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# ON THE FUNDAMENTAL PERFORMANCE SIMILARITIES of FLYBACK AND FORWARD CONVERTERS at HIGH FREQUENCIES

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## ABSTRACT

The flyback converter has been widely held to be a low performance topology, suitable only for low power levels where minimal parts count was desirable. This bias has some justification at lower frequencies, where magnetic flux density is limited by core saturation, but at high frequencies flux density becomes limited by power loss, and the potential efficiency and power density is similar for the two topologies under the same circuit conditions. It will be shown that flyback converters are a potential high performance alternative to forward converters in the frequency range of 100 KHz to 1 MHz and kilowatt power levels.

## GENERATING SIMILAR FORWARD AND FLYBACK CONVERTERS

To establish the basic similarities and differences between forward and flyback converters, <u>we must maximize the similarities</u> in order to find the residual, irreducible differences.

The difference in control loop design and performance between buck and boost topologies has been well documented elsewhere; this paper will deal with the performance of the power circuits. Specifically, we will compare relative losses and stresses in magnetics and other components, from which the relative potential efficiencies and power densities of the two circuits can be deduced. This comparison will be made at high conversion frequencies, which in this context means that flux density in the power transformer core is limited by power loss and not by magnetic saturation.

Making the most meaningful and informative comparison requires a careful choice of those circuit parameters which are to be equivalent, as well as a choice of lower order effects to ignore or defer to later analysis.

We will start with simple circuit models, and include higher order effects by stages. The basic power circuit topologies for forward and flyback converters are shown in Figures 1a & 2a. Both are half wave converters, since there is no flyback equivalent of the full wave forward converter with a single transformer core. Transformer reset circuits, voltage clamps and any snubbers are not shown and are ignored at this time.

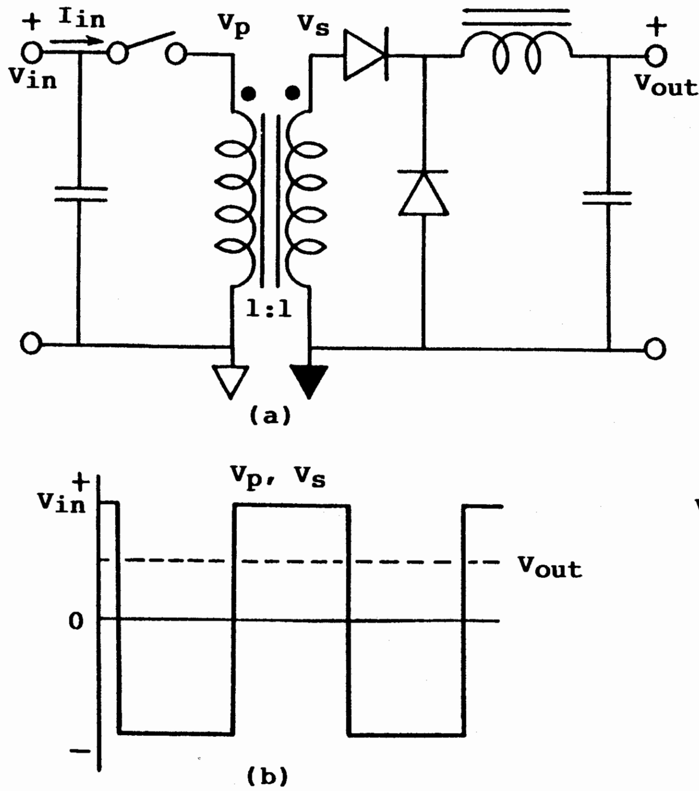



**Figure 1**
**Forward Converter**

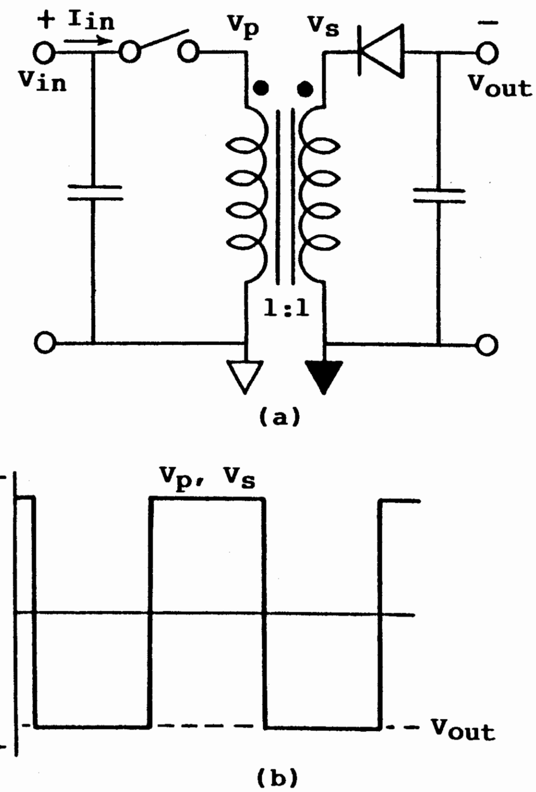



**Figure 2**
**Flyback Converter**

The following assumptions are made for the initial analysis:

1) Both converters have the same fixed input voltage = Vin;
2) Both converters have the same output power = Po;
3) Both converters operate at the same frequency;
4) Both converters operate with a switch duty cycle D = 0.5;
5) Both transformers use the same core (neglecting any air gaps);
6) The transformers have identical windings with a 1:1 turns ratio;
7) The transformers have a single primary and secondary (one magnetic section);
8) The forward converter choke uses the same core as the transformer;
9) The forward converter choke winding is identical to the transformer secondary winding;
10) The inductance of the forward converter choke is made equal to one half that of the flyback transformer (twice the effective air gap);
11) The inductance of the forward choke and flyback transformer are relatively large such that ripple current can be initially ignored;
12) Transformer leakage inductance is zero.

The full reasons for some of these assumptions will become evident later.

There are a number of basic similarities and differences that can be deduced from an initial observation of the circuit operation, without going into a rigorous derivation.

**COMPARISON OF CONVERTER OPERATION**

The primary switch in each converter is open 50% of the time, during which the switch voltage "Vsw" will be twice the input voltage to balance the volt-seconds on the transformer primaries (Figures 1b, 2b). Neglecting losses, the input currents "Iin" and switch currents "Isw" must be equal for the same output power, such that Iin x Vin = Po and Isw = 2 Iin, so the current and voltage stresses on the switches are seen to be similar. The input capacitor AC currents are equal to the switch currents, and thus also equivalent.

Since the same AC voltage waveform is applied to the primary of both converters, the AC flux density and hysteresis losses are the same in each transformer (the presence of DC flux does not affect core loss as long as the peak flux is not near saturation). The transformer primary currents are also equal, so it can be seen that all primary side circuit conditions are identical for both converters under the conditions specified.

The transformer secondary voltage (and current) waveforms must also be the same for both converter types. The transformer reverse voltage is peak rectified in the flyback converter, which at a 50% duty cycle is equal to Vin, and with the 1:1 turns ratio the output voltage is equal to the input voltage. The diode current is equal to the switch current, and the output current is 1/2 Isw for 1-D = 0.5.

In the forward converter, the transformer forward voltage is averaged by the choke to 1/2 Vin, but the current is now equal to Isw. Thus the output of the forward converter is at 1/2 the voltage and twice the current of the flyback converter, or can be considered to have 1/4 the characteristic load impedance. The output polarity is also inverted with respect to the sense of the secondary winding, but this is of little concern in isolated converters.

The single flyback output rectifier has a reverse voltage of 2 Vin, while each of the two rectifiers in the forward converter has a reverse voltage of Vin, and all have a forward current of Isw when conducting. Thus the one flyback diode has a voltage stress equal to the sum of the voltages on the two forward diodes, and the rectifier requirements are similar.

This similarity is more evident if we temporarily change the transformer ratios to produce the same output voltage and current; the flyback rectifier now has the same reverse voltage as each of those in the forward converter, but must handle twice the current. In effect, the two diodes in a forward converter must be placed in parallel when used to produce the same output from a flyback converter in order to keep diode current density constant.

The stress on the output capacitors are not equivalent; the flyback output capacitor has an RMS ripple current equal to the output current, while that of the forward converter has virtually no ripple current due to the large filter choke (assumption 11).

## MAGNETICS WINDING LOSSES

The comparison of the two converters now becomes a little more involved. At high frequencies, skin and proximity effects become significant or dominant conductor loss mechanisms in switchmode magnetics [1][2][3][4]. It is thought by some that in flyback transformers these losses are inherently greater than in forward converters, and further that these eddy current losses cannot be reduced by interleaving primary and secondary windings, as they can in a forward converter transformer. Both of these assumptions will be shown to have little basis in fact.

First, it must be understood that the AC (or eddy current) losses in conductors, both skin and proximity effects, are determined solely by the local time varying magnetomotive force "H" (or the flux density "B", depending on viewpoint).

To find these fields we need to generate a magnetomotive force (MMF) diagram for the windings involved. For ease of comparison we will idealize the H (and resultant B) field along conventional lines; current density is uniform in the conductors, and the magnetomotive force is axially uniform (with solenoidal windings) and varies only with radial position.

From a point of zero MMF, the H field builds up as current carrying windings are crossed, remains constant across interwinding gaps, and drops off as windings with opposite current sense are transversed. The local induced eddy current losses increase with frequency and the square of the field intensity. The actual deviation from the ideal flux distribution will be similar in both transformers, due to the identical windings used, so the approximation is a valid base for loss comparisons.

In Figure 3a the transformer winding currents are shown for the forward converter (current into the winding is positive). When the switch is closed, current Isw flows into the primary winding, and simultaneously the same current flows out of the secondary (neglecting magnetizing currents). At this time the distribution of H in the winding is shown by the solid curve in Figure 3b. Both currents fall to zero when the switch opens, and so does H throughout the windings (dashed curve). The AC component of H is shown in the lower curve.

When the switch is closed in the flyback converter, current Isw flows into the primary, but no current flows in the secondary until the switch opens; at that time the primary current ceases and a current Isw flows into the secondary (Figure 4a). The distribution of H in the windings is shown in Figure 4b with the switch closed (solid) and open (dashed). The value of the current (and hence H) does not change while the switch remains open or closed due to assumption 11. Although the two curves appear different, it can be seen that the distribution and magnitude of the changing or AC component of H, designated $\Delta H$, is identical in the windings of the flyback and forward converter transformers, and thus the induced eddy current losses are equal.

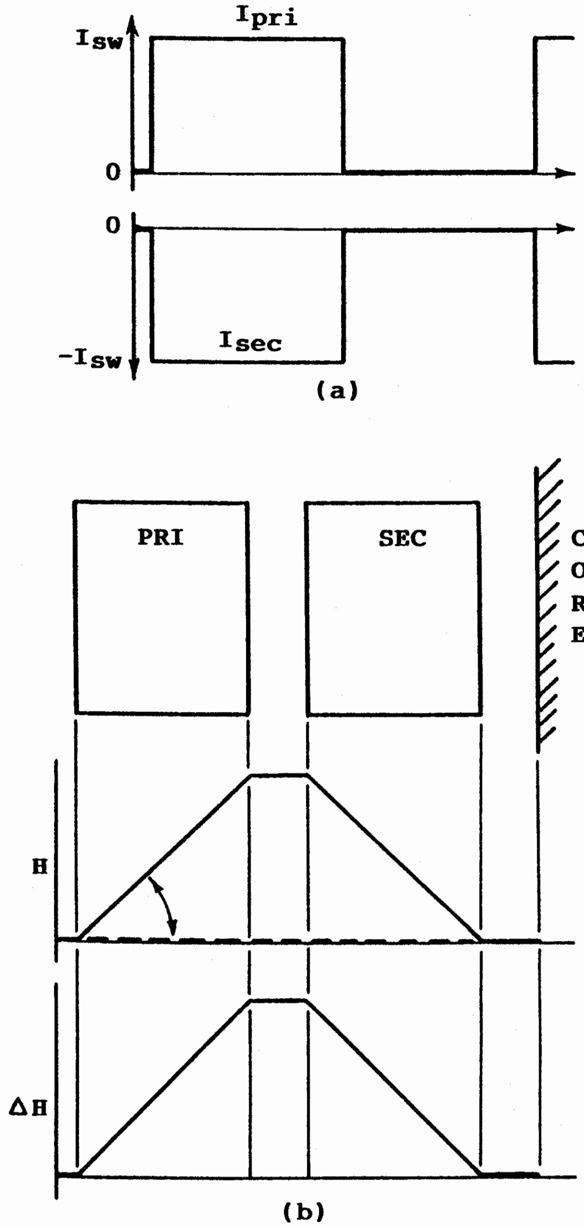



**Figure 3**
**Forward Transformer MMF**

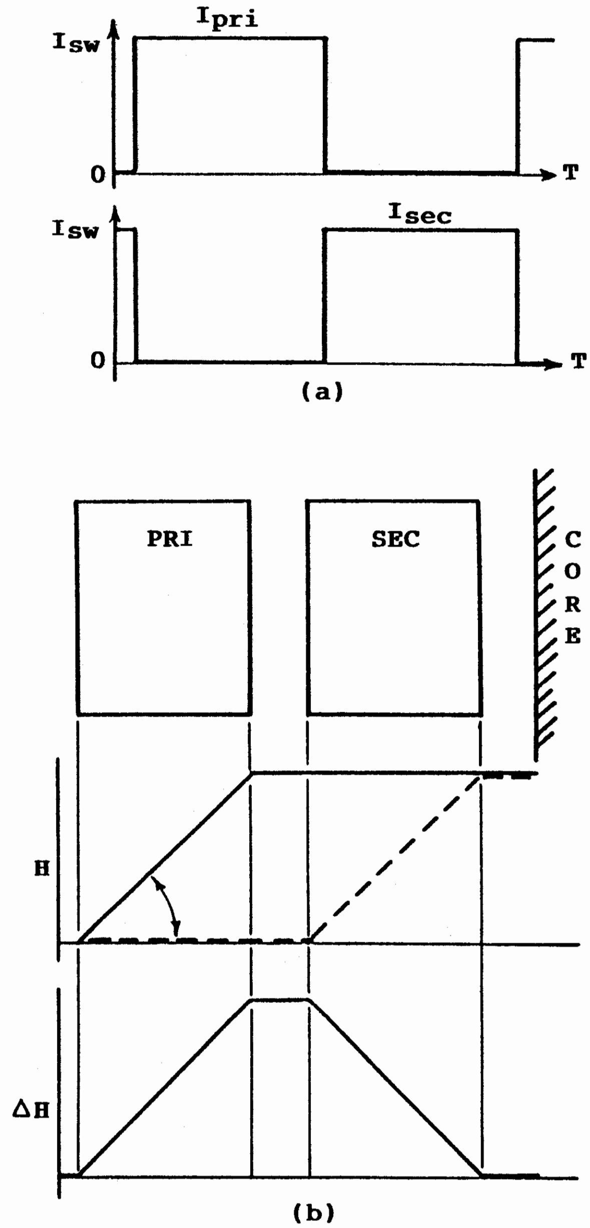



**Figure 4**
**Flyback Transformer MMF**

Further inspection reveals that the flyback secondary current is actually equal to that of the forward converter with an additional DC bias equal to Isw (Figure 5). This DC component generates the additional flux seen in the secondary (Figure 4b), and produces a DC flux in the transformer core (which usually requires an air gap to avoid saturation).

The current in the forward converter choke is also equal to Isw, so the DC MMF in the choke winding is identical to that in the flyback secondary (due to assumptions 8 & 9), but the flux density B in the choke core is only half as much (assumption 10). It begins to appear that the flyback transformer is somewhat equivalent to the magnetic integration of the forward converter transformer and choke into a single structure. We will soon see that this equivalence is more than superficial.

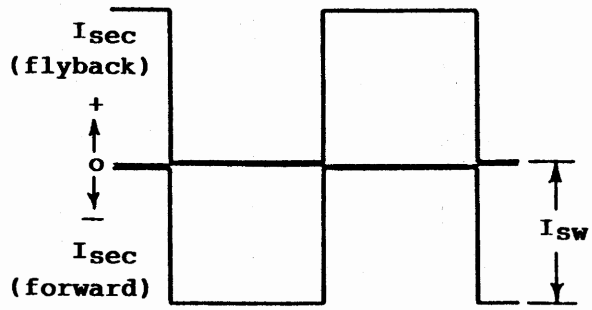



**Figure 5**
**Transformer Currents**

The RMS value of the flyback secondary current is unchanged by the additional DC current; since the RMS currents in both windings of the two transformers are equal, as is the distribution of the dynamic H field, the copper losses in both transformers must be equal. We have already noted that the core losses are equal, so we conclude that all transformer losses will be identical in both converters under the assumed conditions.

There is an additional magnetics loss in the forward converter, however, due to the DC current in the filter choke. This choke also adds volume and mass to the forward converter, so we can conclude that the flyback converter has a potential advantage in size, weight and efficiency over the forward converter, which deserves further exploration.

## EFFECTS OF INTERLEAVING WINDINGS

Next we will change assumption 7 and explore the effects of interleaving the primary and secondary windings in both transformers, while keeping all other conditions the same. This has been done graphically in Figures 6 and 7, for 2 and 3 magnetic sections respectively (labeled MS1, etc). The H field has been plotted with the switch closed (solid line) and open (dashed line) for the forward and flyback transformers.

Inspection reveals that the change in H field for the flyback transformer (Figure 6c, 7c) is algebraically equal to that of the forward transformer (Figure 6b, 7b); thus the induced loss is identically reduced in both types of transformers for the same number of magnetic sections. As before, the additional DC field in the flyback transformer windings and core does not result in any loss.

However, this conclusion is dependent on assumption 11, that the flyback transformer and forward choke inductances are large and ripple currents are minimal.

## EFFECT OF FINITE RIPPLE CURRENT

In practical power converters ripple current is not always of low amplitude relative to full load DC current. In fact there are several valid reasons for minimizing flyback transformer or forward choke inductance and having a large peak-to-peak ripple current:

1) The energy stored in the forward choke or flyback transformer is reduced, allowing a smaller core to be used.

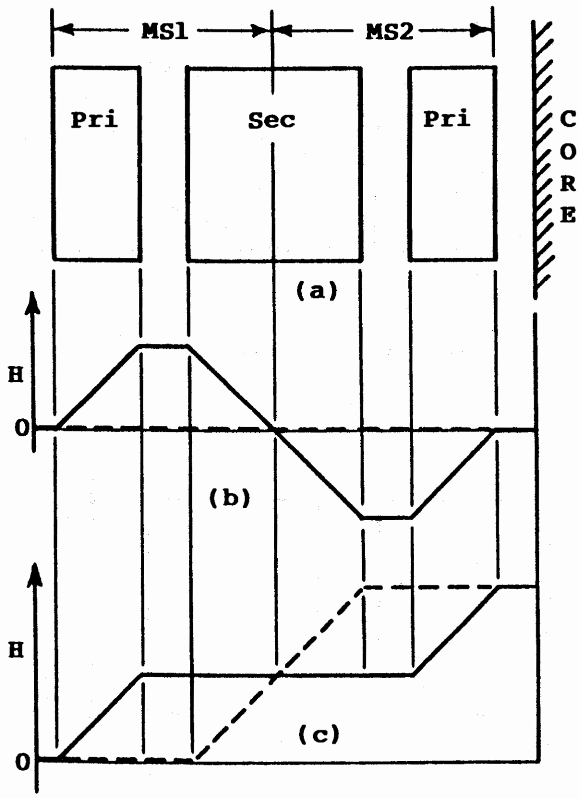



**Figure 6**
**(a) 2 Section Winding**
**(b) Forward MMF**
**(c) Flyback MMF**

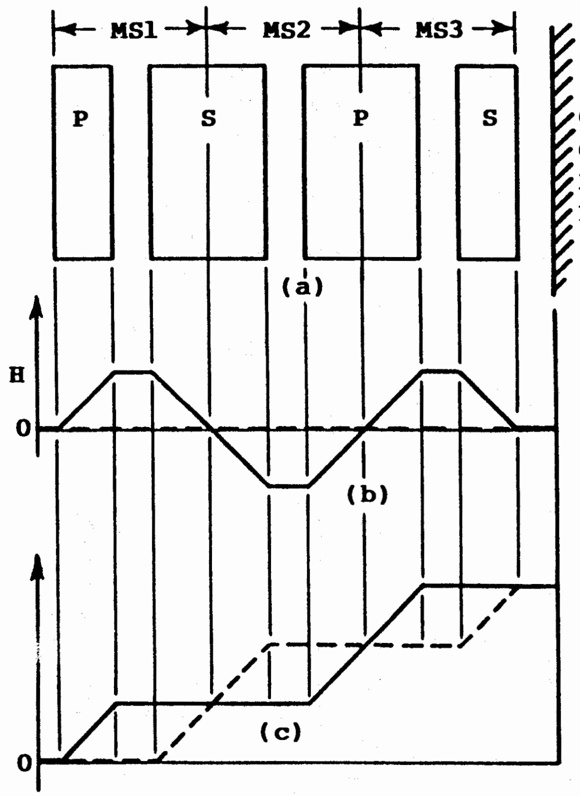



**Figure 7**
**(a) 3 Section Winding**
**(b) Forward MMF**
**(c) Flyback MMF**

2) The transient response to load changes is improved, due to higher current slew rates (this is particularly important when plastic or ceramic capacitors with low stored energy are used in high frequency converters)

3) In flyback converters, operation near or below critical current reduces or eliminates the effect of the right hand plane zero in the control loop.

*(The high ripple increases RMS current and conduction losses in switches, rectifiers, filter capacitors and magnetics windings to some degree. These high ripple currents may require the use of lower resistance semiconductors, modified winding designs, and/or multi-stage filters, and other compromises familiar to the experienced designer.)*

We will now modify assumption 11 to see the effect of a significant ripple current. We will investigate the relatively extreme case of operating at critical current, where the peak-to-peak ripple current is twice the original switch current Isw; the average conduction current is still Isw. (With sophisticated designs, it is practical to operate either flyback <u>or</u> forward converters under these conditions.) The effects of operating with less ripple current will be an intermediate case between these extremes.

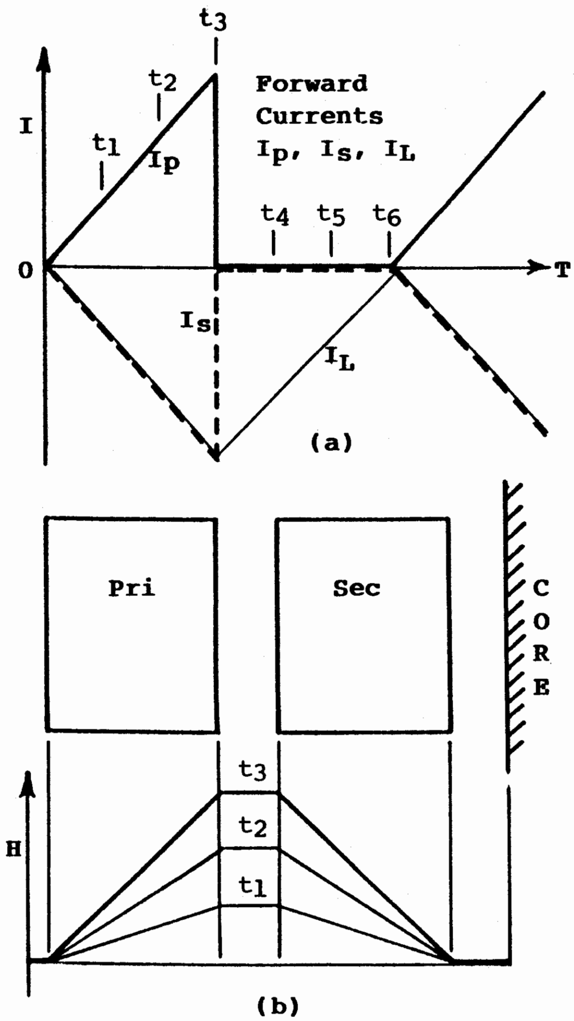



**Figure 8**

**Forward Converter**
**MMF at Critical Current**

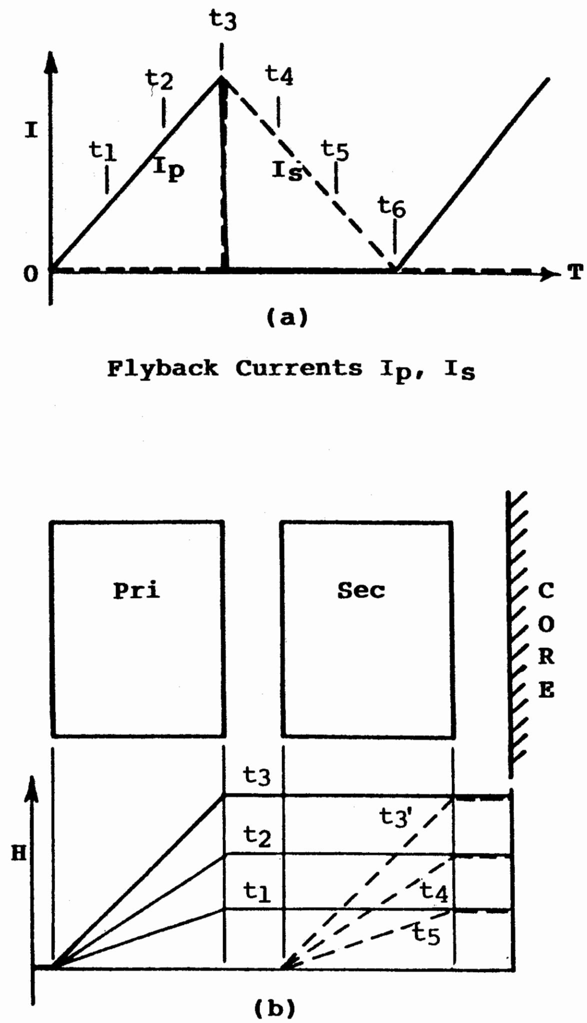
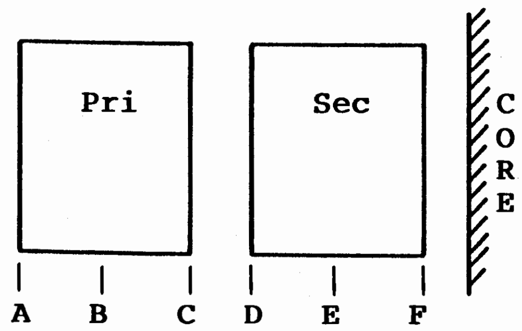



**Figure 9**

**Flyback Converter**
**MMF at Critical Current**

The currents in the primary, secondary and choke of the forward converter are shown in Figure 8a, while the resultant transformer H field is depicted in Figure 8b after time intervals t1, t2 & t3 from switch closure. Instead of the previous step change in H at turn-on, the field builds up continuously until it reaches twice the previous level (with no ripple current), then steps to zero at switch turn-off. *(The eddy current loss has also increased, but this aspect will not be pursued in this paper.)*

The flyback primary and secondary currents are shown in Figure 9a, and the resultant H field in 9b. As the primary current builds, the MMF distribution in the primary is the same as that of the forward converter, but the field builds uniformly in the secondary instead of dropping to zero next to the core. When the switch opens, the current commutates to the secondary and ramps back to zero, with the H field distribution shown in phantom at times t3'(just after switch opening), t4 & t5.



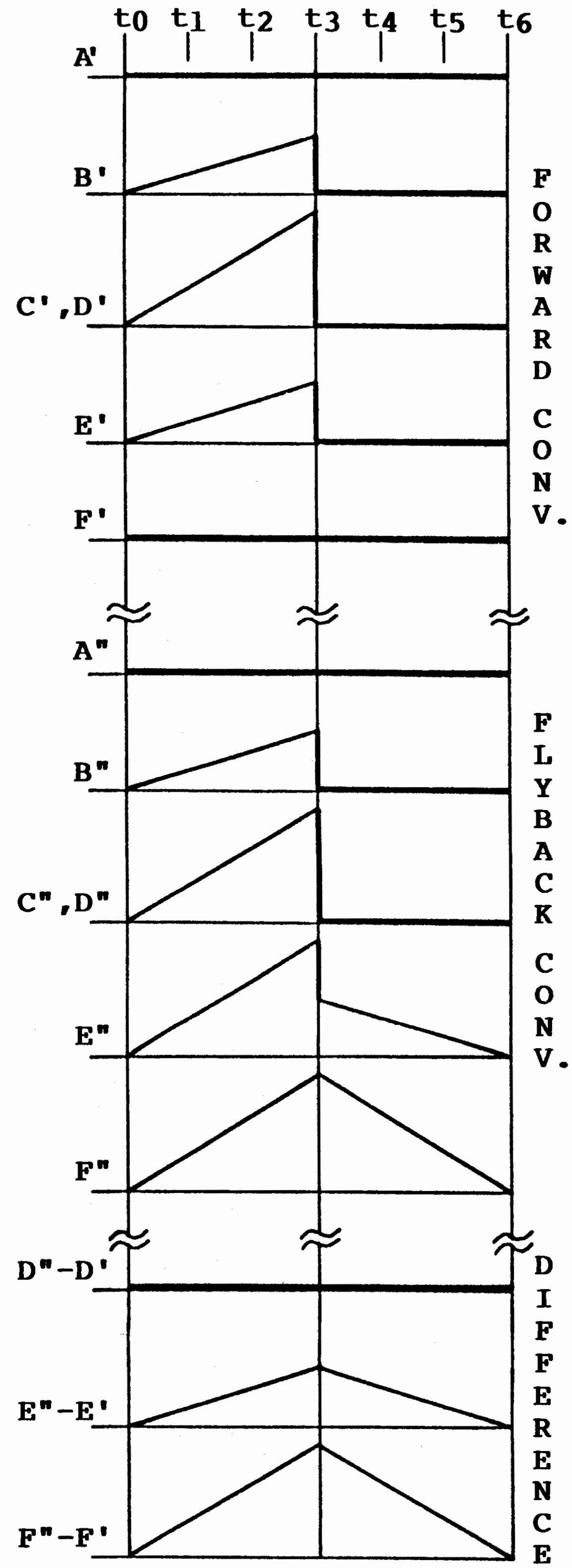

**Transformer Winding**

**Figure 10 a**

In order to help clarify the differences in these two time and space dependent fields, H has been plotted as a function of time for six positions in the windings (Figure 10a), ranging from A' at the outer surface of the forward primary to F' at the inside of the secondary, and A" to F" for the corresponding points in the flyback windings. The fields at C' & D' (and C" & D"), the facing surfaces of the primaries and secondaries, are always equal due to the absence of current between the windings.

The field at A' and A" is always zero, since there is ideally no flux outside a solenoidal winding. *(This is true for an ideal solenoid of infinite length, and is approximately true for a cored transformer where the core acts as a low reluctance return path for the flux between the winding and the core.)*

At positions B' & B", C' & C" in the primaries the time dependent fields remain the same in both transformers, as with no ripple, and also at the outer surface of the secondaries D' & D". The difference between the forward and flyback transformers occurs inside the secondary windings, as seen by the fields at E' vs E" (middle of the secondaries) and F' vs F" (inside surface of the secondaries). The difference in flux (between the two secondaries) vs time has been plotted in the lower three graphs of Figure 10b, where we see a




**MMF vs TIME**

**Figure 10b**

residual field in the flyback transformer secondary that builds when the switch is closed and ramps to zero when the switch is open. The relative intensity increases linearly from zero at D" to a maximum at F".

Although the secondary RMS currents are the same in the two transformers, this residual field causes an increase in eddy current loss in the flyback secondary winding which does not occur in the primary, nor in the secondary of the forward transformer. *(If the primary and secondary were interchanged, the excess loss would remain in the inside winding, now the primary. This asymmetry was noted by Jongsma [3], but he located the extra loss in the outer winding, but neither this conclusion nor the MMF diagrams on which it is based agree with my analysis.)*

It would seem that the total loss in a flyback transformer has indeed been shown to be greater than in forward transformer under practical operating conditions. Strictly speaking, this is true, but if we analyze the H field in the forward converter choke winding, we will find that it is identical in time and distribution to the difference between the fields in the flyback and forward transformer windings.

Thus this loss has not been avoided in the forward converter secondary, but has been transferred to the winding of the filter choke. In addition, the forward converter choke now also exhibits hysteresis core loss due to the AC flux. This flux in the choke core is one half that in the transformer (due to assumption 10), so the core loss is 1/4 or less of that in the transformer, depending on the value of hysteresis loss exponent. Neither this core loss nor the original DC winding loss exists in the flyback converter, so the total magnetics loss is still less in the flyback topology.

**INTERLEAVING WINDINGS WITH RIPPLE CURRENT**

If the forward and flyback transformer windings are again interleaved with multiple magnetic sections and the H fields analyzed, it will be found that there is a component of the dynamic H field in the flyback transformer that is not reduced as it is in the forward transformer. This remaining "choke" field component induces an additional eddy current loss in the flyback windings, which does not exist in the forward transformer.

However, (with assumptions 8, 9 & 10) the AC and DC components of this "choke" field will be found in the winding of the forward converter main filter inductor, and thus the excess winding loss in the flyback transformer occurs in the choke of the forward converter. The choke retains its additional DC winding and core hysteresis losses, so the flyback transformer can be said to integrate the forward converter transformer and choke into a single structure, with some savings in size, weight, core and copper loss.

In any case, this investigation has shown that the high frequency conductor losses in a flyback transformer can indeed be reduced by interleaving primary and secondary windings. This conclusion has been empirically verified by J.P. Vandelac [4] who also, in several conversations, has been of great assistance in clarifying the approach and principles used in this study.

## VALIDITY OF COMPARISON

It may be questioned whether it is valid or meaningful to compare forward and flyback converters under such identical conditions. Forward converters are usually operated well above critical current to minimize ripple, while flyback converters often operate at or below critical current to simplify control, since ripple current in the output capacitor is high in any case. Besides, flyback transformers usually have more turns than forward converter transformers anyway. More to the point, might not the "optimum" design of forward and flyback converters result in quite different designs?

The traditional differences are not really relevant to this study. At lower conversion frequencies flyback transformers may require more turns than full wave forward converter transformers to reduce flux density and avoid core saturation. The use of a high choke inductance in a forward converter to reduce ripple often results in a larger choke than transformer; a more compact and efficient design would use a much smaller choke and a multi-stage filter, resulting in the same or better performance.

As to the question of different design optima for forward and flyback converters, a complete and final answer cannot be given since there are so many different performance criteria which may have to be balanced and traded off in any given application.

However, if achieving a high power density is a significant or dominant consideration, then comparing similar hardware and operating conditions is valid. High power densities imply high operating frequencies [5], and size reductions can be limited by generation of waste heat as well as the volume of the components.

We have not had to consider any of the transformer or choke design details (beyond those in the initial assumptions), or indeed even perform any quantitative calculations, in order to establish the relative performance of forward and flyback magnetics. Thus (at high frequencies) the relative losses must "track" as design parameters change; <u>any optimization of the magnetics for one topology will be applicable to the other.</u>

The flyback topology generally has the edge in total size, weight and loss in power magnetics, but other performance differences and applications criteria may swing the balance either way.

## OTHER FORWARD vs FLYBACK COMPARISONS

### <u>Leakage Inductance Effects</u>

Power transformer leakage inductance can be a problem in both converters. In a forward converter the leakage inductance delays current buildup in the secondary, and can cause problems with slow output rectifiers and voltage spikes on the primary switch. These problems can be overcome with fast rectifiers and voltage clamps, and some forward topologies can return stored leakage inductance energy to the input while virtually eliminating voltage overshoot on the primary switch.

The problem is worse in flyback converters, since some excess primary switch voltage is necessary for the current to commutate from the primary to the secondary. If the voltage is clamped by dissipative means, the power loss can become quite significant as the voltage overshoot is reduced.

If the additional voltage overshoot is limited to a value equal to the theoretical flyback voltage, then the energy loss per switching cycle is twice the leakage inductance energy. An overshoot of 50% of flyback voltage triples the loss, and a 20% overshoot increases the loss to six times the leakage inductance energy.

If efficiency is to be maintained in flyback converters at high frequencies, then either the leakage inductance must be minimized and a significant voltage overshoot allowed on the primary switch, or a non-dissipative voltage clamp must be used which returns the clamp energy to the converter input or output.

## Wide Input Voltage Range

In a conventional forward converter the peak voltage on the primary switch and output rectifiers is proportional to the input voltage. Thus, if the converter is to operate over a 4:1 input voltage range, the switch and diodes must be rated for at least four times the voltage required for a fixed input voltage.

In a flyback converter the peak switch and diode voltages are proportional to the sum of the input and output voltages, which reduces the peak voltage stress when operating over a wide input range. If the switch duty cycle is 50% at the minimum input, then with four times the input voltage the switch and diode will see a peak voltage only 2.5 times higher. This can allow the use of lower resistance FETs or faster rectifiers (or Schottkys instead of epitaxial rectifiers) when a broad input voltage range is required.

The change in switch duty cycle D with input voltage is not the same in forward and flyback converters, which changes the relative RMS and average currents in most power components somewhat. Although the above conclusions based on $D = 0.5$ remain approximately valid, a detailed analysis and comparison may be called for in a specific application.

## Output Filtering

The large ripple current in the output filter capacitor of a flyback converter requires a capacitor with a high RMS current capability. Additional filter stages are usually required to attenuate the output ripple voltage to acceptable levels.

## "One Transformer Turn" Limitations

As a given transformer core is operated at higher frequencies it is usually possible to reduce total transformer loss by gradually reducing the number of turns to keep an optimum balance between core and copper losses. However, when the optimum number of secondary turns reaches one, any further increase in frequency will cause a rise in total loss.

*(Note: this remains true only below a characteristic frequency, just below which the core loss begins to increase rapidly with frequency and above which the permeability decreases with frequency. Near this frequency total loss will always increase with frequency for a given core and material.)*

A flyback transformer produces about twice the output voltage as the same transformer operating in the forward mode. Thus for a given core and output voltage, the "one turn minimum" limitation occurs at a lower operating frequency for the flyback transformer. Combined with the other characteristics of flyback converters (notably output capacitor ripple current) this tends to make them unsuitable for low voltage-high current outputs at high conversion frequencies.

Effects of Core Gaps

The air gap usually required in a flyback transformer can cause significant eddy current loss in any conductors near the gap. This problem does not occur in the forward transformer without an air gap, but it will arise in the filter choke. In either topology the problem of fields generated by discrete air gaps must be addressed when operating at high frequencies with significant ripple current.

## NOTE ON ANALYSIS TECHNIQUE

The analysis technique used in this paper is the qualitative comparison of different power conversion topologies with similarities in construction and operation maximized. This approach can provide some very useful insights to generic traits: for example, that simple buck and boost regulators form a converter class which is inherently more efficient, requires less energy storage, and has lower switch and rectifier volt-amps products for a given output than the class of regulators which include the classical buck-boost, non-inverting buck-boost and Cuk topologies. Furthermore, all these non-isolated regulators are in the same way more "efficient" than isolated converters as a class.

A similar analysis and performance grouping of isolated converter topologies, besides the half wave forward and flyback converters discussed in some detail in this paper, has only begun. If there is sufficient interest, the results of these studies will be published at a latter date.

## SUMMARY

A study of the magnetic and circuit losses in forward and flyback converters has demonstrated two important points:

First, the high frequency conductor losses in a flyback transformer can be reduced by interleaving primary and secondary windings much as they are in the forward converter transformer.

In many respects the flyback transformer is equivalent to the magnetic integration of the transformer and output choke of the half wave forward converter, but with some actual savings in total power magnetics size, weight and loss.

Second, it has been shown that, at high frequencies, efficiency and power density are similar in forward and flyback converters for the same circuit conditions. Power semiconductor voltage and current stresses are also similar.

In a given application the potential savings in power magnetics size, weight and loss with a flyback topology may be offset by higher ripple currents in the output capacitors and the additional filter requirements. Some design sophistication and ingenuity will also be required to overcome problems created by finite transformer leakage inductance and realize the possible advantages of flyback converters at high frequencies and power levels.

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