of a Buck regulator and its control system. The switched voltage, Vpwm, is averaged by the L-C filter. As R1 and R2 approach zero, the conversion efficiency approached 100%. The filter has resistive losses, R1 and R2 that affect the control system. R2, L and the switched voltage at Vpwm determine the ripple at Vo.

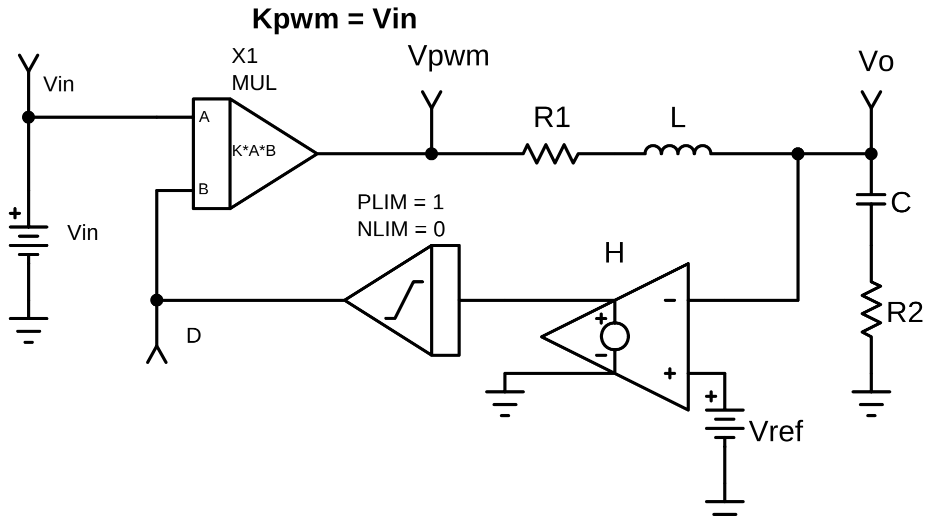

Figure 1, SMPS Buck Regulator Control System.

It's convenient to lump switch resistance, input filter loss and the current sense resistor into R1. Power supplies are generally specified to have some maximum output ripple voltage; usually on the order of 1% RMS of the DC output voltage. As switching frequency increases, R2 (the capacitor effective series resistance, ESR) decreases to some minimum value, then the effective series inductance, ESL (not shown) takes over and the ripple voltage is limited to Vpwm*ESL/L. The frequency defined by ESR/ESL/(2*pi) sets an approximate upper switching frequency where further reduction in L will produce increased output noise. Below that frequency, the inductor needed to meet the output ripple requirement becomes smaller with increasing frequency. Paralleling capacitors will increase this switching frequency limitation.

Increasing the switching frequency reduces size and cost of the inductor, which is a major contributor to overall size and cost. Notice that the series resistance and inductance are qualified with "equivalent". That's because these terms are nonlinear with frequency and are approximated at a specified frequency, usually 100kHz. More thorough SPICE models can account for these nonlinearities; however, the simplified model usually provides acceptable results.

Filter capacitors are characterized mostly by their ESR for power supply filters. Aluminum-Polymer capacitors have about 5 times lower ESR than conventional Aluminum electrolytic capacitors; costing about the same. Similarly Tantalum capacitors have lower ESR; but at a much higher cost.

Notice that neither C nor R1 play important roles in establishing switching frequency and output ripple voltage. R1 is responsible for most of the power loss and should be minimized, while C affects control system performance. The characteristic impedance of the open loop output is given by sqrt(L/C) so that selecting a high value for C is best. Both of these constraints tend to increase cost, calling for some engineering judgment.

PID calculation, A common control system feedback controller is the Proportional-Integral-Differential, PID, controller. It's featured in most digital control applications. That's somewhat different than used in the analog SMPS world where a dual loop controller is used; for example, peak current inner loop and PI outer loop. In this discussion, the optimum PID control coefficients will be derived for the previously describe Buck Regulator.

Plant: By inspection, the plant gain, A, is:

$$A = K_{PWM} \frac{R_2 + \frac{1}{Cs}}{R_2 + \frac{1}{Cs} + Ls + R_1}$$

$$A = K_{PWM} \frac{R_2 Cs + 1}{LCs^2 + (R_1 + R_2)Cs + 1}$$

PID: By definition the feedback gain, H, is:

$$H = P + \frac{I}{s} + Ds = \frac{I}{s}\left(\frac{D}{I}s^2 + \frac{P}{I}s + 1\right)$$

Canceling resonance:

$$\frac{D}{I} = LC$$

$$\frac{P}{I} = (R_1 + R_2)C$$

Then

$$AH = K_{PWM} \frac{I}{s}(R_2 Cs + 1)$$

Solve for PID: Let Fc be the controllers unity gain frequency, assuming R_2C is small, and solve for the PID coefficients:

$$I = \frac{2\pi Fc}{K_{PWM}}$$
$$D = ILC$$
$$P = I(R_1 + R_2)C$$

Next, synthesize the z transform coefficient with T = 1 / (Sample Frequency)

$$\frac{I}{s} = \frac{IT}{1 - z^{-1}}$$
$$Ds = \frac{D}{T}(1 - z^{-1})$$

Then applying the T scaling to I (I=IT) and D (D=D/T) using this scope5 script:

```
kd=12
T=13u
fc=4k
wc=2*3.14159*fc
L=6.8u
C=820u
R1=.1
R2=.02
I=wc*T/kd
D=L*C*wc/(T)/kd
P=(R1+R2)*C*wc/kd
DR=D/P
IR=I/P
```

And finally, the scaling used in the code

p = 206.088m
dr = 4.35897 , d=p*dr
1/ir = 11.9846, I = p / (1/ir)

For those who think T should be ignored, you can see the implementation would fail badly if you forget to account for T!!!

The zero order hold that models the A/D and D/A conversion process has zero gain at the sampling frequency which will mitigate the uncompensated zero at

f=1/(R_2C*2*pi) = 9.7kHz.

If normal Aluminum electrolytic capacitor was used, the zero moves 2kHz, below the cross-over frequency and would significantly destabilize the controller.

The way A and H are presented in the control system equation, the gain from the PWM to the output is 1/H. That's the same form as the output/input transfer function, illustrating the lack of input voltage disturbance rejection.

The response to output disturbance (current step) will be the open loop impedance divided by the excess loop gain at resonance.

dDVout = sqrt(L/C)*1/(1+AH)

Since resonance occurs at the loop cross-over,

dVout= sqrt(L/C)*mag((1 /(1+j2))=40mv/amp

Discussion: Pole-Zero cancellation and derivative feedback terrify control system designers. First, if he pole-zero pairs to be cancelled are under damped (high Q), then component tolerances can cause the poles to occur at a lower frequency than the zeros. That causes a severe phase lag-lead at resonance that could result in low gain instability. When component tolerances are included, the cancellation becomes imperfect and the improvement in output sensitivity from the controller gain is lost. Differentiation amplifies noise. In this case, there is a systematic noise in the output voltage caused by A/D quantizing. The net integration in the H term gives rise to a transfer function from the A/D to the PWM that increases up to Fc. That noise at Fc reduces dynamic range and results in a mini-pwm action near the cutoff frequency in order to average the error to zero. If the bandwidth is increased, the noise becomes higher and looks like a sub harmonic instability. It turns out that selecting a low bandwidth and relatively large capacitor gives acceptable results for PID control.

Using current feedback eliminates the differentiation and produces reduced sensitivity to input disturbance. Current measurements can be subject to EMI generated noise, and for synchronous buck regulators must include both positive and negative current values. Using a Kalman filter approach, the bi-directional current can be inferred using current as a hidden variable in the plant model. Noise is dominated by predictable A/D quantizing. The Kalman approach can be optimized for quantizing noise without the need to continuously update the gain coefficients.