DESIGNING OPTIMAL MULTI-OUTPUT CONVERTERS WITH A COUPLED-INDUCTOR CURRENT-DRIVEN TOPOLOGY

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Paper presented at Powercon 8
Dallas, Texas - April 1981

Abstract

This paper examines the relative merits of several topological approaches to the design of multiple-output switching power converters. A design approach using a coupled-inductor current-driven (CICD) buck regulator topology is developed, with emphasis on the optimal design of the coupled inductor in order to minimize its leakage reactance.

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1. INTRODUCTION

Transformer-coupled boost converter topologies with multiple outputs have inherently good cross regulation. Serious disadvantages of boost-derived converters are their discontinuous output current which puts a great burden on the output capacitors, and the right-half plane zero in their control characteristic when operated in the continuous conduction mode.

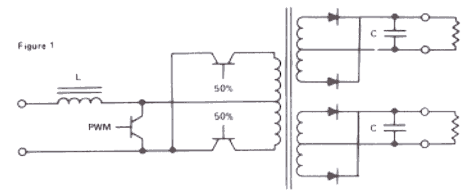
Conventional transformer-coupled buck converters with multiple outputs have poor cross regulation and suffer from current spikes associated with their inherent low impedance (voltage-driven) input circuits.

A unique buck-derived converter topology employing an inductor whose primary winding is on the input side of the transformer and secondary winding(s) on the output side [1],[2] combines most of the advantages of conventional buck and boost topologies with few of the disadvantages of either.

2. TOPOLOGICAL COMPARISONS

2.1 THE BOOST CONVERTER

A transformer-coupled boost converter is shown in Figure 1. It is often implemented by omitting the shunt pulse-width modulated transistor, achieving modulation control by means of deliberate overlap in the conduction of the two push-pull transistors. During the interval of conduction overlap, the transformer primary is shorted, thereby charging the input inductor and decoupling the output.



Output voltages of the boost converter are related to the input voltage as follows:

$$V_{out} = V_{in}(N_S/N_P)/(1-D)$$

Duty cycle, D, is the ratio of the "on" time of the PWM transistor (or time of conduction overlap) to the total period., N_S/N_P is the ratio of secondary to primary turns.

2.1.1 Advantages.

Current limiting. The boost converter is **current-driven** because the series input inductor acts as a high impedance source at high frequencies, thereby providing dynamic input current limiting. This prevents the current spikes that would otherwise be associated with momentary short circuits caused by events such as transformer core saturation, slow rectifier recovery, or transistor conduction overlap.

Continuous input current. Because of the series inductor, input current is continuous with very moderate rates of change. This results in simpler input filtering and reduced EMI.

Low peak voltages. Peak voltages on the transformer secondaries are equal to the output voltages (neglecting rectifier drops and leakage reactance spikes), regardless of the line input voltage level. Excess input voltage is dropped across the input inductor. Transformer utilization is good because maximum volt-seconds are limited and independent of input voltage. Rectifier PIV requirements are only 2 times their respective output voltages, regardless of input voltage.

Good cross-regulation. There are no filter inductors in series with each output to impair cross-regulation. Series resistances of rectifiers and transformer secondaries are usually quite small, but rectifier forward recovery characteristics must be very good. Output voltages will then relate to each other in almost exact proportion to the ratio of secondary turns, virtually independent of load currents, provided leakage inductances between secondaries and wiring inductances are minimized.

2.1.2 Disadvantages.

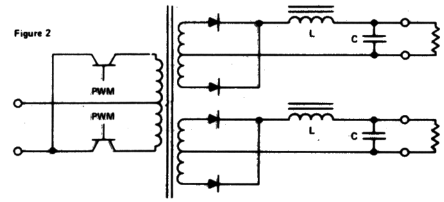
Discontinuous output current. During the "on" time of the shunt transistor, current provided to the output capacitor ceases abruptly. The output capacitor must provide all of the load current during the "on" time, without assistance. This places a severe requirement on both the series inductance and ESR of the capacitor. For this reason, the boost configuration is seldom used in low voltage, high current output applications. During the "off" time, rectifier peak current levels are much larger than DC load current levels. These problems are also characteristic of flyback converters.

Difficult control loop optimization. As Middlebrook has shown, the control-to-output transfer function of the boost converter (as well as the flyback converter) operated in the continuous conduction mode has a right-half plane zero. This causes considerable difficulty in optimizing the feedback loop to obtain the desired gain and dynamic response together with good loop stability.[3]

2.2 THE BUCK CONVERTER

The full-wave center-tap transformer-coupled buck converter with multiple outputs shown in Figure 2 is, together with its half-bridge counterpart, the most widely used configuration for off-line converters.

Figure 2



Output voltages of the buck converter are related to the input voltage and the duty cycle:

$$V_{out} = V_{in}(N_S/N_P)D$$

2.2.1 Advantages.

Continuous output current. The filter inductors provide continuous DC current to the loads. Output capacitors are required to handle only the triangular, small amplitude inductor ripple currents. Capacitor requirements are drastically reduced compared to boost and flyback converters. Also, rectifier peak currents are equal to the DC load currents.

Simple control loop optimization. The buck converter does not have a right-half plane zero in its control characteristic.

2.2.2 Disadvantages.

No inherent current limiting. The conventional buck converter is **voltage-driven** in the sense that there is no high impedance in the input circuit to dynamically limit the input current. Large input current spikes will occur due to the momentary short circuit conditions caused by slow rectifier recovery, insufficient dead-band to prevent transistor conduction overlap, or by transformer saturation.

Transformer core utilization is impaired and leakage reactance is increased by the need to operate the core at lower flux density so as to prevent saturation caused by flux "walking" with asymmetrical drive.

Discontinuous input current. During the "off" time of the switching transistors, input current ceases abruptly. Considerable filtering is required to control EMI.

High peak voltages. Excess input voltage under high line conditions is dropped across the filter inductor. Because this inductor is downstream of the transformer and rectifiers, rectifier PIV requirements are considerably greater than 2 times the output voltage. For example, if the converter is designed to handle a 2:1 variation in input voltage in addition to normal losses, rectifier PIV requirements can easily be 5 to 6 times larger than the output voltage. This means that conventional Schottky rectifiers cannot be used for a 10 volt output.

Full line input voltage is also applied to the transformer primary. This may result in core saturation during start up or after a large step increase in load current, when full duty cycle at high line input voltage applies high volt-seconds to the primary.

Poor cross-regulation. The most critical output is normally used as the basis for feedback control. The other outputs follow along in proportion to the number of secondary turns. DC cross regulation is usually good, provided leakage reactances between transformer secondaries are low.

AC coupling between the various outputs is very poor because of the inductor in series with each output. Step load increases on an uncontrolled output are not communicated to the controlled output. This usually results in severe dropout of the loaded output, slow recovery, and considerable overshoot. In a similar manner, if the controlled output load is suddenly increased, the feedback loop will call for full output in an attempt to maintain its voltage. This will cause all of the uncontrolled outputs to rise substantially above their normal levels.

Loss of DC regulation will occur when the load current of any output falls below the critical current level established by its filter inductor. The output voltage then rises from the average toward the peak value of the voltage waveform at the input to the filter. This can be 2 to 3 times larger than normal. Solutions to this problem, such as switched preloading are either expensive or generally unsatisfactory.

Another technique uses a single large inductor with multiple windings in place of the individual inductors in series with each output.[4] This results in close AC and DC coupling between outputs, and provides good cross-regulation. The turns ratios between inductor windings must be the same as the ratios between their associated transformer secondary windings.

2.3 THE CICD BUCK CONVERTER

A dual-output version of the coupled-inductor current-driven (CICD) buck converter is shown in Figure 3.[1],[2] A single inductor with multiple windings cross-couples the outputs during the "off" time to provide good cross-regulation. A primary winding on this same inductor acts as a high impedance input current source during the "on" times of the switching transistors, hence this configuration is current-driven similar to the boost converter.

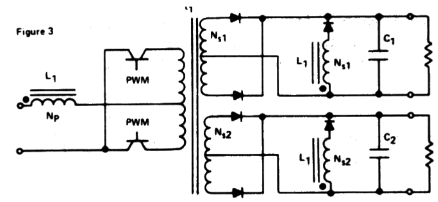


Figure 3

This CICD buck converter topology combines most of the advantages of the conventional buck converter and boost converter, while retaining few of the disadvantages of these other topologies. The input-output voltage relationship is identical to the conventional voltage-driven buck converter:

$$V_{out} = V_{in}(N_S/N_P)D$$

2.3.1 Advantages.

Advantages are discussed in detail in Section 3. They include:

- Input current limiting.
- Reduced semiconductor speed requirements
- Continuous output current
- Moderate output capacitor requirements
- Low rectifier reverse voltages
- Low transformer volt-seconds
- Good cross-regulation
- No right-half plane zero

2.3.2 Disadvantages.

Disadvantages are few:

- Discontinuous input current
- Additional free-wheeling rectifiers
- More complex inductor

The increased costs of the inductor and additional rectifier are offset by reduced rectifier PIV and recovery requirements, reduced peak and average transformer volt-seconds permitting better utilization, and only one inductor instead of several.

Leakage reactances between the coupled-inductor windings are considerably greater than for a comparable transformer. The design of the coupled-inductor to minimize leakage reactances is critical for minimizing voltage spikes and achieving good efficiency and cross regulation.

3. OPERATIONAL ANALYSIS

3.1 TOPOLOGY EVOLUTION

Figure 4 shows the progression of topologies starting with the basic non-isolated buck converter in Figure 4A and ending with its isolated CICD counterpart.

In Figure 4B, a closely coupled bifilar secondary winding with a 1:1 turns ratio is added to the inductor. Since the output ends of these windings are shorted together, there can be no AC or DC voltage differential across the input ends. In the ideal case with no leakage reactance or resistance in the windings. The rectifier can be moved from the primary to the secondary side with no functional change in circuit behavior. Energy storage in the inductor core is related to ampere-turns in the winding. The core does not know or care which winding carries the current.

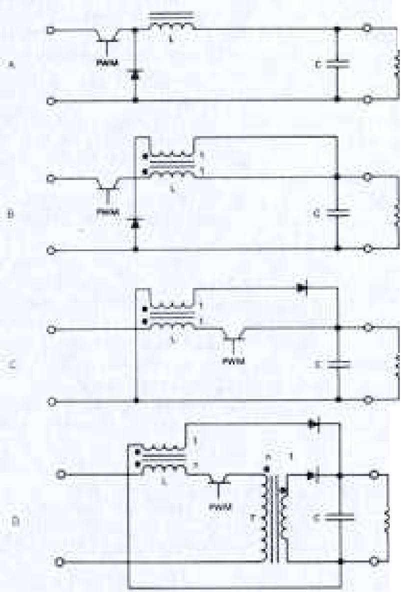
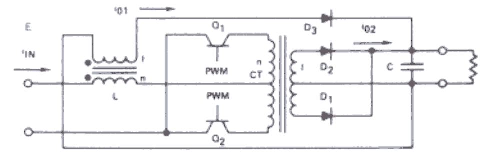


Figure 4

In Figure 4C, the rectifier and transistor are pushed through to the output ends of their respective windings, again without changing any of the current and voltage relationships.

In Figure 4D, a transformer is interposed between the transistor and filter capacitor to provide isolation and permit voltage level changes to optimize the duty cycle. This circuit is a current-driven forward converter, in the buck family.

The additional rectifier in series with the transformer secondary allows the transformer voltage to reverse during transistor "off" time in order to reset the core. Additional means (not shown) must be provided to clamp the reverse voltage excursion to protect the transistor and rectifier. It is essential that the inductor now have the same turns ratio as the transformer, otherwise output current will be discontinuous.



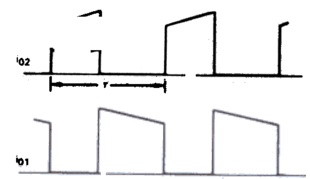
Finally, in Figure 4E, additional windings on the transformer in a push-pull configuration provide for automatic core reset and better utilization of the core. This is the coupled-inductor current-driven converter, demonstrating that it is indeed a member of the buck family.

3.2 BASIC CURRENT WAVEFORMS

Referring to Figures 4E and 5, when both transistors are off, no primary current can flow. Current in the inductor secondary, I_{01}, free-wheels to the output, exactly like any buck converter. The inductor secondary is clamped to the output voltage, causing the current to decrease at a rate equal to V_{out}/L_S.

When either transistor is on, inductor secondary current ceases. Current now flows from the input through the inductor primary and the transformer to the output capacitor as I_{02}. Inductor current increases at a rate equal to $(V_{in} - V_{out})/L_P$.

Figure 5



It can be seen that the sum of currents I_{01} and I_{02} driving the output capacitor is identical to the inductor current in a conventional buck converter. The only topological difference from the conventional buck converter is that during the transistor "on" time the inductor is on the input side of the transformer. This results in a reduction of peak voltages on the downstream components.

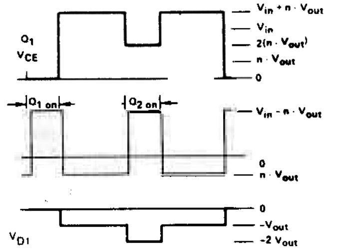
During the "on" time, the inductor primary establishes a high impedance current-driven input circuit, preventing current spikes due to poor rectifier recovery, transistor storage time overlap, or transformer saturation.

3.3 BASIC VOLTAGE WAVEFORMS

3.3.1 ON time.

Referring to Figures 4E and 6, during the "on" time of either transistor, the transformer secondary windings are clamped to the output voltage, V_{out}.

Figure 6



"Off" rectifier D_1,D_2 reverse voltage	$= 2V_{out}$
Primary transformer voltage	$= nV_{out}$
"Off" transistor collector voltage	$= 2nV_{out}$
Inductor primary voltage	$= V_{in} - nV_{out}$
Rectifier D_3 reverse voltage	$= V_{in}/n + 2V_{out}$

Because all of the excess input voltage is dropped across the inductor primary, D_1 and D_2 reverse voltages are less than the conventional buck converter.

When either transistor turns on, rectifier D_3 switches rapidly from forward conduction to reverse blocking, and therefore requires excellent reverse recovery capability. Rectifiers D_1 and D_2, on the other hand, are not critical as to their reverse recovery capabilities, unlike in the conventional buck converter.

3.3.2 OFF time.

During the time that both transistors are off, the inductor secondary conducts and its voltage is clamped to V_{out}.

Inductor primary voltage	$= nV_{out}$
Transformer CT voltage	$= nV_{out} + V_{in}$

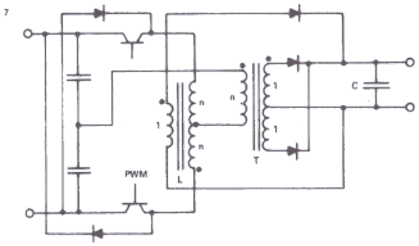
Ideally, there will be no voltage across the transformer windings during the "off" time. Transistor V_{CE} would equal $V_{in} + nV_{out}$, and D_1,D_2 reverse voltages would equal V_{out}.

However, during the initial portion of the "off" period, the transformer magnetizing current will cause the voltage across the transformer windings to flip polarity in order to continue flowing through one of the output rectifiers until the flux goes to zero. Also, voltage spikes caused by transformer and inductor leakage reactances will require clamping.

4. HALF BRIDGE CONFIGURATION

A half-bridge version of the CICD buck converter is shown in Figure 7. A center-tapped inductor primary winding is now required so that unidirectional current flow is maintained in the inductor windings consistent with the DC flux level in the core while the two switching transistors conduct alternately. If the inductor primary was single-ended it would have AC drive and its high impedance would block the transfer of power to the load.

Figure 7



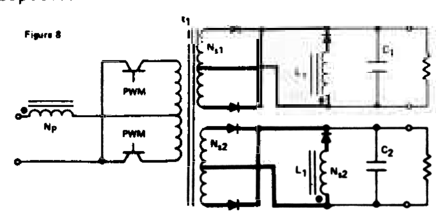
The half-bridge configuration provides a simple way to clamp the voltages across the switching transistors. The two cross-coupled diodes at the input also return the energy stored in inductor and transformer leakage reactances to the input supply rather than throwing it away in a conventional snubber network.

5. MULTIPLE OUTPUT CONFIGURATIONS

5.1 THEORETICAL PERFORMANCE

The basic reason that good cross-regulation is provided between the various outputs of the CICD converter as shown in Figure 8 is that all outputs are essentially driven in parallel at all times during the switching period with very small intervening impedances. The outputs are driven directly from the tightly coupled transformer secondaries during the "on" time, and from the tightly coupled inductor secondaries during the "off" time. At all times, the inductor acts as a high impedance current source, sharing its current among the various outputs (as modified by the turns ratios) in direct proportion to the demands of their respective loads.

Figure 8



Since each output is always driven directly from either a transformer or inductor secondary with only the rectifiers intervening, output voltages must be in exact proportion to the turns ratios of their respective secondaries. If, for any reason, any output momentarily rises above its proportionate level, its source current will cease until the voltage drops back to normal. If, on the other hand, any output momentarily drops below its proportionate level due, for example, to a large step increase in load current, that output will receive **all** the inductor current (as modified by the turns ratios) until it is restored to normal. During that time the other outputs will receive no inductor current until proportionality is restored.

Thus, because of the tight coupling between outputs, the disturbance caused by a step load change on any output is minimal because all outputs share in helping to counter the disturbance. Also, a change applied to an uncontrolled output is coupled immediately to the controlled output so that feedback correction is not delayed.

Because there is a single inductor sharing the multiple load currents, there is no critical inductor current level associated with each individual output, as with the conventional buck converter. However, the total of all loads reflected into the inductor primary must be greater than the critical current level associated with the single inductor primary, otherwise regulation is degraded.

The inductor secondaries may be omitted with slight degradation of cross regulation in auxiliary outputs whose maximum load power is significantly less than the total minimum load power of the supply.

5.2 PRACTICAL CONSIDERATIONS

Rectifier DC drops. Rectifier forward voltage drops will cause a discrepancy between the turns ratios and the ouput voltage ratios. A single rectifier voltage drop will be proportionately larger for a low voltage output than for a high voltage output. This discrepancy must be corrected by adjusting the turns ratios.

Rectifier forward voltage drop temperature coefficients will cause output voltage divergences from the ideal unless the rectifier temperature coefficients are proportional to the voltage levels of their respective outputs. If all rectifiers have the same temperature coefficient, -1.5 mV/ C, all rectifier drops will decrease by 150 mV with a 100 C rise in temperature.

If the feedback controlled output is 5 volts, its associated secondary winding voltages will **decrease** by 150 mV at high temperature. That will cause the voltage of a tightly coupled 15 volt winding to decrease by $150 \times 15/5 = 450$ mV. Since the 15 volt rectifier has also lost 150 mV forward drop, the 15 volt output will decrease by 300 mV total.

Load Regulation. Leakage reactances between primary and secondary windings of the inductor and transformer will hurt overall load regulation. The control feedback loop will usually correct this problem.

Cross regulation. Good cross regulation between the multiple outputs of the CICD buck converter depends entirely on maintaining tight coupling between outputs. **Minimize all intervening impedances!** Realize that the same impedance will have much greater destructive effect on cross-regulation in a low voltage, high current output because of the leverage provided by the turns ratios.

Secondary circuit resistance. Resistance of the secondary windings is usually negligible, but rectifier dynamic resistance will have a definite effect on cross-regulation. This effect will be most noticable with load current changes on the lowest current output.

It is reasonable to assume an increase of 100 mV rectifier voltage drop with a 5:1 change in forward current. If the load current on the 5 volt output increases by a factor of 5, the 5 volt secondary winding voltages must increase 100 mV to maintain the output. This will cause an increase of 300 mV in the 15 volt output, assuming its load current does not change.

Although these relative changes in voltage between outputs due to rectifier dynamic resistance and temperature cannot be reduced, absolute errors in output voltages can be minimized by applying feedback control to an intermediate voltage output, preferably one with relatively constant load.

Rectifier forward recovery. Any rectifier having a poor forward recovery characteristic cannot initially conduct the current intended for its associated output. Because of the tight coupling between secondary windings, current intended for this output will be diverted to the other outputs at the beginning of each switching period for the fraction of a microsecond or more required for forward recovery to be completed.

This will not have a significant effect on cross-regulation if the rectifier with poor forward recovery happens to be associated with a high voltage, low current output. However, if a high current output has its current momentarily diverted to a low current output, the charge delivered to the low current output during each switching cycle may be greater than its load can accomodate. This will charge the output capacitor of the low current output to a voltage much higher than normal. Fortunately, Schottky and UES gold-doped rectifiers normally used in the higher current, lower voltage outputs have excellent forward recovery characteristics.

Secondary circuit inductance. Inductance in series with any of the output rectifiers will cause momentary diversion of current away from higher current, lower voltage outputs (small load resistance) to outputs with higher load resistance, in the same way as with poor forward recovery.

For example, in a 5 volt, 10 ampere output, only 0.2 microhenries series inductance will require 1 extra volt for 2 microseconds to change the current from 0 to 10 amperes at the beginning of each current pulse. This extra 1 volt on the 5 volt secondary will result in an additional 3 volts on the tightly coupled 15 volt secondary. The actual increase in voltage of the 15 volt output depends on the **average** current that is diverted (which is a function of the switching frequency) in relation to the load current of the 15 volt output.

Every inch of wire in series with a secondary winding adds .02 microhenries of series inductance. It is absolutely vital, particularly in the high current, low voltage outputs, to keep the wiring lengths **as short as possible** between secondary windings and the common connection points at the output side of the rectifiers, indicated by the heavier lines in Figure 8.

(Added 8/81): In spite of the best possible techniques for minimizing wiring inductance, it may become the limiting factor rendering the CICD buck converter unsuitable for applications with low voltage outputs greater than 200-300 watts.

Leakage inductance between secondaries causes poor cross regulation in exactly the same way as wiring inductance. However, leakage inductances tend to be proportional to the square of the secondary turns, so that the leakage inductance problem tends to be independent of the output voltage level, but is more of a problem in high power outputs.

(Added 8/81): There are several advantages to building higher voltage auxiliary outputs on top of lower voltage main outputs. Fewer turns are required for the auxiliary outputs, simplifying transformer and inductor fabrication. More importantly, since the low voltage output shares its voltage with the higher voltage auxiliaries, cross regulation is improved. The leveraged effects of rectifier forward voltage drops and temperature coefficients and rectifier dynamic resistance are reduced.

6. ESIGNING THE COUPLED INDUCTOR

6.1 REVIEW OF MAGNETIC PRINCIPLES

6.1.1 Definitions -- CGS Units

	B	Flux density, Gauss
	H	Magnetic field intensity, Oersteds
$B/H =$	μ	Permeability
	A_e	Effective magnetic area, cm^2
	ℓ_e	Mean length of magnetic path, cm
	ℓ_g	Air gap length, cm
	N	Number of turns in winding
	BA_e	Total Flux
	$H\ell$	Total magnetizing force

6.1.2 Basic Equations

Total magnetizing force :

$$H\ell = .4\pi NI \qquad (1)$$

Energy storage / cm^3 (Joules/cm^3) :

$$W/cm^3 = 1/2(BH/.4\pi) \times 10^{-8} \qquad (2)$$

Total energy storage (Joules) :

$$W = 1/2\ LI^2 \qquad (3)$$

$$W = 1/2(BA_eH\ell/.4\pi) \times 10^{-8} \qquad (4A)$$

$$W = 1/2(B^2A_e\ell_e/.4\pi\mu) \times 10^{-8} \quad \text{(mo-perm)} \qquad (4B)$$

$$W = 1/2(B^2A_e\ell_g/.4\pi 1) \times 10^{-8} \quad \text{(gapped)} \qquad (4C)$$

Note: The energy storage formulae above assume a linear B-H relationship, as in an air gap or mo-perm core below saturation.

Two other relationships that are sometimes useful:

Inductance, Henries :

$$L = .4\pi\mu N^2A_e/\ell \times 10^{-8} \qquad (5)$$

Induced voltage, Volts :

$$E = NA_e dB/dt \times 10^{-8} \qquad (6)$$

6.2 MAGNETIC DESIGN APPROACH

The task is to design a multiple-winding inductor with minimum leakage inductance between windings.

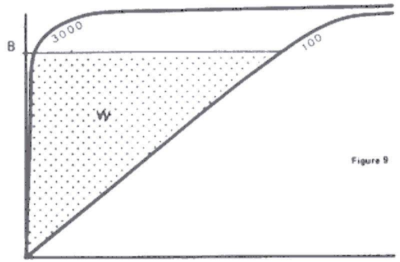
Although the concept of "inductance" is vital for circuit analysis purposes, it is difficult to conceptualize inductance in terms of the magnetic device structure. It is much easier to conceive of flux lines, magnetic force equipotential surfaces and stored energy within the magnetic system and not be concerned about the number of turns required to achieve the desired inductance until the very end of the design process.

Transformers and inductors appear to be very similar devices, but they are fundamentally different. A transformer is intended to couple energy -- energy storage is undesireable. An inductor is intended primarily to store energy.

6.2.1 Energy Storage Criteria.

An efficient inductor for a specific application must be designed to store a maximum energy equal to $1/2\ L I_{SC}^2$ at the threshold of core saturation, where L is the design value of inductance that will result in the desired ripple current, and I_{SC} is the absolute maximum overload current condition.

Figure 9 shows a B-H curve (first quadrant only) for two magnetic core systems, one with high and one with low effective permeability. Energy storage per unit volume is proportional to $\int HdB$ at the operating point on the B-H curve:



$$W/cm^3 = 1/2(BH/.4\pi) \times 10^{-8} \text{ Joules/cm}^3 \qquad (2)$$

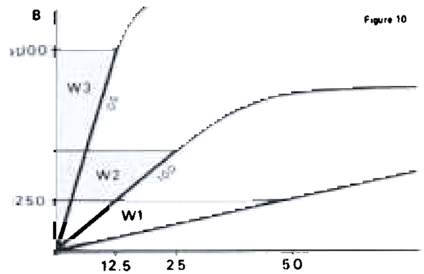
If the cores are operated below saturation flux density, very little energy can be stored in the high permeability core. This is excellent for a transformer, but not for an inductor. It is possible, however, to store a reasonable amount of energy in the lower permeability core. Maximum energy storage capability of a given core system will always occur at flux densities approaching but not exceeding saturation.

It is also evident that by further lowering the permeablity, the energy storage capability per unit volume at saturation flux levels continues to increase. However, the winding will get bigger, so this is the wrong direction to go.

In fact, in a gapped core, virtually all of the energy storage takes place within the gap, where the permeability of air is equal to one. A moly-permalloy powder core also stores all of its energy in a gap, but the "gap" is actually distributed throughout the core within the non-magnetic matrix in which the magnetic particles are embedded. The effective permeability is governed by the matrix-magnetic particle ratio.

6.2.2 Core Selection.

From a wide variety of core permeabilities (or gap sizes) and saturation flux densities, how do we select the best core system and the proper operating point? Figure 10 shows 3 possible core systems and operating points that result in the same energy storage per cm^3.



The results are tabulated as follows:

CORE	Perm.	B	H	Joules/cm^3
1	25	1250	50	249×10^{-6}
2	100	2500	25	249×10^{-6}
3	400	5000	12.5	249×10^{-6}

Assuming these cores are the same size, total energy storage will be the same in all 3 cores. Since magnetic path length will also be the same, Hl is lowest for core #3. For that reason, which will be explained later, it is the best choice.

Core #2 is second best. Note that the permeability of core #2 cannot be increased further without losing energy storage capability, since it is already being operated near saturation.

Core #3 is underutilized. It could store more energy, but we don't need any more. It would be better to increase its permeability and operate at higher B and lower H, like core #2.

6.2.3 The Toroidal Structure.

In order to see why we desire the to operate at high flux density approaching B_{sat} and therefore achieve lowest magnetizing force (by employing the highest permeability or smallest gap we can get away with), we must look at the physical structure of the inductor.

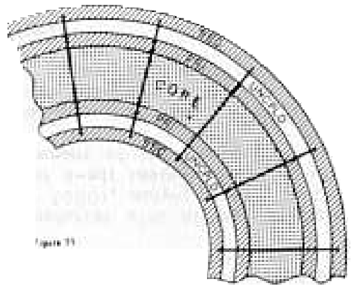


Figure 11 shows a section of an inductor with single primary and secondary windings on a toroidal mo-perm powder core, whose effective permeability is 100. All windings must be distributed evenly around the entire core to achieve good coupling and minimize external fields. Current flowing in the secondary (during transistor "off" time) generates a total magnetic field around the toroid:

$$H\ell = .4\pi NI \qquad (1)$$

The lines extending across the entire secondary coil, cutting through the primary coil and the core represent edge views of the magnetic field equipotential surfaces generated by the secondary ampere-turns. These equipotential surfaces must terminate on the windings which generate them, and the total field is the same inside the core as it is around the air space adjacent to the core, i.e. it is independent of the permeability.

Flux lines (not shown) are always normal to the magnetic field equipotentials. Flux lines must always be closed loops, and can never terminate. Flux lines in the toroidal structure will flow parallel to the core both inside and outside the core. The flux density, B, is proportional to the magnetic field intensity and the permeability:

$$B = \mu H$$

Since the core permeability is 100 and the field intensity is fairly uniform across the entire structure, flux density inside the core will be 100 times greater than the flux density outside the core but inside the secondary (there is zero flux outside the secondary).

__Now comes the moment of truth.__ The flux inside the primary winding represents energy that is 100% mutually coupled between primary and secondary, but the flux between primary and secondary is not linked to the primary and represents secondary leakage inductance.

Now it can be demonstrated why it is desirable to minimize Hl. If we quadruple the permeability of the core and double B (assuming the core does not saturate), H will be halved and the same energy will be stored in the core. However, H outside the core will also be halved, and since its permeability is unchanged (air = 1) Flux density outside the core is also halved. Therefore although the mutually coupled energy within the core is the same, the uncoupled energy between the windings is reduced by a factor of four. The number of turns will be halved to maintain the same inductance with the increase in permeability. The leakage inductance will then be 1/4 its original value.

With 1/2 the original turns, the windings can be slimmed down so that the interwinding volume will decrease substantially, resulting in a further reduction in the leakage inductance.

6.2.4 Completing the Design.

Now that we have a concept of the magnetic system and are hopefully convinced that the core should be operated near saturation at the maximum energy level required for the application, we can use Equation (4B) to calculate the permeability required for a given size core to achieve this objective. With the maximum energy , W, a known value and A_e and l_e for a given core size, set B equal to B_{max} (a little less than B_{sat} for the core material) and solve for the permeability.

For a gapped core system, use Equation (4C). Instead of solving for permeability (which is always 1 in the gap), solve for the gap length, l_g.

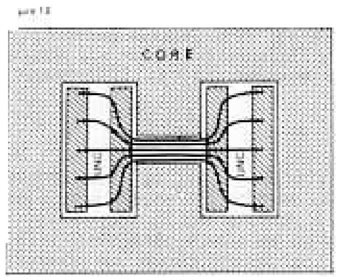
Then, whenever we need to know the number of turns, use Equation (1) as follows:

$$N = H\ell_e/.4\pi I = B_{max}\ell_e/.4\pi\mu I \quad \text{(mo-perm)} \qquad (1A)$$

$$N = H\ell_g/.4\pi I = B_{max}\ell_g/.4\pi 1I \quad \text{(gapped)} \qquad (1B)$$

6.2.5 The Gapped Structure.

Figure 12 shows the magnetic field in a gapped system.



The magnetic field is now severely distorted because the permeability around the loop is not uniform as it was in the toroid. The high permeability of the core (appx. 3000) makes it such a low impedance magnetically that the magnetic field is virtually all dropped across the gap. Because the field equipotential surfaces are so close within the gap (much more crowded than Figure 12 shows), flux density is also much greater within the gap, by a factor of usually more than 100 over the flux density in the interwinding volume outside of the gap. For this reason, uncoupled energy stored in the interwinding volume is much less than the coupled energy stored in the gap, even though gap volume is much smaller.

6.3 DESIGN POINTERS

<u>Minimize Hℓ.</u>

1....Operate close to B_{sat}. Use the smallest gap length or highest possible effective permeability for a given size core.

2....Use magnetic material with highest possible B_{sat}. Ferrite is poorest (B_{sat} of 3000), moly-permalloy is better (6000), but best is silicon-steel 4-mil laminations (15,000). Latter can be used because AC component of inductor current is small.

3....Use the largest size core you can afford in terms of cost and size. Doubled core dimensions (8 times core volume) operated at B_{sat} can reduce leakage inductance 16 times.

<u>Optimize the Winding.</u>

1....For a gapped core, best window shape gives a long, thin winding, which maximizes the interwinding magnetic path length and minimizes the magnetic area. This reduces total flux and decreases H. The larger window associated with a larger core will enable an improved aspect ratio of the winding.

2....Use minimum size wire possible considering winding resistance, power loss and temperature rise permitted. Don't fill the window unless forced to with minimum core size.

3....Spread winding uniformly over entire length available. Use multiple paralleled turns of fine wire laid flat rather than single turns of larger O.D., to get a smooth winding of minimum thickness.

4....Multifilar winding of all secondaries will result in dramatic reduction of leakage inductance between secondaries, permitting excellent cross-regulation. If it is not possible to include the primary because of large turns ratio or insulation voltage requirements, inter-leaving 1/2 primary, secondaries, then 1/2 primary will reduce primary to secondary leakage reactance by almost a factor of 4.

7. CONCLUSIONS

The relative merits of the CICD buck converter, conventional voltage-driven buck converter, and current driven boost converter are summarized in Table I.

(Added 8/81): An example of a 100 watt, multiple output CICD buck converter operating off the rectified 220 volt line is shown in Figure 13.

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3) R. D. Middlebrook and S. Cuk, "A Generalized Approach to Modelling Switching-Converter Power Stages," IEEE Power Electronics Specialists Conference, 1976 Record, pp. 18-34.

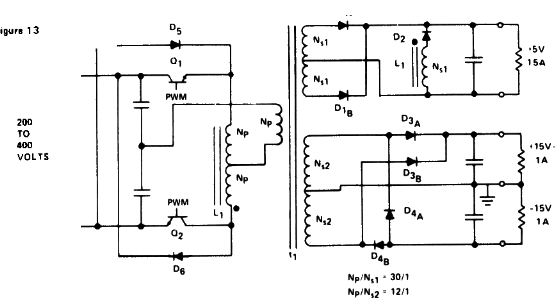
4) H. Matsuo and K. Harada, "New Energy-Storage DC-DC Converter with Multiple Outputs," Solid State Power Conversion, Nov./Dec. 1978, pp. 54-56.

TABLE 1

CHARACTERISTIC	BUCK		
Control Circuit			
Dead Band Control	yes	no	* no
Right-Half Plane Zero	* no	yes	* no
Asymmetry Correction	yes	no	* no
Soft Start	yes	no	* no
Output Current			
Discontinuous	* no	yes	* no
Hi RMS Capacitor I	* no	yes	* no
Severe ESR/ESL Req.	* no	yes	* no
Input Current			
Discontinuous	yes	no	yes
EMI	yes	no	yes
Current Spikes	yes	no	* no
Voltage Spikes	* no	yes	yes
Semiconductor Requirements			
Center-tap Rect. PIV	$2 \cdot V_{in}/n$	$2 \cdot V_{out}$	* $2 \cdot V_{out}$
" " " t_{rr}	Fast	Moderate	* Moderate
Free-wheel Rect. PIV	N/A	N/A	* V_{in}/n
" " " t_{rr}	N/A	N/A	Fast
Rectifier peak I_f	* I_{out}	> I_{out}	* I_{out}
Transistor Vce	$2 \cdot V_{in}$	$2 \cdot n \cdot V_{out}$	$V_{in} + n \cdot V_{out}$
Transistor t_{stg}	Fast	Moderate	* Moderate
Failure Mode			
Transistor/Rect. Open	Catastrophic	Catastrophic	* Reduced Output
Multiple Outputs			
Cross-Regulation	Poor	Good	* Very Good

* = Desirable Attribute

Figure 13



Q_1, Q_2	UMT13005	(400V, 4A, TO-220)
D_{1A}, D_{1B}	USD735C	(DUAL SCHOTTKY, 35V, 16A, TO-220)
D_2	USD735	(SCHOTTKY, 35V, 16A, TO-220)
$D_{3A}, D_{3B}, D_{4A}, D_{4B}$	SES5001	(50V, 2A, 100nS, Axial)
D_5, D_6	1N5617	(400V, 1A, Axial)