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PRESENTS

"MODERN SWITCHED-MODE CONVERTERS WITH INTEGRATED MAGNETICS"

An Application Note

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Abstract

A tutorial overview of the use of integrated magnetic concepts in power converter circuits is presented, including coupled-inductors and new designs wherein all magnetic elements of a converter, including transformers, are blended together on single core structures.

1.0 INTRODUCTION

In the heydays of vacuum tube technology when sizes of electronic systems were measured in terms of room dimensions, their power supplies, or power processing circuits, went unnoticed as their contributions to system volume and weight represented only a small percentage of total size. In most instances, these power supplies were relatively simple rectifier and filter networks preceded by a low frequency power transformer and, if more control on output regulation was needed, a dissipative series or shunt regulator network was added. Power waste was of little concern in these designs, as power supply size could be increased to include heat sinks for removing heat, if need be. For the most part, the added system volume for the often massive heat sinks was of little concern to the designer or to the user.

The tremendous advances made in solid-state electronics in this quarter century, coupled with a more efficiency-minded and cost-conscious society, have caused today's circuit designer to turn to radically different and often unusual solutions to power conversion problems than were followed in the past. Prime examples are the increased use of high frequency switchmode power converters of many varieties, including topologies using resonant power switching approaches.

While a high value of power processing efficiency is most certainly an important goal of a modern power supply system, small size and low cost have higher priorities today, especially in the commercial power supply industries. As a result, power electronics designers are taking careful stock of every engineering method that will help them to achieve these goals.

The current interest in very high frequency resonant converter circuits is a good example of the motivation to reduce power supply size. In theory, such circuits can achieve the goal of smaller size due to the reduction in the physical sizes of energy storage devices, such as power inductors and filter capacitors, required at higher operational frequencies. However, in fact, at these frequencies, the second-order parasitic properties of the energy storage devices now become dominant and their values important in determining conversion characteristics and efficiency.

Another area of renewed interest to reduce the size of the magnetic content of a power conversion network is the concept of *integrated magnetics (IM)* illustrated in Fig. 1. Analogous to the placement of small-signal electronic circuits on a common substrate of silicon, integrated magnetic design techniques are those which result in the blending of *all* or selected magnetic functions of a converter on a common magnetic core structure. If performed thoughtfully, IM design methods can also produce a converter design of reduced electrical stresses on surrounding converter components and, in particular in the instance of inductive IM filter arrangements, reduced conducted ripple current magnitudes on input *and* output terminals of a converter. Once thought to have application in only special converter circuits [1], IM methods have been shown recently [2,3] to be *universal* in this regard, and can be implemented in *any* converter topology, regardless of its complexity.

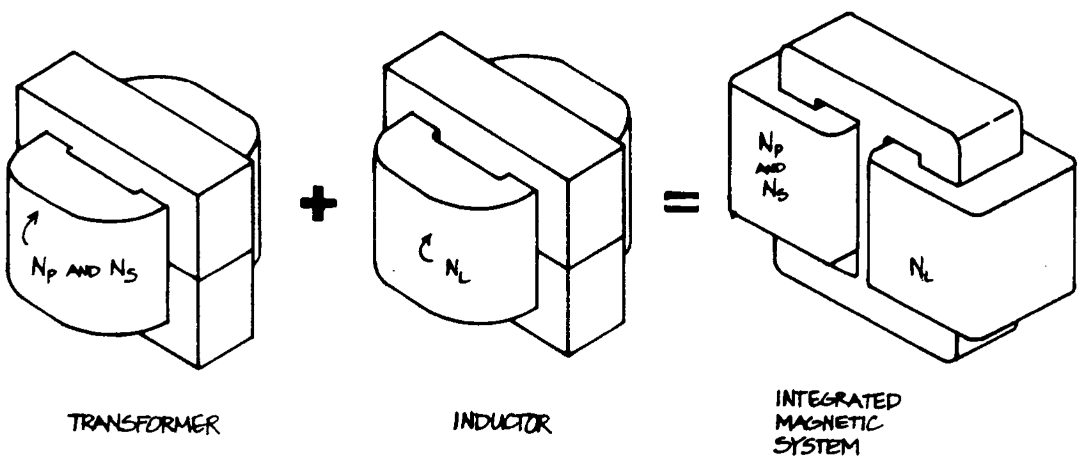


Fig. 1. The concept of integrated magnetics - blending transformer and inductor functions on a single magnetic core.

In this paper, we will explore the concepts of IM, including magnetic coupling of inductors in basic power filter networks, conventional switchmode converters where inductive and transformer functions have been consolidated into a single magnetic component, and recent usage in resonant power conversion circuit topologies.

2.0 COUPLED-INDUCTOR TECHNIQUES

In a broad sense, IM design techniques can be categorized into three different magnetic circuit areas of converters, namely those methods which are directed toward integration of only inductive components, those associated with only transformer functions, and those orientated to blend transformer and inductive functions simultaneously. Of these three categories, integration of inductive functions alone, or *coupled-inductor* arrangements, has been and continues to be the most common technique found in practice today.

2.1 Historical Background

Although the design practice of implementing inductive energy storage elements of particular DC-DC converter circuits [1,5,6] on single magnetic core structures became commonplace in the late 1970's, there is historical evidence of exploration of the advantages of these methods prior to 1928. It was during that year that one of the first United States patent applications dealing with coupled-inductor concepts was filed.

Fig. 2 is a reproduction of the two primary circuit illustrations of the resulting patent [7], issued five years later. The first circuit diagram depicted a typical, or prior art, 50Hz AC-DC power supply design of the period for a basic radio receiver system. As shown, two cascaded LC filters were employed, the first of which included a series LC "trap" filter as a shunt network. This "trap" filter was designed so as to be tuned to the primary harmonic frequency of the transformed and rectified AC line, while the elements of the second stage of the overall filter system selected to reduce the higher harmonics.

In the second part of Fig. 2, the idea of magnetically coupling the two inductors of the first stage filter is illustrated. Here, the inventor proposes to add a second winding to the primary inductor of the first-stage filter, forming a secondary inductance that could replace the inductor of the "trap" filter section described earlier in the first part of Fig. 1. As detailed in the body of the patent, the presence of mutual inductance between the two windings of the new primary inductor could then be used to increase the sensitivity of the filter, making it more effective as a ripple reduction network.

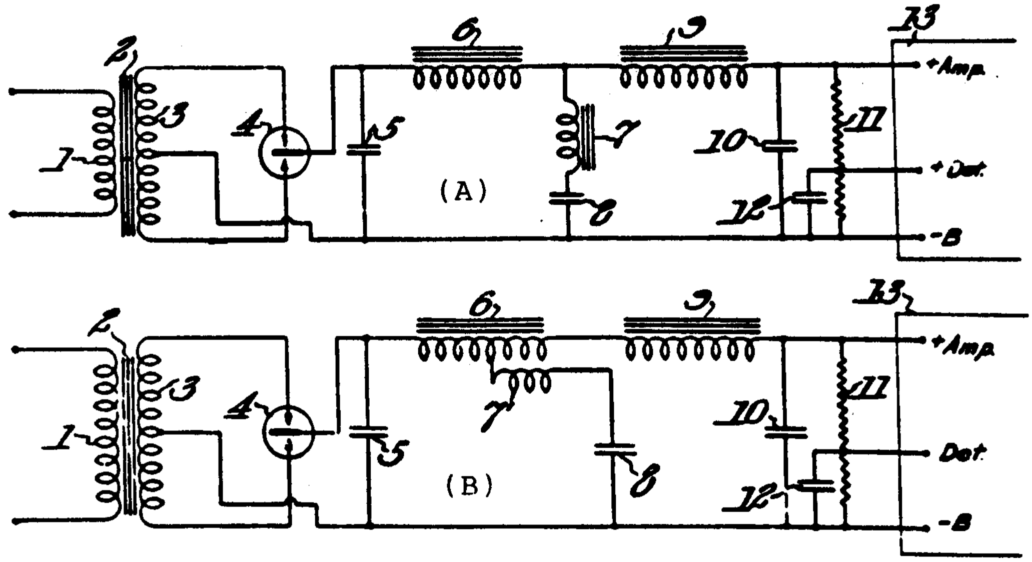


Fig. 2. Early coupled-inductor concepts for power filter networks (same as Figs. 1 & 2 in ref. [7]).

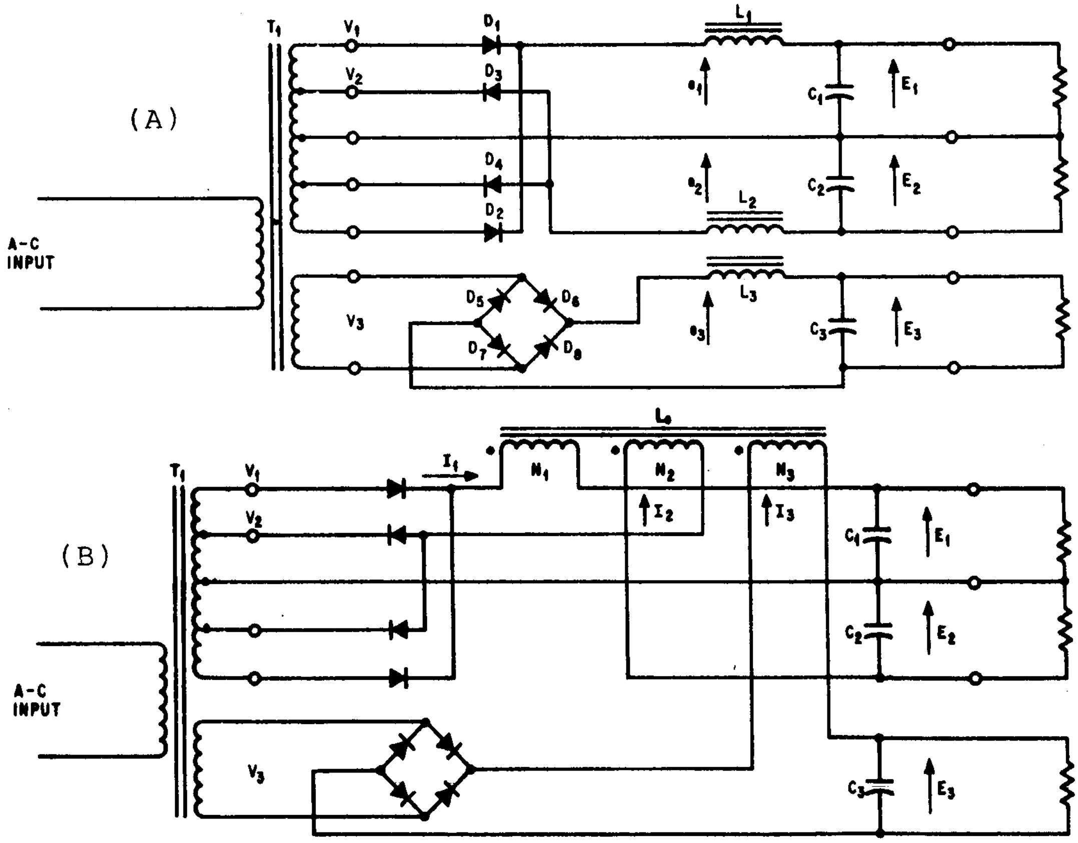


Fig. 3. Coupled output inductors in an unregulated AC-DC power supply (reproduced from ref. [8], pages 94 and 95).

While this early patent dealt with the use of coupled-inductor techniques for increased ripple rejection of a LC filter network, some years later the advantages of this method for blending a multiplicity of filter inductors in a simple AC-DC power supply was explored openly in an ELECTRONICS magazine article [8].

Two of the circuit diagrams of this 1967 article are reproduced in Fig. 3. Part (a) of this figure shows the original power supply circuit, while part (b) is the circuit arrangement after the three filter inductors of the power supply have been mounted on one magnetic core. In this highly informative article, the author noted that, in the circuit of Fig. 3a, the rectified 60Hz AC voltages appearing across each filter inductor are proportional to one another by the corresponding secondary-to-primary turns ratios of the transformer, T1, and that the phasing of these AC voltages, timewise, must always be identical. Based on these conditions, it is shown that the inductors may be combined together on one magnetic core, resulting in the circuit configuration of Fig. 3(b).

Three years later, another article dealing with a passive filter design very similar to that shown in the 1933 patent discussed earlier appeared in an IEEE publication [9]. Here, the author was concerned with the development of a LC filter approach that would permit the use of a much smaller value and size of associated output capacitance, while retaining or improving the desired ripple rejection properties of an equivalent standard second-order filter circuit. Fig. 4 are diagrams taken from this article, illustrating the concept and the passive network suggested to meet these goals. Note the similarities of the circuit of Fig. 4(b) to that shown earlier in the second network of Fig. 2.

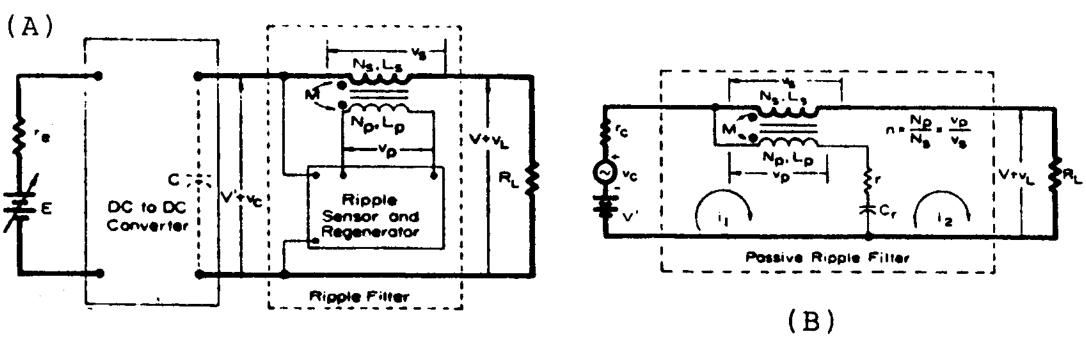


Fig. 4. Another example of coupled-inductor methods to reduce output filter capacitance (same as Figs. 1(b) & 2(a) in ref. [9]).

Beginning in 1975 and continuing even today, patents dealing with the similar (if not identical) idea of coupled-inductors in output filters of certain DC-DC switchmode converters are in evidence (e.g., see references [10] thru [18] noted in this paper).

In addition, there have been many technical papers written on this category of IM over the past ten years, too numerous to reference here. The basic ideas behind the methods of coupled-inductors can be found, however, in publications [1],[5],and [6] as applied to 'Cuk and transformer-isolated DC-DC buck converters, respectively.

2.2 Coupling Fundamentals & Filter Examples

Recently, there have been various attempts to explain the foundations for coupling of inductors in selected converter topologies [1,5,6], using magnetic core properties, air gap lengths of cores, and mathematic modeling variables (such as coupling coefficents and mutual inductance values). However, a more palatable and, for most electrical engineers, a more understandable approach to an explanation employs the use of a basic electric circuit model of a two-winding magnetic component [2]. This familiar model is shown in Fig. 5(A).

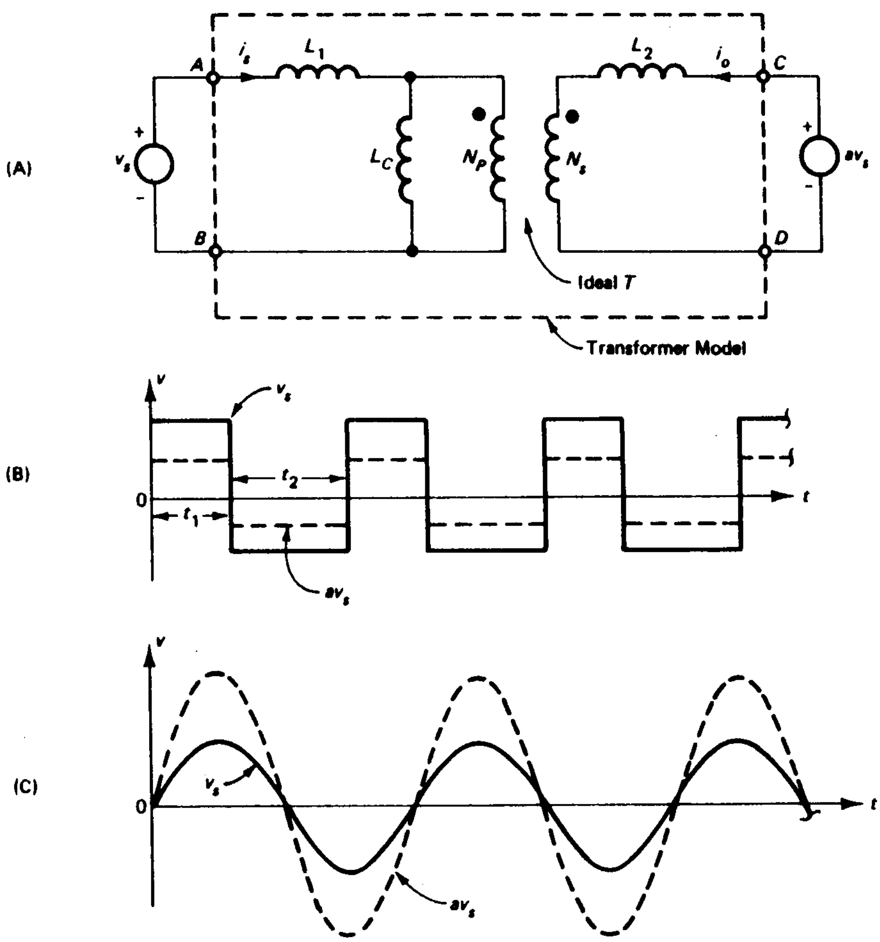


Fig. 5. Two-winding magnetic model (A) driven by proportional voltage sources of different form factors (B) or (C).

We will now assume that each circuit port of the model is being driven by ideal AC voltage sources (v_s & av_s), with amplitudes being proportional *at all times* by a constant defined by "a" in Fig. 5(A). We will also assume that the frequencies and phase relationships of these voltages are *identical*. We will place *no* other restrictions on the characteristics of these voltage sources, such as waveform shapes. They can be of a quasi-squarewave form (Fig. 5(B)), sinusoidal (Fig. 5(C)), or any other shape as long as the restrictions of similar amplitudes, polarity phasings and frequencies are met.

In the model of Fig. 5(A), L_1 and L_2 represent the *leakage* inductances associated with the windings of N_P and N_S turns, respectively, while inductance L_C represents the *self-inductance* value associated with winding N_P. In the case of a two-winding inductor, the value of L_C would be set by the cross-sectional area of the magnetic core of the unit and the length of any air gap introduced into the magnetic path of the core by the designer. If the model was to represent a two-winding transformer, L_C would be of relatively large value, representing the magnetization energy needed to excite the core.

For the winding current directions assumed in Fig. 5(A), we can immediately say, from basic electrical theory:

$$v_s - v_C = L_1 di_s/dt \ , \ av_s - nv_C = L_2 di_o/dt \tag{1}$$

where $n = N_S/N_P$, the *ideal* turns ratio of the magnetic assembly, with v_C as the AC voltage appearing across the core inductance, L_C. This voltage is produced by a summation of the two winding currents as:

$$v_C = L_C d(i_s + ni_o)/dt = L_C di_s/dt + nL_C di_o/dt \tag{2}$$

From the two relationships of (1), it is seen that they can be combined to eliminate the value of v_C. Doing so then yields:

$$(n - a)v_s = nL_1 di_s/dt - L_2 di_o/dt \tag{3}$$

The simple voltage expressions of (1) thru (3) can be used to predict the *effective* inductances seen by each of the two voltage sources of the model, given known values of n and a. A typical example found in practice today is to design the coupled-inductor such that $n = a$, i.e. the proportionality constant between driving voltages is made to match that of the ideal turns ratio of the inductor windings. Substitution of this condition into (1) thru (3) then shows:

$$L_{AB} = L_1 + L_C[1 + (a^2 L_1/L_2)] \quad (4)$$
$$L_{CD} = L_2 + L_C[a^2 + (L_2/L_1)] \quad (5)$$

to be the *effective* inductances presented to the voltage sources v_s and av_s, respectively when $n = a$. Also, the rate of winding current changes with respect to time will then be defined as:

$$di_s/dt = v_s/L_{AB}\ ,\ di_o/dt = av_s/L_{CD}\ ,$$
$$di_s/dt = (L_2/aL_1)di_o/dt \quad (6)$$

From (6), it is seen that the values of leakage inductances play an important role in determining the difference in winding current changes with time for the case of $n = a$. It is also interesting to note that, in this case, if leakage inductance values were *equal*, the effective inductances presented to each of the two voltage sources would also be *equal*, using the relationships of (5) above.

Before continuing our discussions of circuit applications, there is one special case of coupled-inductor design of practical importance. This special term applied to this case is the *zero ripple current* condition [1], i.e. a design where the AC ripple current in one or more windings of a coupled-inductor arrangement is forced to be very close to zero value.

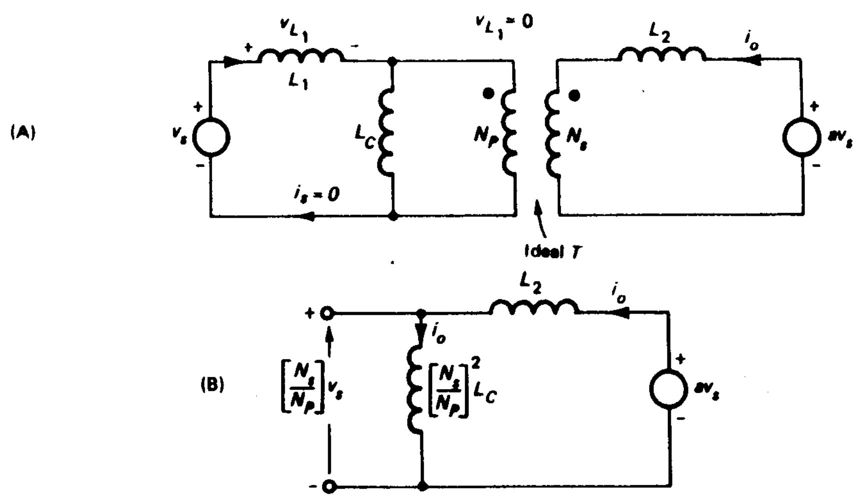


Fig. 6. Circuit model of Fig. 5 (A) reduced to a simple form (B) when i_s is zero (see text).

In Fig. 6(A), the condition of *zero* AC ripple current has been assumed for the current i_s in the electrical model of Fig. 5 for the case of a two-winding inductor. If this AC current is zero, then it follows that di_s/dt also must be *zero* in this instance, and that the AC voltage drop across the leakage inductance L_1 is also *zero*. Given these conditions, the circuit of Fig. 6(A) may then be reduced to that shown in Fig. 6(B). From this very simple network and the voltage conditions shown therein, it can be shown that:

$$L_2 = [(a/n) - 1]n^2L_C \tag{7}$$

Thus, if the conditions stated in (7) are satisfied, there will be *no* dynamic AC current drawn from the voltage source v_s. As far as the ripple current drawn from the source av_s, the use of (7) in the reduced network of Fig. 6(B) reveals:

$$di_o/dt = av_s/(anL_C) = v_s/(nL_C) \tag{8}$$

A similar exercise can be performed, using the circuit model of Fig. 5(A), under the assumption that the AC current i_o has *zero* value. In this case, we find that, for:

$$L_1 = [(n/a) - 1]L_C \tag{9}$$

this current can indeed be reduced to zero, while the rate of change of the other current, i_s, will become:

$$di_s/dt = v_s/(nL_C/a) \tag{10}$$

Fig. 7 illustrates the current waveforms in the model of Fig. 5(A) depicted by the conditions of (7) thru (10), assuming the two voltage sources driving the circuit have quasi-squarewave form factors similar to those diagrammed in Fig. 5(B).

The above exercises on a simple electrical model of a magnetic system serve to demonstrate the feasibility of *magnetically* forcing ripple current in selected windings of a coupled-inductor system to *zero value by design*. Ideally, this implies the effective inductance of the selected winding(s) will have a value approaching infinity. Therefore, if such an inductor were used in a LC filter application, there would be *no* need for an output capacitor to divert inductive ripple currents. One can now appreciate why the coupled-inductor filter arrangements shown earlier in Figs. 2 and 4 can be made, by proper magnetics design, to be more effective than a conventional LC network.

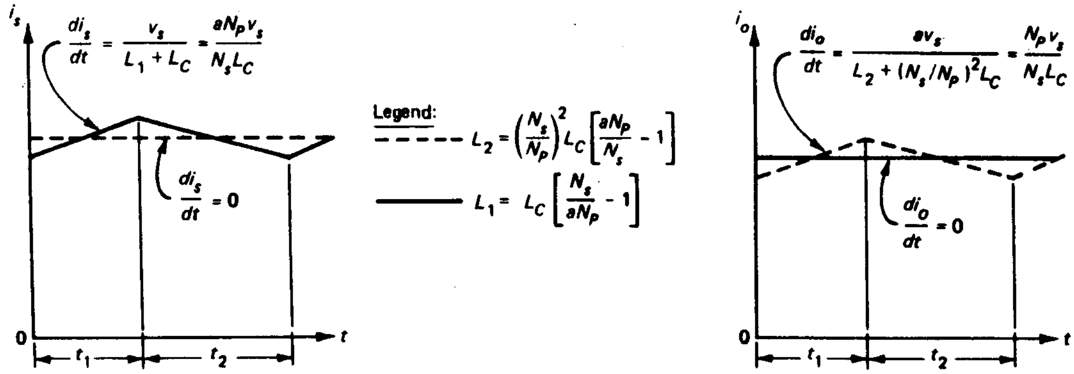


Fig. 7. Ideal terminal currents in the model of Fig. 5(A) under stated L_1 or L_2 values (voltage forms of Fig. 5(B) assumed).

In a practical coupled-inductor system where magnetic ripple control is being exercised, perfect ripple reduction *cannot* be attained. Other parasitics present in the magnetic assembly, not accounted for in the semi-ideal circuit model of Fig. 5(A), are the causes. Such parasitics include capacitances within the magnetic and resistances of windings. However, in a coupled-inductor destined for use in a high frequency converter, the actual values of these parasitic elements are usually made to be small by design. Therefore, while perfect ripple current reduction is not possible, significant reductions in the order of 100 or better can be achieved in a practical coupled-inductor design where leakage inductance values can be consistently controlled. It is also possible, where leakage inductances are small, to add *external* inductances inline with windings for ripple current control [2,14,16].

Simultaneous ripple control of *more* than one winding current by another winding of a magnetic assembly is also feasible, if the core of the assembly has more than one major flux path associated with it. An example of this capability is shown in the circuit of Fig. 8.

Here, a three-winding coupled-inductor arrangement is used in the output filter networks of a three-output "push-pull" DC-DC converter. In this diagram, D denotes the duty cycle of conduction of either of the primary switches of the converter (S1 or S2) over a switching period, T_s. Note that the dynamic voltages appearing across each of the three windings of the magnetic will be quasi-squarewave, similar to those indicated in Fig. 5(B), and their amplitudes will differ, ideally, by the secondary-to-primary turns ratio of the converter's transformer, T1. The core used for the inductor here would be typically an E-E or E-I in form, with a winding bobbin placed on each of its three magnetic path "legs".

Following a similar procedure of forming an electric circuit model for this three-winding inductor system, relationships between the turns ratios of T1, winding turns of the inductor and air gap lengths can be formulated [2]. From these equations, gap lengths can be found that will permit winding N of the inductor assembly to control and reduce the AC ripple currents of windings N_1 and N_2 significantly *without* upsetting the desired DC-to-DC conversion action of the overall circuit to any of three outputs!

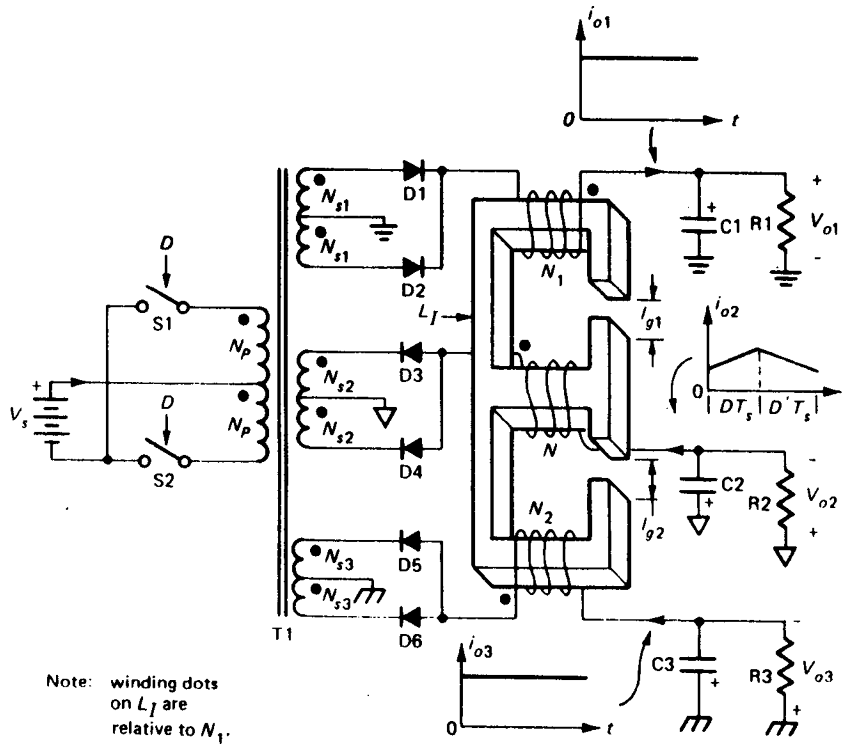


Fig. 8. Achieving ripple current control of two load currents simultaneously in a multi-output converter.

2.3 New Coupled-Inductor DC-DC Switchmode Converters

In converter circuits where dynamic inductive voltages are *not* proportional under all operational conditions, the designer can resort to the use of magnetic cores with *multiple flux paths* to find a way to combine the inductances. In many cases, however, by the reorientation of the functions of the inductances within the overall converter circuit while retaining the desired filter function of each inductance, it may be possible to then place them on a magnetic core with only one major flux path.

For example, the power inductor of a simple DC-DC buck converter and the inductance of a LC filter network preceding the converter do not have AC voltages that are dynamically proportional in either amplitude or in time. In a dual fashion, the AC voltage of an inductor in a second-stage filter network tied to the output of a DC-DC boost converter is not dynamically similar in form to the power inductor of the converter section. These circuits and filters are shown in Figs. 9(A) and 9(B).

Recognizing that the functions of the inductances in the filter circuit portions of the designs of Figs. 9(A) and 9(B) are to eliminate the pulsating currents either at the input or the output of the contemporary converter topology, it is possible to change the location of these inductances to retain their filter functions, while causing their AC voltages to now become proportional in the desired manner for coupling. Figs. 9(C) and 9(D) are diagrams of the resulting buck [19] and boost [20] designs, respectively.

Note that the reorientation process has *not* added any more circuit components to either circuit. Also, examination of the various currents and voltages seen by the power switches or diodes shows that *no* stress levels have been changed. From a "black-box" standpoint, both of these new coupled-inductor designs retain the desired buck or boost voltage transfer functions of the original designs of Figs. 9(A) or 9(B), including the highly desirable *non-pulsating* input *and* output current waveform shapes.

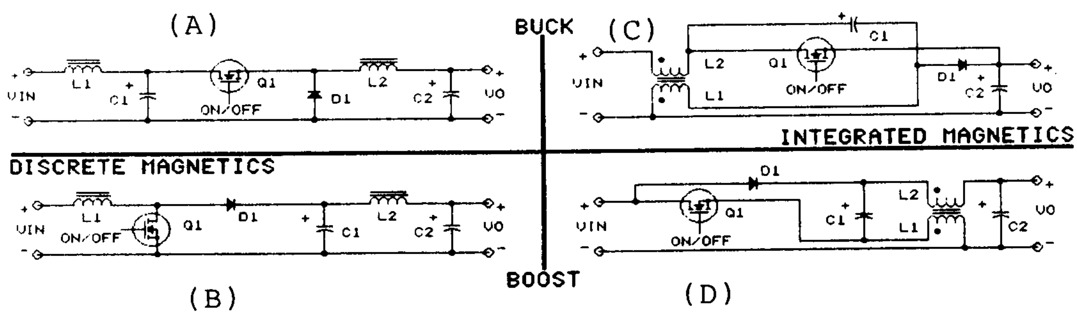


Fig. 9. Conventional buck (A) and boost (B) DC-DC converters with input/output filter networks, and new designs (C,D) that permit coupling of filter and power inductors.

The inductor voltages in either of the two new designs are now in 1:1 proportion (i.e., $a = 1$), and are in phase relative to polarity for each switch state (ON or OFF) Therefore, the magnetic design techniques described earlier to attenuate ripple currents in either inductances of both designs can be easily applied, if desired.

3.0 IM TRANSFORMER ARRANGEMENTS

In the category of IM dealing with the blending of transformer functions, here again we find much historical evidence of use in the past. Perhaps the best example of IM transformers is the classic example of the three-phase transformer. While it is possible to use three separate magnetic core assemblies, one for each phase, to implement a three-phase transformer system, a more efficient approach is to use a single E-E or E-I core, and to place primary and secondary windings of each phase on each of the three "legs" of the core. Analysis of the phase relationships of the three fluxes then shows that they must sum to *zero* value in the center "leg" of the core. Thus, a *single* core arrangement with *less* material can replace the functions provided originally by a *three* core system.

Recently, designers have found techniques where the integration of transformer functions on a single magnetic core will eliminate the need for an added converter circuit! An excellent example of this possibility is the converter design illustrated in Fig. 10 [2].

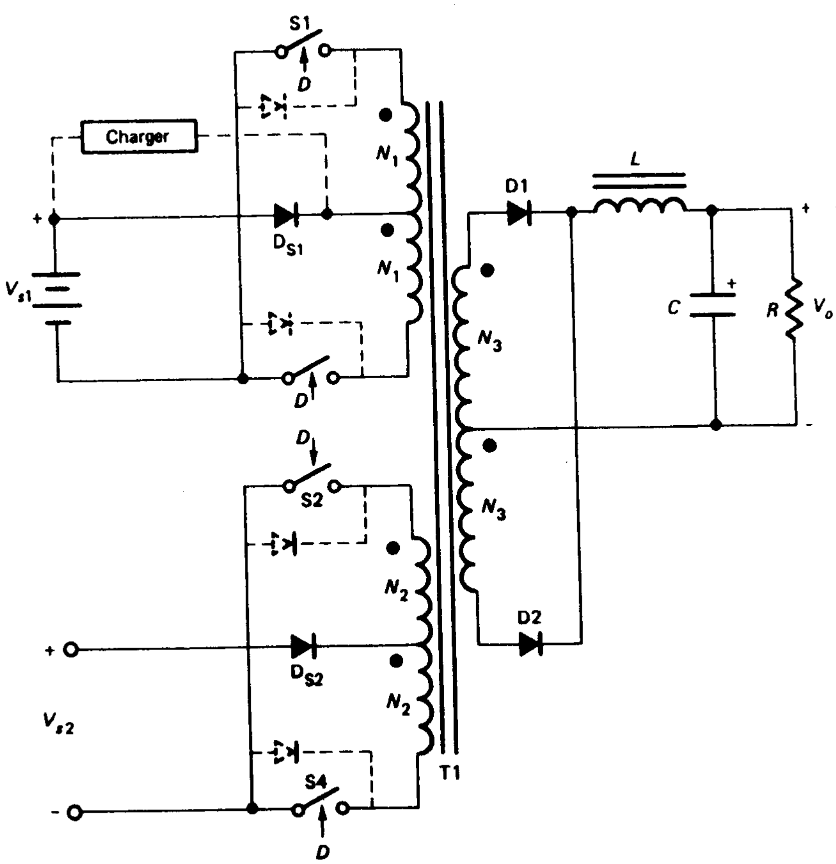


Fig. 10. Integrating a UPS battery backup feature into a DC-DC power converter circuit.

Traditionally, in power supplies where a low voltage battery backup is required to sustain outputs during losses of primary DC voltage, a second converter is added to boost the battery voltage to a level comparable to the primary voltage value. The primary voltage and the boosted battery potential are then "ORed" together at the input to the primary power converter circuit of the supply. If the battery is rechargable, it is not uncommon for the power supply to have a charger circuit that operates from the primary DC source when it is present.

In contrast to adding a second converter for boosting battery potential, the design of Fig. 10 uses a second primary winding on the main converter transformer of the power supply to provide a battery interface. In this illustration, V_{s2} represents the main DC power source, with V_{s1} as the battery. Windings labeled as N_1 associated with the main transformer, T1, are the added primaries.

During normal operation, simple level-detecting electronics (not shown in Fig. 10) examining the voltage value of V_{s2} for proper range prevent the two power switches tied to windings N_1 from operating, while allowing the two switches associated with windings N_2 to assume converter control. Assuming that the turns ratio between N_2 and N_1 has been selected such that the battery diode D_{S1} is reverse-biased, the added primary windings will then act as *secondaries*, allowing the charger electronics to operate, using the primary power source to provide recharge energy.

Under conditions of a loss or out-of-range primary DC power, the level sensors monitoring this voltage are then designed to route main converter control signals to the upper switches tied to windings N_1 during the next switching period of the converter circuit. At the same time, the main switches tied to windings N_2 are deactivated completely. These actions then permit the battery to supply converter power thru windings N_1, diode D_{S1}, and the associated power switches. Note that diode D_{S2}, in line with the primary power source, now prevents the source from loading the converter transformer while the battery windings are in control.

Not only does the design scheme illustrated in Fig. 10 permit the elimination of a battery-boost converter, switchover from the primary source to its battery backup, or vice versa, can be made to be very fast. Since the switchover can be done in essentially one switching period of the converter, transfer times in microsecond intervals for high converter frequencies can be accomplished by this scheme!

The design of Fig. 10 is a particularly vivid example of how the integration of transformer functions of separate converter systems can produce an overall power supply of significant lower cost, weight, and size while, at the same time, enhancing performance.

4.0 COMPLETE IM CONVERTER CIRCUITS & SYSTEMS

Prior to 1984, it was commonly believed that only special DC-DC circuit arrangements [1] were of forms where their transformer *and* inductive elements could be integrated on a common magnetic core structure. These particular circuits were those in which dynamic voltages across each magnetic element, whether it be a transformer winding or a power inductor, remain dynamically proportional under all possible operational conditions. Recalling our earlier discussions of basic requirements for magnetic coupling of inductors, it follows then that integration of *all* magnetic components of these special converter topologies will follow the same design rules, and that the resultant IM assembly could use a magnetic core that possesses only one major material flux path, if desired.

However, if a core with more than one major flux path can be used, such as an E-E or an E-I structure where three paths are possible, then it is possible for the magnetic components of *any* DC-DC converter circuit or system to be blended together. IM versions of *all* converter topologies [2], regardless of conversion characteristics or complexity of inputs and outputs, become readily feasible.

Examples of multi-path IM technologies are two U.S. design patents issued to the IBM Corporation [21] in the early 1970's for variations of "push-pull" quasi-squarewave converters of the buck variety. The advantages of multipath IM relative to control of inductive ripple currents have been demonstrated by 'Cuk [1,13,24] in his converter arrangements since 1981. Even more recently, two other U.S. patents associated with potential IM methods for selected buck and coalesced flyback-forward converter circuits [22,23] have been awarded.

4.1 Quasi-Squarewave Switchmode Converters

As is the case for conventional buck or boost-derived DC-DC switchmode converters, magnetic flux levels within the cores of contemporary inductors or transformers are *not* dynamically equal for all possible operational conditions. However, under specified turns relationships between inductor and transformer windings, it then becomes possible to develop a multi-path IM assembly that will produce the desired flux amplitudes and directions required in the inductive and transformer elements when the IM is installed in the user converter circuit.

For example, consider the single-switch IM DC-DC converter design of Fig. 11(A). This topology is termed a "current-fed" DC-DC converter circuit by many designers, while others refer to it as a "single-ended Weinberg" converter in recognition of the engineer who first noted the advantages of this topology in a 1974 European research publication [25].

If the turns ratio equality stated in the equation below Fig. 11(A) is satisfied by design, this converter will act as a "buck-derived" topology, with *ideal* voltage/current transfer characteristics *(continuous mode)* given by:

$$V_o/V_s = I_s/I_o = nD, \quad n = N_s/N_p \tag{11}$$

with D as the duty cycle of conduction of power switch $S1$ over a single switching period T_s of the converter. The main advantage of this particular converter over a conventional "forward" converter circuit is the reduced OFF voltage stress that can be achieved on its output diodes and its power switch element.

(A)

(B)

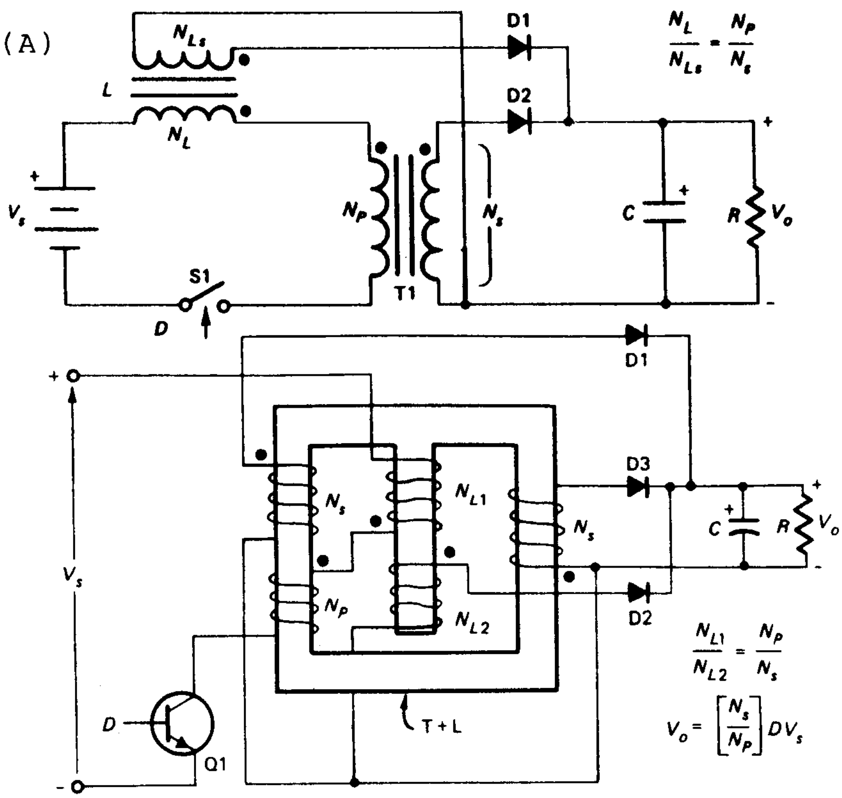


Fig. 11. A buck-derived Weinberg converter (A) and a three-bobbin integrated-magnetic version (B). NOTE: Reset winding for T1 in (A) not shown.

For the two states of the power switch in the design of Fig. 11(A) (i.e., ON or OFF), an evaluation of the voltages appearing across each of its two magnetic components will show that the potentials are *not* dynamically proportional for all possible values of the switch duty cycle, *D*. However, the locations of the inductor and transformer windings within the converter circuit as shown do permit inter-relationships to be developed between the various winding potentials for the times when switch *S1* is ON, and when it is OFF.

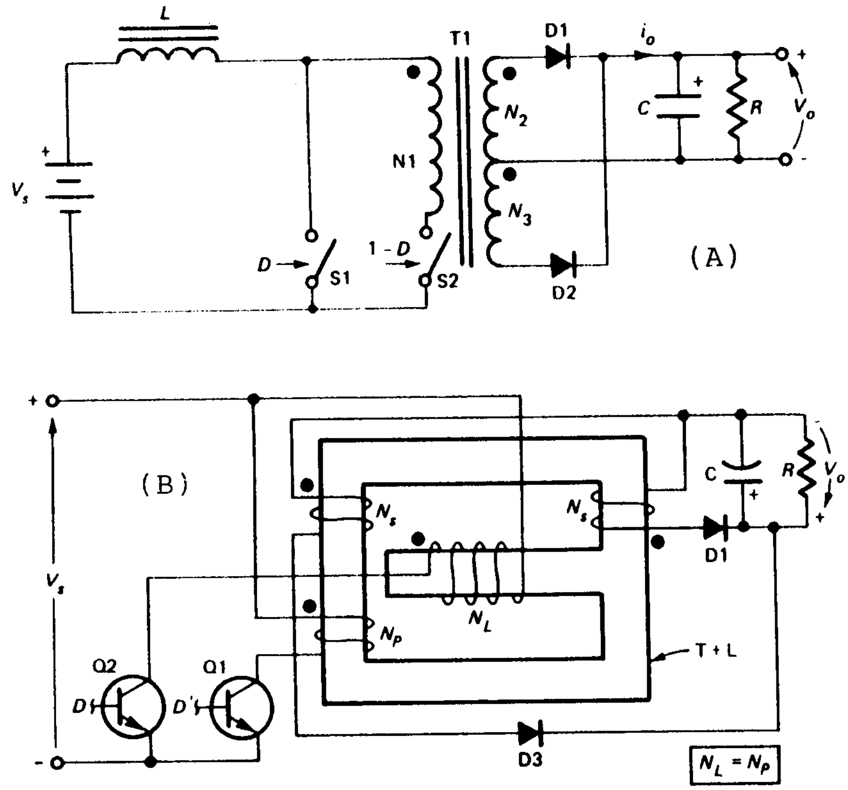
With these potential relationships in hand, and the use of basic magnetic laws that define flux change as a function of voltage, time and winding turns, a magnetic core structure can then be formulated for each converter state. The voltage relationships, as such, serve as "road maps" for the required flux paths within the structure and where windings must be placed to produce the required flux changes.

Once the magnetic system has been developed from the above exercises, all that remains is to add diode and switch elements to the windings of the IM to cause the required flux changes to occur for each converter state.

Following this deductive path of reasoning, the IM topology of Fig. 11(B) can be found [2] for the converter design in Fig. 11(A). Note that a three-path magnetic core structure is used (i.e., an E-E or E-I core set). The two "outer" legs of the core serve as common flux paths for both inductive and transformer actions, while the air-gapped "inner" leg is used for inductive energy storage alone. Under the same restrictions of equal turns ratio between inductive and transformer windings as noted in Fig. 11(A), the IM design of Fig. 11(B) has the *same* voltage and current transfer ratio as was stated earlier in (11) for the original discrete-magnetic version. Note that, in Fig. 11(B), the outer-leg winding associated with diode *D3* is used to provide a means for removing magnetization energy from transformer actions whenever the power switch *Q1* is turned OFF. In this IM example, this energy is shown absorbed by the output of the converter via *D3*. This winding could be attached to the input of the converter by moving the cathode connection of *D3* to the positive terminal of the incoming voltage, if so desired, with a corresponding change in the turns of this winding to assure core reset action during each switching period.

Development of an IM concept, following the deductive methods outlined earlier, is applicable to *any* converter circuit, regardless of its transfer characteristics. For example, the two-switch *boost* converter of Fig. 12(A), with ideal I/O transfer expressions given by:

$$V_o/V_s = I_s/I_o = n/D', \quad n = N_2/N1, \quad D' = 1 - D \qquad (12)$$



Note: Dots relative to N_P.

Fig. 12. A boost converter (A) and a three-bobbin integrated-magnetic version (B) where $N_L = N_P$.

for a continuous mode of energy storage, can be placed into an IM form as diagrammed in Fig. 12(B). Here, it is necessary to place an additional design restriction of $N_L = N_P$, requiring inductive turns to equal those of the primary transformer winding. Note that transistor switches *Q2* and *Q1* provide the functions of *S1* and *S2*, respectively, in Fig. 12(A), and that $N_P = N1$, $N_S = N_2 = N_3$. Also, diode *D3* performs the same reset control function for the IM as does diode *D2* in Fig. 12(A).

It must be emphasized that the deductive synthesis process can yield more than one IM circuit possibility [3,4], with some variations being less complex in windings required or in the number of wiring bobbins needed for the IM. For example, the IM versions shown in Figs. 11(B) and 12(B) will require an IM assembly with *three* wiring bobbins, one per core leg. There are other versions, however, that would require only *two* bobbins.

An example of a two-bobbin IM assembly is shown in Fig. 13(B), with its discrete-magnetic (DM) version in Fig. 13(A). One will recognize the DM topology as a buck-derived *forward* DC-DC converter, a favorite circuit used in 100-200 watt DC power supplies today. Note that, in the IM version in Fig. 13(B), one outer leg of the magnetic core serves as the inductive storage area, while the other is used for transformer actions. The center, or inner, leg in this design is then used as a common magnetic path for both transformer and inductive fluxes. The use of an outer core leg for inductive windings permits the window areas available in the core structure to be used more effectively and, in cases where a *custom* core design can be considered, these areas can be *adjusted* for a smaller overall magnetic assembly!

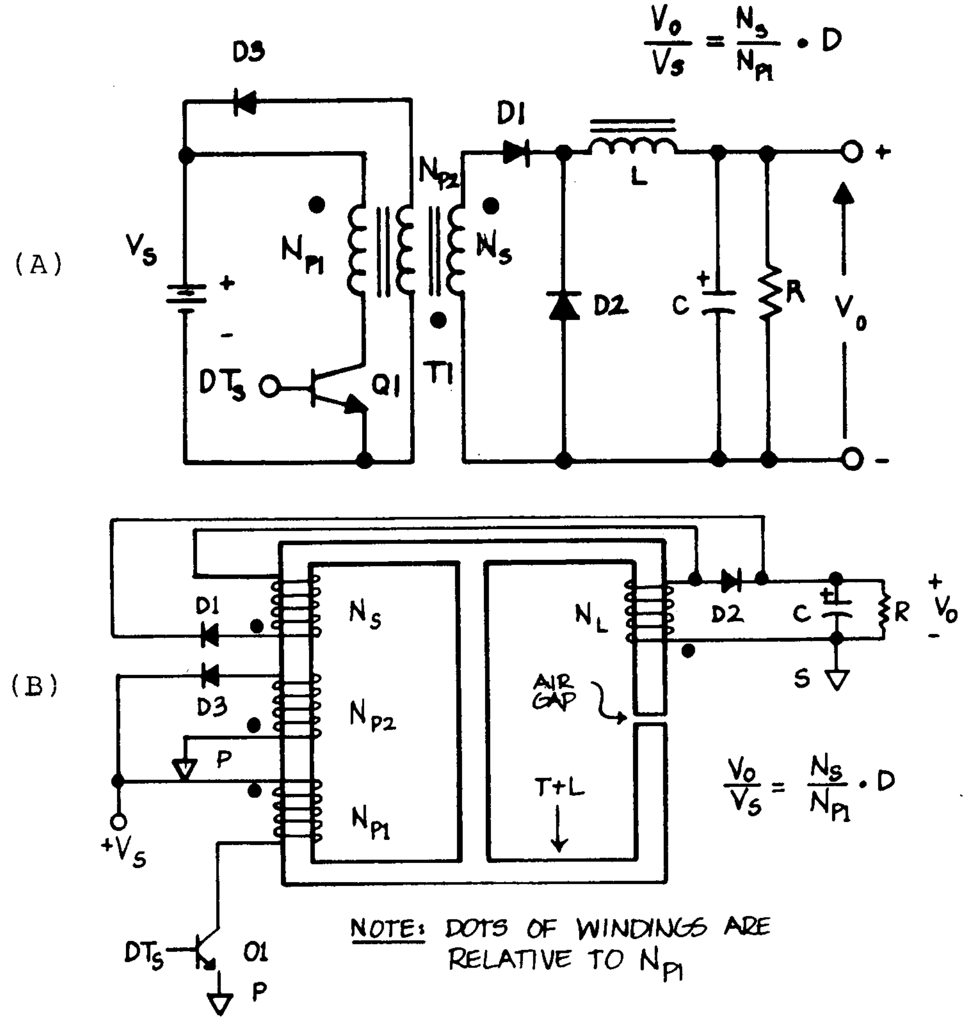


Fig. 13. A forward converter (A) and a two-bobbin integrated-magnetic version (B) where $N_L = N_S$.

4.2 Switchmode Resonant IM Converters

Recalling our earlier discussions of the basic requirements needed to place inductive elements on a single core structure, it follows that the techniques of coupled-inductor design can be readily applied to corresponding magnetic elements of *quasi-resonant* and *fully resonant* DC power converter circuits. Also, because of the general nature of the deductive methods outlined above for development of complete IM approaches for conventional "quasi-squarewave" converter circuits, it should be apparent that paths of similar thought can be used for finding IM versions of resonant converter topologies. This line of reasoning is further reinforced by recent research showing definite correlations between basic "quasi-squarewave" converter circuits and quasi-resonant designs that could be derived from their topologies [26].

A recent example of IM methods being applied to resonant power converter circuits is the design shown in Fig. 14 [27]. This converter, using zero-current quasi-resonant *(ZCQR)* power switching techniques, was derived from the transformer-isolated version of the basic 'Cuk converter circuit [1]. Although the switching method has been altered to achieve turnon or turnoff of the main switch element Q of the converter under ZC conditions, the proportionality of dynamic voltages across each of the three filter inductances (L_1, L_2' and L) is still retained in the quasi-resonant circuit version shown in Fig. 14. Therefore, magnetic coupling of all filter inductances becomes possible, following the basic rules described earlier for this category of IM.

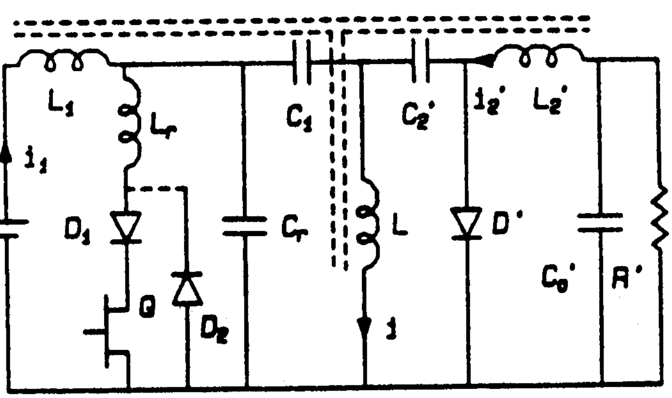


Fig. 14. Integrated-magnetics Zero Current Quasi-Resonant (IM-ZCQR) 'Cuk topology (same as Fig. 3 in ref. [27]).

This ZCQR 'Cuk converter is a good example of the use of IM methods which, when implemented using multi-path single-core arrangements, permits control and reduction of input and output ripple current magnitudes. In addition, the circuit is intended to be operated at a very high switching frequency, which implies the physical size of the IM assembly can be made to be small. Prototype designs, using this converter at a 50W output power level and a switching frequency ranging from 420 to 850KHz, have been reported [27] to achieve power densities of 27 watts per cubic inch of packaging space (excluding heat-sinking), using conventional packaging techniques and discrete electronic components.

Another example of a unique resonant power converter using IM technology is shown in the patent diagram of Fig. 15 [28]. In this design, a multi-path core is used to provide the desired relationships between transformer and inductive elements. Windings *P* and *S* are the primary and secondary of the transformer portion of the IM. The remaining winding *T*, wound on the air-gapped leg of the core, and capacitor C_R form a parallel resonant network, whose resonant frequency value sets the operational frequency of the overall conversion action of the circuit.

Working in conjunction with the leakage inductance associated with the primary winding *P*, the resonant winding *T* induces a flux characteristic in the primary that roughly approximates a sinusoid which, in turn, requires the primary winding current to assume a similar characteristic whenever the power switch *12* is closed. The

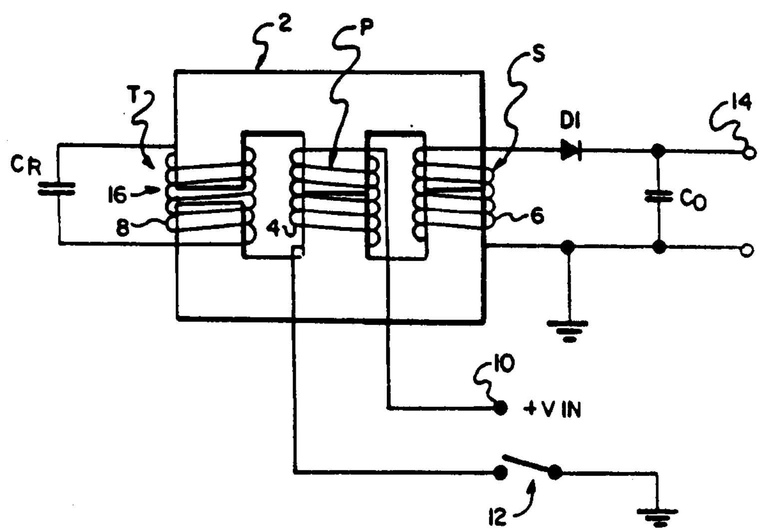


Fig. 15. One embodiment of an integrated-magnetic resonant power converter (same as Fig. 1 in ref. [28].).

design of Fig. 15 is of the ZCQR variety of resonant topologies, in that the current level in the power switch *12* will be zero at the instant of turnon or turnoff by any controlling regulation network. Diode *D1*, in conjunction with the secondary winding *S*, provides a means of developing the desired DC output voltage from the converter at point *14* whenever the primary winding *P* is active. Capacitor C_O is used to reduce dynamic ripple voltage on the output terminal to a desired level and, as such, provides energy for any load tied to the converter whenever diode *D1* is in a non-conducting state.

As far as reported performances of the IM design in Fig. 15, related patent documents [28] describe a specific application where a nominal 100 watt power delivery at a 5VDC output potential was achieved by a related circuit embodiment. Here, an operational tank resonant frequency of about 30KHz was set by design. Input voltage value was typical of that derived from off-line rectification methods, or approximately 320VDC.

Finally, an advantage stated for the resonant IM circuit of Fig. 15 is the possibility of pulse-width-modulation (PWM) control of switch action, since the operational frequency is fixed to a great degree by the resonant tank system formed by capacitor C_R and the inductance of winding *T*. Such PWM control of this resonant design then removes typical conducted noise problems that would be introduced by frequency modulation control approaches.

5.0 CONCLUSIONS & REMARKS

Integrated-magnetic circuit approaches, such as those presented herein, offer practical and viable solutions to reductions in power converter size, weight and cost. The techniques are *universal* with regard to application, as they may be used to combine and blend the many inductive and transformer functions of *any* converter topology into a magnetic structure where only a *single* core will be needed.

In this paper, a broad overview of IM concepts, principles and converter circuit applications has been presented. While the benefits of IM are just recently beginning to be appreciated by power supply engineers, the methods have roots in filter designs extending in time back to the first quarter of this century. Once thought to be applicable to only special converter circuit arrangements, recent research has shown the principles of IM can be used in *all* converters to great advantage.

Research in this unique and fascinating field of magnetics design is continuing, with each advance bringing us one step closer to achieving optimum power processing circuits with a magnetic content of minimum dimensions, cost and number.

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