Introduction to Forward Power Converters Utilizing Active Clamp Reset

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Hello, and welcome to today's National Semiconductor's online seminar. I'm Cory Coronado (phonetic) and I'll be your host. Before we begin, I'd like to go over the operation of the seminar interface. Slides will appear in the upper right section of your interface. If you wish the slides to be larger, click the "Enlarge" button. Slides will automatically advance or you can click on the slide title in the list on the left to jump to that place in the presentation. At the bottom of the interface is an interactive Web browser set to the Web page containing additional resources for this seminar. Questions may be submitted at any time and the presenter will respond via email. To ask a question, click the "Ask a Question" button, then fill in the form that appears and submit the form. Finally, National Semiconductor owns and is responsible for all the content in this seminar. Today's topic is Introduction to Forward Power Converters Utilizing Active Clamp Reset. Today's seminar will be given by Bob Bell, Applications Engineering Manager. Welcome, Bob. Please go ahead with the seminar.

BELL:

Great, Cory. Thanks a lot. I appreciate the opportunity to be here today and talk about this topic. It's quite a popular topic for DC to DC power converters, which are using more and more of this active clamp reset. So today I'd like to take this opportunity to introduce this topic for those who are new to the idea and maybe new to power electronics and then a little bit later on we'll have a little bit more advanced material. Those who have been doing this for a while will also see some second order more interesting challenges. So with that said, it won't necessarily be a fit for everybody. Some folks may find this a little bit redundant in the beginning and then the same folks maybe later on will be thinking that some of this stuff is a little bit advanced. So again, we're going to try to do our best to cover a lot of ground today. We have about 40 slides, and we're going to start right now.

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Outline:

Buck & Forward Topology Introduction

Active Clamp Theory of Operation

Transformer Design Example

Voltage Mode and Current Mode Control

Features and Challenges for Active Clamp Design

**Introduction to Active Clamp Controllers**

Interleaved Active Clamp

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The outline of the presentation is going to be, we're going to start with the buck and forward topologies. We're going to introduce those, look at the similarities between them, we're going to looks at the active clamp theory and how that pertains to the forward topology to reset the transformer. We're going to do a design example, looking at a transformer for a demo board that we recently did, a 3.3 volt output with a 48 volt input. We're going to look at control aspects, both the voltage mode and current mode control. We're going to introduce those topics and look at some of the differences between the two. We're going to look at some of the features and challenges associated with this active clamp design, and we're going to introduce some controllers that National Semiconductor has specifically designed for the active clamp topology, and then finally we're going to wrap up looking at interleaved active clamp topology where we basically are taking two converters and we're going to interleave them together to effectively double the output power.

Buck Derived Power Converter Topologies
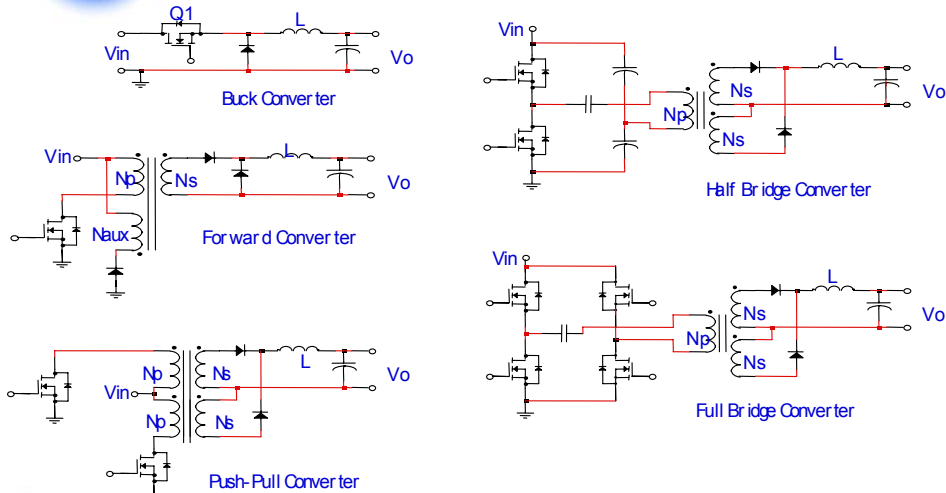
Buck Converter

Forward Converter

Push-Pull Converter

Half Bridge Converter

Full Bridge Converter

Shown on this slide is what we call buck derived power converter topologies and you'll notice that they all have one similarity and that's basically on the right side is an LC output filter and what we call the free-wheeling diode associated with that. And so when you look down the list here, you'll see that each of those share those three components configured in the exact same way on the output. In a general sense, the output power capability increases as you work your way down the list or down the columns and then back up again. So we have the buck converter and we're going to look at that in the next slide in more detail. You'll notice on that particular converter the input and output grounds are common and they're one and the same and there's no transformer, just the output inductor. We look at the forward topology, which we're going to spend a lot of time looking at today. You'll see that that incorporates a transformer, but again, the output is still the same. We have the free-wheeling diode and the LC output filter. And so basically the balance, the push-pull, the half bridge, and the full bridge are essentially just the same in that they reconfigure the transformer some and perhaps in a later presentation we can talk about those, but for today we're going to start with the buck converter, and we're going to move on to the forward in much more detail.

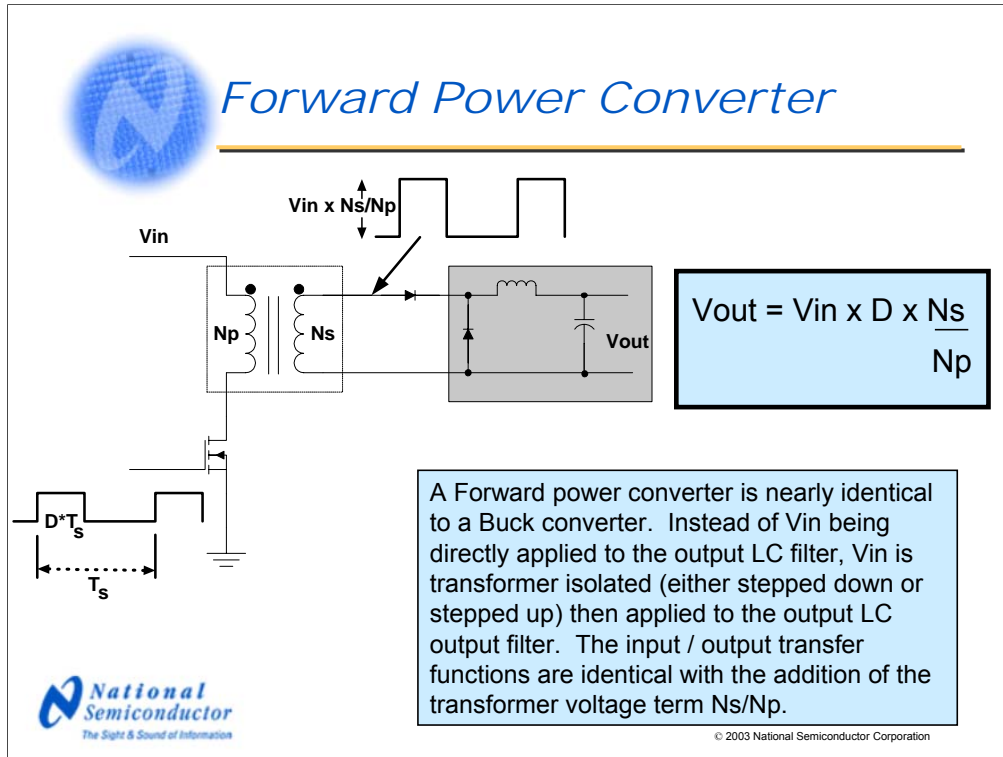**Buck Regulator Power Stage Waveforms and Characteristics**

Vout = Vin x D
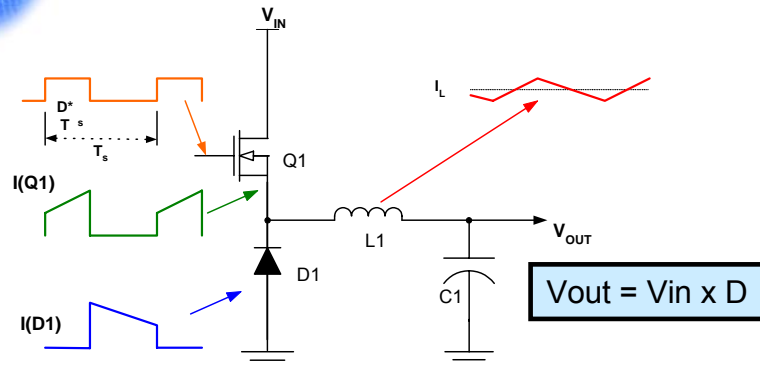
- Non-Isolated Grounds
- Voltage Step-down Only
- Single Output Only
- Very High Efficiency

- Low Output Ripple Current
- High Input Ripple Current
- High Side (Floating) Gate Drive Required
- Large Achievable Duty Cycle Range
- Wide Regulation Range (due to above)

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The characteristics of a buck regulator, hopefully many of you have seen this before, is basically it's used for step-down type of purposes. Very efficient for the input and output. The output is always lower than the input. The grounds, as I said before, are one and the same. They have to share the input and output ground. The efficiency is very high and because of the way the LC output filter is configured on the output, basically the output is fed by the inductor, which can be viewed as a current source. The output ripple voltage is generally quite low. That said, the input ripple current tends to be quite high because if you look at the characteristics of the buck current in the switch, you can see that it's a pulsing current, so the input ripple current can be quite high.
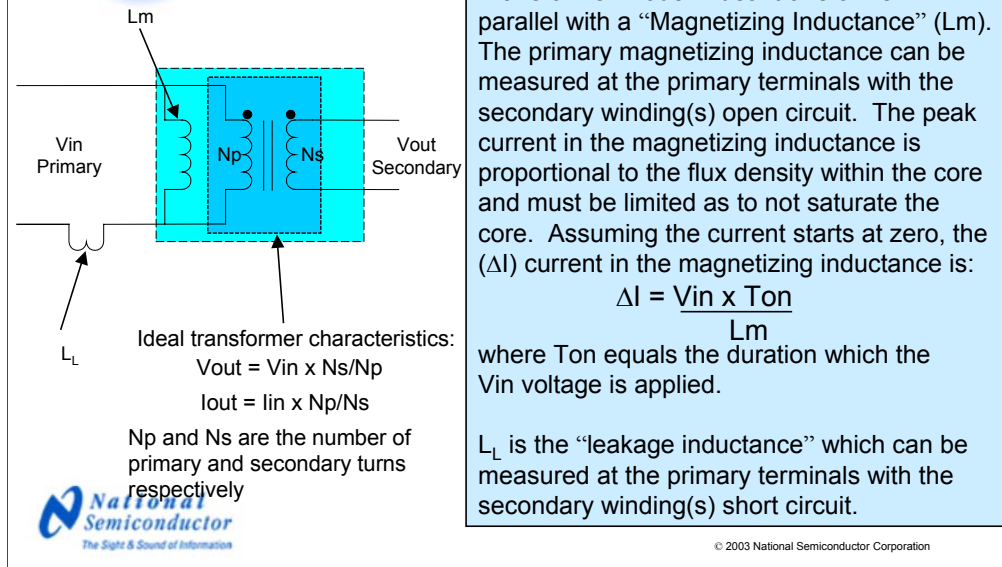
Forward Power Converter

Vin x Ns/Np

Vin

Np   Ns

Vout

$$Vout = Vin \times D \times \frac{Ns}{Np}$$

D*T$_s$

T$_s$

A Forward power converter is nearly identical to a Buck converter. Instead of Vin being directly applied to the output LC filter, Vin is transformer isolated (either stepped down or stepped up) then applied to the output LC output filter. The input / output transfer functions are identical with the addition of the transformer voltage term Ns/Np.

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So another important point is that the output voltage is quite a very simple equation. The Vout equals the input voltage times D, where D is the duty cycle of the main modulating switch Q1, and that's shown there on the slide as well.

Now the important point I want to bring across now is we're going to look at a forward power converter and if you look inside the shaded box, it's identical to the buck regulator. And so instead of applying the input voltage directly to that, you know, when the buck switch turns on and applying the input voltage directly to that LC output filter including the diode. So the forward power converter, it's almost the same except we have a transformer there this time. So what's going to happen now is that when you turn the switch on, which is ground referenced, it's going to act as a short circuit. It's going to apply the input voltage to the primary Np of the transformer and then assuming just for sake of discussion that the turns ratio of this transformer was one to one, so when the input voltage is applied to the primary winding, you'll get an exact copy of that voltage on the secondary side. And that's identical to what a buck regulator is. One of the advantages of the forward topology is that you can change the turns ratio to make the effective duty cycle bigger and you can also get isolation from the input to the output grounds, which is also a very important feature.

And also another thing you can do is you can actually take the turns ratio and actually do a step up equivalent as well. So for a forward power converter, it's not mandatory that the output voltage be less than the input voltage. And so the characteristics of the transfer function of the forward power converter is the output voltage Vout equals Vin times D, just like the buck regulator except now we have a new term which is Ns over Np. And Ns represents the number of secondary turns and Np represents the number of primary turns, where sometimes the two together is referred to as the turns ratio or more often than not it's a step down ratio.

Transformer Model

Transformer model: Ideal transformer in parallel with a "Magnetizing Inductance" (Lm). The primary magnetizing inductance can be measured at the primary terminals with the secondary winding(s) open circuit. The peak current in the magnetizing inductance is proportional to the flux density within the core and must be limited as to not saturate the core. Assuming the current starts at zero, the (ΔI) current in the magnetizing inductance is:

$$\Delta I = \frac{Vin \times Ton}{Lm}$$

where Ton equals the duration which the Vin voltage is applied.

L_L is the "leakage inductance" which can be measured at the primary terminals with the secondary winding(s) short circuit.

Ideal transformer characteristics:

Vout = Vin x Ns/Np

Iout = Iin x Np/Ns

Np and Ns are the number of primary and secondary turns respectively

The next area I'd like to introduce in a pretty good amount of detail is a model for the transformer. Since we're going to be talking about a forward power converter and resetting it. (the term "reset" we'll introduce in a minute). We have to have a more clear understanding of a model and we have to have a model for a transformer that we understand and that we believe in.

Shown here on this slide is a model. It's on the inner, darker blue box is what would be the transformer model from an idealized first order point of view. And that's the model you probably learned in school and it basically means, as I talked before, that if you put a voltage on the primary, on that Np, assuming a one-to-one turns ratio, you'll get the exact same voltage on the secondary side. The current is transformed in a similar manner. You know, put one amp in, you get one amp out, again assuming that the turns ratio is the same.

And so if you look at that equation that's down beneath there, you can see that you can change the turns ratio and get a step up or step down. And then, of course, the current changes accordingly. Obviously, the input power and the output power have to be equal. So that's hopefully a fairly easy to understand first order model, and that's very accurate. It works very well.

Now we're going to start looking at a second order model. What is actual when you go to build the transformer, you have to build this transformer on a core and what happens is that as you – if you take an inductance meter and you measure the primary inductance with the secondary open circuit, you're going to measure an inductance. And that inductance is referred to as the magnetizing inductance and that can be measured, as I said, with the secondary open circuit and an inductance meter, connected to the primary and you're going to measure this magnetizing inductance. The way to model that is directly and parallel with the primary. And so when you look at the current that goes into the primary, there's actually going to be two terms. There's going to be the one that's associated with the ideal transformer that flows into the Np and there's going to be a second current that's added or in parallel with that goes through the magnetizing inductance.

So to understand this, there's an equation that I have here. You can predict what the delta ' I ' or what the increase in magnetizing current, the current in the magnetizing inductance is simply the input voltage or the voltage applied to the primary times the on-time or the quantity of time at which you apply this voltage divided by the magnetizing inductance value. So, every time you apply a voltage to the primary, you're going to see this small increasing current. The reason I say small is because more often than not, for forward power converters, the magnetizing inductance is quite big, in the order of hundreds of microhenrys.

A third term that we're going to add to our transformer model is what's referred to as leakage inductance and this inductance is shown in the lower left part of our model and that's referred to as LL. The basic idea of leakage inductance is that that's a term that is the inductance that doesn't couple from the primary to the secondary, so hence the term leakage inductance. And that, too, is quite easy to measure and because of the definition, you can almost intuitively understand how you would measure it. The way to measure the leakage inductance is to short out the secondary winding with a hard, short circuit and because you shorted the secondary, you've essentially shorted out the primary as well because of the "ideal coupling." And whatever inductance you measure then into the primary with the secondary shorted is going to be your leakage inductance.

So now we have this model. We have the ideal part, which is pretty easy to understand with the turns ratio and the input/output voltage and corresponding current is shown on the slide. We have the magnetizing inductance and a model for that and a way to measure it and we have our leakage inductance. So we're going to take this model and we're going to apply it to our forward power converter, we're going to see one of the challenges that we have in front of us immediately and then we're going to solve that challenge with the active clamp.

Forward Converter Primary Waveforms
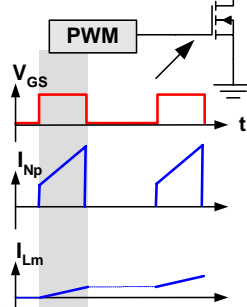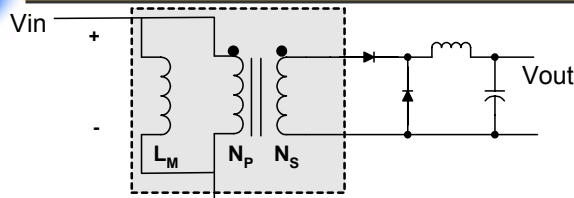
So again, here's just a piece of our forward power converter. This time I've put in – I guess we'll call it the second order model where we have our ideal transformer and we have our parallel magnetizing inductance. And so again, when we have our pulse width modulator and it's going to turn on their modulating switch Q1, the current through the primary, the main part, the ideal part is shown in the slide as well. And then remember, now we're going to start to have this magnetizing current that's going to flow as well through the magnetizing inductance, and when the switch turns on, we're going to have a polarity of plus to minus as shown across the primary winding and we're going to have an increasing magnetizing current that's going to ramp up linearly, to some value, maybe in the order of 100 milliamps.

Now at the conclusion, when you turn off the switch, that current will have reached a peak level and we have the equation for that on the previous slide, you have to do something. You have to force that current to return back to where it was initially before you start another cycle. And shown on the dash line on this particular slide was just suppose you did nothing where you didn't give any reverse voltage across this inductance to force that current back down to a reset level. If the current just stayed at the same level until the next cycle, then it would start to ramp up again and then eventually that current would just continue to ramp up and you have what's referred to as a saturation of the transformer.

So for a forward power converter there is this overhead that's required, and it is basically to push that magnetizing current back down to zero or at the very least it has to be pushed back down to the initial value of what was flowing at the previous cycle. So it doesn't necessarily have to go to zero. It just has to start every cycle at the same place so that the delta current or the net effect or no DC terms. There's lots of ways of thinking about this, but the circuit that we're going to introduce now is a method in which to force that current back to the initial starting point.

Forward Converter with Reset Capacitor and Switch

Vin

Vout

L_M N_P N_S

PWM

+ V_C −

V_{GS}

t

I_{LM}

Vc > Vin
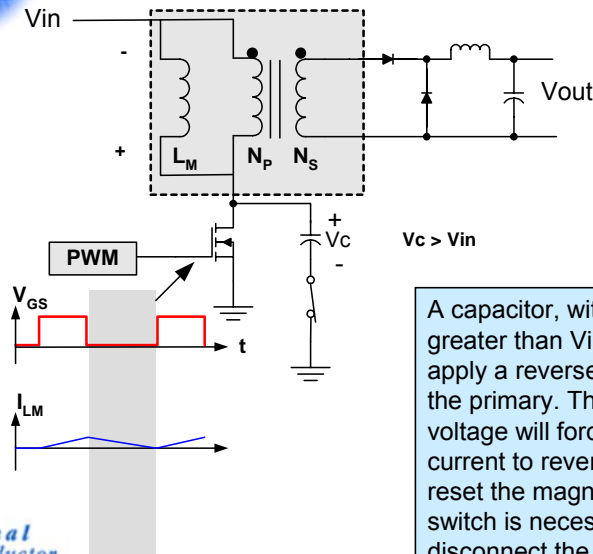
A capacitor, with a voltage greater than Vin, can be used to apply a reversed voltage across the primary. The reversed voltage will force the magnetizing current to reverse slope and reset the magnetizing current. A switch is necessary to disconnect the capacitor while the main switch is active.

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And so shown here is just the theoretical idea here that again when you turn the main switch on, you have the ramping up current that was shown on the previous slide in the gray area. Now, just suppose that we had this capacitor that was charged up to some value and that value was bigger than Vin and just supposed it was twice Vin for the sake of discussion. And if we had a switch in series with the capacitor that was shown, that when the main switch turns off, if we closed our reset switch, we'll call it, and we basically applied that Vc, that capacitive voltage across the drain of the main switch, if that Vc voltage was doubled, for the sake of discussion, Vin. Then we'd have a Vin potential shown on the slide as plus to minus across the magnetizing inductance and if the quantity of time was the same, say 50 percent duty cycle, then that delta 'I' would be the same and the current would ramp back down to right back where it began on the previous cycle. The reason you need the switch is because just prior to turning on the main power switch again, you need to open up that switch because you wouldn't want to discharge your capacitor through the main switch. So the operation of the reset switch will be that it will close immediately after the main switch opens up. Then it would open up just prior to the main switch turning on again for the next cycle. So that's the first order way in which we are going to reset it and we're going to now look at some ideal real circuits that can be used to implement this.
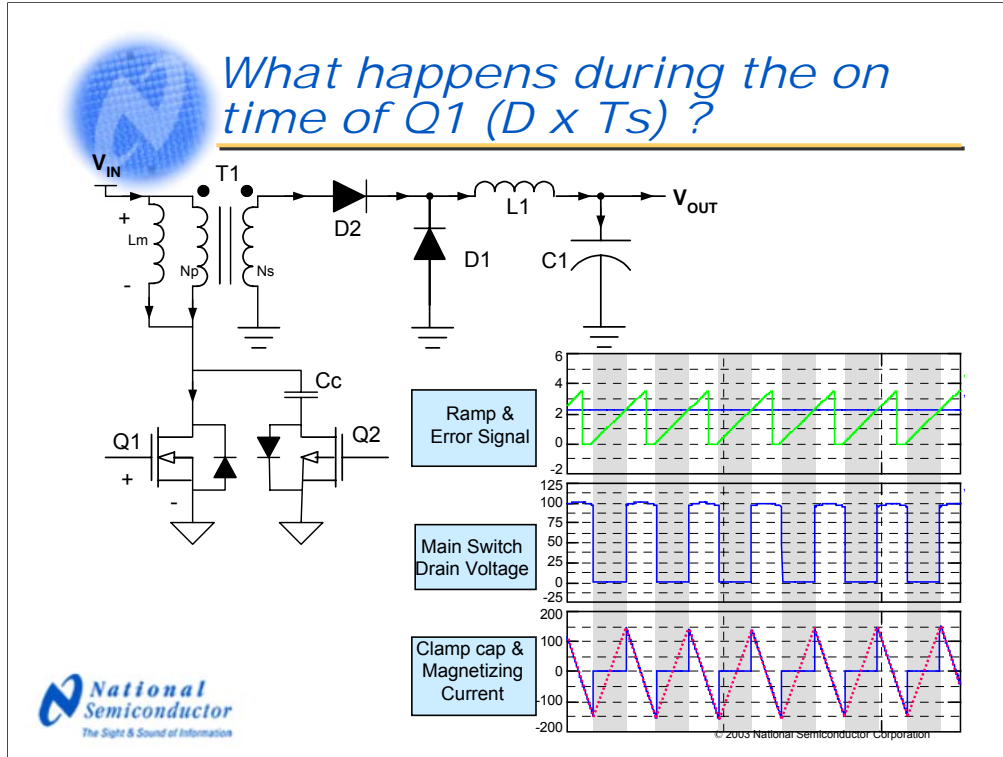
Early Forward Power Converter

 A Forward topology based power converter is very similar to a hand water pump.
• Pushing the water pump lever down, is similar to the main power switch turning on.
• The power stroke of the water pump, provides water (electrons) to the output.
• Following the power stroke, after the pump lever has been fully depressed, it is necessary to reset the pump by raising the lever again. This raising of the pump lever is similar to resetting the transformer.

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I know early on when I was thinking about how this works, one of the analogies I came up with was basically a hand water pump, and it's not very different from an analogy point of view of what a forward power converter is, in that, you know, when the main switch turns on, that's analogous to when you've pushed down the pump and you're basically providing electrons or water, you know, to the output. And that's a wonderful thing that works as you're applying it. But you're not going to be ready for the next cycle until you lift the handle back up, and that's basically resetting the pump and that's definitely analogous to how the transformer works in that you have to reset the transformer, get it back to the starting place before you can apply the power stroke again. So I hope that's a reasonable way to think about it for you.

V_IN T1

+ Lm Np Ns D2 L1 V_OUT

- D1 C1

Cc

Q1 Q2

+

-

Ramp & Error Signal

Main Switch Drain Voltage

Clamp cap & Magnetizing Current

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So now we're going to basically just say everything I've just said all over again, but we're going to do it with a little bit more practical circuits and some more waveforms to go with it. And we're going to break it down into about three or four sections and we're going to go through each time interval as shown in the gray. We're going to basically move these gray lines across and talk in detail about what's happening during each of these periods of time.

So the main power pump section, if you will, is comprised of when the Q1 turns on, you know, for the duty cycle that we talked about earlier for the transfer ratio. That's going to apply plus the minus across the primary winding as shown. You're going to get the secondary to be forward biased during that time. You're going to be providing power to the output LC. Diode D2 will be on and it's basically the power stroke, but also during that time you're going to have this magnetizing current that's going to be rising and increasing and that's shown in the lower waveforms in the dashed red line. The blue line in the lower waveform is the actual capacitor current or the switch current of Q2. So during that time the switch is open so that there's zero current on the blue trace. So another way of looking at this, too, is that the magnetizing current is going to be the sum of what goes through the Q1 when Q1 is on and it'll flow through the capacitor when Q2 is on. So the net of those two will be this triangle you can see where the positive slope will be the red portion and the negative slope will be the blue portion which will actually be going through Q2, or said differently, the red portion goes through Q1 and the blue portion go through Q2.

So again, during this on time, the point we're trying to make with this slide is that the current is rising, the magnetizing current is rising, it's flowing through Q1 and it's shown in the lower waveform.
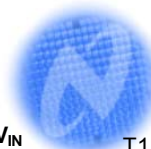
Core resetting during the off time, Part I
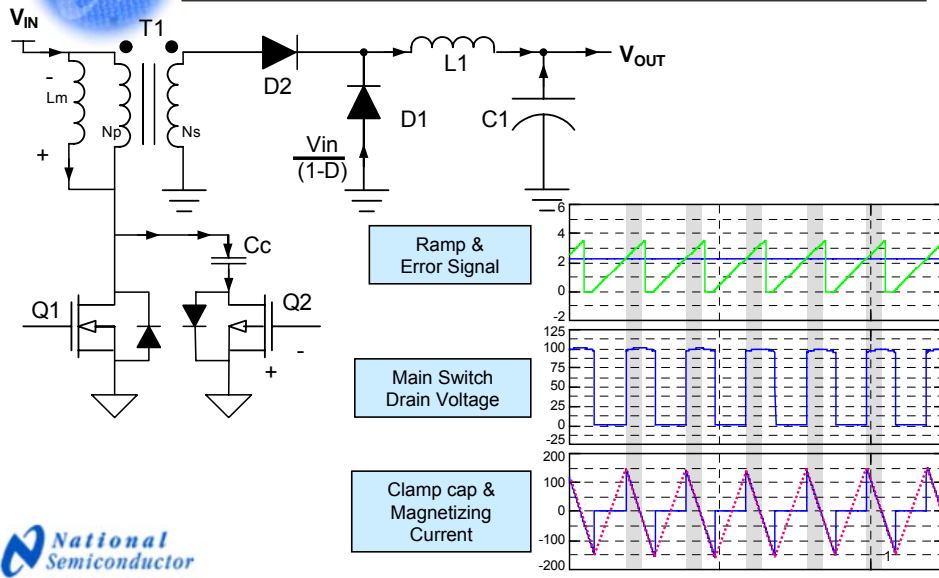
Now at this point in time, the main switch has turned off. The Q2 switch is going to turn on immediately after Q1 turns off. The magnetizing current, because it comes from an inductor and wants to continue to flow, that's what inductors do, the current will continue to flow down in the same direction that it was flowing. The instant that Q1 turns off, whatever magnetizing current was flowing through Q1, that exact same current, the same value will start to flow through Q2. Now because the clamp capacitor CC, the value of voltage across primary is going to be Vin divided by 1 minus D. That's an equation that we're going to look at in more detail later, but that value is much bigger than Vin and that's going to actually have a different polarity across the magnetizing inductance that's shown. It's basically reversed from what it was before and that's why the slope of the magnetizing inductance is going to change from a positive one to a negative one. You can see over that time the current is being sourced into the capacitor. The main switch drain voltage at that time is actually identical to the capacitor voltage and you can see that it has a small increasing voltage across the capacitor, and that's because you are sourcing current into the capacitor at this time. This is only a piece or half of the off time. On the next slide we're going to look at how it continues for the remainder of the cycle.
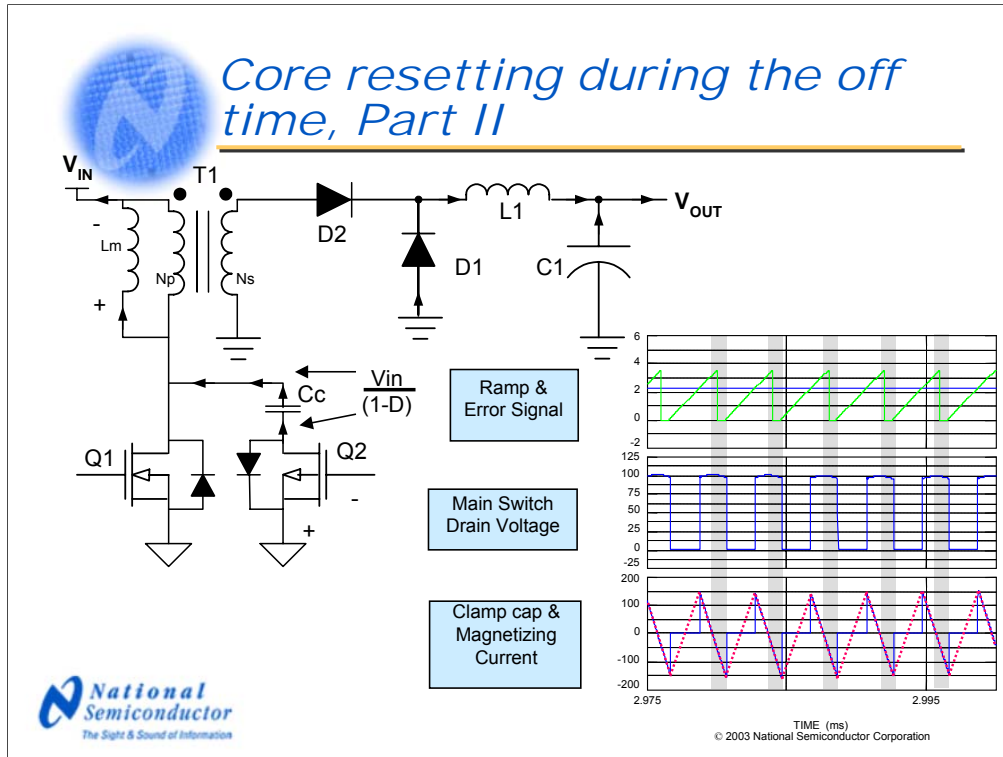
So again, the capacitive voltage, although changing a little bit, it's still relatively flat during this off time and so what we call reverse voltage across the magnetizing inductance is still relatively the same. It's still the polarity that's shown from the previous slide. But you'll notice now the magnetizing current has reached zero and gone through zero current. And at that point in time, the polarity across the inductor is the same and the voltage across the capacitor is the same that the current will actually start to reverse and start to flow upwards through the magnetizing inductance and back out to the input voltage. So that's basically a way of saying that because this current is flowing back into the source, that you're actually recycling the energy or pushing it back into the source and by virtue of doing this, you're also resetting the transformer at the same time.

So the beauty of this whole thing is that the magnetizing current or the energy from the magnetizing inductance is captured in the capacitor and it basically filled that capacitor up a little bit higher and then it eventually sources it back to the input voltage. And because of the charge balance of the capacitor and what we call a volt second balance of the primary inductance, you end up with this balance where the capacitive voltage will find a DC level that equals that Vin divided by the 1 minus D and by definition the magnetizing current will be reset back to its original value.

P-Channel Gate Drive

A P-Channel active clamp switch is commonly used, due to ease of drive. The controller active clamp output is in phase and overlaps the the main output. A capacitor and diode circuit invert the clamp switch drive signal, creating a negative Vgs with dead time.
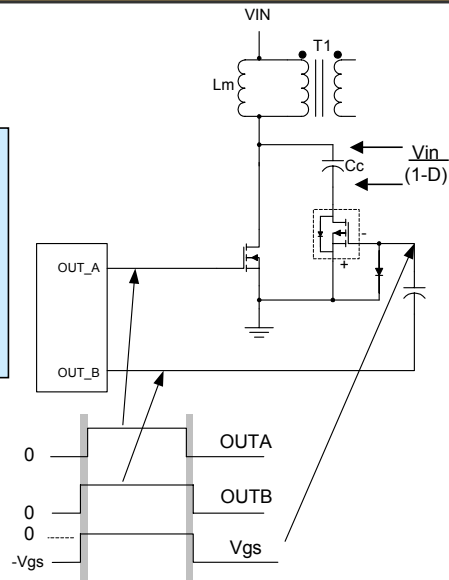
So what was shown on the last slide was an implementation of our reset capacitor and what was used as a P-Channel switch and so you might look at that P-Channel switch initially and say, well, gee, how does that really work? It seems a little odd. And it's actually quite clever. What it's being implemented as, it's sort of upside down compared to what a P-Channel is normally used as. And because it's upside down, it's a reversal of voltage and a negative voltage is going to be required to be generated. And a quite clever way to do that is shown in this particular slide where the P-Channel device is connected as shown. The controller, wherever that may come from, whatever is controlling the active clamp switch, remember it wants to turn the active clamp switch on when the main switch turns off. And so by using the capacitor and diode, basically if you look at the waveforms as shown here, when the B output goes high, the gate voltage will be clamped by the diode to 0.7 volts and then when the B output goes low, you'll end up with a negative voltage at the gate of the P-Channel switch, which will actually turn it on, so that an inversion takes place and you also generate this negative voltage that's required.

So it's quite an economical, easy way to implement the clamp switch with the P-Channel device. Of course, one of the limitations on it, you may know, P-Channel devices are only available with voltage ratings somewhat limited. They're not really available in much higher than 150-200 volt rating, but at the same time if you can live within those guidelines, you can get a very effective, easy to use clamp switch. And so to actually get dead time, since you don't want both of those switches on to be at the same time, you actually have to have your output B output overlap or envelope the main output as shown in this slide.

Active Clamp Using an N-Channel Clamp Switch
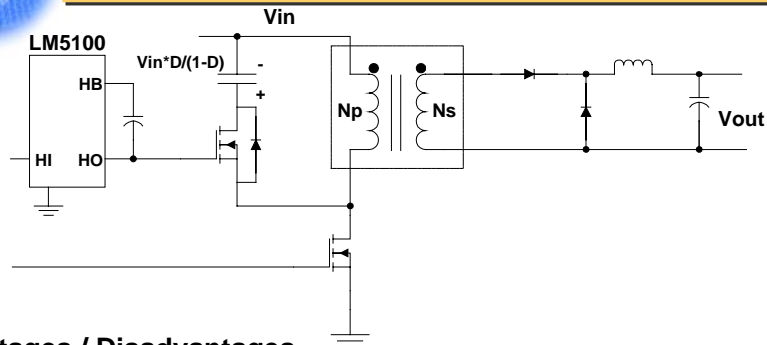
Advantages / Disadvantages
- N-Channel devices are less expensive, there is a limited selection of high voltage P-Channel devices available.
- Lower clamp capacitor voltage rating for N-Channel applications (lowered by Vin)
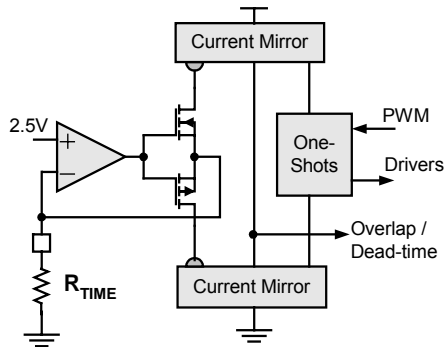- Level shifted gate drive is necessary to drive the N-Channel switch (Driver IC or Gate drive transformer)

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All that said, this whole implementation can be addressed with an N-Channel switch. Suppose you had an offline-type application with a very high input voltage, a P-Channel wouldn't be available, you could potentially use an N-Channel switch. You'd configure it as shown in this slide. The advantage of this approach is that, you know, N-Channel devices are less expensive than P-Channel devices and there's more available to choose from. Also, for the clamp capacitive voltage, if you go around the loop, if you will, in the previous configuration, you had a finite voltage that needed to be generated across the primary winding, plus you had Vin, where in this case when you're doing the actual resetting, you just need to have the voltage necessary to reset the transformer. This is because the capacitor is connected directly across the primary winding in this case. So the voltage necessary or the voltage rating necessary for the clamp capacitor is actually less with the N-Channel implementation.
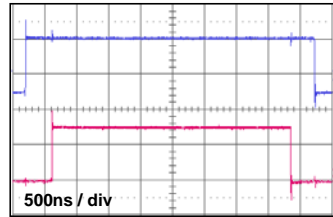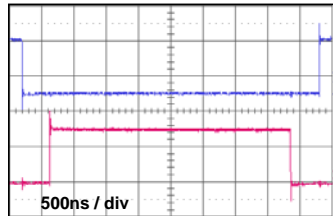
Shown on this slide is the timing of real waveforms from the output of a given controller, you could see for the P-Channel timing the main switch is the lower one, the red waveform as the signals leave the controller. You can see the P-Channel control switch or control waveform, actually overlaps the main output. Whereas, on an N-Channel timing, you need dead time between the two waveforms. The reset control was high, then it goes low, then there's a dead time where neither switch is on and then the main switch turns on. Over on the left side of the slide is a simple block diagram of the timing internal to one of our controllers where it can detect and configure itself for either of these types of applications. And we'll see later on, you want to have the ability to set this dead time. At the end of the day you really just want to minimize it as small as possible, which is dependent upon the propagation of the given FETs and gate driver that you have in your circuit.

Self-Driven Synchronous Rectifiers

Utilizing the active clamp approach allows relatively easy implementation of self driven synchronous rectifiers.

Vin

Np   Ns

Vout

2 1 µs 5.0 V

1 1 µs 5.0 V

LeCroy

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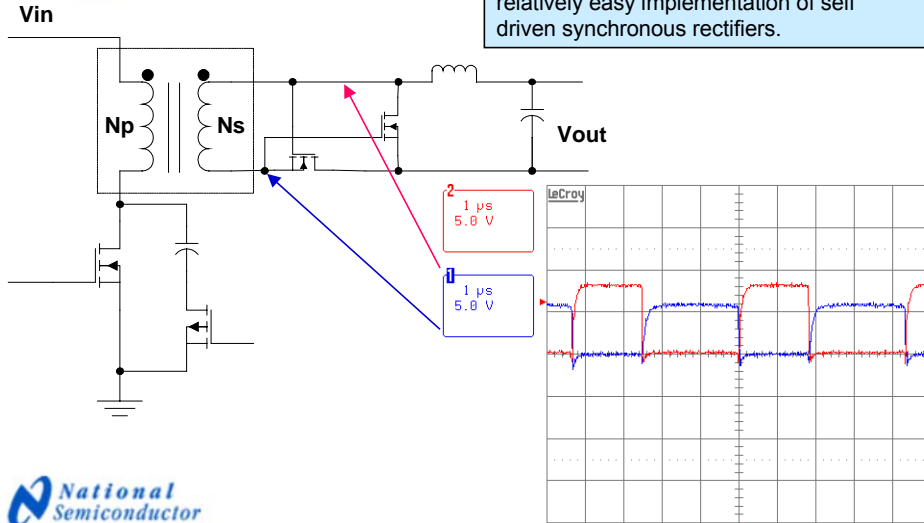
One of the advantages of the active clamp topology is that it lends itself well to self-driven synchronous rectifiers. Synchronous rectifiers are not unique to the forward topologies. They are used pretty much in every topology these days where high efficiency is required, particularly with higher currents and lower output voltages. And as you can see here are some real waveforms from a power converter and it shows the given node, where those waveforms were taken. You can see, it's almost as simple as shown here that you can connect your two FETs in this synchronous fashion and they can be driven directly from the winding. It depends upon what your output voltage is, you might have to put a couple of RC delays in there if there's some overlap still present. But in this particular case, that wasn't necessary and because of this 3.3 Volt output you could connect the synchronous rectifiers as simply as shown here and get quite a good efficiency improvement without a lot of additional circuitry.

So in summary, the advantages of the active clamp topology are that it basically provides what we refer to as an optimum reset of the core in that the voltage across the clamp capacitor finds the exact voltage that's necessary to reset the core. Because the polarity of the magnetizing current is both negative and positive, you're actually using two quadrants of the magnetic core, and we're going to talk about that in the slides up ahead. The minimum stress on the primary MOSFET allow the potential for cost savings, by not having to buy more voltage rating than is necessary. The leakage inductance energy is re-circulated. We didn't talk about that very much, but when the main switch does turn off, the leakage energy also flows into the capacitor and it is also sent back to the input voltage during the clamp reset time. Any energy in the leakage inductance is also recycled and there's no voltage spikes (on the FET drain) since they are basically clamped as well. Ease of self-driven sync rectifiers as described previously and the ability to have some zero voltage switching, particularly at the main switch turn off. We'll talk about zero voltage switch in the following slides.

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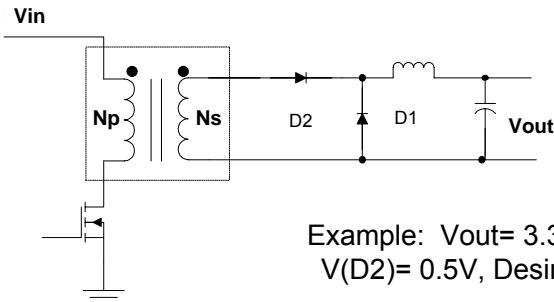The Sight & Sound of Information

TRANSFORMER DESIGN

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Next we're going to go through a transformer design with an example. If you don't follow it all as we go through, you can take your time and look at it afterwards.

Transformer Design
Picking the Turns Ratio

Vin

Np **Ns** D2 D1 **Vout**

Example: Vout= 3.3V, Vin(min) = 33V,
V(D2)= 0.5V, Desired D(max) = 0.65

$$\frac{Ns}{Np} = \frac{\dfrac{Vout}{D} + V_{(D2_{ON})}}{V_{IN}}$$

$$\frac{Ns}{Np} = \frac{\dfrac{3.3}{0.65} + 0.5}{33} = 0.1689 = \frac{1}{5.9}$$

Selected Np : Ns turns ratio 6 : 1

In the very beginning we talked about the ideal forward transfer function; Vout equals Vin times D times Ns over Np. If you take that same equation and reshuffle it around, you'll see that you get the equation as shown here where we're going to solve for what we'll refer to as the turns ratio, or Ns over Np. So Ns over Np, the turns ratio, equals the equation shown, which requires the output voltage, we're going to use the maximum duty cycle that we want, we're going to pick a voltage drop across the diode, D2, and we're going to use the minimum input voltage, Vin.

Now we're going to go through an example here where Ns over Np is going to be calculated for Vout equal 3.3 volts. That's going to be our output voltage. A very practical limit for a maximum duty cycle for active clamp is something in the order of between 0.55 and 0.65, maybe as high as 0.7 or 70 percent duty cycle. For mathematical purposes, we'll take 0.65, which is the same as saying a 65 percent duty cycle is what we're going to pick for our maximum duty cycle at low Vin our smallest Vin operating point. The real lowest Vin is 36 volts, but from a design point of view, we're going to put some margin on and use 33 volts. We put those numbers in the equation, we're going to see that basically we end up requiring a six to one turns ratio.

So that means that we need some integer number of six to one of the turns ratio, for the primary winding and the secondary winding. So we could have 12 and 2, we could have 6 and 1, 18 and 3, so on and so forth. We'll see in the next equation there are some bounds to where and how many turns we wan to use. So hopefully this is fairly clear this is an easy way to predict the required turns ratio for an Ns and Np.

Transformer Design
Selecting the Core Size

Delta flux swing (ΔB) for a magnetic core:

$$\Delta B = \frac{Vin \bullet \overline{D} \bullet 10^8}{Fsw \bullet Np \bullet Ae}$$

Core loss is proportional to ΔB and Fsw
Set ΔB for < 2000G

Continuing our example . .Set ΔB to 2000G,
rearrange the equation to solve for Ae,
pick primary turns (Np) and solve for the necessary
minimum core area (Ae).
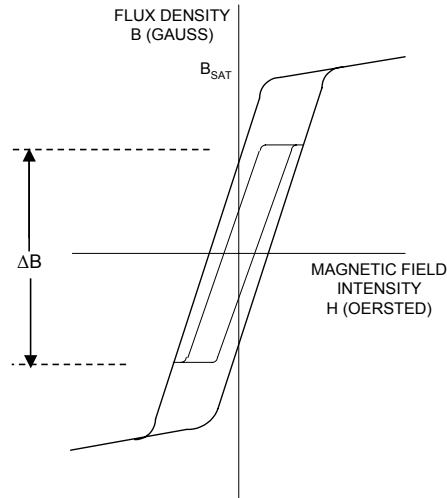Np must be 6. .12..18 etc

$$Ae_{(min)} = \frac{33 \bullet 0.65 \bullet 10^8}{250\,K \bullet 12 \bullet 2\,K}$$

For switching frequency (Fsw) = 250kHz,
Np=12 turns
Ae(min) = 0.358 cm²
Core area $\cong$ 0.5 cm² would be good choice.

FLUX DENSITY
B (GAUSS)

B_{SAT}

ΔB

MAGNETIC FIELD
INTENSITY
H (OERSTED)

The next thing we're going to look at is what's referred to as the flux swing for the core. And this in some ways is analogous to our magnetizing current. Our magnetizing current is proportional to the flux swing. Shown on the right of the slide is a BH curve, you may have seen this in some earlier classes. The flux density is proportional to the volt x seconds. The slope of this is curve is proportional to the inductance and we have the ampere x turns over on the right. In order to predict the flux swing 'delta B' equals our minimum input voltage or maximum duty cycle, similar to the last set of equations, times this constant of 108 divided by our switching frequency divided by the quantity of turns on the primary divided by this term called AE. AE is the cross sectional area of the core. So as you can see, if you want to make your flux swing small, you need to make your transformer big. So that can kind of counteract what you really want to do in the real world. From practical experience I've found that keeping the flux swing to something in the order of about 2,000 gauss is a good place to be. And we'll see later on that the larger the flux swing, the higher the core losses, you certainly can't have the flux swing all the way up to the saturation level shown there at BSAT. And for most transformers, that saturation level is somewhere in the order of about 3,500 gauss, absolute above zero . But remember this time we're talking about delta B and we're going through two quadrants, so it's plus and minus 1,000 or a delta of 2,000 for what we're talking about right here.

And so if we rearrange this equation and we set it for, and solve for AE, as a minimum area and then we plug in our numbers again, we're going to have 33 volts or 0.65 or 65 percent duty cycle and our constant of 108. Early on in this design we decided we were going to operate the converter at 250 kHz, but that's something that can be changed but there's always more tradeoffs associated with that. Np was picked for 12 turns and remember that with 12 turns on the primary, that's going to force two turns on the secondary. There are other solutions, but it's going to change the area. You could have six turns, for example, but because it's in the denominator, it's going to drive much larger core. So you can iterate this. Using this equation, there is not just one output. So with that said, in general magnetics design tends to be an iterative type of process and that's why it's oftentimes not understood clearly.

Back to where I was, I was saying that Np equals 12 and if we limit the 'delta B to 2,000 gauss or 2K as shown here, that sets us up for a 0.35 cross-sectional area. And so when you go shopping around then for a core, you have to be sure that you have at least that area before you move on to the next step. As you're looking at the manufacturer's data books and for our given design, we picked about 0.5 as our cross-sectional area.

Next step: Calculate the **RMS** primary and secondary currents,
select a wire size, calculate the conduction power losses.

$$Ipri_{(AVG)} = \frac{\overline{Pout}}{\underline{Vin} \bullet \underline{Eff}}$$

$$Ipri_{(PEAK)} = \frac{Ipri_{(AVG)}}{\overline{D}} \qquad\qquad I\sec_{(PEAK)} = \overline{Iout}$$

$$Ipri_{(RMS)} = Ipri_{(PEAK)} \bullet \sqrt{\overline{D}} \qquad\qquad I\sec_{(RMS)} = I\sec_{(PEAK)} \bullet \sqrt{\overline{D}}$$

For our 100W, 3.3V example: Ipri(rms) = 4.2A and Isec(rms) = 24.2A
Using Coilcraft B0357 as an example Rpri = 55mΩ then Ppri = I² x R = 1Ω
Rsec =1mΩ the Psec =I² x R = 0.6W

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The next thing to do is to calculate the RMS current, both the primary and the secondary. I won't go through a lot of detail here since we're starting to run a little bit short of time, but the average current of the primary is the max output power at min input, but divided by your minimum efficiency as well. The peak then is the average, primary average divided by the duty cycle and then the RMS is times the square root of that duty cycle. Also for the secondary, it's only an approximation, but the secondary peak current is the same as the output. Again, that's an approximation. And then you can get the RMS again by multiplying by the square root of the duty cycle.

And so when you start to design your quantity of currents, you calculate the resistance and then you can do the I2R times the resistance, and you can predict your what they refer to as conduction losses for your currents. So for the primary, for our example 100 watt power converter, we have about one watt, and for the secondary we have 0.6 watts.

Transformer Design Summary

Turns ratio Np:Ns - 6:1,
Actual turns selected - 12 turns primary and 2 turns secondary
Core cross sectional area (Ae) = 0.5 cm^2

Primary resistance 55mΩ
Secondary Resistance = 1mΩ
Conduction losses ~ 1.6W
Core loses can be estimated from the core loss graphs supplied by the core mfr.

For example:
Bpk=1KG (use peak, not peak to peak),
Fsw=250kHz
From graph PL = 300 mW/cm^3
Core Volume (Ve) = 1.5cm^3
Pcore = 450mW
Total transformer losses ~ 2W

CORE LOSS vs. FLUX DENSITY
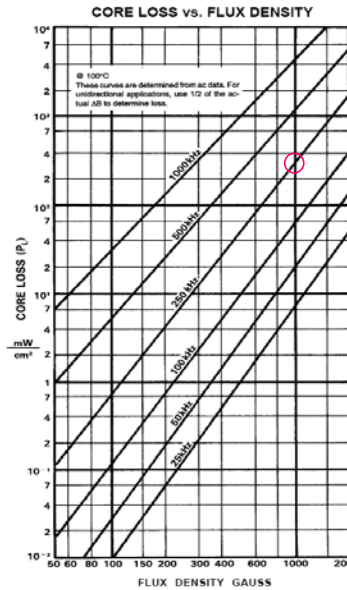
To do a little bit of summary here, we have our turns ratio of six to one. We actually picked 12 and 2, our cross-sectional area is .5 cm square. Our conduction loss is 1.6 watts.

There's also another term that we need to look at called core losses, and these can be predicted from the manufacturer's empirical data that they provided. So we're not going into it too much, we're just going to use our example. Basically you look across the bottom. One thing to remember on these particular data graphs provided, is that instead of using the peak-to-peak, the plot assumes that it's a peak-to-peak, so you just use the peak value.

For our example, we go across to 1,000 gauss. We said we're going to run a 250 KHz. We look at that line and then you can see the red circle that's on the chart. And that's going to give us this constant that is 300 milliwatts per centimeter cubed. To get the actual loss, you multiply by the volume of the core which you've selected, in this case it's 1.5 cm cubed and you end up with 450 milliwatts of core loss. So you end up with about 2 watts of loss in the core total.

And as I said this is an iterative process, you may go through this several times trying to optimize your transformer and with a little bit of practice you can be designing transformers in no time.

Duty Cycle & FET_Vds vs Vin
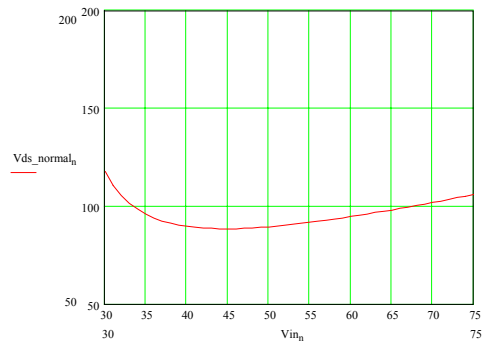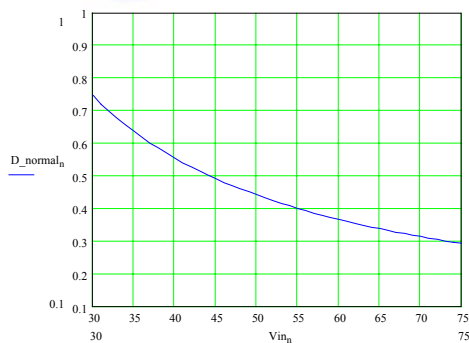
$$D = \frac{Vout + V_{(D2_{ON})}}{V_{IN}} \times \frac{Np}{Ns}$$

$$Vds = \frac{Vin}{1-D}$$

Nominal case

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I included the next slide because I though it was interesting. We talked about what the ideal duty cycle is for the forward converter and I basically just plotted that as shown on the left. It agrees with what we said we were saying, that we had 33 volts, we were going to have about 0.65 for our duty cycle, and sure enough, that's shown here. We also talked about the voltage across the reset capacitor during the active clamp reset time The drain is source voltage across the main switch and the P-Channel reset switch is the same as a capacitor voltage, that equals Vin divided by 1 minus D. When you plot that (Vin / (1-D), you can see that's a little bit interesting and it has a dip in the center and then it kind of starts to rise at both ends. The important point here is that when you're designing your transformer and the point I was trying to make before about why you don't want to have to have the duty cycle get too big is that if the duty cycles starts to get big, the reset clamp capacitor voltage has to get big, in fact, very quickly. You can see that on the left side (of the right plot) that if you were to have a bigger duty cycle, the reset capacitive voltage could get quite large quickly.

Voltage Feed-Forward PWM and Volt×Second Clamp
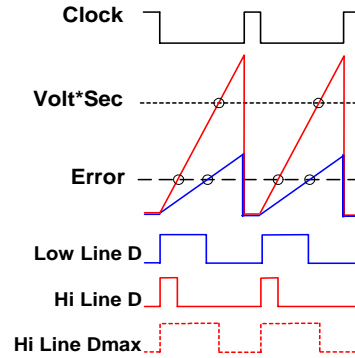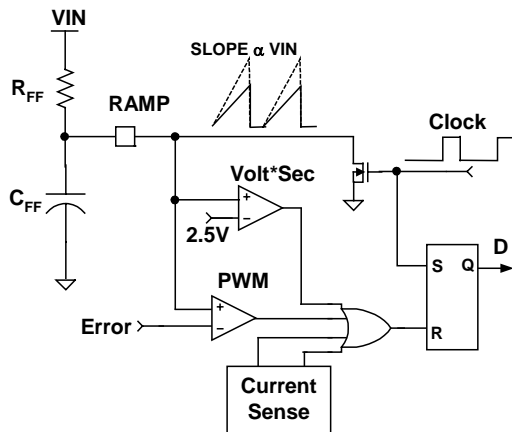
$$Vout = Vin * D * \frac{N_S}{N_P}$$

Another point about the duty cycle too is that everything we talked about was all steady state. In reality, you know, if you're operating in a low line condition and the duty cycle is quite large, if you have a transient load, the duty cycle will need to get bigger to accommodate that, to help the current traverse to a new output loading level. And so with that sometimes it's necessary to bound the max duty cycle so it doesn't get too big. There's a number of ways to do that, shown here's what's referred to as a volt x second clamp. You can read about this in some of the data sheets, it's quite nice because it's adjustable because it's a volt second clamp as the input voltage goes down, the clamp moves along with it.
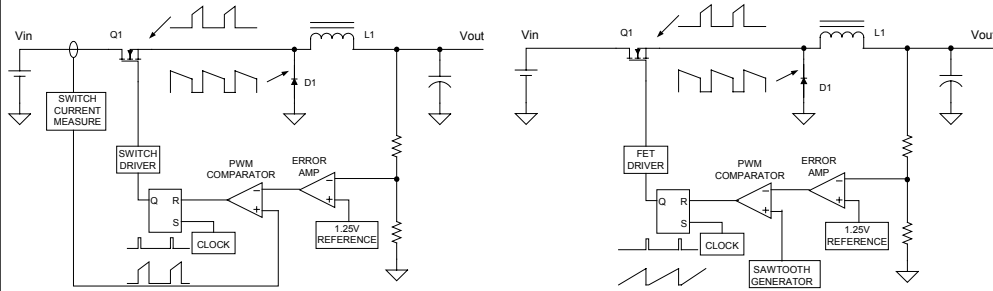
Current Mode / Voltage Mode Buck Regulators

FIGURE1
BUCK REGULATOR WITH CURRENT MODE CONTROL

FIGURE2
BUCK REGULATOR WITH VOLTAGE MODE CONTROL

Now, I'm going to touch on a topic of control, current mode and voltage mode. Shown here are both simplified and buck regulators showing a current mode control and a voltage mode control. They're basically the same in the sense that they look at the output voltage with an error amplifier. They compare that to a reference. They send that error signal to a pulse with modulator. The place where they start to differ, is in the pulse width modulator which compares this error signal against a ramping signal to modulate the duty cycle. For voltage mode, it uses a fixed ramp. Sometimes the ramp's slope is proportional to the input voltage, but it's a fixed ramp. Where current mode control actually samples the inductor current, measured at the main switch Q1 current when it's on. By its very nature that has a ramp included. Current mode uses that for the modulation ramp. That method provides a number of benefits that we're going to see on the next slide.

Current Mode Advantages / Disadvantages

ADVANTAGES
- Single pole system. The current loop forces the inductor to act as constant current source.
- Current mode control remains a single pole system regardless of conduction mode (continuous mode or discontinuous).
- Inherent line feed-forward since the ramp slope is set by the line voltage.
- By clamping the error signal, peak current limiting can be implemented.
- Ability to current share multiple power converters

DISADVANTAGES
- Often susceptible to noise on the current signal.
- As the duty cycle approaches 50% current mode control exhibits sub-harmonic oscillations. A fixed slope ramp signal (slope compensation) is generally added to the current ramp signal.
- The open loop output impedance is a current source feeding the filter capacitor and load. At low frequency this impedance is relatively high leading to poor load regulation for a given bandwidth.

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For current mode control, because it uses the current ramp, it actually forces the output to look like a current source and makes basically an inner current loop. The output has just basically a single pole, which is the RC, of the output capacitor and the load resistance is a single pole system. Current mode stays a single pole system whether it's in continuous or discontinuous, so it makes the compensation a lot easier. It also has inherent feed-forward, as the input voltage changes the duty cycle adjustments continuously without any change in the error signal. Also, by clamping the error signal, you can get a peak current limit.

One of the main disadvantages of current mode control is that because you have to measure that current, particularly at smaller duty cycles, it can be quite susceptible to noise and the modulation can be erratic at times. And also, current mode control is susceptible to what's referred to as a sub-harmonic oscillation. But that only occurs at duty cycles greater than 50 percent and that can be compensated with a slope compensation ramp. There's a lot of references to this material that can be read on the Web and in various books.

Voltage Mode
Advantages / Disadvantages

ADVANTAGES

- Greatly reduced noise susceptibility using a fixed ramp signal.
- The open loop output impedance is an LC filter. At low frequency this is essentially the inductor resistance in parallel with the load, which is relatively low.
- No slope compensation required.

DISADVANTAGES

- Loop compensation is more difficult. In continuous conduction mode, the system contains two poles. In discontinuous conduction mode, the system contains one pole. This requires more complexity and compromises in the compensation network. Also the loop bandwidth must be well above the output filter resonant frequency.
- Feed-forward and current limit are not inherent and must be implemented in the controller.

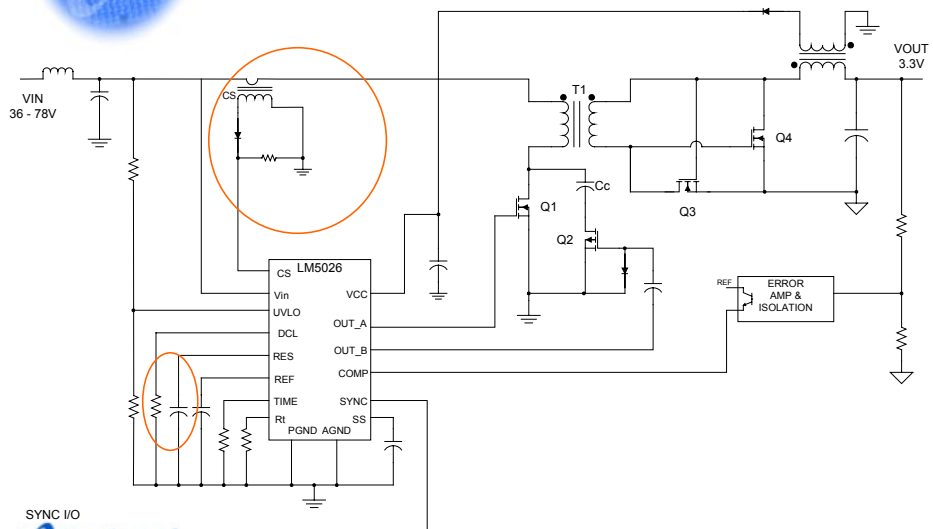**National Semiconductor**
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For voltage mode control, one of the main advantages, as least as I see it, is greatly reduced noise susceptibility, because you don't need to measure that current ramp every time to create a duty cycle. That's an important benefit, for sure. Also, there's no slope compensation required. One of the disadvantages is that the compensation is more difficult, it's basically a double-poled system. It changes with whether you're continuous or discontinuous. You can get feed forward with some additional circuitry.
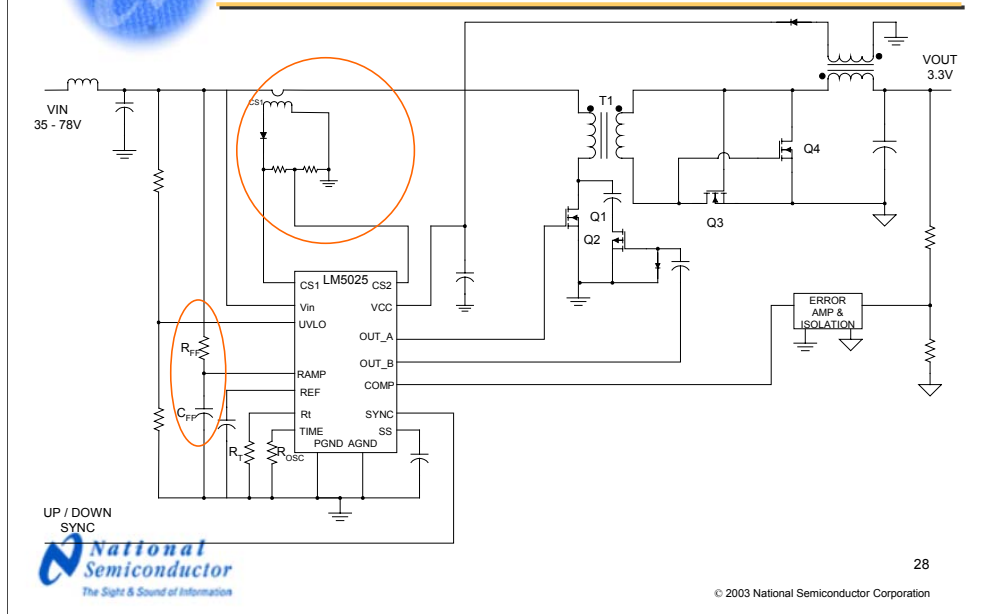
At this time, I wanted to look at some block diagrams, maybe expand our view, of a forward power converter. This slide is still simplified, but as a lot more content than what we had seen previously. You can see we still have our main switch Q1, we have our main transformer T1, we have our reset P-Channel reset capacitor, and we have the capacitor and diode to drive in series with the reset cap, self-driven synchronous rectifiers and our output LC filter. Now we can introduce our controller, you can see here, we have our LM5025 controller. This particular controller is a voltage mode and the circle shows the differences between the voltage mode and the current mode.

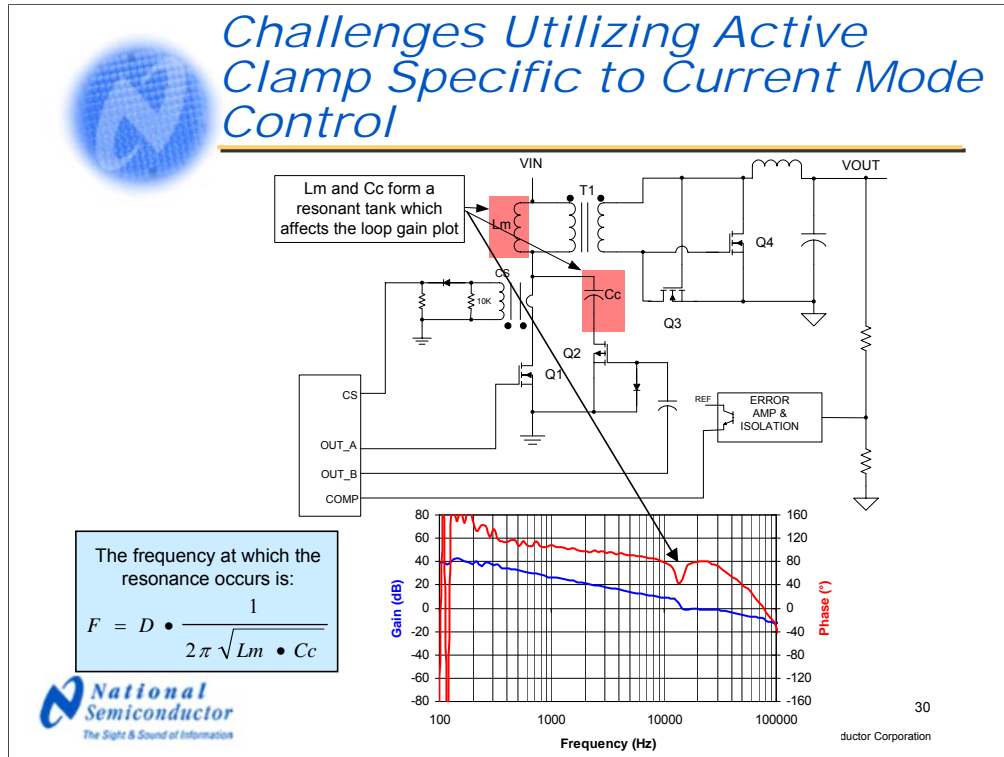LM5026 Active Clamp Forward Converter with Current Mode Control

Shown here is the current mode controller, the LM5026. The power stages are identical and you can see here the subtle differences of the way in which the current information is fed back and the over-current protection scheme.

Challenges Utilizing Active Clamp Specific to Current Mode Control

Lm and Cc form a resonant tank which affects the loop gain plot

VIN
T1
Lm
Cc
VOUT
Q4
Q3
Q2
Q1
10K
CS
OUT_A
OUT_B
COMP
REF
ERROR AMP & ISOLATION

The frequency at which the resonance occurs is:

$$F = D \bullet \frac{1}{2\pi \sqrt{Lm \bullet Cc}}$$

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Gain (dB) — Phase (°) — Frequency (Hz)

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One of the more interesting things that I learned recently, was regarding one of the challenges, that is unique to the current mode control. It's shown here on the bottom of the slide, shown is a bode plot of the loop gain for a power converter with current mode control and you can see that funny notch that happens at about 15 KHz. We went looking for the origin of that notch and what we attributed it to was the resonance associated with the magnetizing inductance and the clamp capacitor. This is only true for current mode control, we never saw this before on voltage mode control. The reason is because with voltage mode control, there is no inner loop that has basically feedback that can create or change the duty cycle on a cycle-by-cycle basis. Voltage mode control is driven primarily or strictly from the ramp which doesn't change on a cycle-by-cycle basis and an error signal which changes very slowly. Only for the current mode control will you see this type of phenomenon and when you go to predict where the frequency of this notch is, it's fairly close to what you would think. It was just one over 2 times the square root of LC, but not exactly there. It's actually shifted by a percentage that is proportional to the duty cycle (D). The reason that you can think about that is because, you know, this resonance only occurs for part of the cycle. That was an interesting but not necessarily immediately intuitive thing that we learned (about current mode control in active clamp). This notch isn't necessarily very desirable. It can definitely get in the way with your loop compensation and cause some instability.
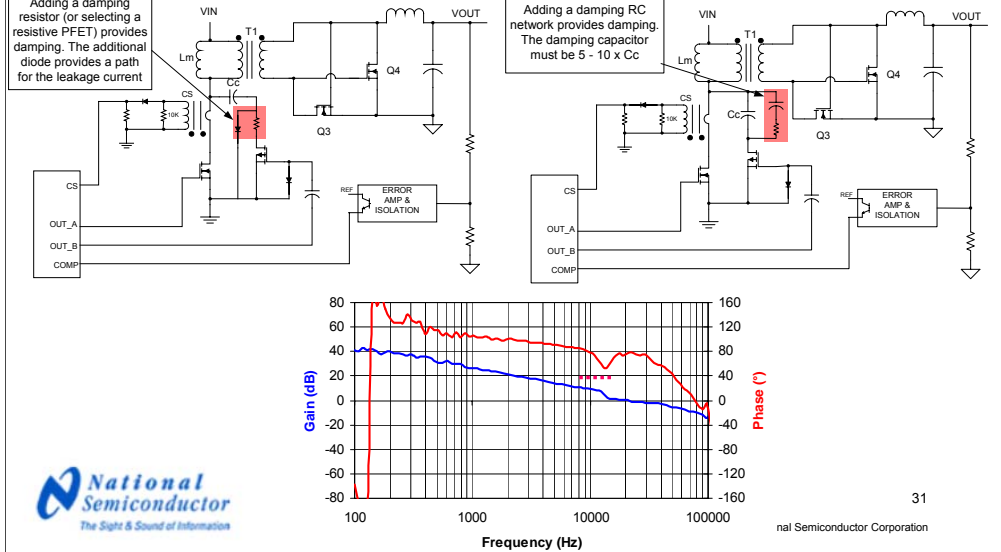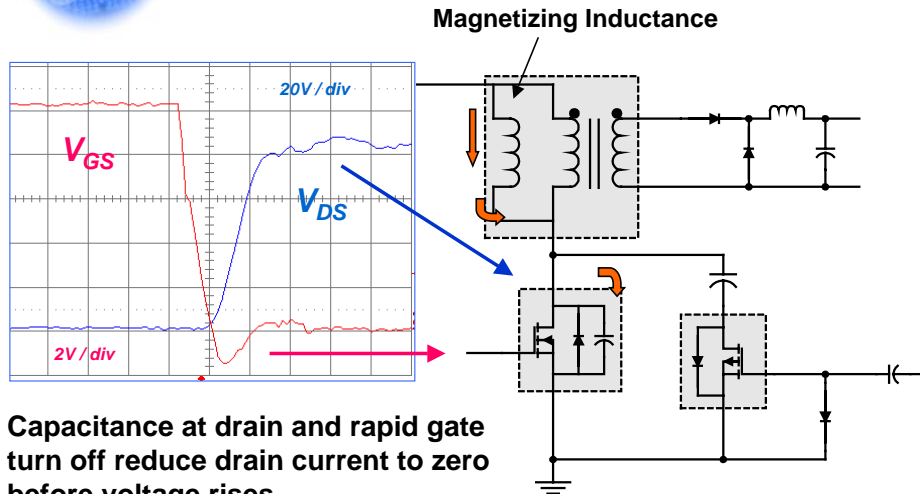
On the next slide we can look at several solutions. The one on the right is what's referred to as a parallel damping type scheme where we have an RC, where the damping capacitor is much, much larger than the clamp capacitor and then a resistor in series to provide actual damping. Another way would be a series damping, where the resistor is in series with the P-Channel switch. The problem with the resistor in series is that a large voltage spike can occur due to the leakage current. By putting a diode across there, you can clamp that spike and make this approach effective as well.

Shown below is the improved version, with the series damping, but I do believe you can get even more effective damping with the parallel damping but the physical size would probably be larger.

Main Switch Zero Voltage Turn-off Transition

Magnetizing Inductance

20V / div

V_{GS}

V_{DS}

2V / div

Capacitance at drain and rapid gate turn off reduce drain current to zero before voltage rises.

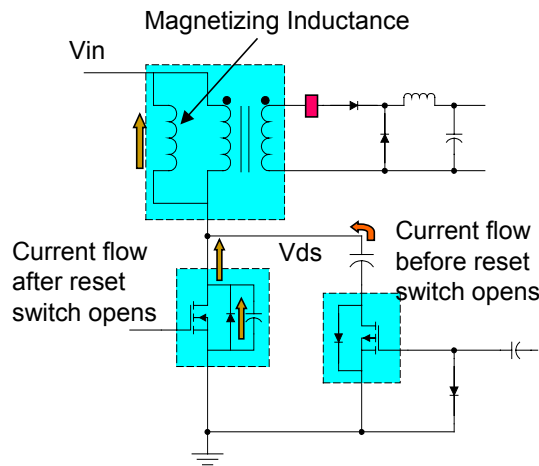National Semiconductor
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I wanted to look at some zero volt switching that does occur with the active clamp topology. It's not strictly limited to the active clamp. What's show here is that when the main switch is on, and all the current is flowing through the main switch. When the gate voltage drops, if you visualize the gate voltage as analogous to the channel current, then if you can turn the gate off very, very quickly then basically you stop the flow of current through the main MOSFET before the voltage gets a chance to rise. The reason why the voltage doesn't rise quickly is because of the capacitance associated with the FET itself. So you actually get a zero volt switching event here, and that's going to save a good bit on efficiency.

Zero Voltage Switching of Main FET Turn-on

- Following the turn-off of the reset switch, the magnetizing current forward biases the main switch body diode. At this time the main switch Vds is approximately zero. If the gate is activated before the magnetizing current decays, zero volt switching can be achieved.
- As Vds falls below Vin the secondary forward biases, a small delay implemented with a magnetic bead may be necessary.

Magnetizing Inductance

Vin

Current flow after reset switch opens

Vds

Current flow before reset switch opens

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To some extent it's at least theoretically possible that you can get a zero volt switching for the main FET turn on. The idea here is that when the active clamp switch is turned on, a current is flowing up through the capacitor, through the magnetizing inductance and up and back to the Vin source. The idea here is if you turn off the switch, you immediately stop the flow of current through the capacitor. The current flowing up through the magnetizing inductance wants to continue to flow and it will tend to pull the voltage down on the drain and you could potentially get a zero volt switching there. In a practical application it's hard to do, since the secondary side starts to affect the zero volt switching. Sometimes additional circuits are required to actually implement this or sometimes you get just a quasi zero volt switching in the sense that part of the transition drops with zero volts and then the rest of the event, has the current flow at the same time.

LM5025/25A Active Clamp Voltage-Mode PWM Controller

Features:
- Internal 14V to 100V Start-up Bias
- Voltage Mode Control with Feed-Forward
- Internal 3A Compound Gate Driver
- Programmable Line UVLO with Adjustable Hysteresis
- Dual Mode Over-Current Protection
- Programmable Overlap/Deadtime
- Programmable Volt x Second Clamp
- Programmable Soft-Start
- Leading Edge Blanking
- Single Resistor Programmable Oscillator
- Frequency UP / DOWN Sync
- Precision 5V Reference
- Thermal Shutdown

Packages: TSSOP16 and LLP16 (5 x 5 mm)

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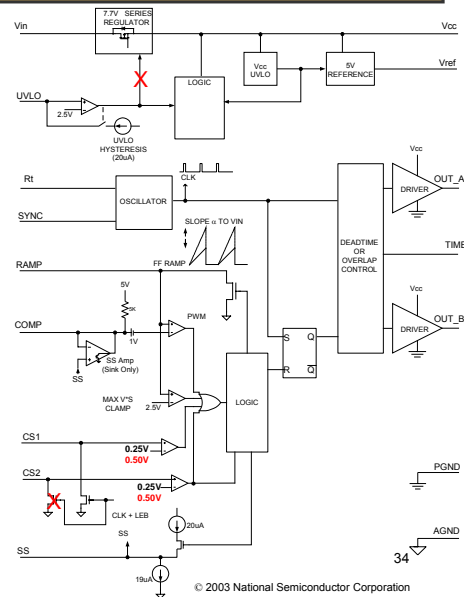
Next I want to let you be aware that there are a number of controllers that National Semiconductor has available designed explicitly for the active clamp control. The first series is the LM5025 series. This actually is the LM5025 base part. There's also LM025A and a LM5025B. For sake of time I'm not going to go through all of these. You can read the feature list yourself. The LM5025 is for voltage mode control. It's a full featured part. It has robust gate drivers and basically has everything that you need for voltage mode control to implement the active clamp topology. Some customers have specific requirements that we've accommodated on the A and the B versions. No one part is better than the other. They're all just basically the same with small subtle differences in thresholds and basic functionality.
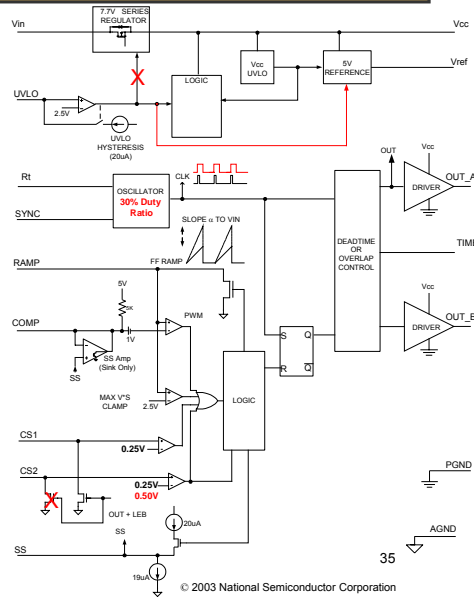
LM5025/25B Active Clamp Voltage-Mode PWM Controller

Features:
- Internal 14V to 100V Start-up Bias
- Voltage Mode Control with Feed-Forward
- Internal 3A Compound Gate Driver
- Programmable Line UVLO with Adjustable Hysteresis
- Dual Mode Over-Current Protection
- Programmable Overlap/Deadtime
- Programmable Volt x Second Clamp
- Programmable Soft-Start
- Leading Edge Blanking **(CS1 Only)**
- Single Resistor Programmable Oscillator
- **Reduced Max Duty Cycle (70% typ.)**
- Frequency UP / DOWN Sync
- Precision 5V Reference
- Thermal Shutdown

Packages: TSSOP16 and LLP16 (5 x 5 mm)

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What's shown here is the LM5025B.

LM5025 / LM5025A / LM5025B Active Clamp Controller

The LM5025A and LM5025B are functional variations of the LM5025 active clamp PWM controller:

Feature	LM5025	- A	- B
CS1 Current Limit Threshold	0.25V	0.5V	0.25V
CS2 Current Limit Threshold	0.25V	0.5V	0.5V
CS2 Filter Discharge (LEB)	Yes	No	No
Vcc disabled by UVLO	Yes	No	No
Vref disabled by UVLO	Yes	No	Yes
Maximum Duty Cycle (typ.)	85%	85%	70%

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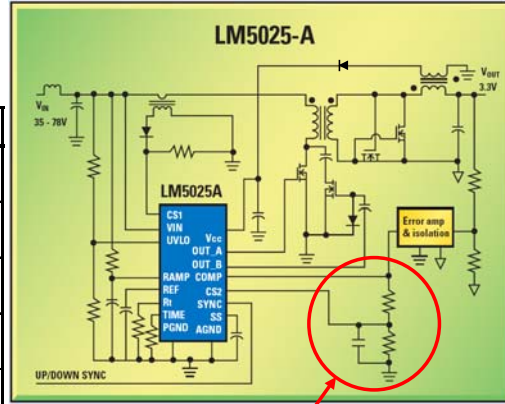
Hiccup mode triggered by excess COMP voltage after filter delay

36
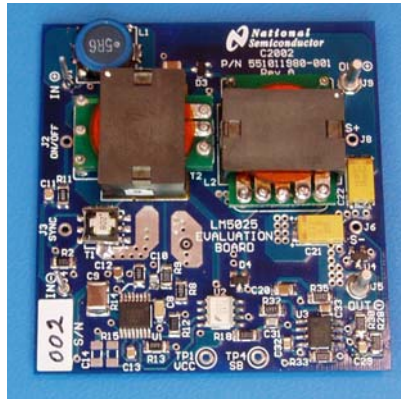
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Here is the summary slide of the base part, the LM5025, the dash A and the dash B.

LM5025 Active Clamp Forward Demo Board

Performance:

Input Range: 36 to 75V

Output Voltage: 3.3V

Output Current: 0 to 30A

Board Size: 2.3 x 2.4 x 0.5

Load Regulation: 0.2%

Line Regulation: 0.1%

Line UVLO, Current Limit

Measured Efficiency: 90.4% @ 30A

92.5% @15A

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Here's a photograph of a demo board that we did utilizing the LM5025. You can see the efficiency numbers are quite good. Shown here is the transformer that we designed in our previous design example. The efficiencies came out to just over 90 percent at full load at 30 amps and 92.5 at 15 amps.

LM5025 Active Clamp Forward DC-DC Converter (3.3V @ 30A)
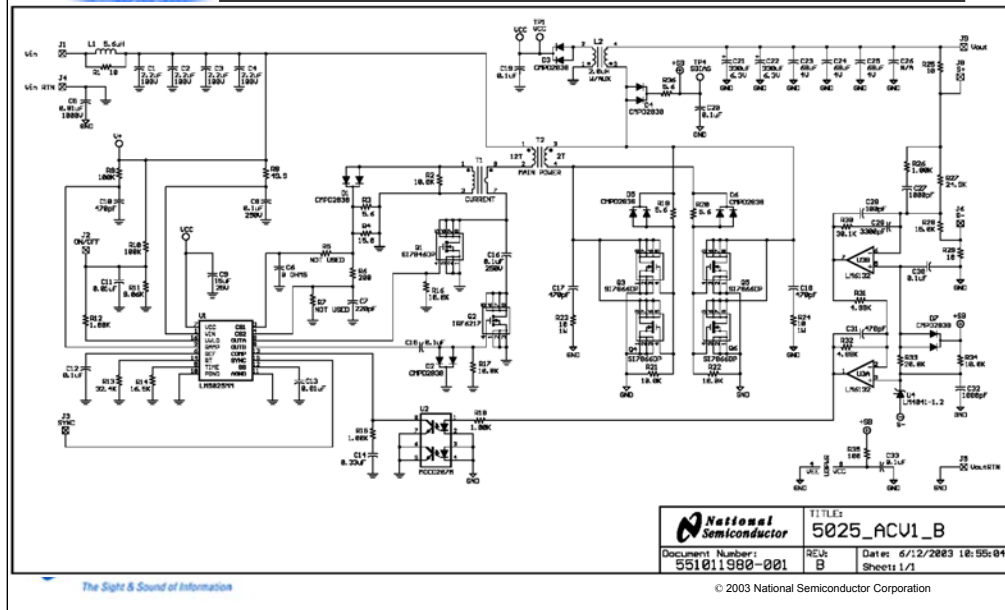
Shown here is the complete schematic for that particular demo board. As you can see, it's not a particularly expensive solution, had the very high efficiency solution. It's the best of both worlds in that it's not a particularly costly solution, at the same time it's high performance.
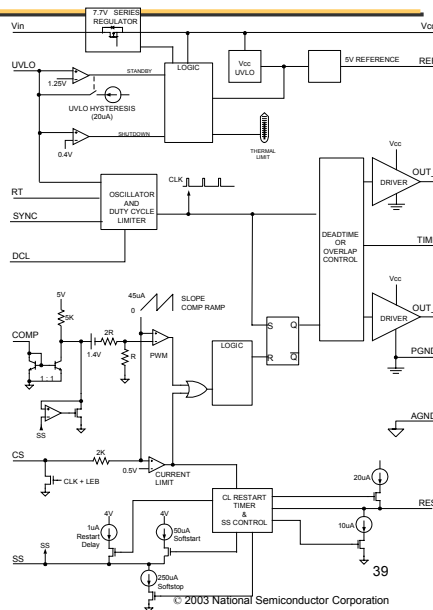
LM5026 Active Clamp Current Mode PWM Controller

Features:
- Internal 14V to 100V Start-up Bias
- Current Mode Control with Slope Comp
- Internal 3A Compound Gate Drivers
- Programmable Line UVLO with Adjustable Hysteresis
- Dual Mode Over-Current Protection with Delay
- High Bandwidth OPTO Interface
- Programmable Overlap Timer
- Programmable Max Duty Cycle Clamp
- Programmable Soft-Start / Soft-Stop
- Leading Edge Blanking
- Single Resistor Programmable Oscillator
- Oscillator Sync I/O Capability
- Precision 5V Reference
- Thermal Shutdown

Packages:
TSSOP16, LLP16
(5 x 5 mm)

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I wanted to make you aware that we've recently released an LM5026 controller which is another controller which works on the current mode control principle. There's a number of improvements that's within the LM5026, but the primary difference is that it works on current mode control rather than voltage mode control.
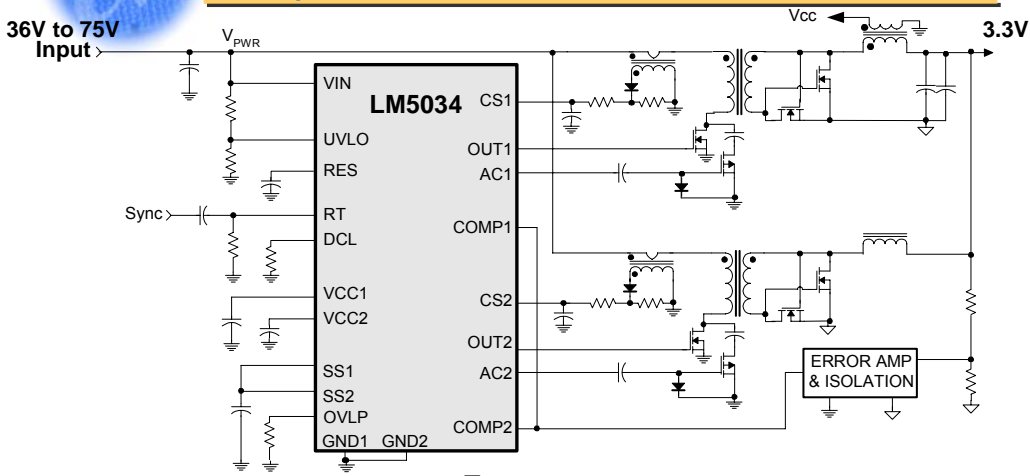
LM5026 Active Clamp Forward DC-DC Converter (3.3V @ 30A)
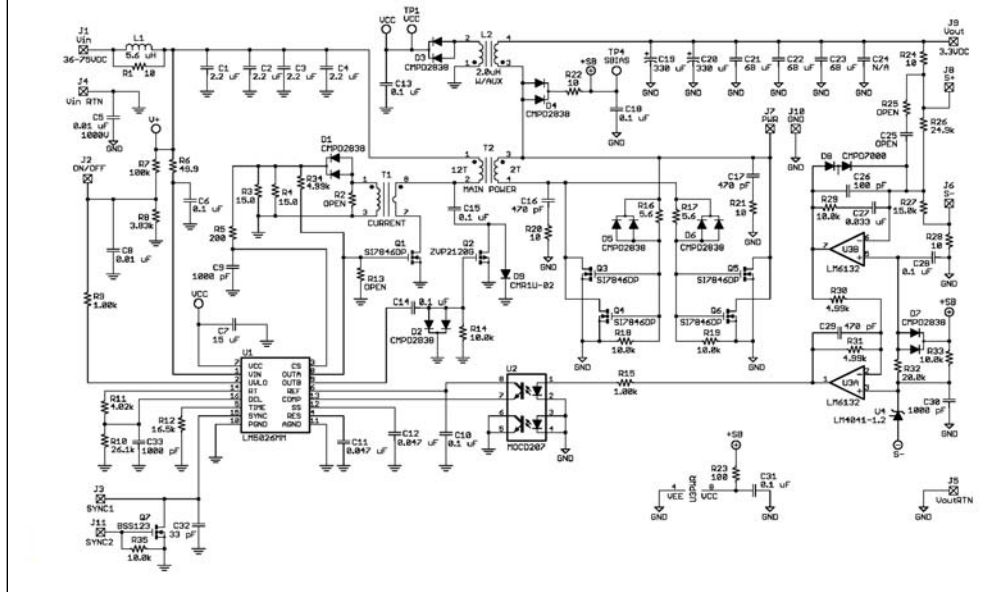
Here's the schematic for the LM5026 demo board. Again, it's virtually the same as the LM5025, the same power stage, same magnetics.

LM5034 Forward Active Clamp Controller, Single Interleaved Output

Features
- 3.3V, 60Amp output
- Interleaved operation reduces need for input and output filtering
- The power stresses are spread over two channels which increases reliability.

I also want to take this opportunity to share with you another controller for a related topology. It's actually the same topology, basically just an extension, and it's what we refer to as an interleaved power converter. We have this LM5034, which one controller allows you to create two active clamp forward power stages. Each one operates at 180 degrees out of phase with the other. The benefit of this is that the input and output filtering can be reduced dramatically since each channel operates 180 degrees out of phase. It effectively doubles the operating frequency as seen by both the input and the output filters. And it also spreads the power out amongst the two channels so it's not unusual to use something like this shown here with the 3.3 volt output where you can get, 60 or 100 amps of output current still at very high efficiency and a very reasonable material cost.
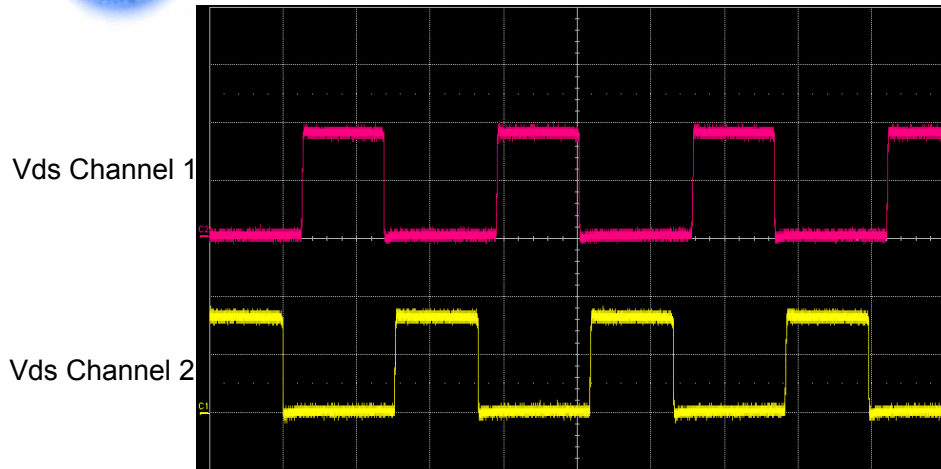
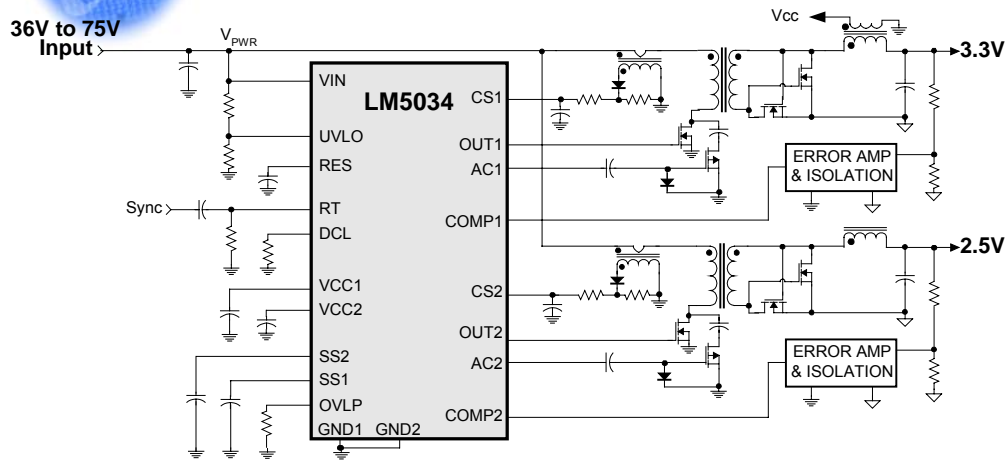Shown here are some waveforms for the interleaved. These are the drain waveforms in particular. As the voltage drops, that's when the switch for the main primary, main switch is turning on. As you can see, each of these are 180 degrees out of phase.

LM5034 Forward Active Clamp Controller, Dual Output Converter

36V to 75V Input

V_{PWR}

Vcc

LM5034

VIN
UVLO
RES
RT
DCL
VCC1
VCC2
SS2
SS1
OVLP
GND1   GND2

CS1
OUT1
AC1
COMP1

CS2
OUT2
AC2
COMP2

Sync

3.3V

2.5V

ERROR AMP & ISOLATION

ERROR AMP & ISOLATION

Features
- 3.3V, 30Amp and 2.5V, 30Amp outputs
- Interleaved operation reduces need for input filtering
- Two regulated and protected outputs from one controller

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The LM5034 can also be used for two separate independent separately regulated outputs. As shown here you have a 3.3 volt and a 2.5 volt output. They're both controlled with the one controller and you still get the benefit of reduced input filtering in this case since both channels draw from the same input capacitor. Because they're 180 degrees out of phase, that reduces the filtering needs on the input as well.
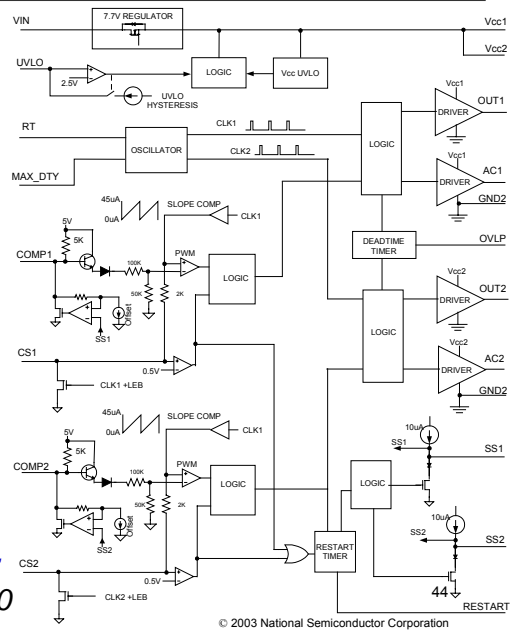
LM5034 Dual Interleaved Current Mode Controller w/Active Clamp

Features
- Two Independent I-Mode Controllers
- Single Interleaved or Dual Outputs
- Active Clamp Gate Drivers
- Compound 2.5A Main FET Drivers
- Single Resistor Deadtime Control
- Internal (100V) Start-up Regulator
- Single Resistor Oscillator Setting
- Synchronizable
- Line UV Lockout Comparator
- Independent Adjustable Soft-start
- Dual-Mode Over Current Protection
- Slope Compensation
- Compound Gate Drivers
- Thermal Shutdown

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Package:
TSSOP-20

VIN — 7.7V REGULATOR — Vcc1 — Vcc2
UVLO — 2.5V — UVLO HYSTERESIS — LOGIC — Vcc UVLO
RT — OSCILLATOR — CLK1 — CLK2
MAX_DTY
LOGIC — DRIVER — Vcc1 — OUT1
DRIVER — Vcc1 — AC1 — GND2
DEADTIME TIMER — OVLP
45uA — 0uA — SLOPE COMP — CLK1
5V — 5K — COMP1 — 100K — PWM — LOGIC
50k — Offset — 2K — SS1
CS1 — 0.5V — CLK1 +LEB
LOGIC — DRIVER — Vcc2 — OUT2
DRIVER — Vcc2 — AC2 — GND2
45uA — 0uA — SLOPE COMP — CLK1
5V — 5K — COMP2 — 100K — PWM — LOGIC
50k — Offset — 2K — SS2
CS2 — 0.5V — CLK2 +LEB
RESTART TIMER
10uA — SS1 — SS1
LOGIC
10uA — SS2 — SS2
44 — RESTART

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Shown here is the block diagram on the feature list for the Lm5034 controller.

LM5034 Interleaved Active Clamp Forward Demo Board

Performance:
Input Range: 36 to 75V
Output Voltage: 3.3V Single
            or 3.3V & 2.5V Dual
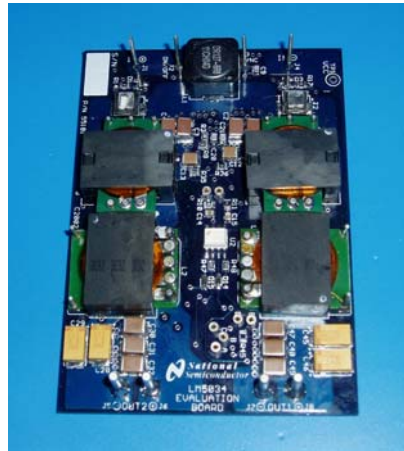Output Current: 0 to 60A
Board Size: 3.4 x 2.4 x 0.5 in.
Load Regulation: 0.2%
Line Regulation: 0.1%
Short Circuit Input Current: 150mA
Line UVLO, Current Limit

Measured Efficiency: 93%

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Shown here is the demo board for the LM5034. This particular one can be configured as a single 3.3 volt output up to 60 amps or it can be reconfigured, same board, for a 3.3 and a 2.5 volt output each with a 30 amp capability.

Additional Design Resources:

"*Active Clamp Resets Transformer in Converters*" byBob Bell, *Power Electronics Technology*, January 2004

http://powerelectronics.com

High Voltage Regulator Products

http://www.national.com/appinfo/power/hv.html

Power Management Product Information

http://power.national.com

In closing here I wanted to let you know there are some additional design resources. I realize and I apologize for a lot of material in a fairly short period of time, but there are additional places to look. I wrote an article back in January of 2004 in the *Power Electronics Technology Magazine* you can take a look at. The National Website has additional design resources that you can go take a look at and all the datasheets for the parts I have shown you today.

And so with that I'm going to turn it back over to Cory and I appreciate your time and thank you very much.

Thank you, Bob, and thank you, everyone, for joining us for this seminar, Introduction to Forward Power Converters Utilizing Active Clamp Presets, brought to you by National Semiconductor. This concludes today's online seminar. When you close your seminar window, a survey will appear. Please fill out and submit the survey form. Your answers to this survey will help us in the development of new products as well as future seminars. Thank you for attending and good day.