

**A PULSE-WIDTH-MODULATED CONTROLLED-TRANSFORMER
POST REGULATOR**

by

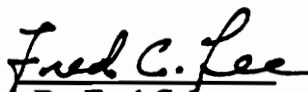
Ning Sun

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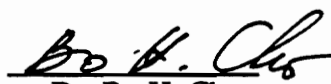
APPROVED:



Dr. Dan Y. Chen, Chairman



Dr. Fred C. Lee



Dr. Bo H. Cho

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Committee Chairman, Dr. D. Y. Chen

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(ABSTRACT)

The theory of operation of a controlled transformer is described. A PWM controlled transformer is proposed and implemented in a forward converter to provide post regulation. Experimental results are presented to verify the new control scheme. Overall efficiency of 82%-86% is achieved in a 200khz, 500-watt, 5v-output off-line regulator. A discussion of design issues of the controlled transformer is also presented.

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Chapter 1

INTRODUCTION

Post regulators are often needed in multiple-output switching power supplies. Commonly-used post regulators include linear regulators and magnetic -amplifier regulators (magamps). A linear regulator is only used in very low-output-current application because of its low efficiency. A magamp regulator is widely used because of its efficiency, simplicity, and reliability . However, A conventional magamp requires the use of a square-loop magnetic material of a high degree of squareness in order to achieve high conversion efficiency and large regulation control range [1, 2, 3]. To maintain high degree of squareness, an ungapped magnetic structure , usually a tape-wound core, is required. Tape-wound cores are generally expensive and require costly winding production method. Further, the high-frequency (>200KHz) core losses of existing tape-wound cores are usually prohibitively high. This prevents application of magamp in high-frequency power conversion.

Recently, a modified magamp circuit was reported that allows the use of soft ferrite, a material of low B-H squareness, as the saturable reactor and the circuit still maintains high conversion efficiency and large regulation control range [4, 5, 6]. The use of soft ferrite is made possible in this circuitry by using a parallel reset circuit. However, there are still several disadvantages of using this scheme for certain applications. For high-output-voltage applications, the saturable reactor must be designed to block full output voltage to achieve

full-range current limit. This fact, plus the fact that the number of turns of the saturable reactor is normally kept small to minimize power loss, usually leads to a large reactor size. For high-output-current applications, to avoid large control circuit losses of the reactor, the turns ratio between the control winding and the current-carrying winding must be large. High turns ratio leads to high voltage across the control winding in the feedback circuit. Therefore, isolation is normally needed, which defeats one of key advantages of magamps. A controlled-transformer regulator which is the focus of the thesis is developed to alleviate the above problems associated with magamps. The idea of using a controlled transformer for voltage regulation is not new [7]. In this type of regulator, relative inexpensive ferrite cores can be used. However, the power conversion efficiency reported is relatively low [8]. In the present thesis, a pulse-width-modulation (PWM) control scheme is proposed and experimentally verified. This scheme significantly increases the conversion efficiency of the circuit. Furthermore, this scheme makes the controlled-transformer regulators amenable to production standardization, which is not practical for magamps.

In the thesis, Chapter II reviews the theory of a controlled transformer. Three ways of implementing the controlled transformer are described. One of these three possible implementation is chosen later to illustrate the use of a controlled transformer for regulating a power converter output voltage. A linear control scheme of a controlled transformer is described. However, this control circuit results in a large power loss. A new pulse-width-modulated (PWM) scheme is proposed in chapter III for the control of the controlled transformer to achieve converter output voltage regulation. This control scheme leads to significant overall loss reduction of the power converter. A two-transistor forward converter is employed for the illustration of using controlled transformer for output voltage regulation.

A comparison of the controlled transformer regulators and the modified magamps is also given in this chapter. Chapter IV describes the implementation and the experimental results of a 200-khz, 500-watt 5v output regulator using the PWM controlled transformer. Design issues of the controlled transformer are also discussed. Chapter V concludes the research work and describes possible future research directions. Because of many parameters involved in the thesis, a glossary of symbols is also given in Appendix I. Appendix II gives a detailed calculation of power loss distribution in every component. In Appendix III, DC control characteristics of PWM-controlled transformer is analyzed and the compensator design examples are given too. Appendix IV has a list of a detailed calculation of control current of PWM-controlled transformer.

Chapter 2

CONTROLLED TRANSFORMER

In this chapter, review of the basic operating principle of a controlled transformer will be given in section 2.1. Several ways of implementation will be described. In section 2.2, an application of the controlled transformer to a power converter will then be described. A forward converter configuration regulator using a controlled transformer is used for illustration. In section 2.3, a comparison is made between a magamp regulator and a controlled transformer regulator.

2.1 Operating Principle of Controlled Transformer

Fig. 2.1 depicts the concept of a controlled transformer. The main power is transferred from the primary winding to the secondary winding. The control winding is used to control the amount of power being transferred. The voltage waveforms of the primary and the secondary winding illustrate the basic regulating principle of a controlled transformer. There are several ways to implement this basic concept. They are discussed in the following two subsections.

2.1.1 Controlled Transformer Using Single Core

Controlled-Transformer

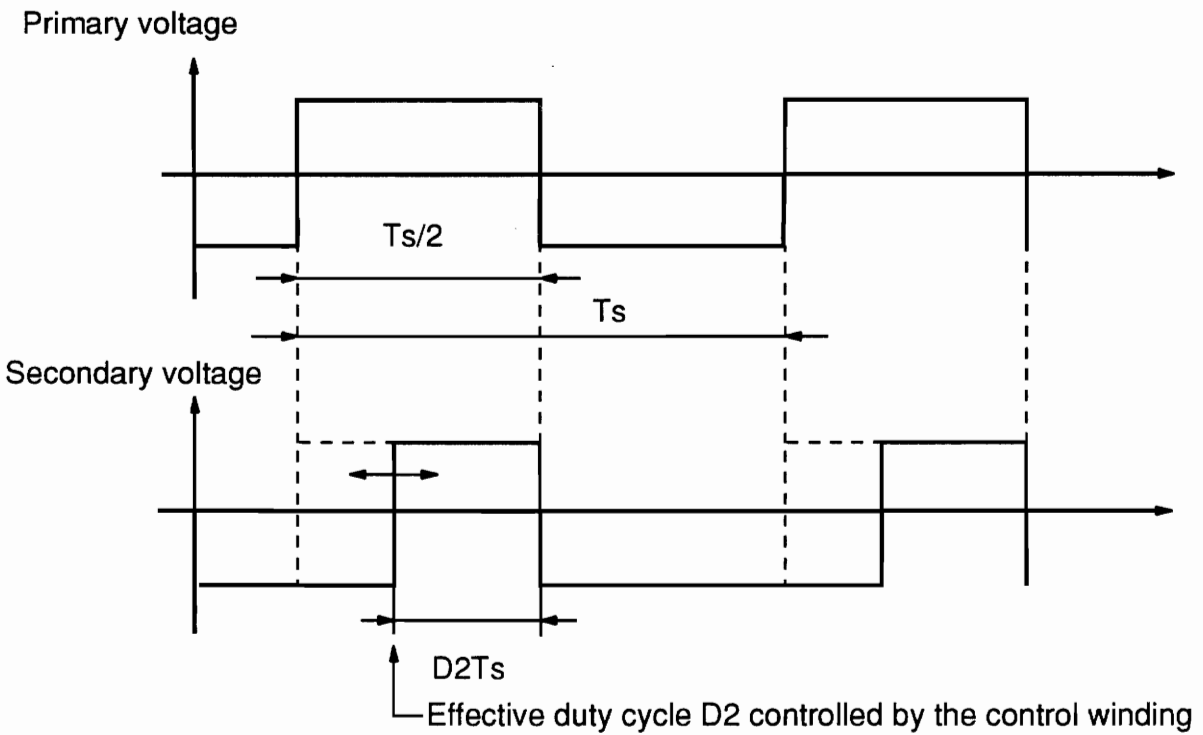
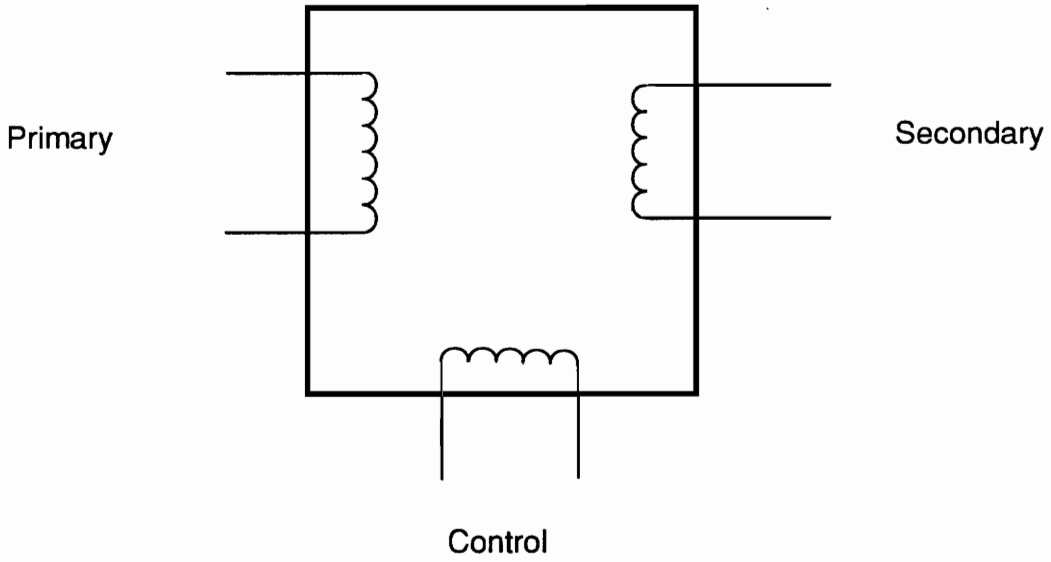


Fig. 2.1 Conceptual controlled transformer and voltage waveforms

Fig. 2.2 shows the diagram of an E-core controlled transformer. The primary winding and the secondary winding are wound on the outer legs and the control winding is wound on the central leg. The magnetic flux generated by the primary voltage is partially coupled to the secondary and partially diverted through the central leg. By varying the magnetic reluctance of the central leg, the coupling between the primary and the secondary can be controlled and the secondary power can be regulated. The effective magnetic reluctance of the central leg can be electrically controlled by varying the resistor R_c . When R_c is small, a large current is induced in the control winding and sets up a field opposing the field generated by the primary current. This effectively increases the magnetic reluctance of the central leg and forces more flux coupling between the primary and the secondary windings and therefore more power is transferred from the primary to the secondary. Under the theoretical case of $R_c=0$, the voltage on the control winding is clamped at zero volt and there is no flux flowing through the central leg and all the flux of the primary is coupled to the secondary, similar to a conventional transformer. As the resistance R_c increase, more flux diverts through the central leg and less flux enters the secondary leg due to the decreased reluctance of the central leg. When R_c goes to infinity (open circuit), then a relatively large percentage of flux generated by the primary winding flows through the central leg and small portion of total flux enters secondary flux path since the central leg has shorter flux path and lower reluctance. Therefore, little power is transferred from the primary to the secondary and output voltage is reduced. Under this condition, the flux division between the two paths depends on the core geometry and the secondary load.

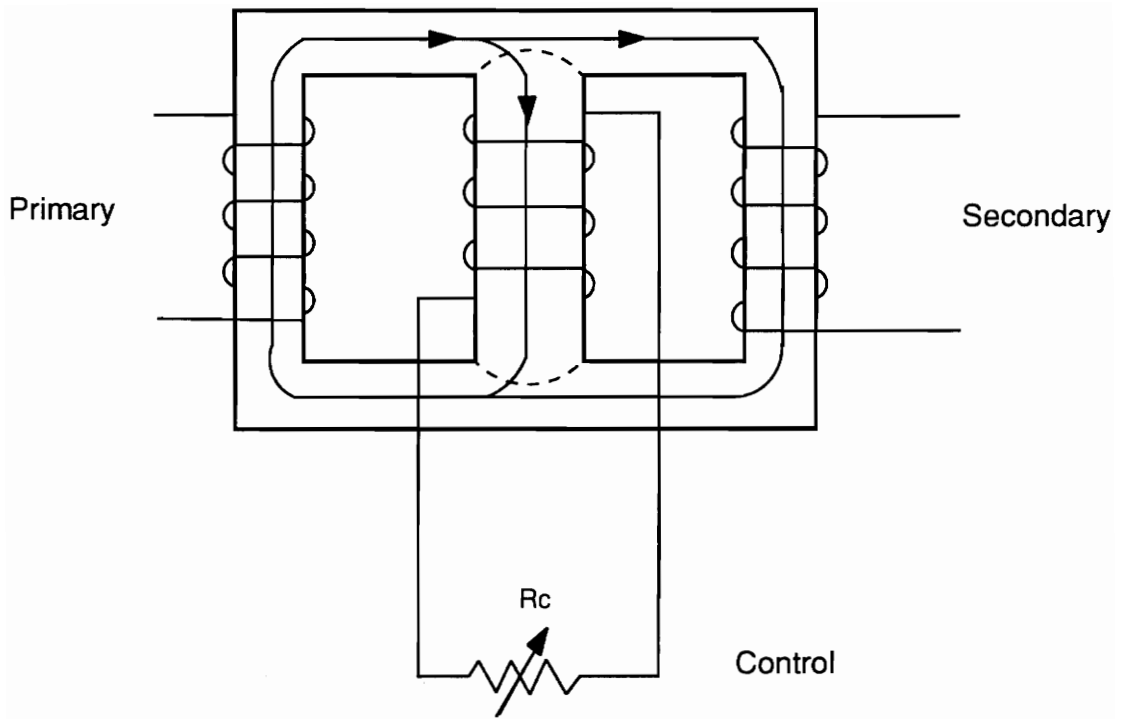


Fig. 2.2. Controlled transformer using single core.

While this implementation works in principle, it is limited practically by the relatively poor coupling between the primary and the secondary when a power transfer is needed. The reason is that the primary winding and the secondary winding are widely separated. As a result, leakage inductance are large, which limits large output power and reduces the efficiency of the regulator. To provide good coupling between the primary winding and the secondary winding, two-core implementation of the controlled transformer is described in the next section.

2.1.2 Controlled Transformer Using Two Cores

There are two ways to implement the controlled transformer concept using two separate cores, depending upon the winding configuration of the primary winding. They are discussed separately in the following

Parallely-connected control transformer

Fig. 2.3 shows the winding configuration of a parallely-connected controlled transformer. One of the core is used as the power transformer T_1 core and the other core is used for the control core T_c . The word "parallely-connected" refers to the fact that the primary winding is wound on both the power-transformer core and the control core. The secondary winding is wound only around the power-transformer core and the control winding is wound only around the control core. The principle of operation of this implementation is the same as that described for the single-core controlled transformer which is described in section 2.1.1. In the present implementation, a better coupling between the primary and the secondary can

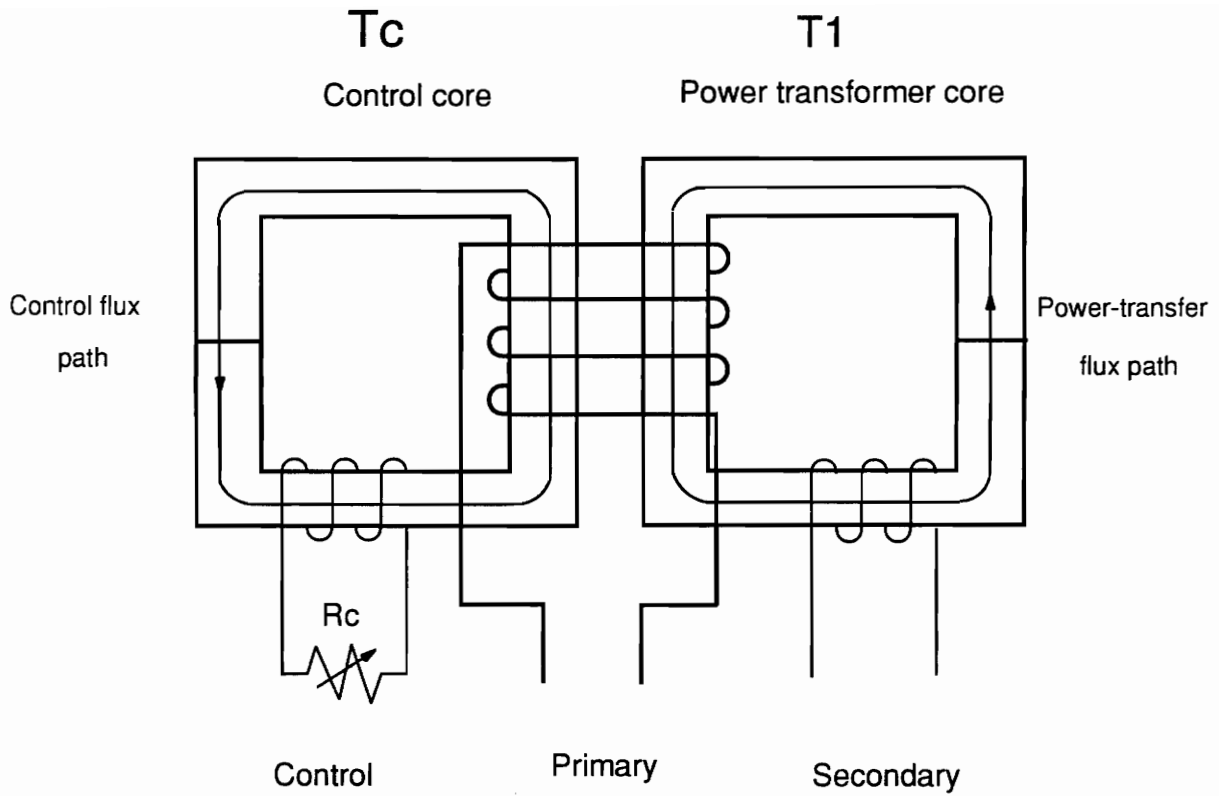


Fig. 2.3 Winding configuration of a parallelly-connected controlled transformer.

be achieved. The saturation of the control core does not impede the flux coupling between the primary and the secondary even when the control core is saturated because the power-transfer flux and the control flux are separated. Nevertheless, primary winding and secondary winding cannot be interleaved, the eddy current loss and leakage are still substantial and the power converter yields relatively low efficiency. Also, this structure is not amendable to volume production because the primary winding needs to be wound on both cores. To amend this situation, a serially-connected controlled transformer configuration is described below.

Serially-connected controlled transformer

Fig. 2.4 shows the winding configuration of a serially-connected controlled transformer, in which the primary winding is wound on each core and then connected in series. When the control winding is shorted, the primary winding voltage is applied mainly across the primary of the power transformer and power transfer occurs easily. When the control winding is open (R_c is infinite), the power transfer is essentially blocked by the large magnetizing inductance of the control core. Therefore, by controlling the resistance R_c , power transfer from the primary side to the secondary side can be controlled and the output voltage can be regulated. Similar to a parallelly-connected controlled transformer, this configuration allows the control core to saturate without reducing the coupling between the primary and the secondary winding. Furthermore, the two windings can be interleaved in this construction to achieve even better coupling than the parallelly-connected controlled transformer. This eventually leads to better power efficiency especially at large output power. For this reason, serially-connected configuration will be used in the discussion for the remainder of the

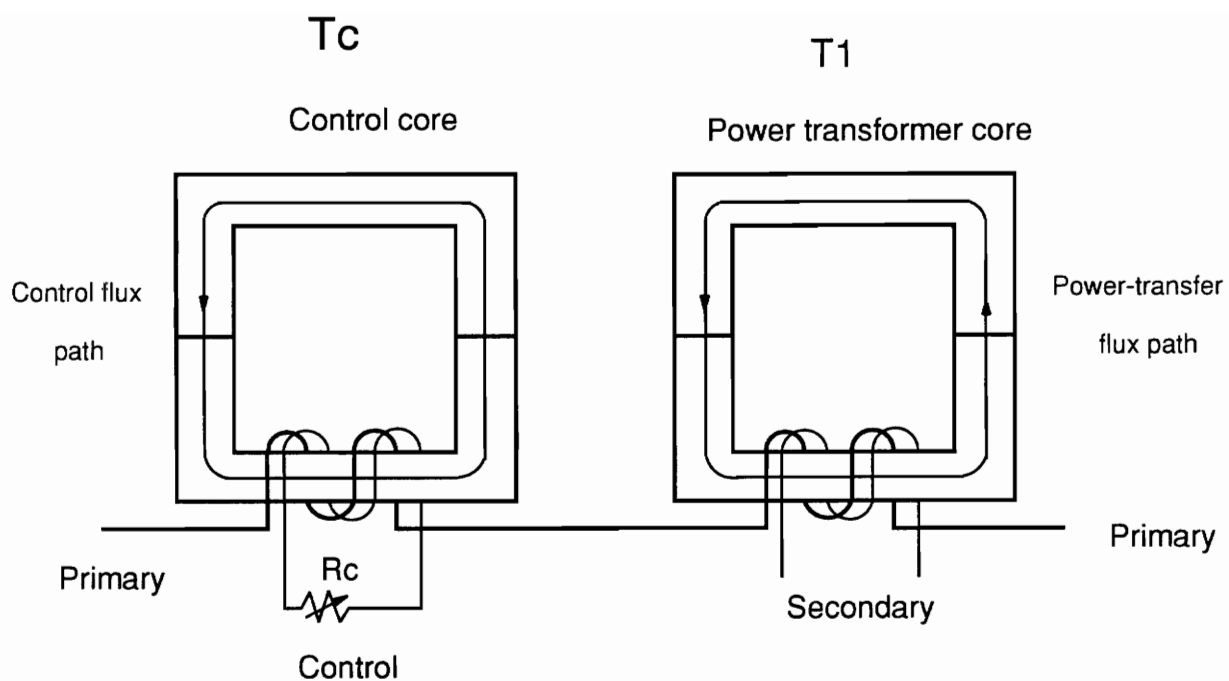


Fig. 2.4 Winding configuration of a serially-connected controlled transformer.

thesis. In these two subsections, it can be seen that the power transfer between the primary and the secondary side of the controlled transformer can be regulated by controlling resistance R_c . This kind of controlled transformer is called linearly-controlled transformer. In Chapter 3, a PWM-controlled transformer which is the main focus of this thesis, will be described, in which the resistance R_c is replaced by a MOSFET switch S3.

2.2 Converter Regulator Using Linearly-Controlled Transformer

In this section, a serially-connected controlled transformer is used in the illustration of a controlled transformer working in conjunction with a forward converter regulator. Fig. 2.5 shows a two-transistor forward converter configuration with a controlled transformer for regulating the output voltage. The duty cycle of the two switches Q1 and Q2 can be fixed or controlled by other means, but the function of the controlled transformer is to regulate the output voltage V_o against the load variation by controlling the resistance (R_c) across the control winding. The resistance R_c is achieved by using a MOSFET S3. By controlling the effective resistance of S3, the duty cycle of the secondary voltage of power transformer can be controlled and the output voltage can be regulated. The details of the circuit operation are described in the following. Reader should refer to Fig. 2.6 for the theoretical waveforms of the circuit. Fig. 2.7(a)(b) shows the characteristics of the B-H loops of both control core and the power core. This figure is used very often in the illustration.

Operation under heavy load condition

t_0 - t_1 (T_c blocking period): During this period, the main switches Q1 and Q2 are "on" and

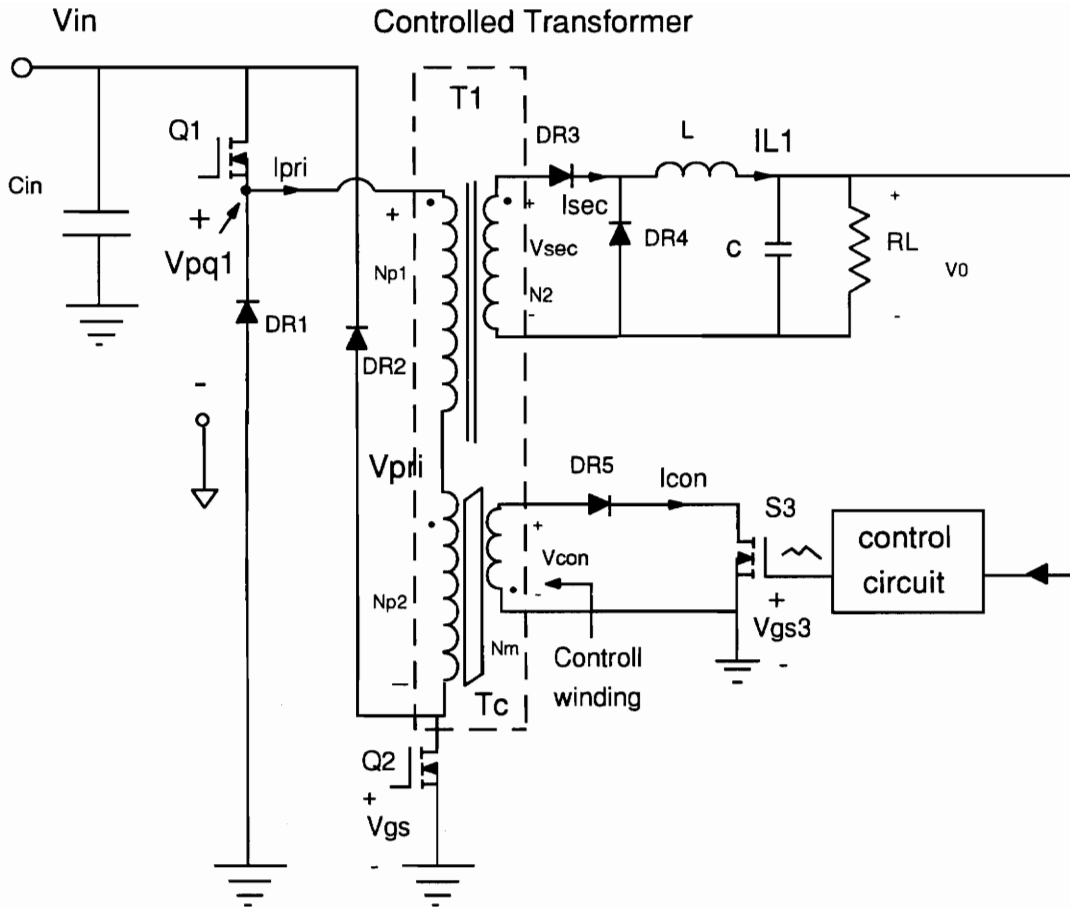


Fig. 2.5 Forward converter with a serially-connected controlled transformer using linear control.

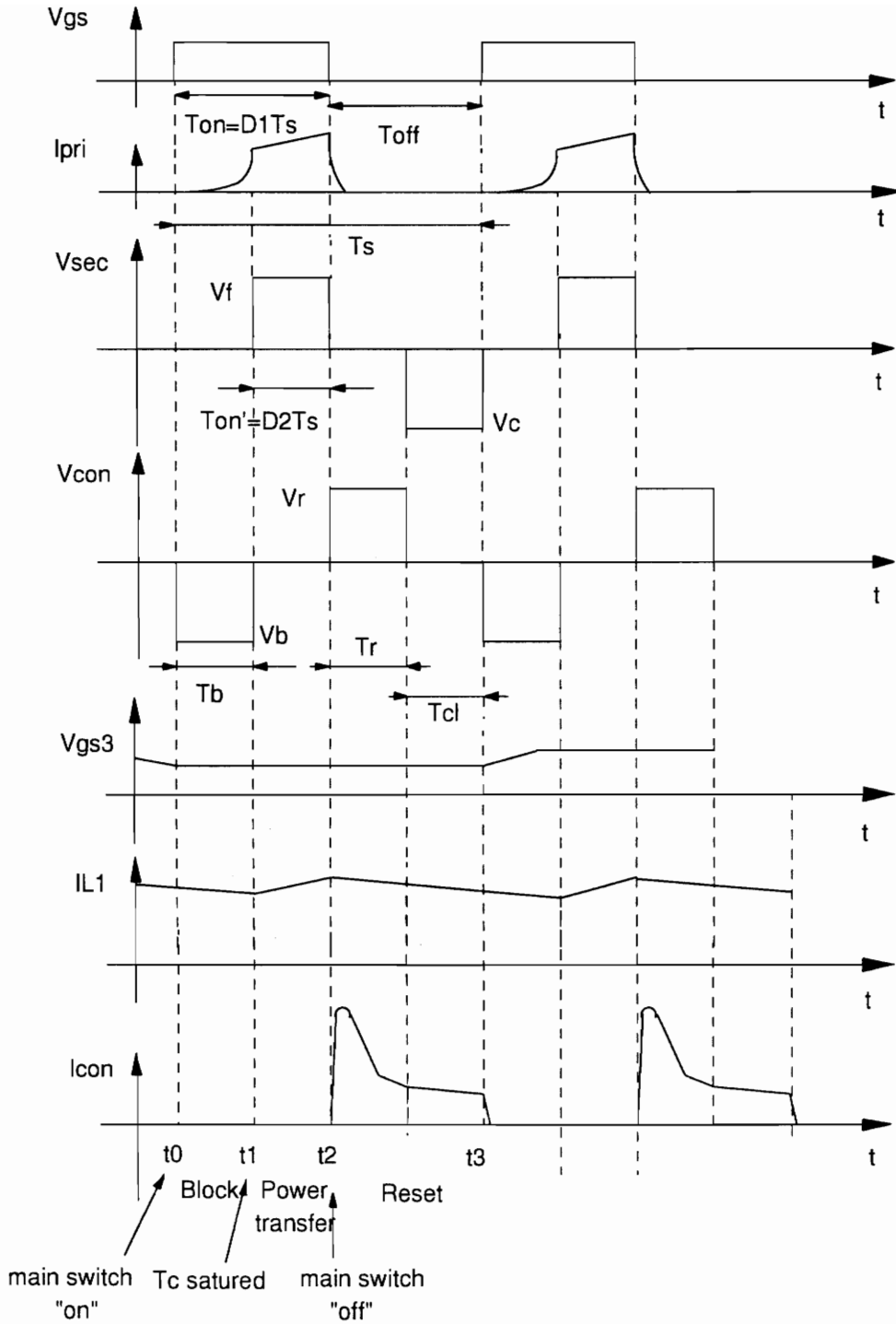


Fig. 2.6 Theoretical key waveforms of the circuit under heavy load by linear control.

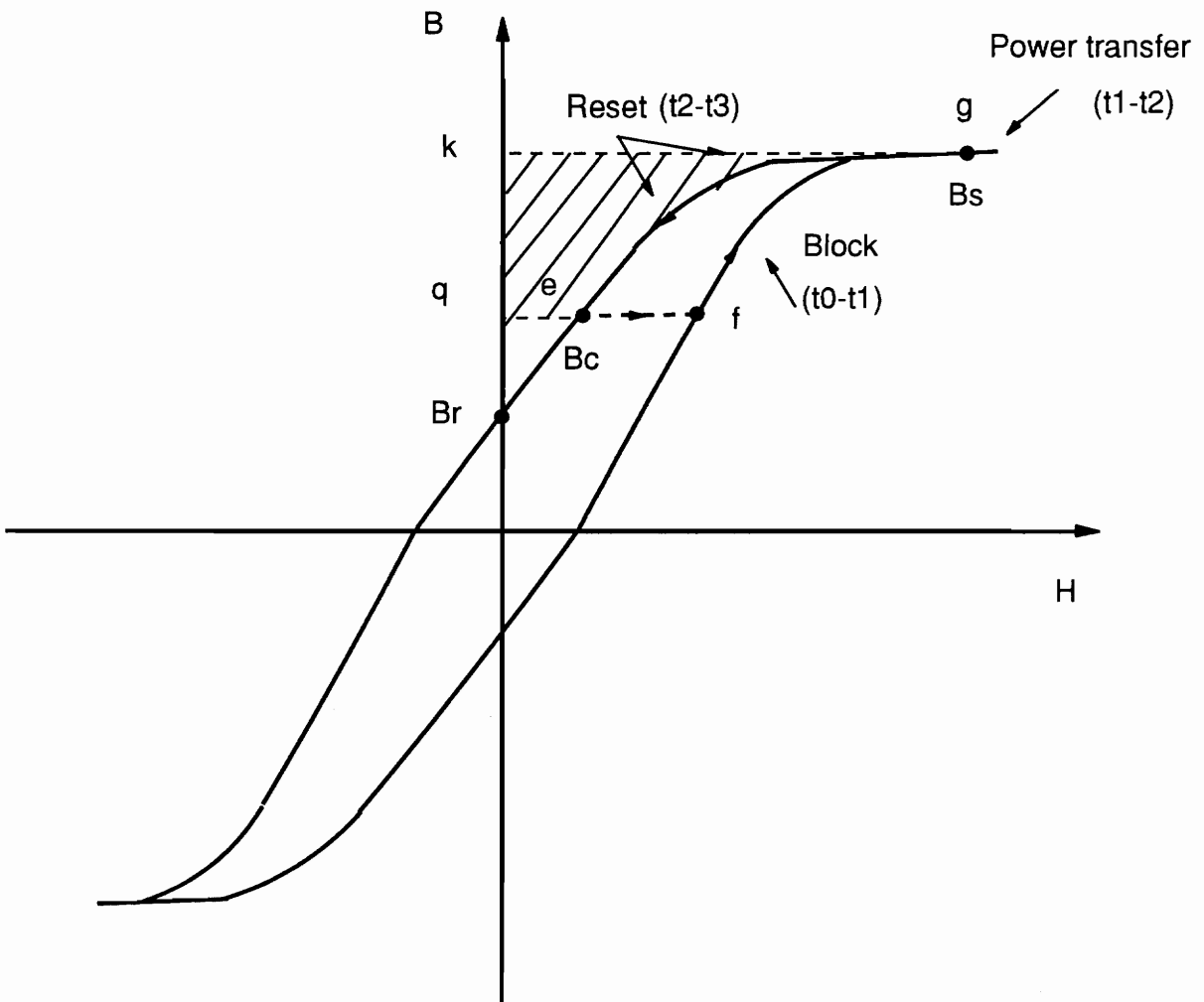


Fig. 2.7(a) Operation of the B-H loop of the control core T_c (by linear control).

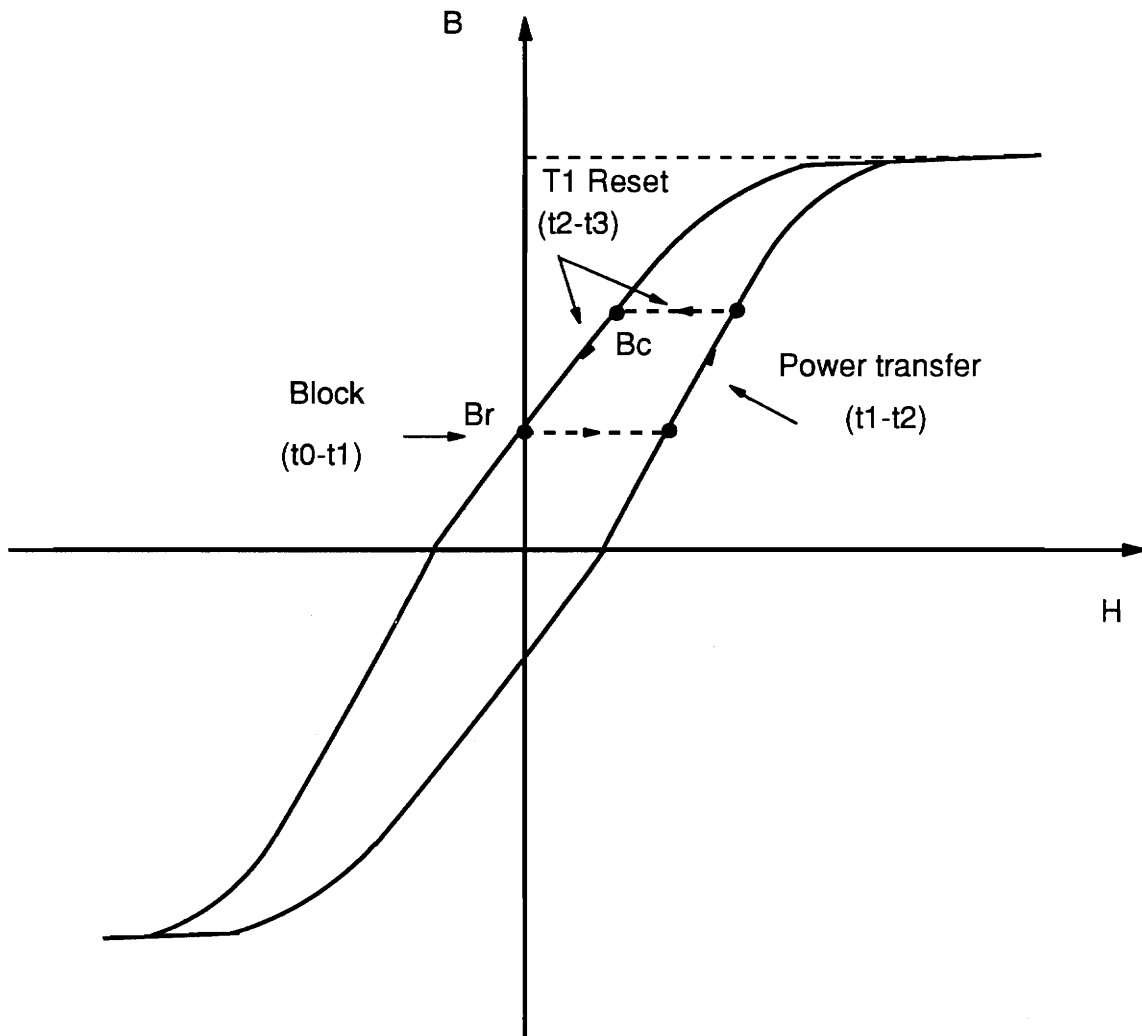
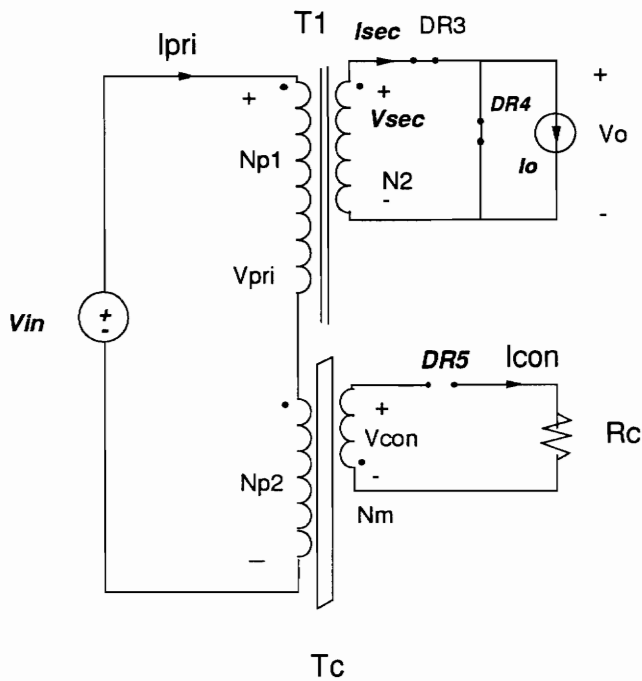


Fig. 2.7(b) Operation of the B-H loop of the power core T1 (by linear control).

DR5 is "off". The equivalent circuit is shown in Fig. 2.8. The primary current I_{pri} is the magnetizing current of the control core T_c . In a normal design, the magnetizing current of T_c is very small compared to load current. Therefore, the secondary current I_{sec} of the power transformer is very small or could be zero, depending upon the magnitude of the magnetizing current of T_c . DR4 has to conduct to support output load. In any event, the secondary winding of the power transformer T_1 is essentially shorted and the primary winding of T_1 cannot maintain any voltage. Therefore, the primary voltage is blocked by the primary winding of T_c and the flux density of T_c driven by this voltage moves from point f (in Fig. 2.7(a)) toward saturation. During this period of time, there is no power transfer to the output.



Flux density in T1 stays essentially constant

Flux density goes up on the vertical portion of B-H loop of Tc

Fig. 2.8 Equivalent circuit of converter during blocking period to-t1 (linear control)

t1-t2 (Power transfer period): As the flux density in the control core moves beyonds point g, T_c becomes saturated. Now, the primary voltage is mainly applied to the secondary winding and the current is forced to flow in DR3 to supply the load current. When I_{pri} becomes large enough, secondary current I_{sec} conducts through DR3 and therefore DR4 cuts off. Power is then transferred from the primary side to the load. During this period of time, flux density in T_c stays in saturation until the main switches Q1 and Q2 are turned off. Fig. 2.9 shows the equivalent circuit for this period.

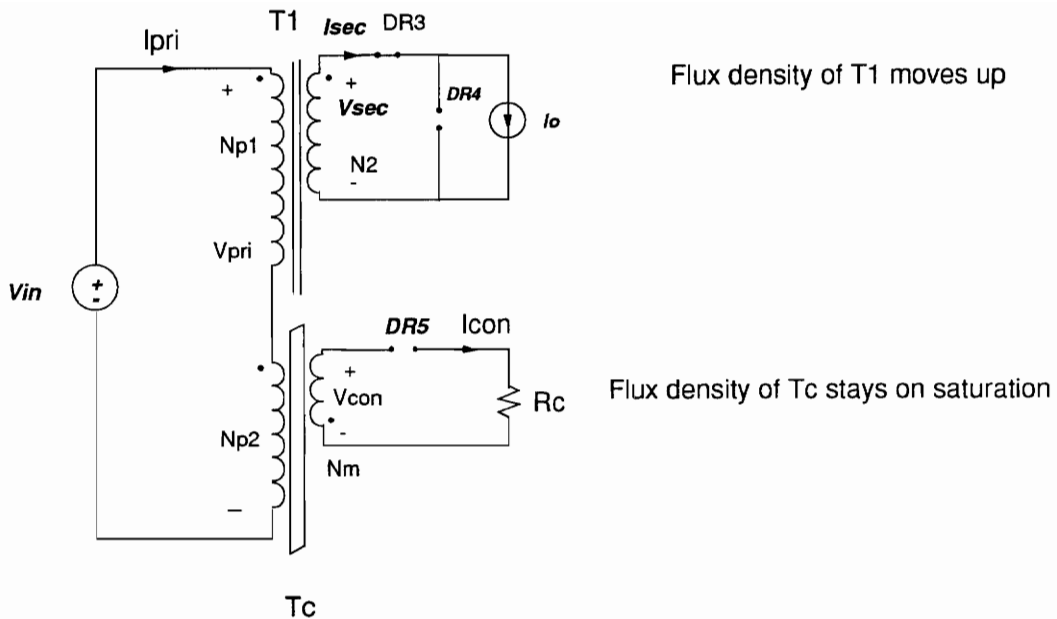
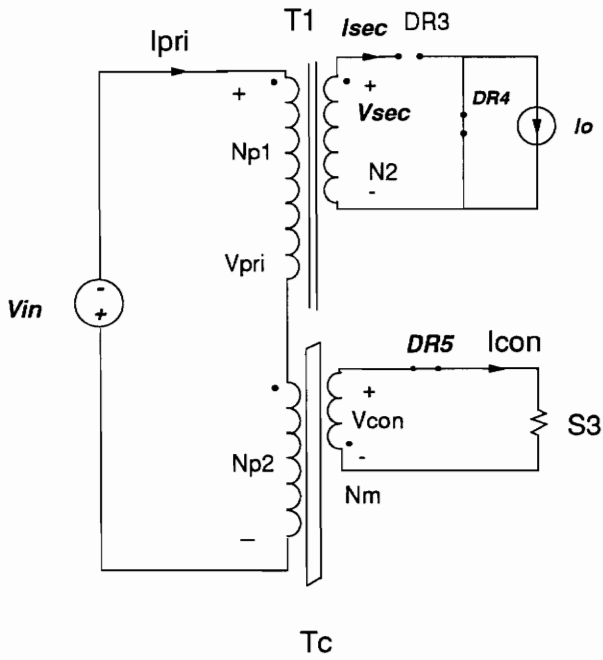


Fig. 2.9 Equivalent circuit of converter during power transfer period t_1 - t_2 (linear control).

t_2 - t_3 (Reset period): The power transfer period ends when the main switches turn off. Continuity of MMF forces reset diode DR1 and DR2 to conduct and the voltages on all the windings change polarity, which takes the control core out of saturation and reset both cores. DR3 cuts off and DR5 conducts. Fig. 2.9 shows the equivalent circuit for this period. The voltage across the control winding resets the flux density (in the control core) to point e. The amount of reset is determined by $V_{con} \cdot t_{off}$, where V_{con} is dependent on the effective resistance of S3 (i.e. R_c) and I_{con} depends on the load current and core material. A detail

analysis of control current is given in section 3.5 .



Flux density of T1 reset to B_r

Flux density of Tc resets to point e

Fig. 2.10 Equivalent circuit of converter during reset period t_2 - t_3 (linear control)

Operation under light load condition

Refer to circuit diagram of Fig. 2.5 and waveforms of Fig. 2.11 for the explanation of operation of linearly-controlled transformer regulator under light load. At very low output power the secondary and control voltage do not look like square waveforms since the primary voltage is divided by the two windings N_{p1} and N_{p2} unlike the case of heavy load. The reason is that DR3 could conduct and DR4 does not conduct during t_0 - t_1 period and the secondary winding voltage is not shorted, so flux changes occur in both control core and power core, and the primary voltage is distributed on both cores, as shown in Fig. 2.11. The same thing happens again during t_1 - t_2 since primary current is too small to saturate the control core. The shape of voltage waveform is also affected by effects of parasitic capacitors, snubbers. Equivalent circuit of controlled transformer under light load is shown in Figure 2.12. L_c is the equivalent inductance of the control core and L_s is the equivalent inductance of the power core. The reset flux density of control core is controlled by feedback control circuit. Therefore, L_c can be varied and the distribution of the primary voltage on L_c and L_s can be controlled. By controlling the voltage across L_s , output voltage of the converter can be regulated. Operation of controlled transformer under light load is like combination of variable inductor and power transformer.

Output voltage regulation

According to the above description, the feedback mechanism for regulating the output voltage can be explained as follow. Referring to Fig. 2.6, the output voltage is equal to the average of the positive-going V_{sec} waveform, i.e. $V_o = V_{sec} \cdot (T_{on} / T_s)$, where V_{sec} is the

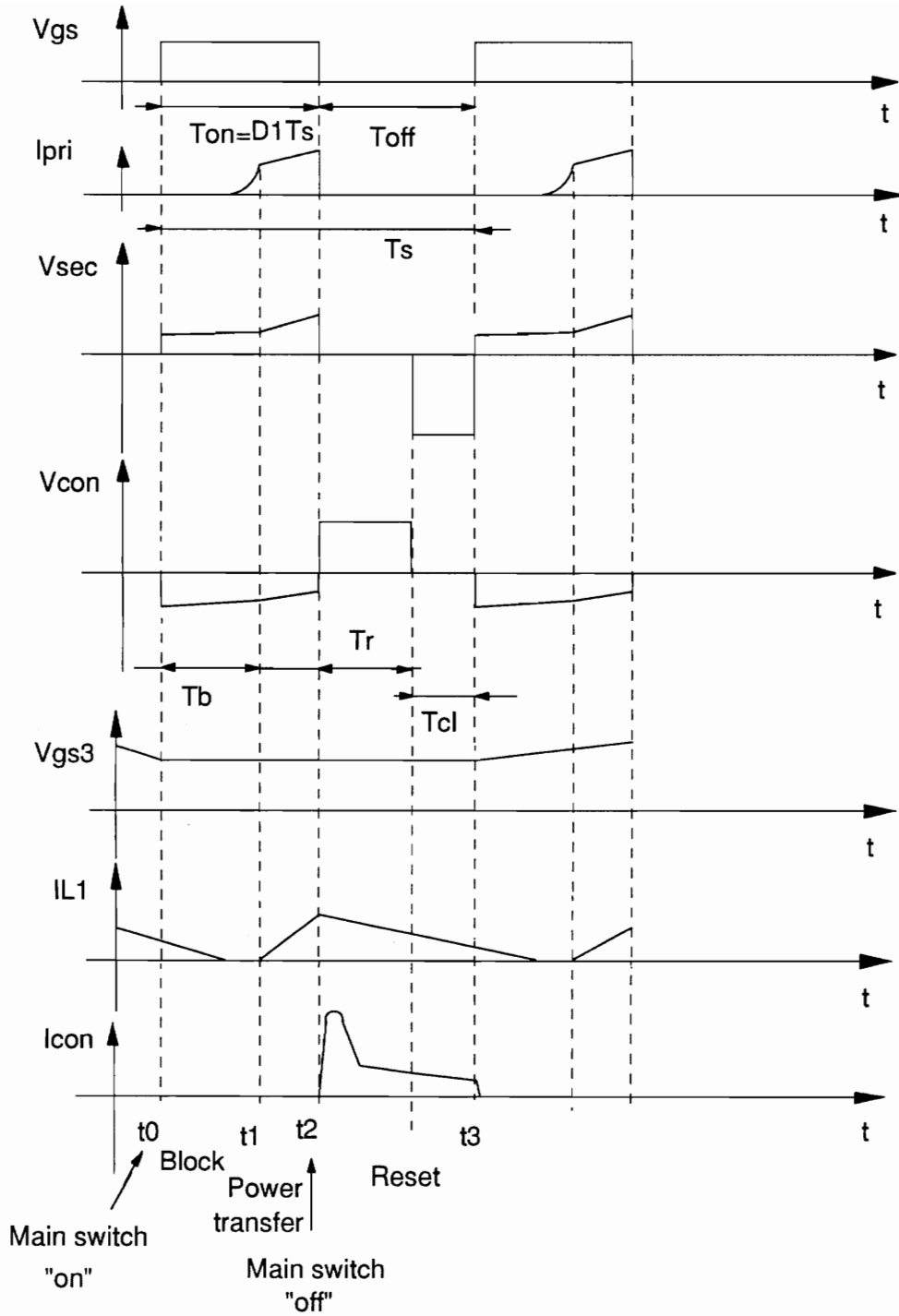


Fig. 2.11 Theoretical key waveforms of the circuit under light load by linear control.

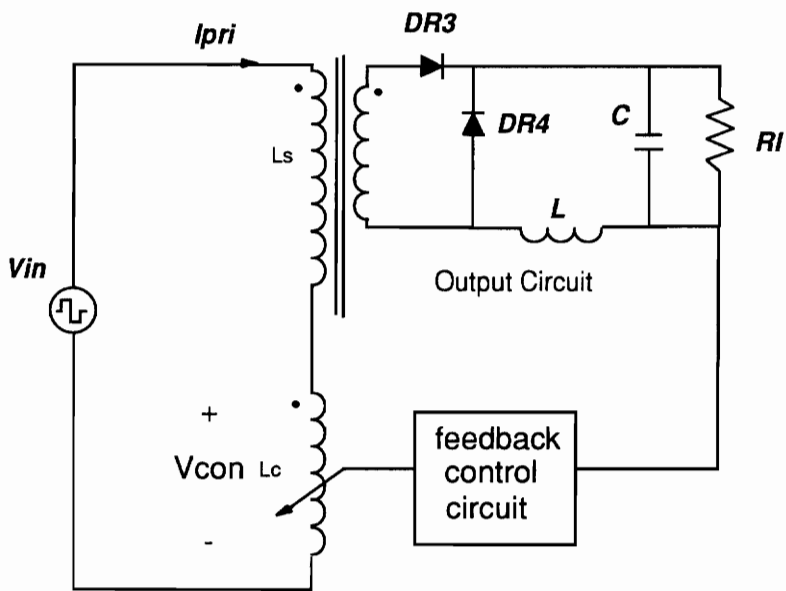


Fig. 2.12 Equivalent circuit of controlled transformer.
under light load

amplitude of the positive-going secondary voltage and T_{on}' is the power transfer period. The duration of T_{on}' (the difference of the main switches on time T_{on} and the duration of T_b) depends on the amount of flux reset occurred. The larger the reset flux, the longer T_b will be. The amount of flux reset, however, depends the voltage developed across the control winding N_m during the reset period. By controlling the effective resistance of transistor S3, the voltage across N_m can be controlled. A sequence of action is described in the following to explain the control mechanism for regulating the output voltage. Assume V_o exceeds the desired the voltage level. The error amplifier produces an error signal which reduces the gate drive signal and increases the effective resistance of S3, which in turn causes a larger voltage across the control winding during reset period. The voltage resets the flux of the control core further than before. This makes T_b longer in the following cycle which reduces T_{on}' . As a result of this action, V_o is pulled back to the desired value voltage again and regulation is achieved.

2.3 Comparison of Controlled-Transformer and Conventional Magamp Regulator

Both magamp regulators and controlled-transformer post regulator can be used in multiple-output switching-mode power supplies. The similarities and the differences between these two types of post regulator will be discussed in the following.

Similarities

- 1) Both are capable of providing independently regulated output voltages. Relatively good regulation can be achieved for multiple-output power supplies.

- 2) Both can be operated with local feedback to regulate the output voltage. Local feedback do not need to cross the isolation boundary which is a significant advantage in mass production.
- 3) Individual and/or collective shut down of outputs are possible.

Differences

- 1) A magamp regulator normally requires the use of a square-loop core for efficient operation [1]. Therefore, toroidal tape-wound cores of nickel-iron alloys and amorphous cores are used. These cores are, generally speaking, relatively expensive and cannot be operated at very high frequency ($>200\text{KHz}$) because of excessive core losses at such frequencies.
- 2) A controlled transformer, however, does not require the squareness of the core loop for efficient operation. Therefore, two-piece soft ferrite core (such as E-E core) can be used. The cost of ferrite cores is generally lower, but more importantly the winding cost can be drastically reduced in a high volume production situation. This flexibility also allows the possibility of extending the operating frequency above 200KHz . A detailed description of the desired core characteristics for controlled transformer is given in section 3.3. Using linear control, a relatively large power loss occurs in a controlled transformer. Therefore, a linearly-controlled transformer regulator compares unfavorably with the magamp. Unless power conversion efficiency in a controlled trans-

former can be significantly improved, it is not really a viable approach. This leads to Chapter 3 in which an efficient control way for controlling the controlled transformer is proposed.

Chapter 3

VOLTAGE REGULATOR USING CONTROLLED TRANSFORMER BY PWM CONTROL

Voltage regulator using linearly-controlled transformer works, but yields low efficiency due to relatively large power dissipation in the control circuit. In this chapter, a voltage regulator using controlled transformer by a PWM control will be presented. The principle of operation of a PWM-controlled transformer and the implementation of feedback control circuit are described. Smaller control power and higher efficiency of regulator can be achieved when the control core is reset by PWM control instead of linear control. Design consideration and design example for controlled transformer are given.

3.1 Theory of Operation of a PWM Controlled Transformer Regulator

Figure 3.1 shows the functional diagram of a two-transistor forward converter configuration with a PWM controlled-transformer. The main difference between this circuit and the linearly-controlled transformer regulator shown in Fig. 2.7 is the manner the switch (S3) is controlled. In Fig. 2.7, S3 is controlled by the feedback signal to function in the active region behaving as a variable resistance; and in Fig. 3.1 S3 is pulse-width-modulated by the feedback signal. In other words, switch S3 is either "on" or "off" with controllable duty

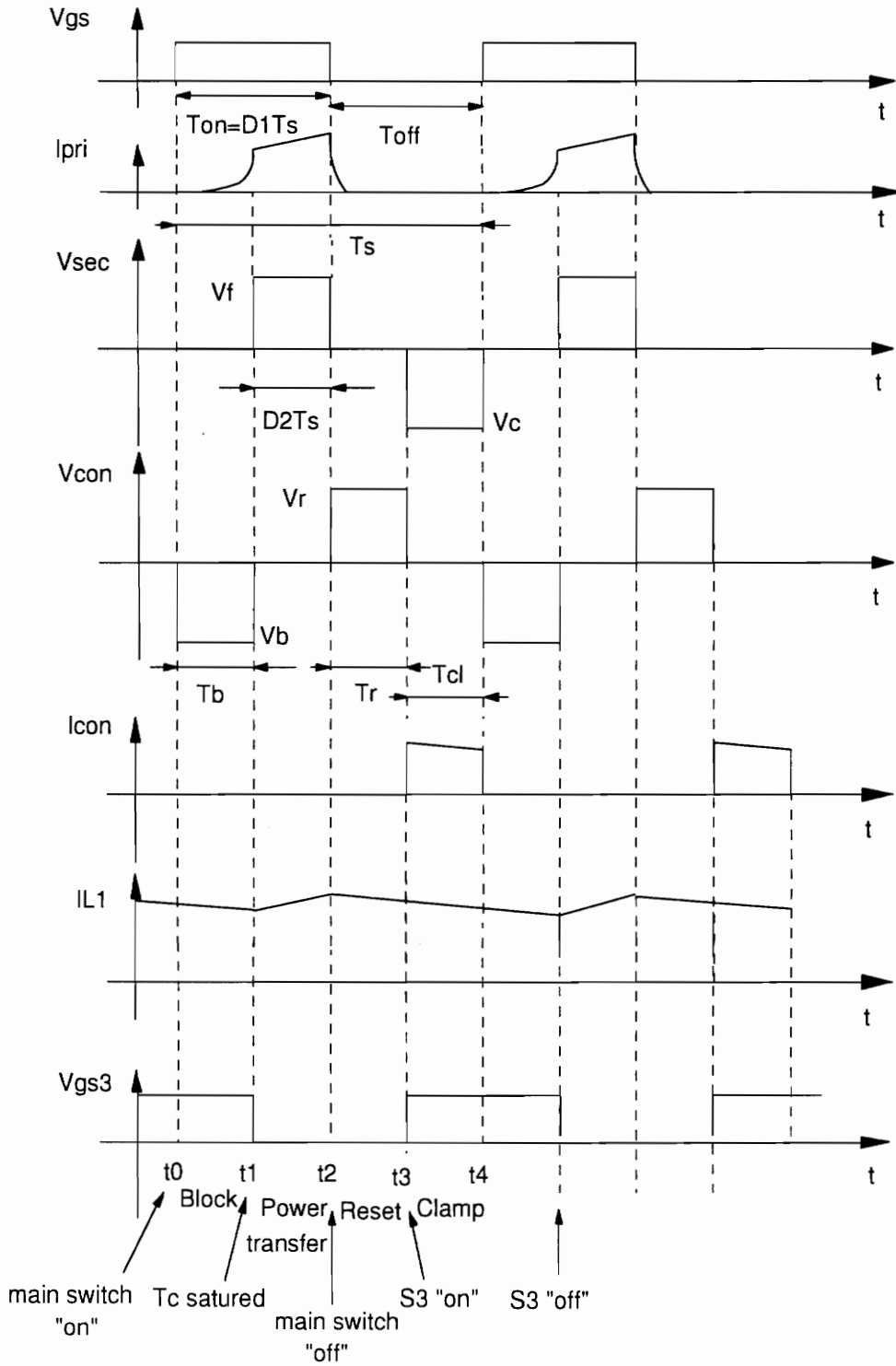


Fig. 3.2 Theoretical key waveforms of the circuit under heavy load by PWM control.

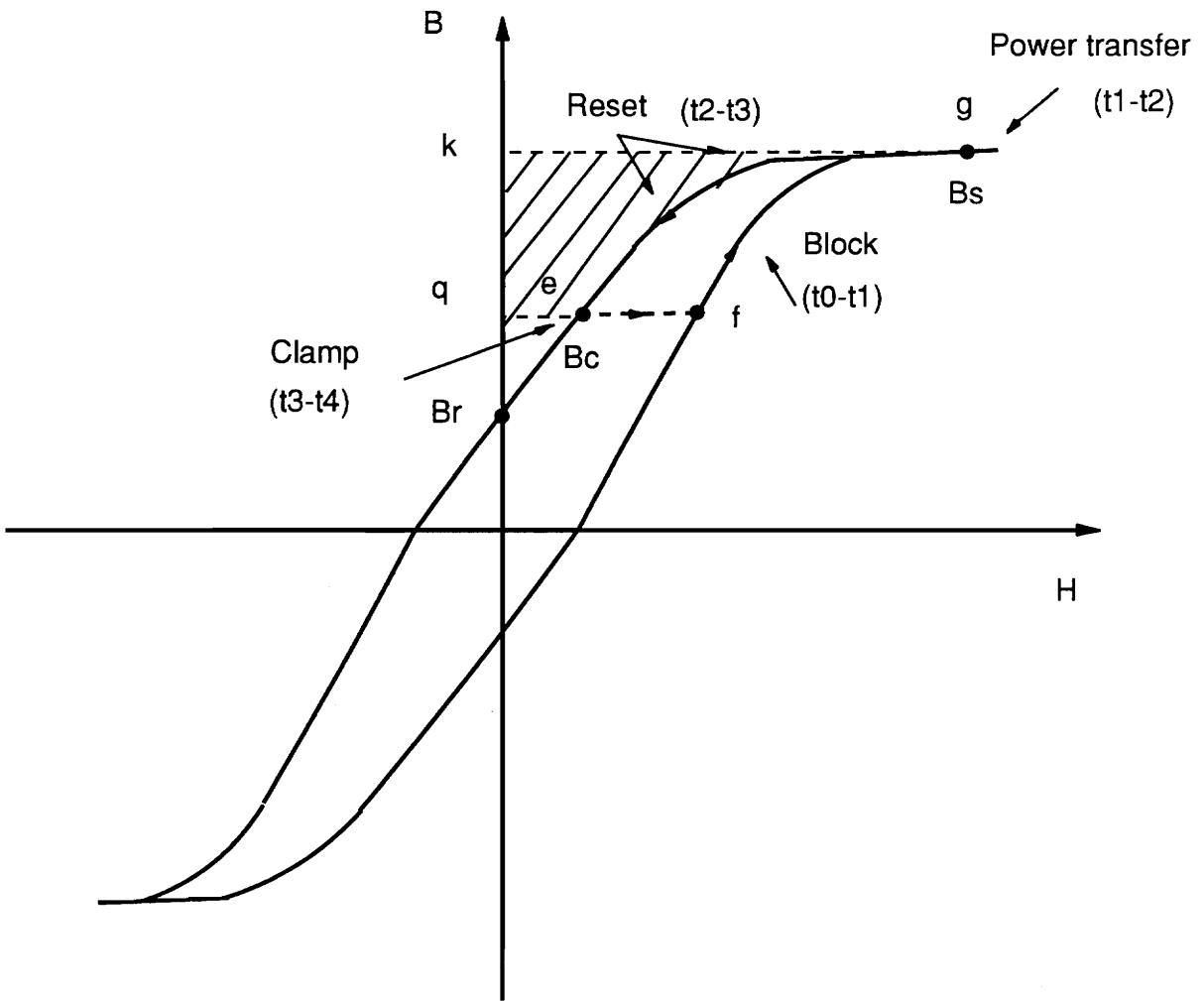


Fig. 3.3(a) Operation of the B-H loop of the control core T_c (by PWM control).

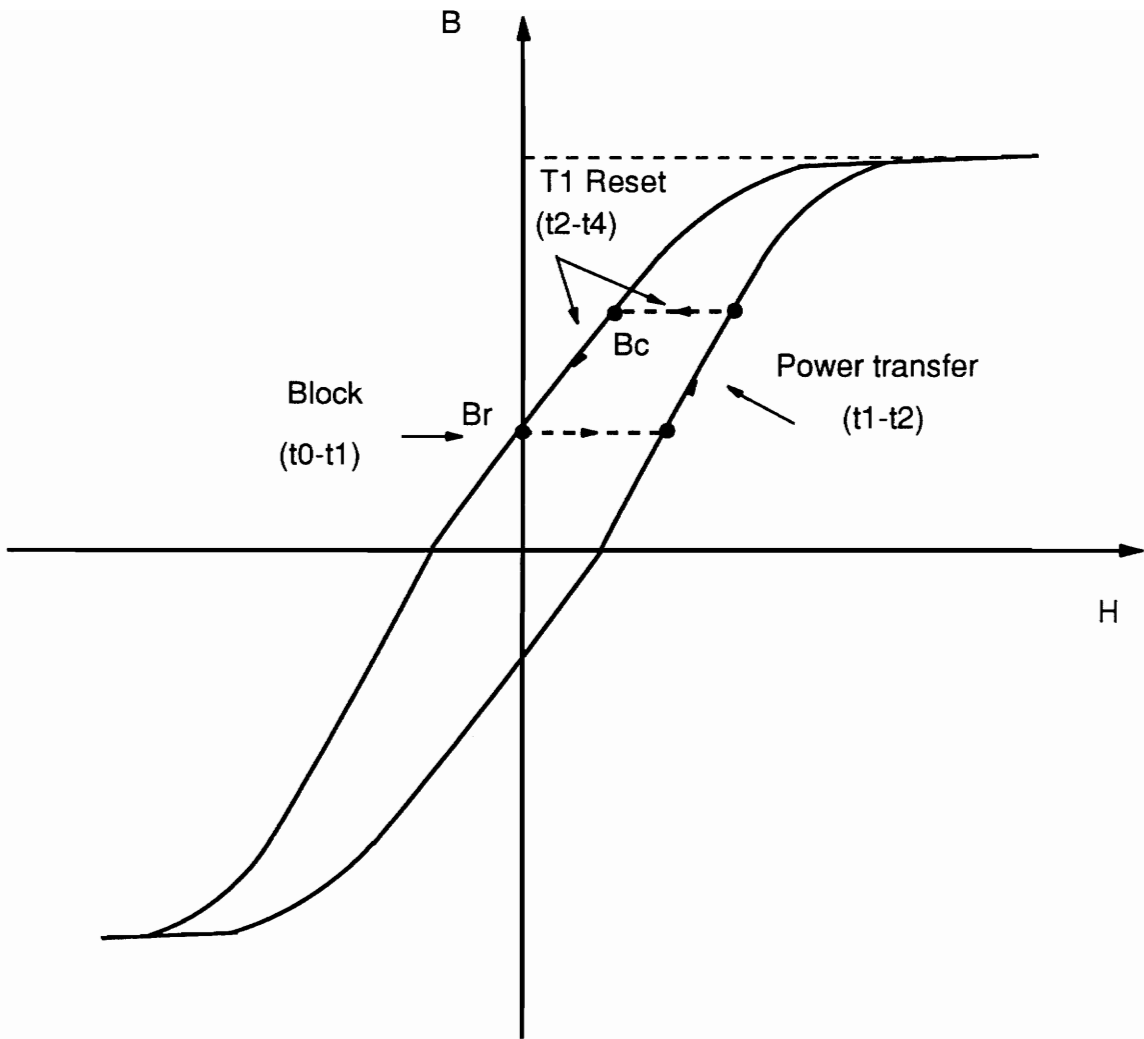


Fig. 3.3(b) Operation of the B-H loop of the power core T1 (by PWM control).

cycle. By controlling the turn-on time of S3, output voltage can be regulated. The "synchronous circuit" block in Fig. 3.1 is used to derive a synchronous signal from the controlled transformer to drive S3. Since a serially-connected controlled transformer is a better choice as described in chapter 2, it will be used to explain the basic principle of operation when a PWM controlled transformer is used with a converter regulator. The operation of a PWM controlled transformer regulator is similar to that of a linearly-controlled transformer regulator as described in chapter 2. However, there is a basic difference in the reset mechanism. Four operation stages exist within one switching cycle: blocking period, power transfer period, reset period and clamp period. The detail description will be given in the following. In the description, the readers should refer to Fig. 3.2 for the waveforms and Fig. 3.3(a)(b) for the B-H trajectories of both T_c and T_1 whenever necessary.

t_0 - t_1 (T_c block period): During this period, main switches Q1 and Q2 are "on" and DR5 is "off". The operation of the circuit during this period is identical to that of the linearly-controlled transformer. For the convenience of readers, the same description of the operation is given below. The equivalent circuit is shown in Fig. 3.4. The primary current I_{pri} is the magnetizing current of the control core T_c . In a normal design, I_{pri} is very small compared to load current. Therefore, the secondary current of the power transformer I_{sec} is at best very small or could be zero, depending upon the magnitude of magnetizing current of T_c . DR4 must be "on" to maintain output load current. In any event, the secondary winding of the power transformer T_1 is essentially shorted and the primary winding of T_1 cannot maintain

any voltage. Therefore, the primary voltage is blocked by the control primary winding of T_c and the flux density of T_c driven by this voltage moves from point f (in Fig. 3.3) toward saturation. During this period of time, no power is transferred from the input to the output.

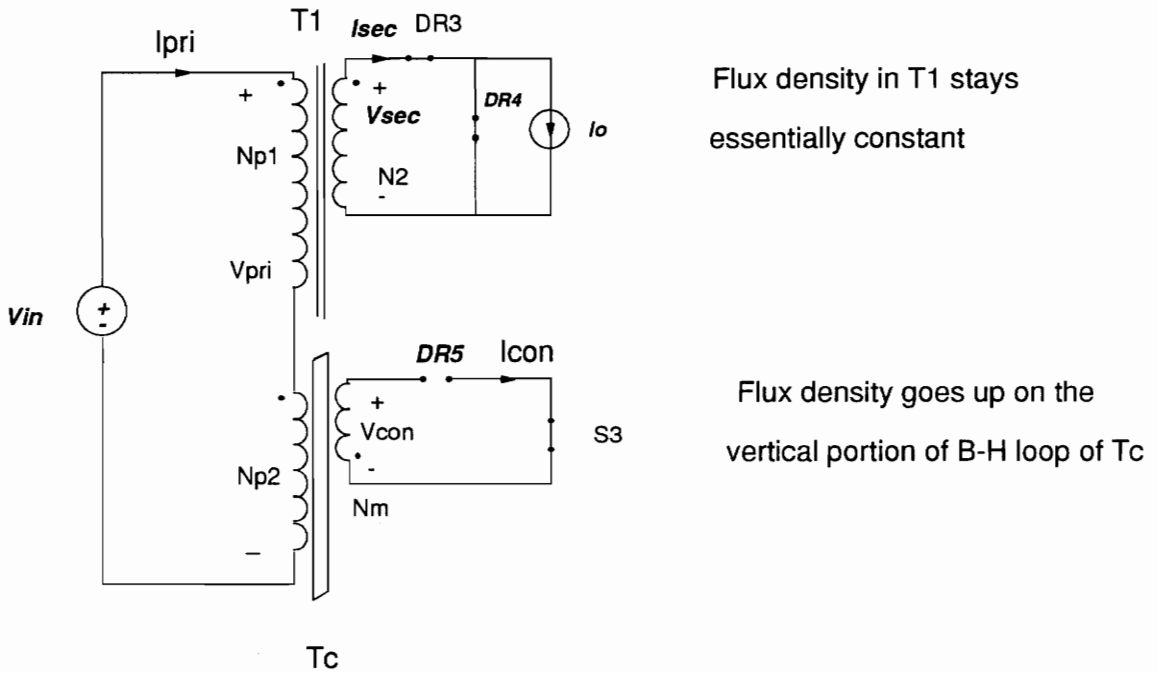


Fig. 3.4 Equivalent circuit of converter during blocking period t_0-t_1 (PWM control).

t_1-t_2 (Power transfer period): During this period, DR5 is "off" and, therefore, the operation of the circuit is also identical to that of the linearly controlled counterpart. When the control core becomes saturated, the magnetizing current becomes large. When I_{pri} becomes large enough, secondary current I_{sec} conducts through DR3 and forces DR4 "off" and power

transfer occurs. The input voltage is essentially sustained by the primary winding N_{p1} of power core. Fig. 3.5 shows the equivalent circuit for t_1 - t_2 period. During this period of time flux density in T_c stays in saturation until the main switches Q1, Q2 are turned off.

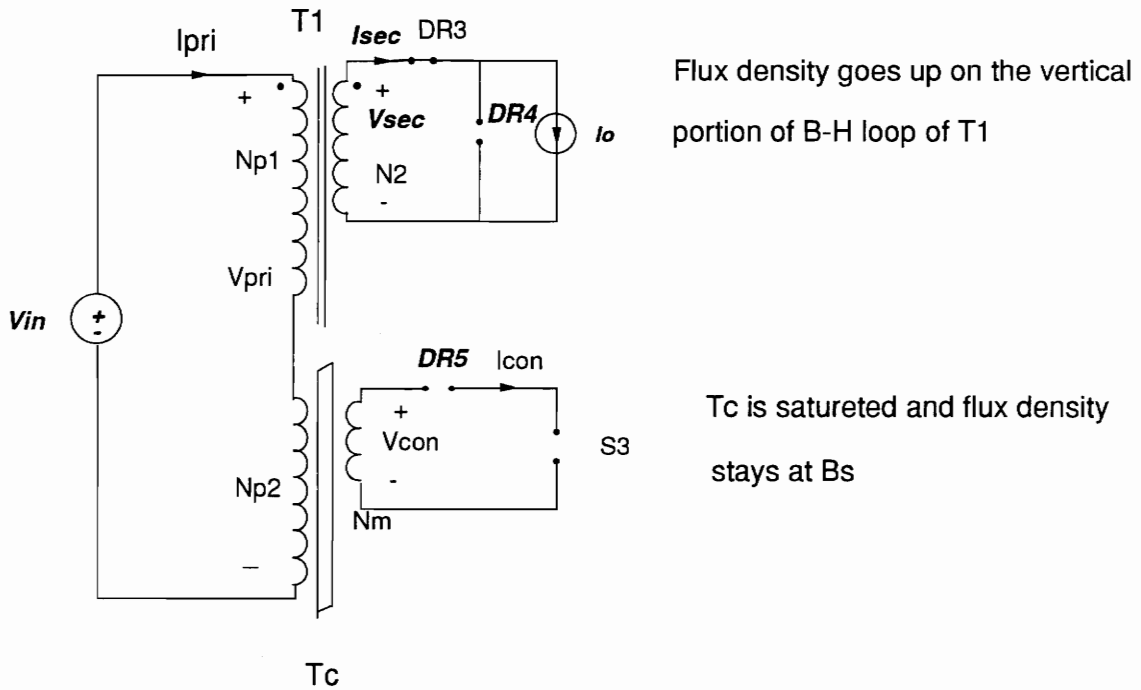


Fig. 3.5 Equivalent circuit of converter during power transfer period t_1 - t_2 (PWM control).

t_2 - t_3 (Reset period): During this period of time, there is big difference between the PWM-controlled and the linearly-controlled scheme. When the main switches Q1 and Q2 turn off and the two reset diodes DR1 and DR2 conduct, the voltage on the control primary (N_{p2}) reverse polarity which takes T_c out of saturation and the flux moves toward remnant point and S3 is still "off". Fig. 3.6 shows the equivalent circuit for the reset period. During

this period of time, both T1 and T_c are being reset. The stored energy in both T1 and T_c are being returned to the input source. This is the main reason why a PWM-controlled transformer is more efficient than a linearly-controlled transformer. In the latter, this part of energy is being dissipated mostly by the variable resistor S3 during the reset period. Magnetic field strength H of both T_c and T1 is determined by I_{pri} and the turns N_{p1} and N_{p2}. However, before the flux density in T_c is reset to remnant flux density B_r, S3 is turned on by the feedback signal which leads to the next operational period.

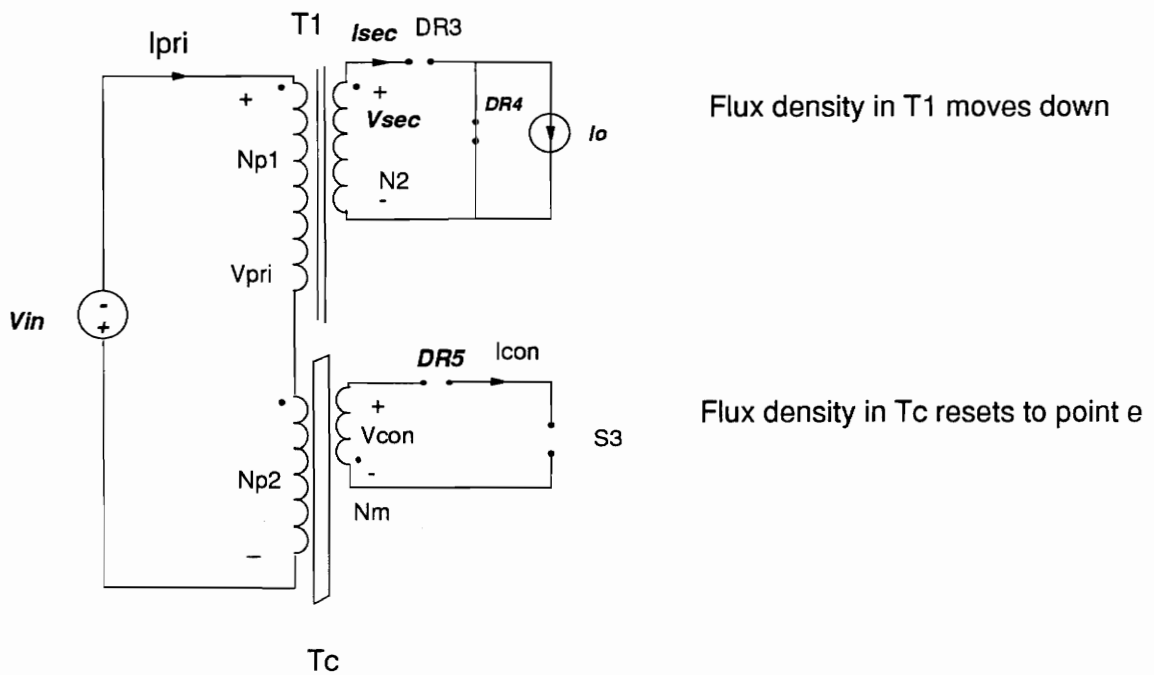


Fig. 3.6 Equivalent circuit of converter during reset period
t2-t3 (PWM control).

t3-t4 (Clamp period): There is no such period in the linear control case. In the PWM-controlled case, at t3 which is determined by the feedback control circuit, S3 is turned on and a short circuit is applied across the control secondary winding as shown in equivalent Fig. 3.7. Since the voltage across the control secondary winding is essentially equal to zero, flux in T_c stays at point e. The negative primary voltage is blocked by N_{p1} that resets T_1 . The current flowing in the control winding is determined by the primary current of T_1 and the magnetizing current of T_c . During this period of time, T_1 is reset to B_r and the flux density in T_c is kept at point e determined by feedback control circuit. When Q1 and Q2 turn on again, the initial flux density in the control core will rise from the point e toward saturation and the control core is back to blocking period as described earlier.

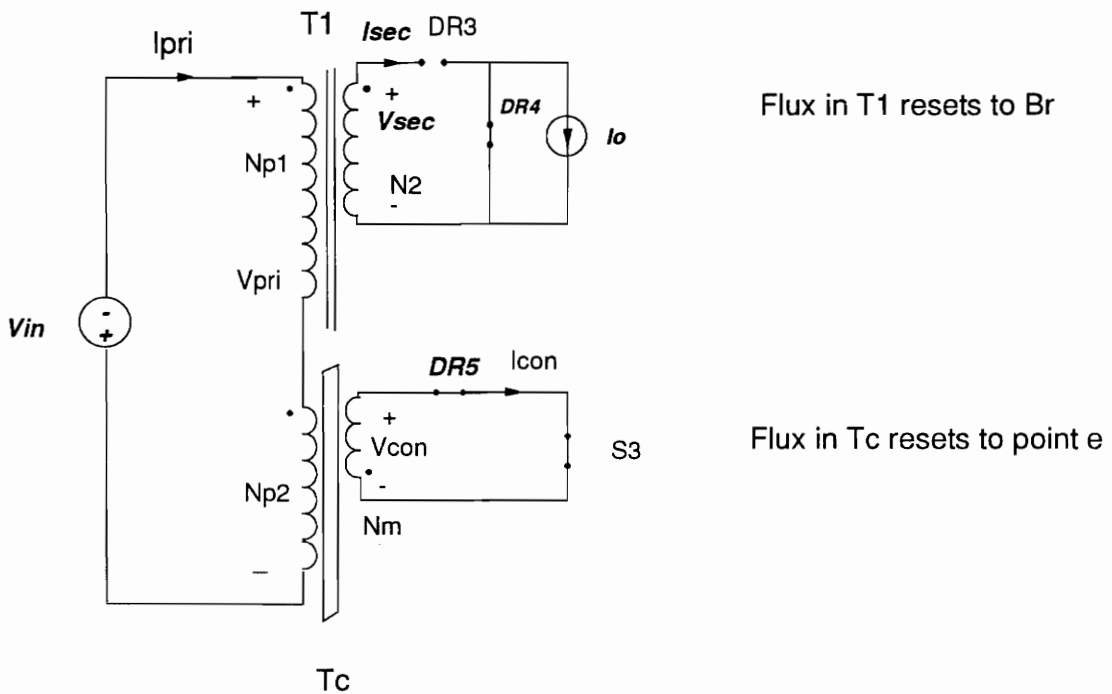


Fig. 3.7 Equivalent circuit of converter during clamp period t_3 - t_4 (PWM control).

Operation under light load condition

Operation of PWM-controlled transformer regulator under light load is slightly different from that under heavy load. Refer to the functional circuit diagram of Fig. 3.1 and the waveforms of Fig. 3.8 for the explanation of the operation of PWM controlled transformer regulator under light load. The voltage waveforms on the secondary winding and the control winding do not look like square waveforms during t_0 - t_2 when the load is very light. The

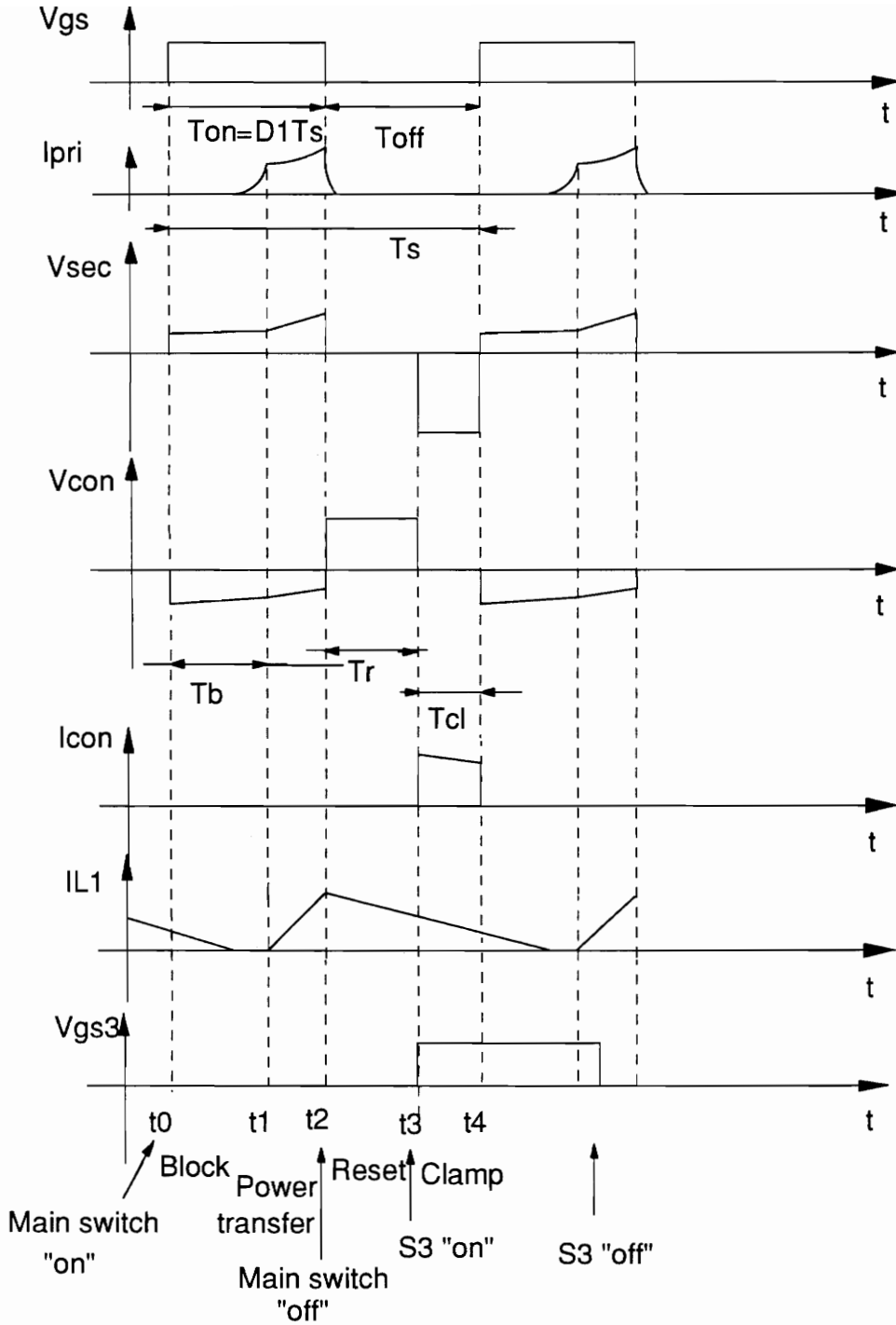


Fig. 3.8 Theoretical key waveforms of the circuit under light load by PWM control.

reason is that input voltage is divided by two cores, unlike the case of heavy load. The explanation of this case is that DR3 could conduct and DR4 does not conduct during t_0 - t_1 period and the secondary voltage cannot be clamped to the ground any more, so flux changes occur in both control core and power core, and the primary voltage is distributed on both cores, as shown in Fig. 3.8. The same thing happens again during t_1 - t_2 since the primary current is too small to saturate the control core. The primary voltage V_{pri} is divided by two primary winding N_{p1} and N_{p2} since DR4 is not conducted and the secondary winding is not shorted, which is different from heavy load case. Another reason is that the shape of waveforms is affected by effects of parasitic capacitors, snubbers. Equivalent circuit of controlled transformer is shown in Figure 2.12. L_c is the equivalent inductance of the control core and L_s is the equivalent inductance of the power core. L_c can be controlled by feedback control circuit and the secondary voltage on L_s is controlled too and therefore, output voltage of the converter can be regulated. Therefore, operation of controlled transformer under light load is like combination of variable inductor and power transformer.

In summary, the main-switches "on" initiates the blocking period and flux moves toward saturation. When the control core become saturated, power transfer between the primary side and secondary side starts. When the main-switches are turned off by gate drive, T_c core resets. Sometime during main switch "off" period, depending on feedback control, S3 is turned on, which stops the flux motion of the control core and T_1 core resets. By controlling the timing of S3 turn-on, the duration of the blocking period and the duty cycle of the secondary voltage is indirectly controlled and therefore, the output voltage is regulated. The detail of PWM control circuit will be described in the next section. It's to be noted that for

two-transistor forward converter, the maximum duty cycle is limited by 50% and therefore, with controlled transformer, the duty cycle of the PWM transfer period is less than 50%. In a single-transistor forward converter with separate reset winding, the duty cycle can be designed to operate at more than 50% if necessary.

3.2 Description of Control Circuits

While the controlled-transformer alone can be used to regulate the output voltage against variation of input voltage and output power, it is often more efficient to have a feedforward control circuit as well. Feedforward control is a commonly-used scheme to provide constant volt-second across the primary winding. i.e to keep $V_{pri} \cdot t_{on}$ a constant, where t_{on} is the "on time" of the two switches Q1 and Q2. In this way, the efficiency is improved and there is no need for isolation in the feedback circuit. However, the main focus of this section is the feedback control of the controlled transformer and the feedforward control will not be discussed.

3.2.1 Implementation of Feedback Control Circuit

Figure 3.9 shows a completed circuit diagram of two-transistor forward converter using a PWM-controlled transformer. The feedback control circuit consists of three parts of circuits, synchronous circuit, error circuit and drive circuit, as shown in Fig. 3.9. Synchronous circuit

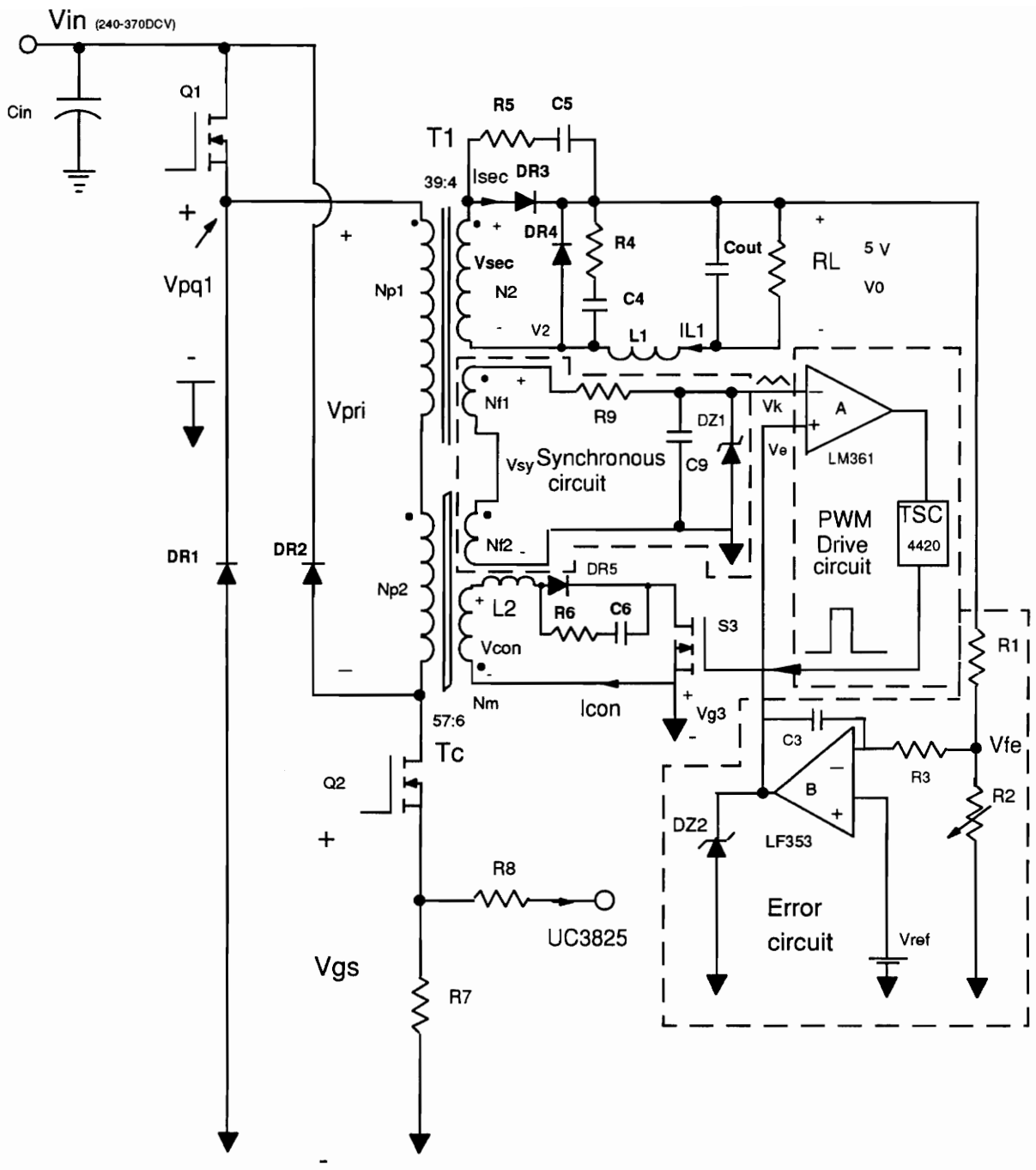


Fig. 3.9 Schematic of forward converter with controlled transformer using PWM control.

provides synchronous signal with main switches for control of S3; error circuit produces amplified error signal by comparing the sampled output voltage and the reference voltage; and the drive circuit produces the PWM signal to drive the power MOSFET S3. The detail of each part of the feedback control circuit is described as follows:

Synchronous circuit

In order to obtain a synchronous signal for control circuit, two sampling windings, N_{f1} and N_{f2} , were used. One is wound around the power core and the other one wound around the control core. Then opposing polarities of two windings are connected together. The number of turns of sampling windings should satisfy the following equation:

$$\frac{N_{f1}}{N_{f2}} = \frac{N_2}{N_m}$$

The voltage sum V_{sy} of sampling windings N_{f1} and N_{f2} is in phase with primary voltage V_{pri} and is proportional to V_{pri} . A RC integrator is then used to produce a triangular waveform V_k for PWM control. A zener diode DZ1 is used to limit the triangular voltage within the input voltage rating of the comparator. The time constant of the RC integrator is usually chosen greater than T_s (the switching cycle). The triangular waveform V_k is then used as one of the inputs to the PWM drive circuit.

Error circuit

Error circuit consists of sampling resistors R1 ,R2 and reference voltage source V_{ref} , an operational amplifier and feedback compensating elements R3 and C3 and a output voltage limiting zener DZ2 shown in Fig.3.10. The output of the error voltage is fed to the input of the PWM control circuit.

PWM drive circuit

A high speed comparator LM361 is used as a PWM modulator. PWM signal is generated by comparing the triangular signal V_k (output of the synchronous circuit) and the error signal V_e . An IC driver TSC4420 is used to boost the current capability to drive MOSFET S3.

Operation of feedback control circuit

Fig. 3.10 shows the waveforms associated with the control circuit. The purpose of the feedback control circuit is to vary the duty $D_2(=T_{on}'/T_s)$ to keep the output voltage at a desired value. From the top two waveforms, it can be seen that $T_{on}'=T_{on}-T_b$ where T_b is the blocking period.

From the V_{con} waveform, volt-seconds balance in the control core is:

$$T_b = V_r \frac{T_r}{V_b} \quad (3-1)$$

Then, power transfer period T_{on}' can be expressed:

$$T_{on}' = T_{on} - V_r \frac{T_r}{V_b} \quad (3-2)$$

Equation (3-2) leads to Eq.(3-3)

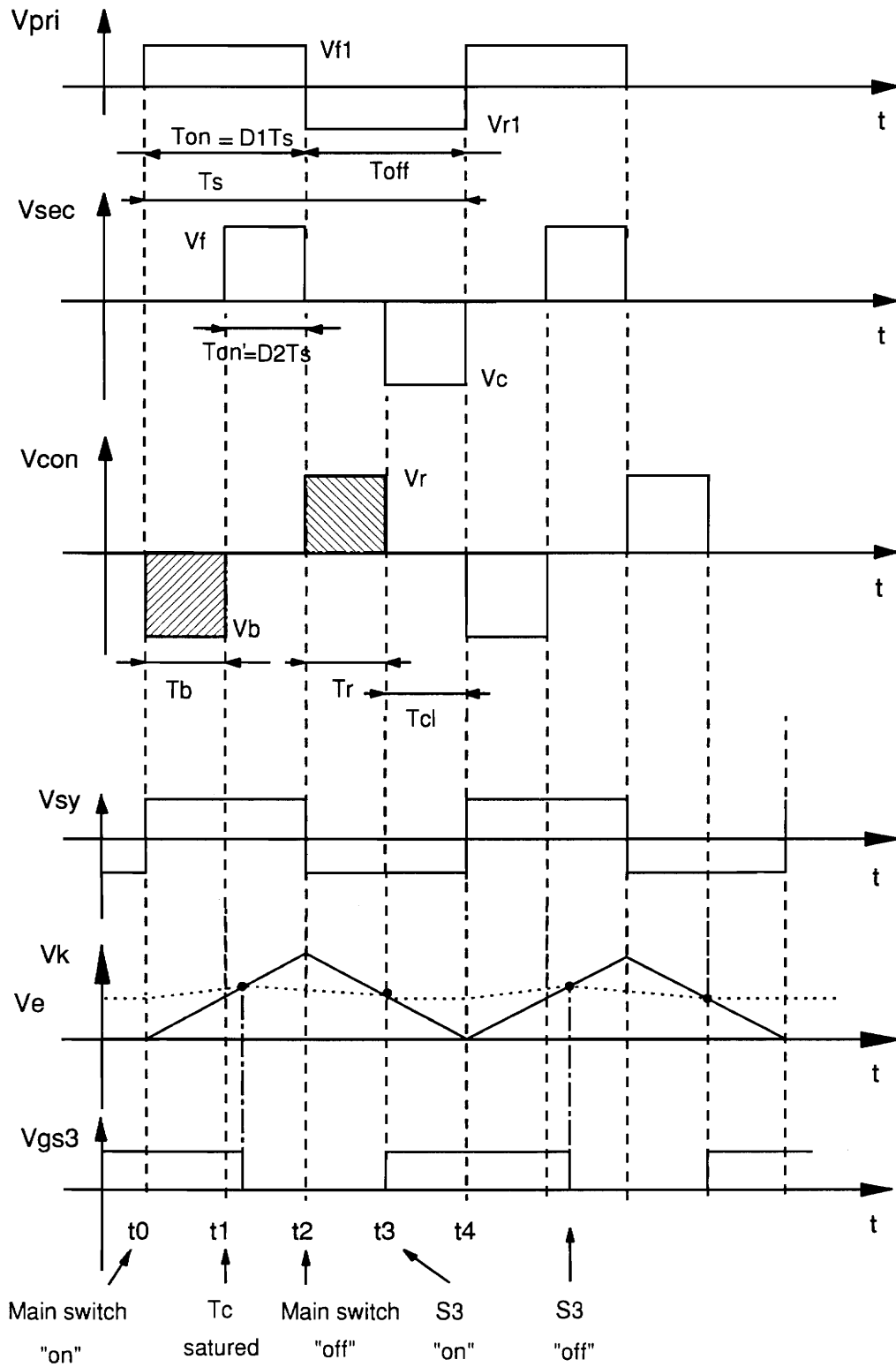


Fig. 3.10 Theoretical operating waveforms of the main and control circuit.

$$D_2 = D_1 - \left(\frac{T_r}{T_s} \right) \cdot \frac{V_r}{V_b} \quad (3-3)$$

Where D_1 is determined by the main switch duty cycle. V_r and V_b are respectively the voltage across the control winding during the reset and the blocking period. They are fixed and usually equal. D_2 is controlled by controlling T_r/T_s . This is achieved by controlling the instant t_3 when S3 is turned on. From the bottom two waveforms, it can be seen that S3 is turned on when the feedback error voltage V_e intersects with V_k on the declining side of the triangular waveform. The sequence of control circuit action is described in the following when the output voltage deviates from the desired voltage. Assume V_o exceeds the desired regulated voltage level. The error voltage V_e should increase which leads to later turn-on of S3, as can be seen from the V_k and the V_e waveforms. This increases T_r and reduces D_2 which pulls V_o back to regulated voltage. It is noted that S3 conducts only during clamp period(t_3-t_4). Therefore, the longer the clamp period, the larger the loss in the control circuit. However, the longer the clamp period, the shorter the blocking period and therefore, the smaller the flux swing of the control core. This makes control core loss smaller.

3.3 Design Consideration for Controlled Transformer

A controlled transformer consists of a power transformer and a control core and winding. The efficiency of the controlled transformer is greatly influenced by magnetic core material and winding configuration. In this section, design consideration for controlled transformer will be given and the selection of magnetic core material for controlled transformer will be discussed. Finally, a design example of controlled transformer will be given in section 3.4.

Turns-ratio , leakage and flux density of power transformer T_1

For the application of high frequency and high power conversion, design of transformer is very important to the efficiency and size of power converter. It's important to reduce the leakage inductance of the transformer. To reduce leakage inductance, the primary and the secondary should be placed closed to each other and also the number of turns should be as small as possible. However, the minimum number of turns is primarily determined by the maximum allowable flux density which determines the core losses. To grantee satisfactory operation of regulator and flexibility of manufacture, both the power core and the control core should be capable of taking all the input voltage-seconds. In practice, the volt-seconds from the primary is always shared to some extent by two core pairs, so the real flux change could be less than the calculated value. But under heavy load case, the volt-seconds must be taken by either core. This was verified by the experiments and explained in theory of operation of PWM-controlled transformer.

Core geometry and winding configuration of power transformer T_1

For high power application, U core or E core is often chosen due to their good heat dissipation capability. Toroidal core can also be used to reduce the leakage inductance of control winding but it is difficult to wind. To make low-leakage and small-eddy current transformer, it is necessary to keep the separation between the windings as small as possible. However, safety requirements limit the minimum separation. The flux coupling can be improved by using interleaved windings. The use of foil windings has a lower self inductance compared with a wire winding. This is especially true for a low number of turns of transformer. Copper losses in the transformer windings can be minimized by shortening the winding length, using

litz wire and/or foil, and placing the windings far from strong magnetic fields (gaps). Due to the skin effect, the diameter of wire or the foil thickness should not be much greater than twice the skin depth.

3.3.1 Desired Features of Magnetic Core Material for Control Transformer

The selection of core for T_c material affects the control range and the magnitude of control current, which indirectly affects efficiency. Several key features of magnetic core material T_c are discussed as follows:

Permeability

Fig. 3.11 shows two B-H loops of equal saturation level (B_s). Loop1 is with high permeability and loop2 is with lower permeability. In the following discussion, assume that both cores are wound with the same number of turns and have the same physical size. From the discussion in chapter 3 (specially in Fig. 3.3(a)), to achieve the same output voltage level, the same flux reset ($B_s - B_c$) is required for both loop1 and loop2. Furthermore, the flux density is clamped at B_c during the clamping period. However, the magnetic field H for loop1 and loop2 is quite different even for the same B_c , as shown in Fig. 3.11. For clamping current (current through the control winding) is $H \cdot l_{m2} / N_m$ where l_{m2} is the mean magnetic path length of control core and N_m is the number of turns of control winding. It's obvious from the above discussion, the core with larger permeability yields less clamping current and less control circuit loss. It's important to point out that the duration of the clamping period is a significant portion of the whole cycle, especially at heavy load.

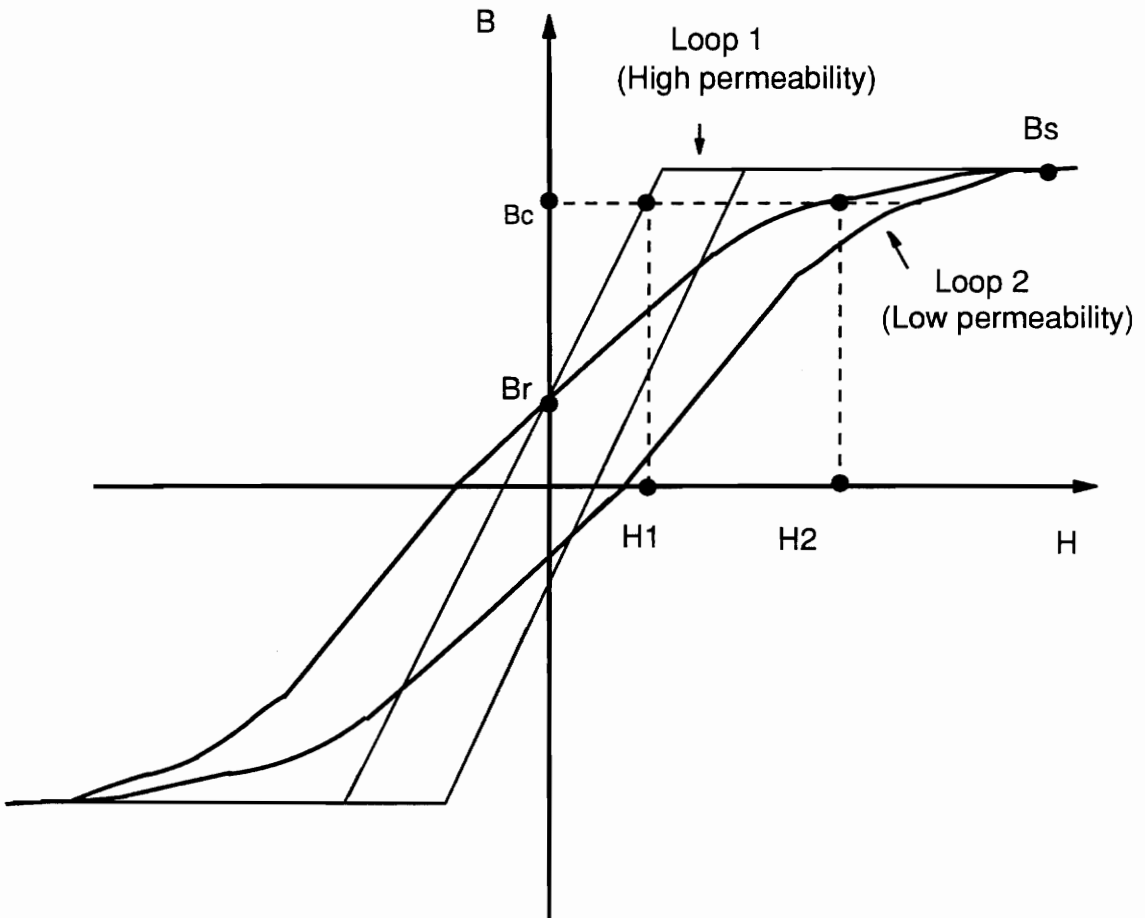


Fig. 3.11 High permeability and low permeability B-H loops.

Squareness

Squareness of a B-H loop is defined as the ratio of B_r/B_s . Fig. 3.12 shows an example. For a conventional magamp regulator, high squareness of core loop is desirable [9]. In the case of control transformer regulation, it's just the opposite and the reason is explained in the following.

Figure 3.13 and 3.14 shows the operating loops of the control core and power core under different load conditions. For heaviest load condition, the clamping period is longest and the blocking period is shortest and the control flux change is smallest. For medium load and lightest load, the operation of B-H loop is shown in Fig. 3.13(b) and (c). Depending on design, when the load is very light, the control core may not be saturated. Since the control core flux is allowed only in the first quadrant as discussed in chapter 3 (especially in Fig. 3.3), the maximum allowed flux variation is $B_s - B_r$. Therefore, to have a large load control range (still maintain regulation), a core with low squareness should be selected.

Sharpness

Fig. 3.15 shows the high sharpness and low sharpness B-H loops of control cores. It can be seen that the magnetizing current of low sharpness control core is larger than that of high sharpness control core for the same flux density. When the reset flux density B_c is high, the difference between the magnetizing current of the control core using the high sharpness and low sharpness core material is significant. It is noticed that B stays in B_c for the entire clamping period. Therefore, the difference of the control current and the control circuit loss is significant. Especially, at heavy load B_c is closer to B_s and the difference is even higher.

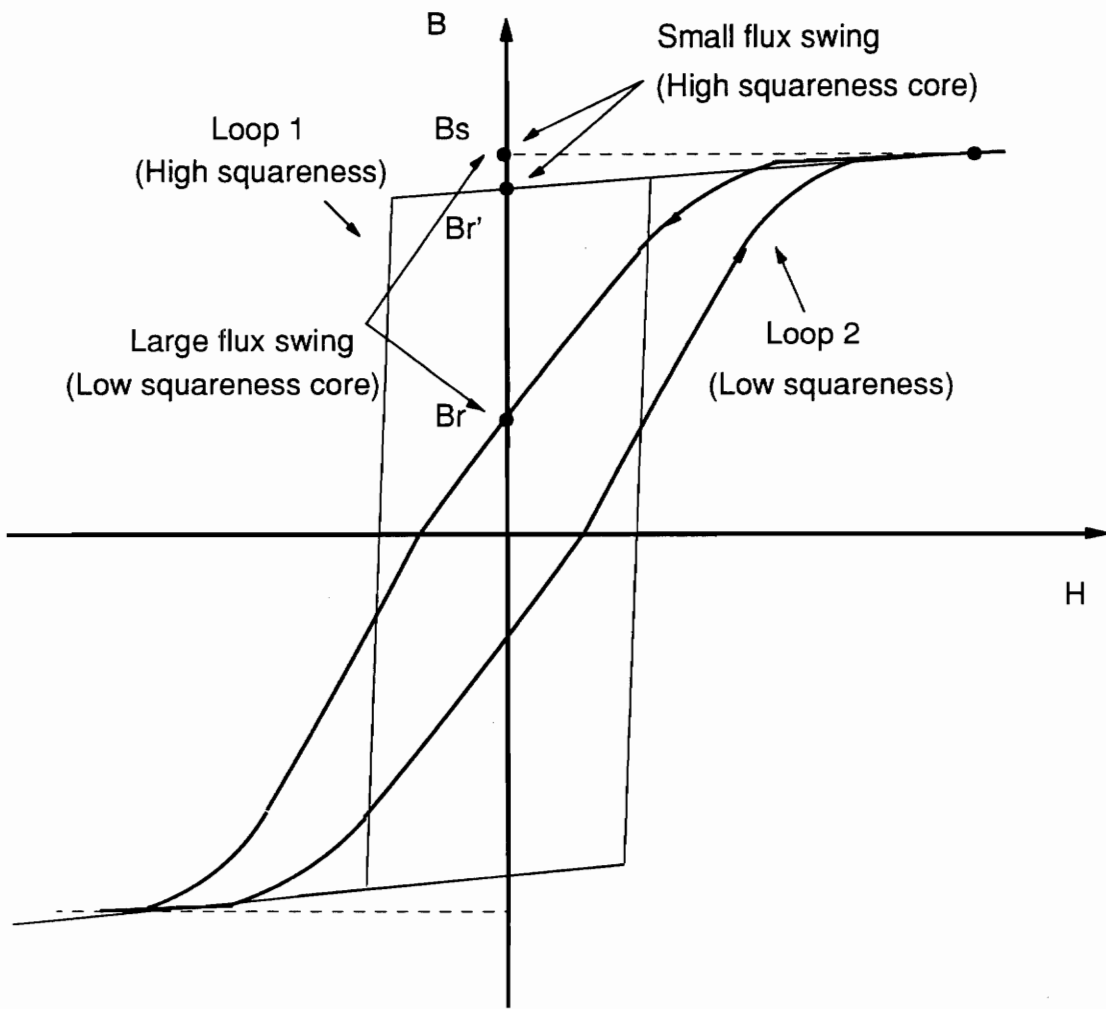
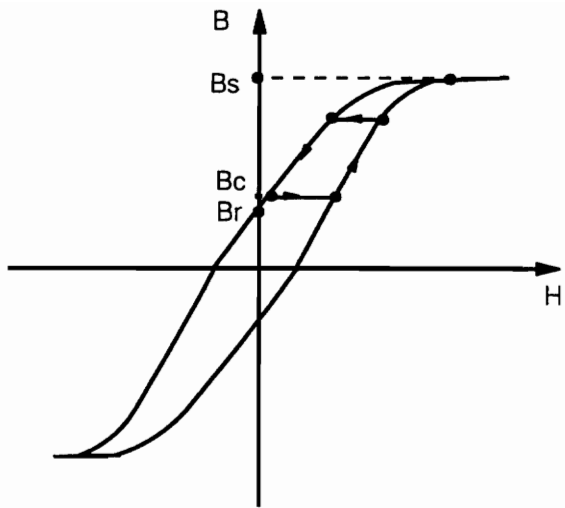
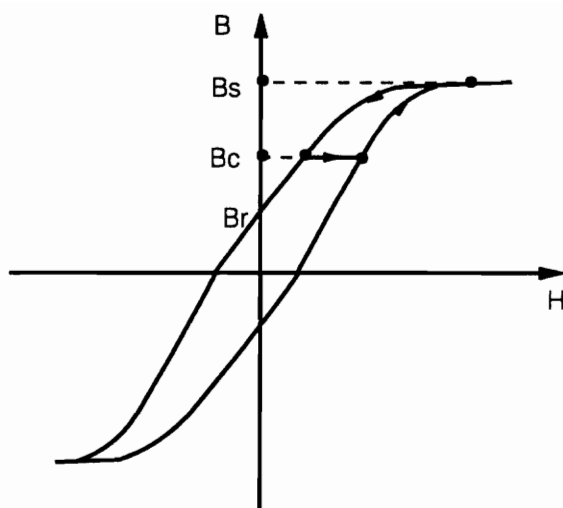


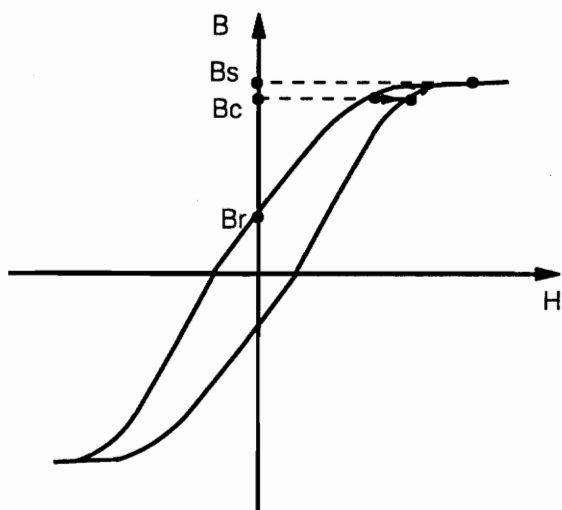
Fig. 3.12 High squareness and low squareness B-H loops.



(a) Light load

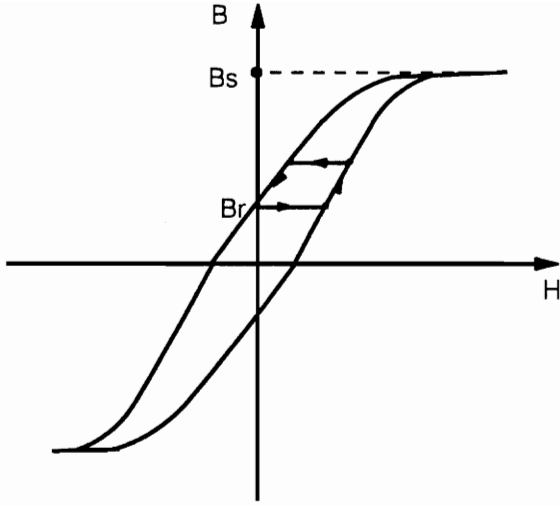


(b) Medium load

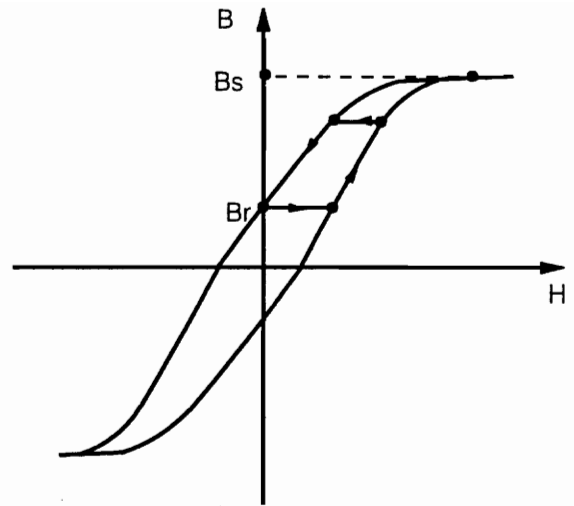


(c) Heavy load

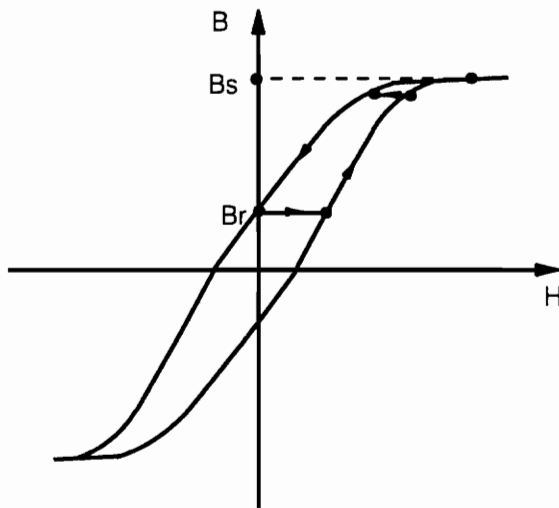
Fig. 3.13 B-H loops of the control core T_c under three operating conditions.



(a) Light load



(b) Medium load



(c) Heavy load

Fig. 3.14 B-H loops of the power core T1 under three operating conditions.

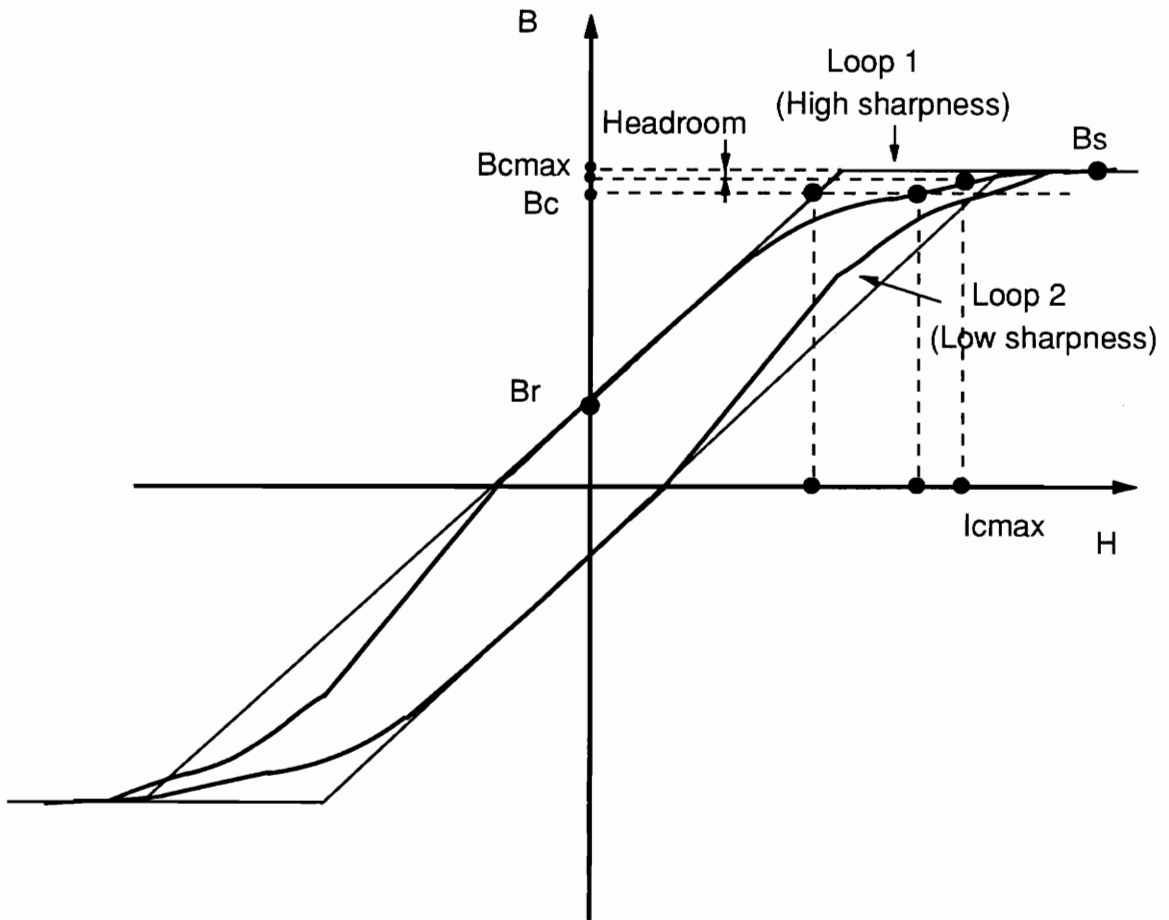


Fig. 3.15 High sharpness and low sharpness B-H loops

As we discussed before, the larger load, the larger the reset flux density B_c and the worse the low sharpness control core. Therefore, to reduce control current and improve conversion efficiency, the high sharpness core is desirable.

Core loss at high frequency and high temperature

Soft ferrite core has smaller core loss than square B-H loop core at high frequency and high temperature. Therefore, soft ferrite is better suite for high frequency and high power conversion.

Summary

Ideal magnetic material of control core has the following features:

- 1) High permeability
- 2) Low core loss at high frequency and high flux density and high temperature
- 3) Low remnant flux and high saturation flux and low squareness
- 4) High sharpness

From the above discussion, Soft ferrite cores fit the desired characteristics.

Power core

The requirement of power core of controlled transformer is identical to that of transformer in the conventional forward converter. Magnetic material of power core should have low core loss and high permeability at high frequency, high flux density and temperature to increase efficiency of regulator. To reduce the size of transformer, the saturation flux density should be high. Core material of power core and control core is listed in section 4.1.

3.4 Design Example of Controlled Transformer

An example is given in this section to illustrate the procedure for designing a controlled-transformer for a forward converter regulator. The converter specification is given as follows:

$$f_s=200\text{khz} \quad V_o=5.0\text{v} \quad P_o=500\text{w}$$

$$V_i=240.0\text{v} \quad F_w<0.4$$

3.4.1 Design of Power Transformer

(1) Determine turns ratio N_{p1}/N_2

The turns ratio of forward converter transformer can be expressed as follows:

$$\frac{N_{p1}}{N_2} = \left(\frac{V_i}{V_o'} \right) \times D_2$$

Where N_{p1} is the number of turns of the primary winding, N_2 is the number of the turns of the secondary winding. D_2 is the duty cycle of the secondary voltage. V_i is the input voltage.

$$V_o' = V_o + \Delta V_o$$

Where V_o is the output voltage. V_o' is the average value of the secondary voltage. ΔV_o is the voltage drop on the main switches, rectifiers and leakage inductance. Due to the large output current, $\Delta V_o = 2.5\text{v}$ is chosen in this converter design to keep the output voltage constant at full load range.

Two design requirements of controlled transformer

Maximum duty cycle of the secondary voltage and the maximum allowed control current.

For heaviest load, $D_{2\max} = D_1 - \Delta D$

D_1 is the duty cycle of the main switches and ΔD is defined as the headroom or duty cycle loss. This duty cycle loss is caused by the difference between the maximum reset flux density B_{cmax} and saturation flux density B_s . B_{cmax} is limited by the maximum allowed control current $I_{\text{con(max)}}$. The duty cycle loss of controlled transformer is shown in Fig. 3. 15. The headroom ΔD can be calculated as follows:

$$\Delta D = N_{p2} \frac{A_e \Delta B_{hr}}{V_i T_s}$$

Where N_{p2} is the number of turns of the primary control winding. A_e is the cross section area of the control core. $\Delta B_{hr} = B_s - B_{\text{cmax}}$

Assume the maximum control current is less than 10A, the maximum duty cycle $D_{2\max}$ is not less than 35%.

Then the following two design inequalities must be satisfied:

$$0 \leq I_{\text{con(max)}} \leq 10\text{A}$$

$$35\% \leq D_{2\max} \leq 50\%$$

Take $D_2=31\%$ in this design, then:

$$\frac{N_{p1}}{N_2} = D_2 \left(\frac{V_i}{V_o'} \right) = 0.31 \left(\frac{240}{7.5} \right)$$

Take $\frac{N_{p1}}{N_2} = 10$

Note: it is necessary to check the above two inequalities after the design of controlled

transformer is completed.

(2) Determine wire size for the primary and the secondary windings

The wire size of the primary and the secondary windings are chosen by the maximum current (rms value) flowing in two windings.

$$I_{pri(rms)} = \sqrt{D_2} \frac{I_o}{10} = 5.57A$$

175/44 Litz wire 2 strands

$$I_{sec(rms)} = \sqrt{D_2} I_o = 55.68A$$

5 mil thick copper foil 4 strands

(3) Calculate $A_e A_w$ requirement

$$D_{1max} = 0.95 \times 0.5 = 0.475$$

$$\Delta B = V_i \frac{(D_{1max} T_s)}{N_{pl} A_e} \times 10^8$$

$$A_e = \frac{32.79}{N_{pl}}$$

To avoid excessive core loss, the flux density is chosen as:

$$\Delta B = 1840 \text{Gaus}$$

(4) Select a core with $A_e A_w \geq 1.2 \text{cm}^2$ (requirement)

U-U core Magnetics J material is used

$$A_e = 0.78 \text{cm}^2$$

$$A_w = 1.68 \text{cm}^2$$

$$N_{pl} = \frac{254 \times 0.475 \times 5 \times 10^2}{1840 \times 0.78} = 39T$$

$$N_2 = \frac{N_{pl}}{10} = 4T$$

(5) Check $F_w \leq 0.4$

$$F_w = 0.33 < 0.4$$

3.4.2 Design of Control Transformer Tc

(1) Volt-seconds of control primary winding

The requirement of volt-seconds of control core depends on application. If output shutdown by T_c is required, then the control core must be designed to block all of the input volt-seconds. In this power converter, the control core is required to take all of input volt-seconds.

$$\text{Volt-seconds of control core} = 240 \times 2.375 \times 10^{-6} = 603 \times 10^{-6} (\text{v.s})$$

(2) Determine turns-ratio N_{p2}/N_m

The control core and winding in this circuit is required to support all the input volt-seconds. To simplify the design of controlled transformer, the number of the turns of the primary control winding could be the same or greater the number of the turns of primary winding of power transformer if the cross section area of control core is chosen close to that of power core. Turns-ratio of control winding is determined by the voltage on the control secondary winding for safety reason and associated with current in the control circuit. If turns-ratio is large, the control current is smaller, however, the voltage on the control secondary winding

is high, which is not desirable to safety. If turns-ratio is small, the voltage on the control secondary winding is small but the control current is larger and the larger loss in the control circuit. Also, the transient response of the control loop is influenced by the number of the turns of the control secondary winding. The larger the number of the turns of control secondary winding, the slower the transient response. For these reasons, turns-ratio of control core is chosen as the same as that of power core and the number of the turns of the control secondary is a little bit large than that of the power core.

$$\text{Take } \frac{N_{p2}}{N_m} = \frac{N_{p1}}{N_2} = \frac{39}{4} = 9.75$$

$$\text{Choose } N_{p2} = 57$$

$$N_m = 6$$

$$\frac{N_{p2}}{N_m} = 9.65$$

(3) Determine wire size for control winding

The primary current of the control primary winding is the same as that of the primary winding of power transformer, therefore, the wire size of the control primary winding could be the same as that of power transformer.

$$N_{p2}=57 \quad 175/44 \text{ Litz wire } 2 \text{ strands}$$

According to estimation of control current in section 3.5, the following wire is chosen for control secondary winding:

$$N_m=6 \quad 1 \text{ mil copper foil}$$

(4) Select control core $A_e A_w \geq 1.2 \text{ cm}^2$ (requirement)

$$\Delta B = \frac{603 \times 10^{-6}}{N_{p2} A_e} \times 10^8 \leq 2000G$$

$$A_e \geq \frac{603 \times 10^2}{1840} \times 57 = 0.575cm^2$$

Take $A_e = 0.78cm^2$

TDK H5B T28-16-13 is used as control core.

(5) check $F_w < 0.4$

$$F_w = 0.2 < 0.4$$

(6) Check two design inequalities for controlled transformer

Assume $I_{con(max)} = 10A$

$$H_{max} = \frac{0.4\pi I_{con(max)} N_m}{l_{m2}} = \frac{0.4\pi \times 10 \times 6}{13.8} = 5.5Oersted$$

$$B_{cmax} = \mu_2 H_{max} = 700 \times 5.5 = 3850G$$

$$\Delta B_{hr} = B_s - B_{cmax} = 4200 - 3850 = 350G$$

$$\Delta D_{bl} = 57 \times 0.78 \times \frac{(350)}{240 \times 5 \times 10^{-6}} \times 10^{-8} = 12.5\%$$

$$D_{2max} = D_{1max} - \Delta D_{bl} = 0.5 \times 0.95 - 0.125 = 35\%$$

Therefore, two design inequalities for controlled transformer are satisfied and the design of controlled transformer is successfully completed.

3.5 Analysis of Control Current

The control current is important to the efficiency of converter. The smaller control current, the smaller control power and the higher the efficiency. In this section, the factors that determine control current are identified and an expression for the control current is derived for the PWM-controlled transformer.

3.5.1 Calculation of control current of PWM-controlled transformer

To calculate the control current, it is necessary to analyze the MMF of a controlled transformer before and after the main switches turn off and the reset process of controlled transformer. When main switches are turned off and S3 is still open, no control current exists in the control circuit. When S3 is turned on, control current starts flowing in the control circuit during clamp period (t3-t4). Fig. 3.16 shows current waveforms related to control current by PWM control. To simplify the analysis, assume no inductor in the control circuit. Refer to the equivalent circuit of PWM-controlled transformer during power transfer period (t1-t2) in Fig. 3.4, MMF for both cores during power transfer period is:

$$N_{p1}I_{pri} - N_2I_o = N_{p1}I_{m1} \quad \text{power core} \quad (3-4)$$

$$N_{p2}I_{pri} = N_{p2}I_{m2} \quad \text{control core} \quad (3-5)$$

Where I_{pri} is the primary current.

I_{m1} is the magnetizing current of T_1 .

I_{m2} is the magnetizing current of T_c .

The magnetizing current of the control core is the primary current, which could be very large under heavy load condition. After main switch Q1 and Q2 are turned off, the secondary

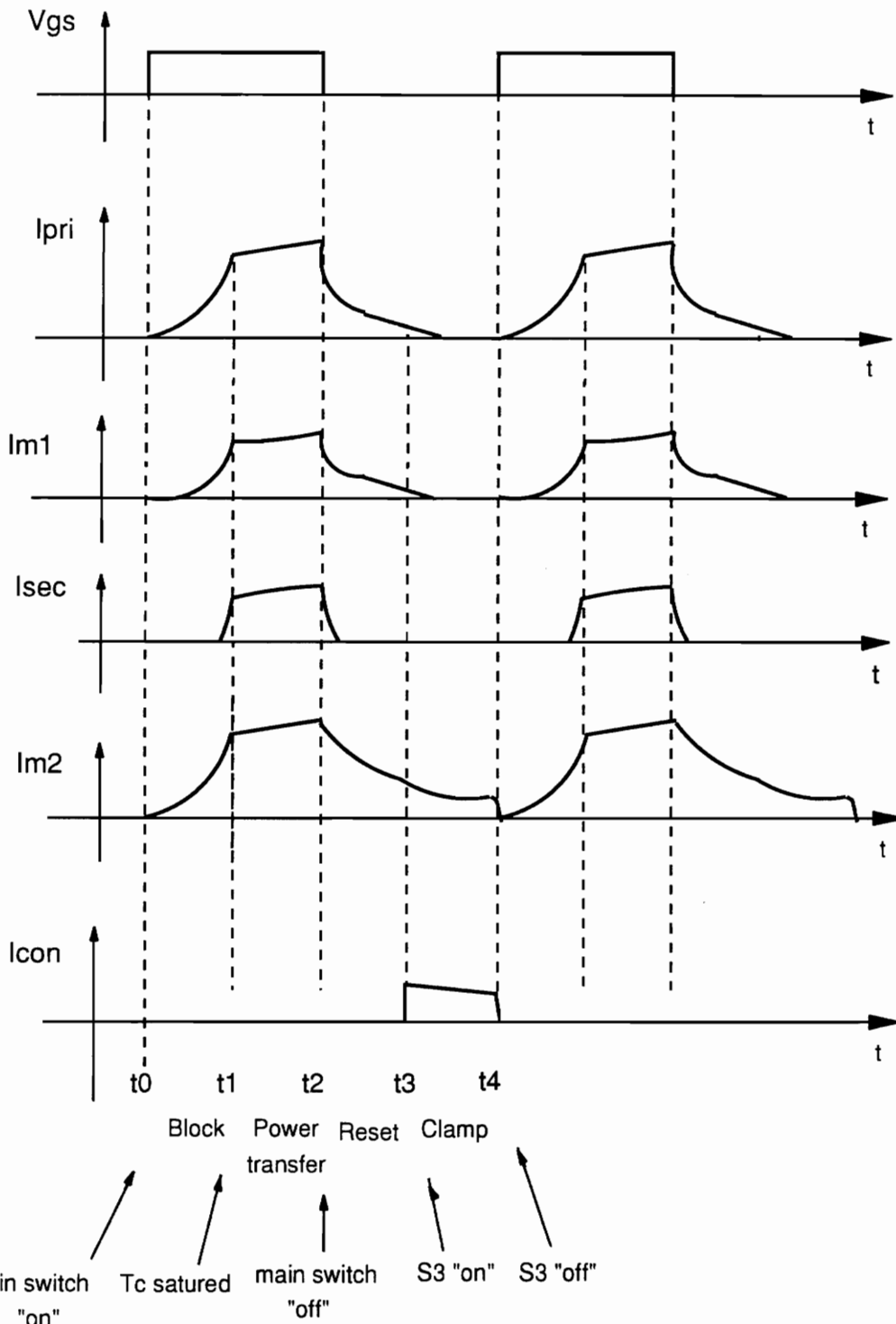


Fig. 3.16 Current waveforms related to control current for PWM control.

current in the DR3 drops fast to zero and the current in the DR4 keeps increasing until it reaches the output current. Switch S3 is not yet turned on and the control current is still zero. Equivalent circuit during this period t2-t3 is shown in Fig. 3.5.

MMF for both cores during reset period (t2-t3) is:

$$N_{p1}I_{pri} = N_{p1}I_{m1} \quad \text{power core} \quad (3-6)$$

$$N_{p2}I_{pri} = N_{p2}I_{m2} \quad \text{control core} \quad (3-7)$$

Combining eq. (3-6) and eq. (3-7), then:

$$I_{pri} = I_{m1} = I_{m2} \quad (3-8)$$

The primary current decreases quickly and become the magnetizing current of the power core after the secondary current I_{sec} is zero. The magnetizing current of the control core is the magnetizing current of the power core, which is small in normal power transformer design. Current waveforms are shown in Fig. 3.16. During t2-t3 period, both the secondary circuit and the control circuit are open and the primary current is magnetizing current of the power core and control core. Therefore, after main switches are turned off, the magnetizing current of the control core is only maintained by the primary current and no current in the control circuit since MOSFET S3 is still open during t2-t3. The primary current has to return to the source and decrease quickly. Therefore, most of energy in the cores is sent back to the source rather than dissipated in the control circuit during the reset period of controlled transformer. Flux of T_c goes down from point g to point e, as shown in Fig. 3.17. At t3 instant, S3 is turned on and the control winding is shorted, so the flux density of the control core is clamped at point e. There is a current circulating in the control winding to keep flux at point e through clamp period t3-t4 period until next switching cycle starts.

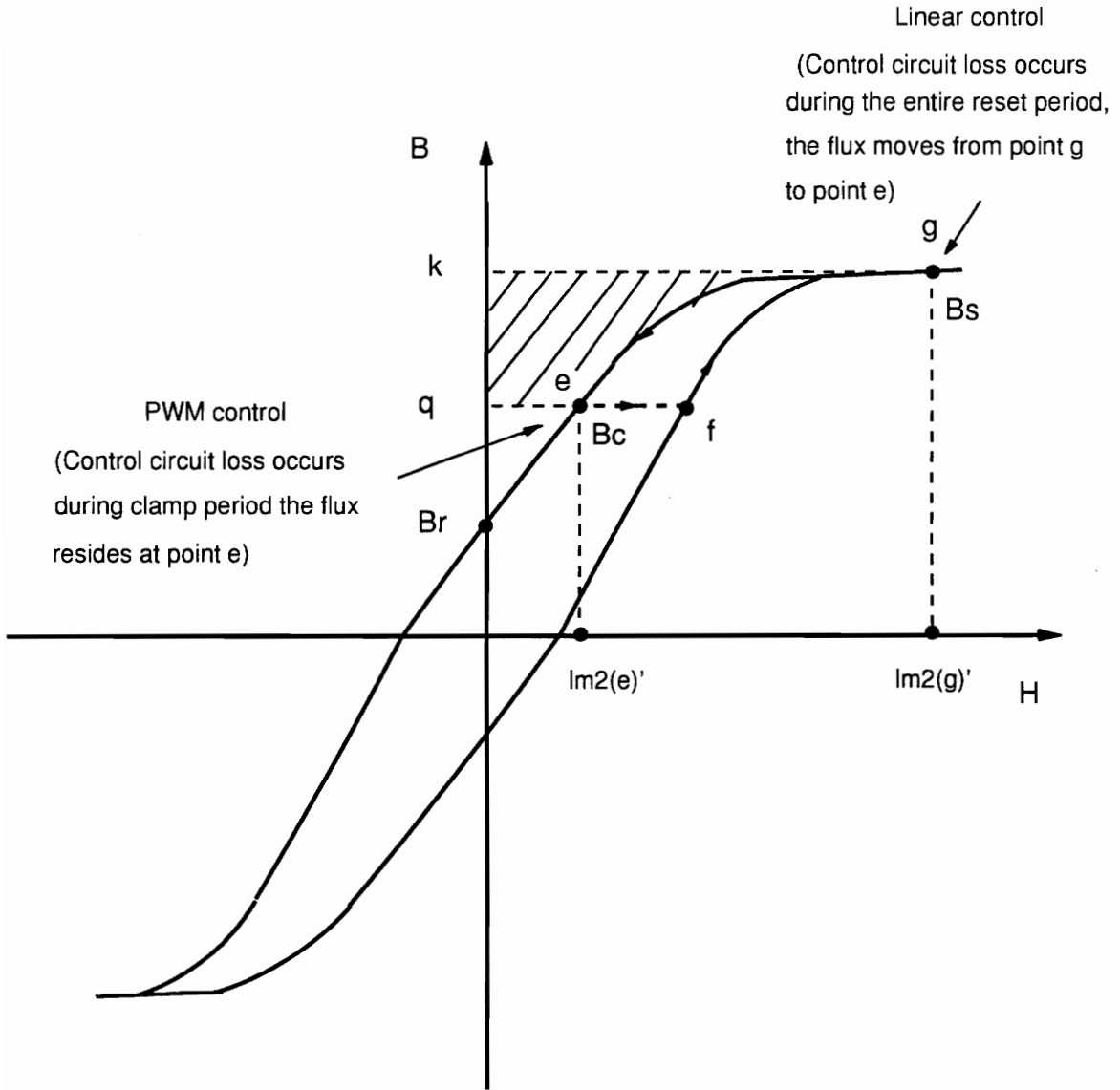


Fig. 3.17 Magnetizing current of T_c for PWM and linear control

Equivalent circuit during clamp period t3-t4 period is shown in Fig. 3.6.

MMF for both cores during t3-t4 is:

$$N_{p1}I_{pri} = N_{p1}I_{m1} \quad \text{power core} \quad (3-9)$$

$$N_{p2}I_{pri} + N_m I_{con} = N_m I_{m2}' \quad \text{control core} \quad (3-10)$$

$$I_{con} = I_{m2}' - \frac{N_{p2}}{N_m} I_{pri} \quad (3-11)$$

Where I_{con} is the control current in the control secondary winding.

I_{m2}' is the magnetizing current of T_c reflected to the secondary side.

The primary current drops very fast after the main switches are turned off and becomes nearly zero after reset period t2-t3. Therefore, if this primary current is ignored, control current at this point can be approximately expressed as:

$$I_{con} = I_{m2(e)'} \quad (3-12)$$

Where $I_{m2(e)'}$ is the magnetizing current of control core in the control secondary winding corresponding to the flux density at point e shown in Fig. 3.15.

To derive an expression for the control current, the Farady's law and Ampere's law are used.

The output voltage can be expressed as:

$$V_o = \left(\frac{N_2}{N_{p1}} V_i - I_o (R_{on}' + R_w) - V_{kp} \right) D_2 \quad (3-13)$$

The duty cycle of the secondary voltage is:

$$D_2 = \frac{V_o}{\left(\frac{N_2}{N_{p1}} V_i - I_o (R_{on}' + R_w) - V_{kp} \right)} \quad (3-14)$$

Where R_{on}' is the total equivalent on-resistance of the main switches reflected to the secondary side.

R_w is the equivalent resistance of the windings of transformer and inductor and circuit wires.

V_{kp} is the voltage drop on the leakage inductance of transformer and parasitic resistance and inductance, which is the function of load current. The larger load current, the larger this voltage drop.

The blocking time of T_c is:

$$T_b = (D_1 - D_2)T_s \quad (3-15)$$

The voltage on the control winding during blocking period is:

$$V_{con} = \frac{N_m}{N_{p2}} V_i \quad (3-16)$$

According to Farady's law, the voltage V_{con} also can be expressed as:

$$V_{con} = \frac{N_m \Delta B A_e}{T_b} \quad (3-17)$$

Where V_{con} is the voltage on the control secondary winding.

The flux density swing in the control core is:

$$\Delta B = \frac{T_b}{N_m A_e} V_{con} \quad (3-18)$$

Substituting (3-16) and (3-17) into (3-18), then:

$$\Delta B = \frac{(D_1 - D_2)T_s}{A_e N_{p2}} V_i \quad (3-19)$$

The magnetizing current of the control core during the reset time is given by ampere's law:

$$Hl_{m2} = N_m I_{m2}' \quad (3-20)$$

$$H = \frac{B}{\mu_2} \quad (3-21)$$

The magnetizing current in the control secondary winding side is:

$$I_{m2}' = \frac{B l_{m2}}{N_m \mu_2} \quad (3-22)$$

The flux density can be expressed as:

$$B = B_s - \Delta B \quad (3-24)$$

Then the magnetizing current in the control secondary winding side is:

$$I_{m2}' = \frac{l_{m2}(B_s - \Delta B)}{N_m \mu_2} \quad (3-25)$$

Substituting eq. (3-25) into (3-12), control current I_{con} is:

$$I_{con} = \frac{l_{m2}(B_s - \Delta B)}{N_m \mu_2} \quad (3-26)$$

Substituting (3-26) and (3-19), Control current I_{con} is:

$$I_{con} = \frac{l_{m2}}{\mu_2 N_m} \left(B_s - (D_1 - D_2) T_s \frac{V_i}{N_{p2} A_e} \right) \quad (3-27)$$

Substituting (3-14) into (3-27), therefore, leads to the general expression of the control current.

General expression of the control current is:

$$I_{con} = \frac{l_{m2}}{N_m \mu_2} \left(B_s - \frac{D_1 T_s V_i}{N_{p2} A_e} + \frac{V_i T_s}{N_{p2} A_e} \cdot \frac{V_o}{\left(\frac{N_2}{N_{p1}} V_i - I_o (R'_{on} + R_w) - V_{kp} \right)} \right) \quad (3-28)$$

It can be seen from eq. (3-28) that the control current increases as the output current increase. The reason is that the voltage drop on R'_{on} and R_w and V_{kp} are significant and output voltage decreases when output current is large. To keep output voltage constant, the smaller blocking time is required and duty cycle D_2 need increase. The flux density of the control core has to be biased closer to saturation and the magnetizing current of the control core is larger, therefore, control current is larger. It also can noted from eq. (3-28) that the high permeability of control core, and shorter magnetic path and more turns of control winding can reduce the control current. The theoretical calculated and experimental results of control current are given in Appendix IV.

3.5.2 Explanation for Low Control Loss in a PWM Controlled-Transformer

Refer to Fig. 3.16 for the following explanation. During the reset period, control circuit of T_c incurs no power loss because S3 is still "off". Flux moves from point g to point e and the stored energy in T_c (the shaded area kgfe) is return to the input source, as shown in Fig. 3.17. Control circuit loss incurs only when S3 is turned on during clamp period. During this period the flux resides at point e and the current flows in the control circuit. However, because of low H are associated with point e, the control current is low and therefore, low control circuit loss occurs. In the case of linear control, however, control circuit loss occurs for the entire reset period when the main switches are turned off. The flux moves from point

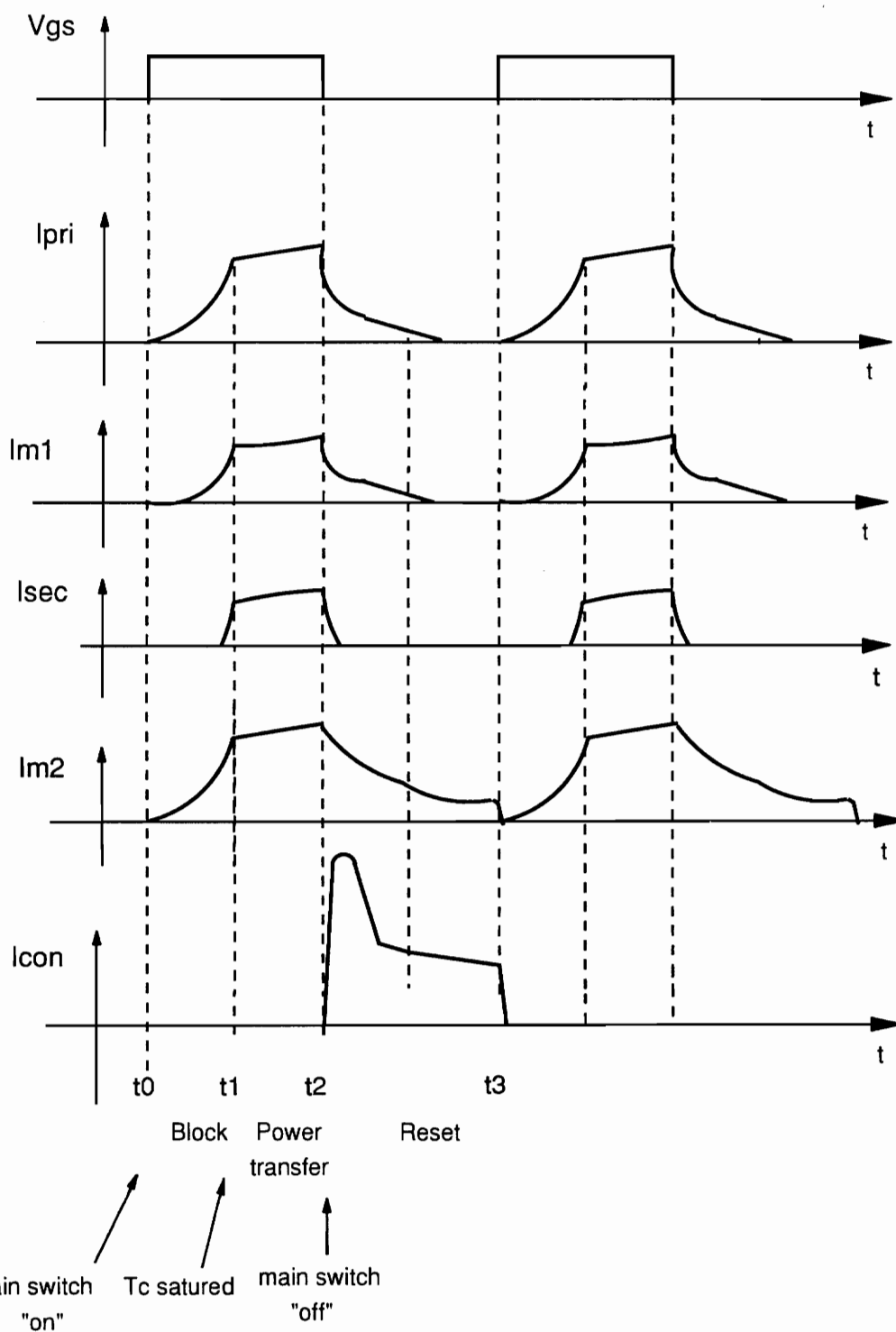


Fig. 3.18 Current waveforms related to control current for linear control.

g to point e. Relatively large current, corresponding to point g to point e, flows through the control circuit for the entire reset period. The energy stored in the T_c , as shown in the shaded area kgeq in the Fig. 3.17, is dissipated in the control circuit. This results in a large power loss in the control circuit. Fig. 3.18 shows the key waveforms for a linearly-controlled transformer.

3.6 Comparison of Controlled Transformer and Modified Magamp

From the previous discussion, the advantages and the disadvantage of controlled transformers compared to the modified magamps are summarized as follows:

Advantages:

- (a) Inexpensive two-piece soft ferrite cores can be used for controlled-transformer regulators, which reduces core cost and winding cost. This statement also applies to the modified magamps reported in [4, 5, 6] but not to the conventional magamps.
- (b) The design of the control portion (T_c) of controlled transformer can be standardized, which is a significant advantage. In other words, the same T_c design can be used for 5V or 12V or 15V output of reasonably wide range of output current. This is neither true for the conventional nor for the modified magamps. Because of large variation of the load current level and the saturable core volt-second blocking capability, it is not economical to standardize the saturable reactor for magamps. For the controlled transformer, however, the same volt-second blocking capability is required, regardless of the output voltage level. Therefore, reactor volt-second can be standardized. Also, in off-line step-down applications, the primary current is much reduced and it's not a

big sacrifice to use a standardized primary conducting wire size for different output current applications. The control winding current of T_c is also much reduced compared to the load current, and therefore the control circuit of T_c can also be standardized with economy.

- (c) Since the regulation of controlled transformer is done on the primary side rather the secondary side as the magamps do, there are advantages of avoiding the problems of the high-output-voltage and the high-output-current applications as described in the chapter I, such as large reactor size and requirement of isolation in the feedback circuit. Also, placement of a saturable reactor on the secondary side increases the physical loop area of schottky switched current. For high current application, this generates significant losses and EMI noises.

Disadvantage:

For multiple-output switching power supplies, each output requires one controlled transformer, which consists of two transformers (T_1 and T_c) that meet both the isolation and the safety rules. In a magamp regulator, it requires only one transformer and as many reactors as the number of the outputs. For high power applications, however, this is not really a disadvantage for the controlled transformer regulator, because it is often preferable to have several smaller transformers rather than a large one, anyway.

Chapter 4

EXPERIMENTAL VERIFICATION

To verify the theory and the design described in Chapter 3, a 200kHz, 500-watt, 5v-output regulator using a controlled transformer is built and tested. In this chapter, the experimental results obtained are presented. Results include the efficiency of the circuit, the estimation of power losses, and the control performances. The focus of effort is on the PWM-controlled transformer regulator but a linearly-controlled transformer regulator is also built for making a comparison.

4.1 Description of Converter Hardware

The converter was designed to meet the following specifications:

Input Voltage V_i : 240V-380VDC

Output Voltage V_o : 4.9V-5.1V regulated

Maximum output current I_o : 100.0A

Switching frequency: 200KHz

Fig. 4.1 shows the complete circuit diagram and Table 4.1 gives the component list. Fig. 4.2 shows the photograph of the circuit.

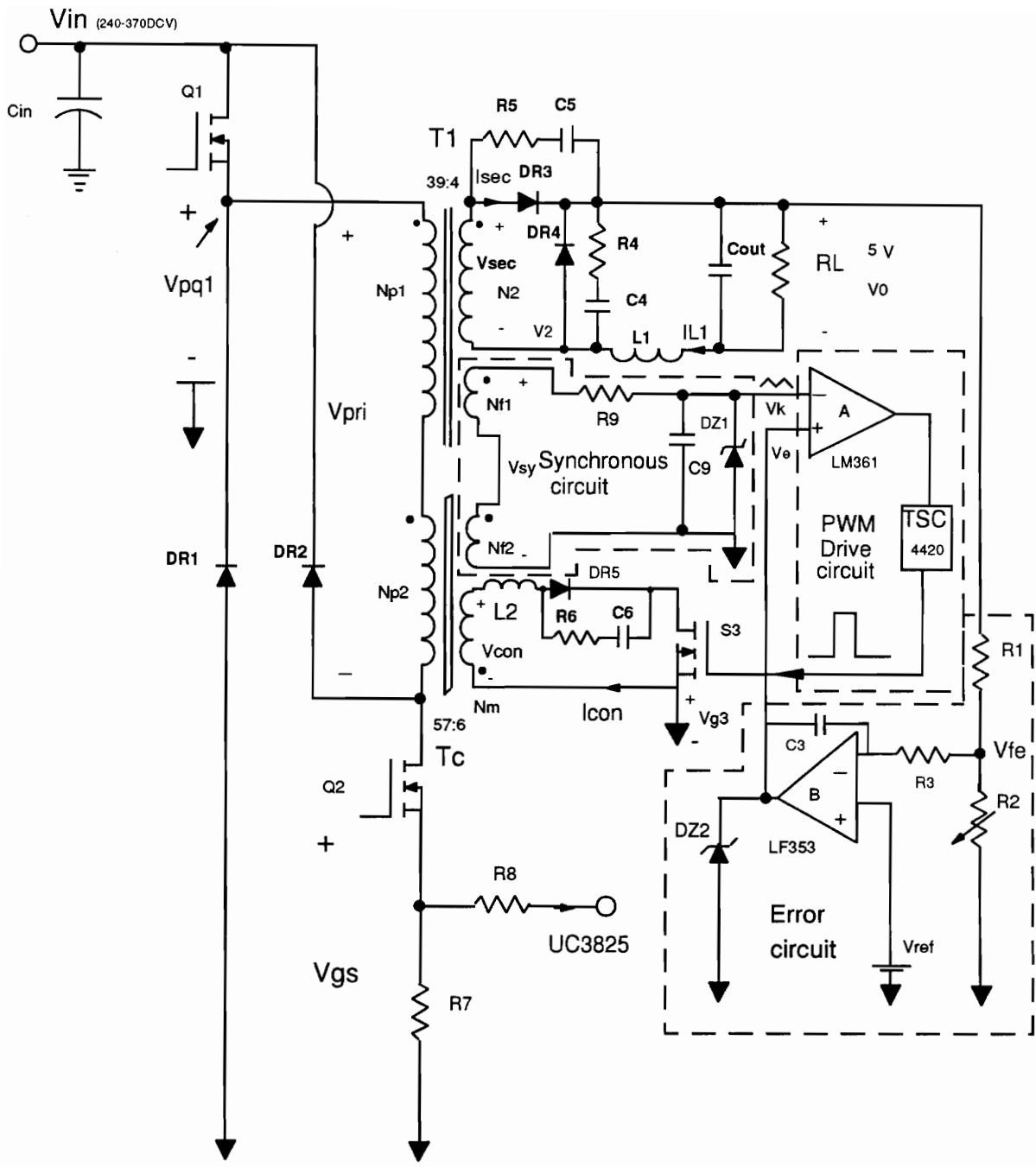


Fig. 4.1 Schematic of forward converter with controlled transformer using PWM control.

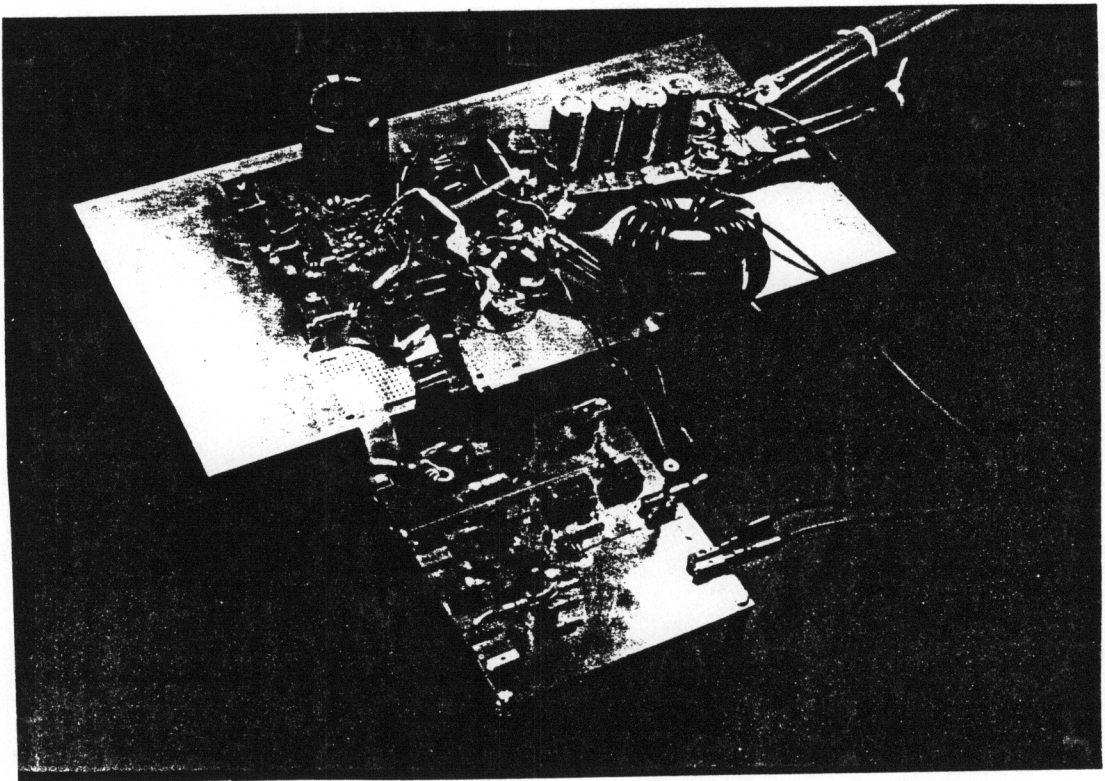


Fig. 4.2 Photograph of the a 200khz, 500w, 5v-output
forward converter using controlled transformer

Table 4.1 Component List

MOSFET switches

Q1 and Q2 - MOSFET IRFP450; S3 - MOSFET IRF540

Rectifier diodes

DR1 and DR2 - UHVP406; DR3 and DR4 - 200CNQ045; DR5 - 8TQ100

DZ1 and DZ2 - 6.2v Zener diode;

Resistors

R1 4.7kohm; R2 10kohm; R3 4.7kohm; R4 18ohm; R5 18ohm; R6 9.1ohm;

R7 .04ohm; R8 1.0kohm; R9 5kohm;

Capacitors

C9 1000pf; Cout 4000uf; C3 .033uf; C4 3300pf; C5 3300pf; C6 1000pf;

C_{in} 220uf (450v)

Inductors

Inductor L1 3.8uh (2/MPP55716); Inductor L2 200nh (MPP55280)

Primary control circuit

PWM signal modulator UC3825

Feedback control circuit

Comparator LM361; Error amplifier LF353; Driver TSC4420

Controlled-transformer(T₁ and T_c)

Table 4.2 summarizes the design parameters of both T₁ and T_c. Table 4.3 summarizes the material characteristics of the two magnetic material used in converter. Fig. 4.3 shows the winding construction of the transformer.

Table 4.2 Parameters of controlled transformer

Power transformer(T1)					
Material	Winding	Turns	Strands in parallel	Wire	Core
J Magnetics	Primary	39	2	175/44 Litz	U-U
	Secondary	4	4	5 mil copper	
Control core and winding(Tc)					
Material	Winding	Turns	Strands in parallel	Wire	Core
H5B TDK	Primary	57	2	175/44 Litz	Toroid T28-16-13
	Secondary	6	1	1 mil copper	

Table 4.3 Magnetic Material of Controlled Transformer

	magnetic materials	
	J	TDK H5B
initial permeability	5000	5000
saturation flux density (Bm) Gausess at 10 Oerteds	4300	4200
Coercive force(Hc)Oesteds	0.1Max	0.1
Relative loss factor	20Max 100khz	<40 (100khz)
Curie temperature (c)	140	>130
practical frequency range	<0.5MHZ	<0.3MHZ
Remanence B (G)	500	450

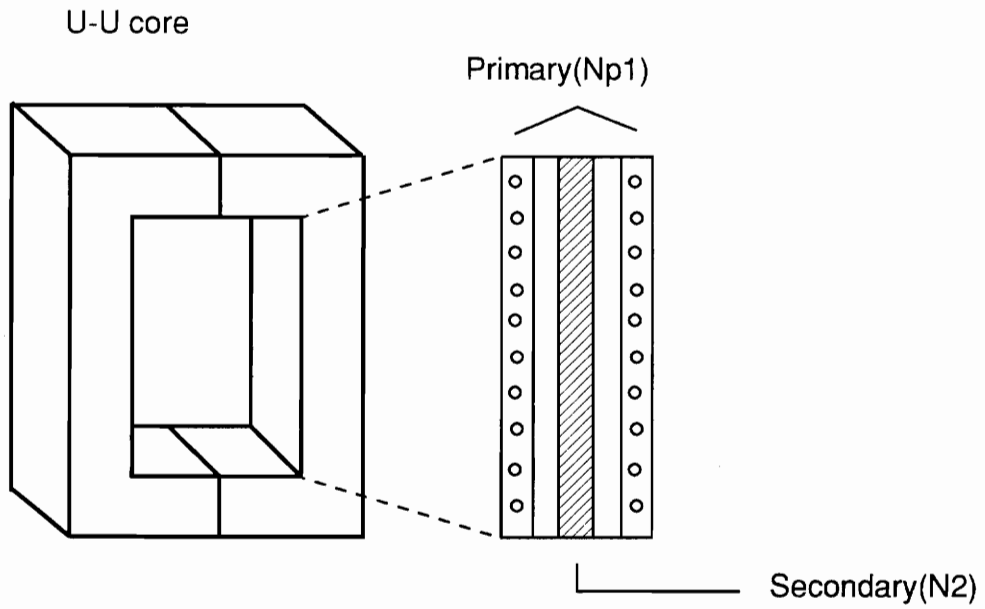


Fig. 4.3(a) Power core geometry and winding layout.

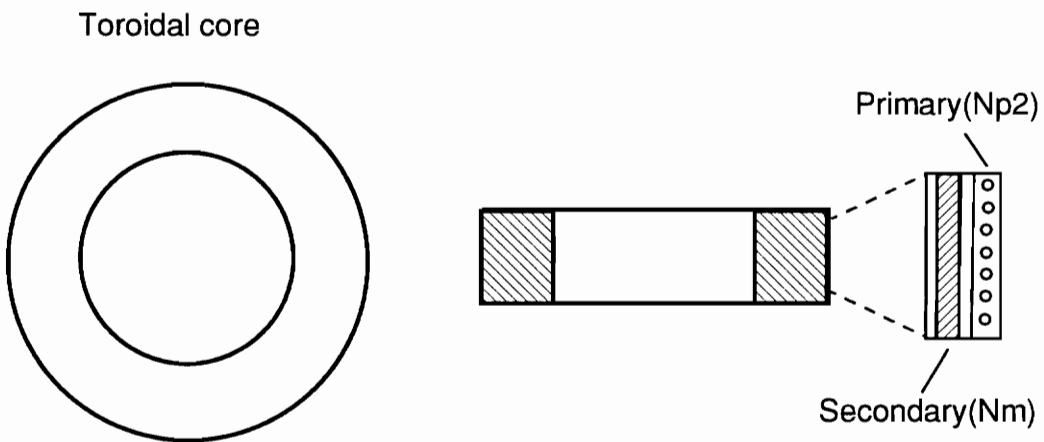


Fig. 4.3(b) Control core geometry and winding layout.

4.2 Experimental Waveforms

Figures 4.4, 4.5, and 4.6 show the circuit waveforms for the three different output current conditions of $I_o=4.0A$ (light load), $40.0A$ (medium load), and $100.0A$ (full load) when input voltage is $240v$. The definition of waveforms is explained as follows:

V_{gs} is the gate-to-source voltage of the main switches. The duty cycle of the main switches is fixed at 50%. V_2 is the voltage of the output inductor. V_{con} is the voltage of the control secondary winding. V_{gs3} is the gate-to-source voltage of S3 switch . The duty cycle of S3 is controlled by feedback circuit. V_{pql} is the switch Q1 source-to-ground voltage. I_{con} is the control current of the control secondary winding.

The voltage waveforms of the control winding and the secondary winding almost are square wave when the load is not very low. As shown in Fig. 4.4 and 4.5, the larger the output current, the smaller the blocking time T_b and the larger duty cycle D_2 . The output voltage is therefore regulated. Controlled transformer functions like a PWM modulator. It can also be seen that the current I_{con} in the control winding depends on the load current. The larger the load current, the larger the control current. Under very light load condition, as we explained in chapter 2, the voltage waveforms V_{con} and V_2 Fig. 4.6 are less like square wave. Since the primary current is too small to saturate the control core under very light load condition, the voltage on the primary side is divided between the power transformer primary winding and control primary winding and the power converter works in the discontinuous mode. Regulation of output voltage is achieved by varying inductance of the control winding in the discontinuous mode. Controlled transformer functions like variable inductor. Reg-

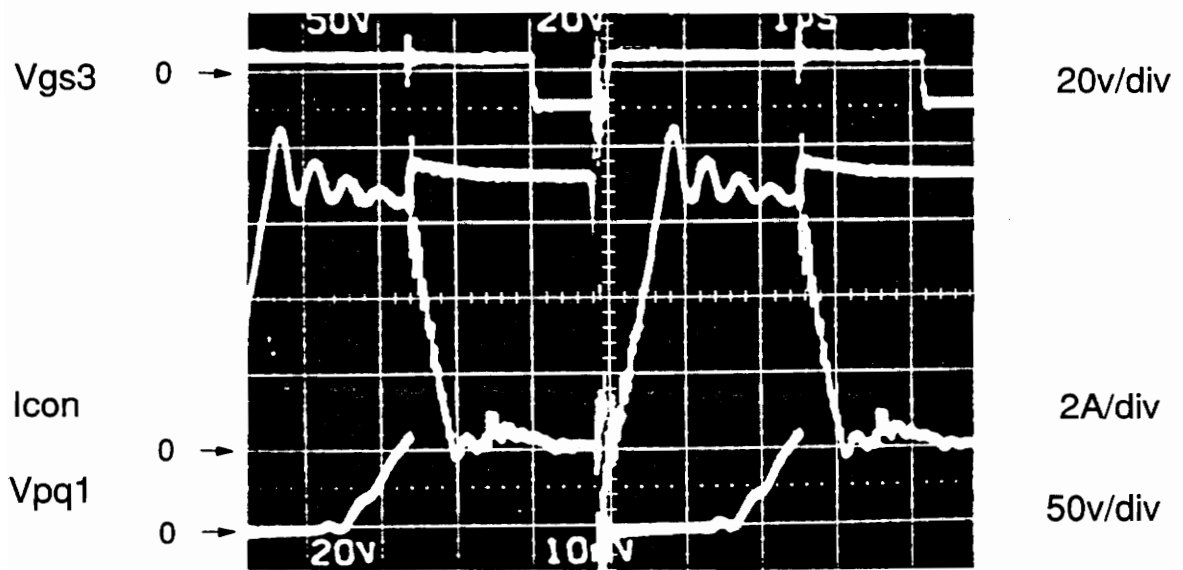
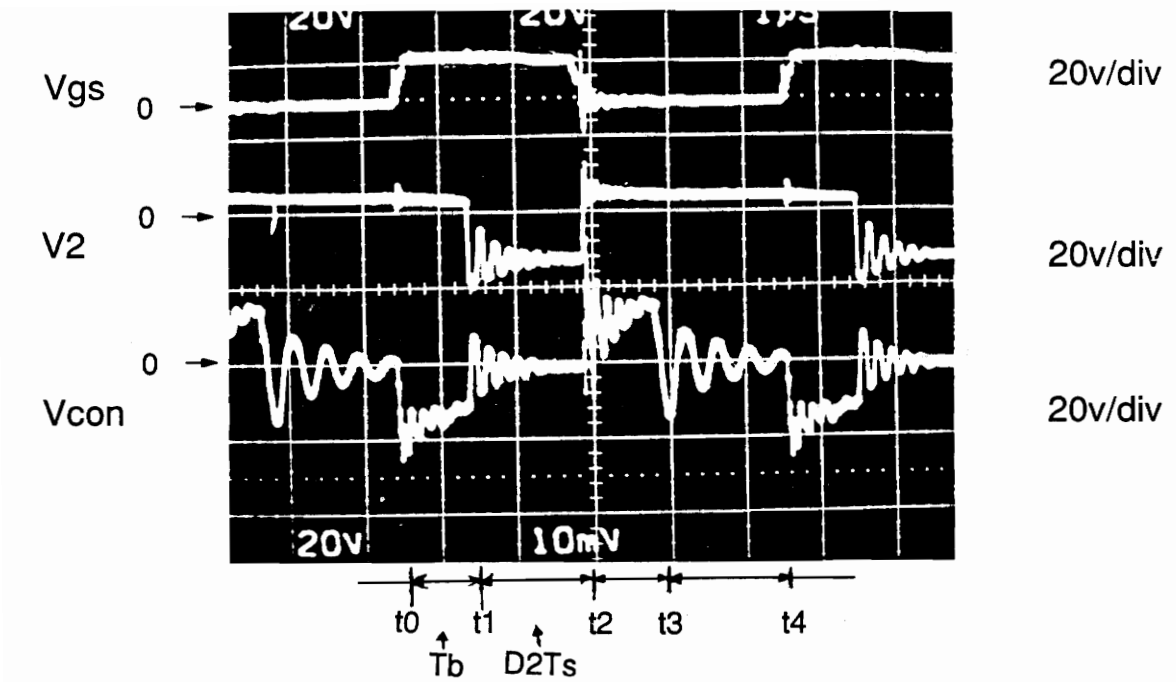


Fig. 4.4 Experimental waveforms of forward converter
 with PWM-controlled transformer
 Load Current $I_o=100A$ (continuous mode)

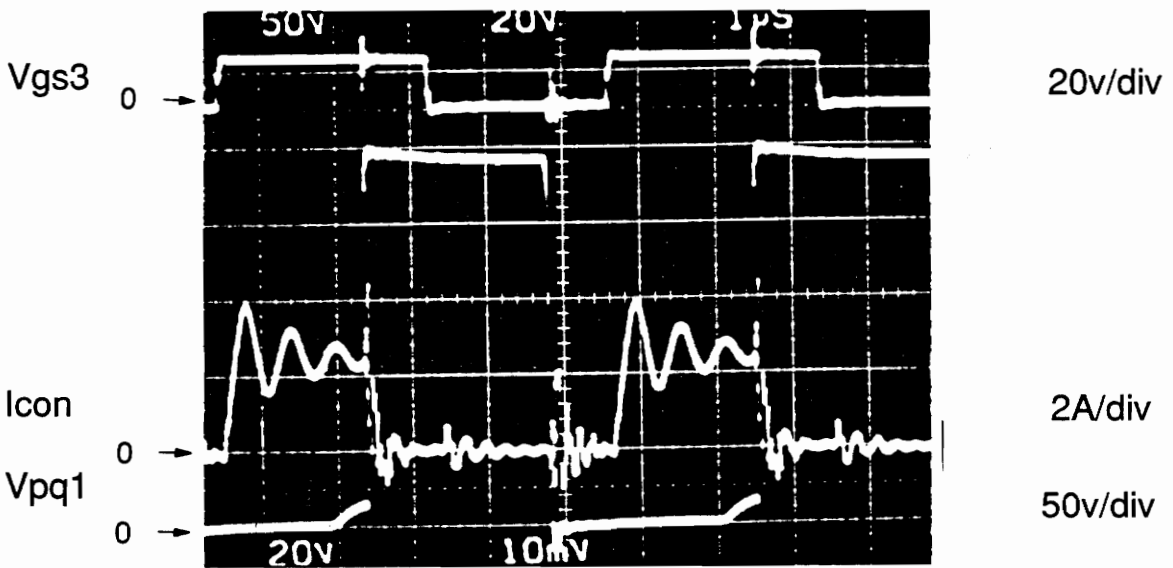
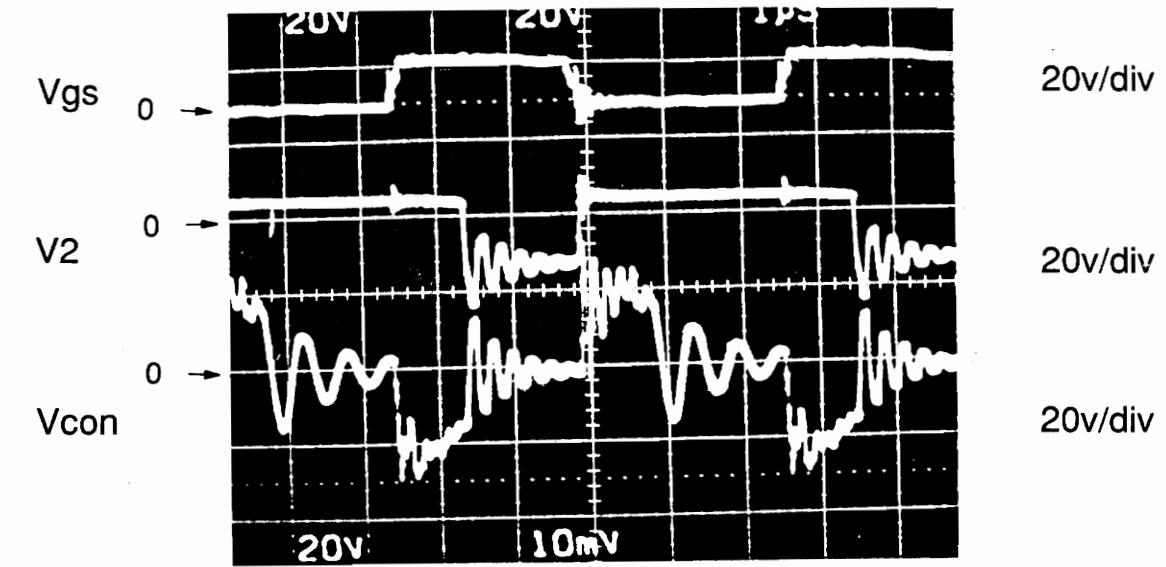


Fig. 4.5 Experimental waveforms of forward converter
 with PWM-controlled transformer
 Load Current $I_o=40A$ (continuous mode)

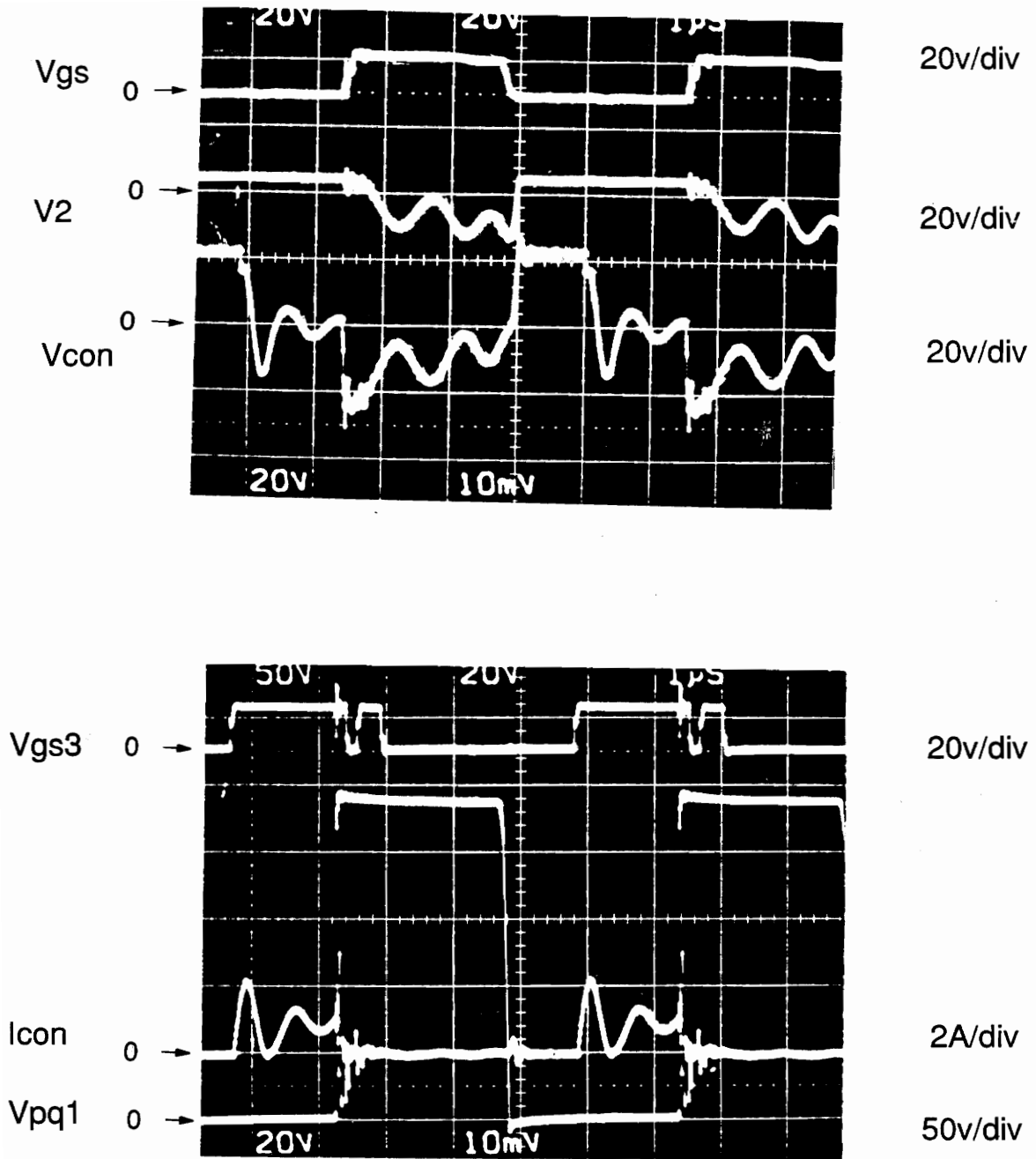


Fig. 4.6 Experimental waveforms of forward converter with PWM-controlled transformer
Load Current $I_o=4A$ (discontinuous mode).

ulation is still achieved in this mode. From gate voltage waveform of V_{gs3} and control winding voltage V_{con} , it was observed that the turn-on time of S3 is varied with the load. The larger the load current; the earlier the S3 turns on and the larger the clamp time T_{cl} ; the smaller blocking time T_b . This leads to a larger duty cycle of the secondary winding voltage and therefore the output voltage is kept constant. Under the heavy load condition, V_{pql} starts to rise before Q1 and Q2 switches are turned on. This is because the large voltage drop on the two MOSFET switches when the output current is large. If the smaller on-resistance MOSFET is chosen, this voltage V_{pql} rise is much smaller. The ringing of V_{con} is caused by a resonance of inductor L2, the magnetizing inductance of the control core and MOSFET drain-to-source capacitance. The ringing on the secondary side of the transformer is caused by the resonance of power core magnetizing inductance with the rectifier capacitance. A small inductor L2 can be used in series with DR5 to reduce ringing frequency and ringing amplitude of I_{con} .

4.3 Conversion Efficiency

Figure 4.7(a) shows the overall efficiency of the converter when the input voltage is 240v. Figure 4.6(b) shows the overall efficiency of the converter when the input voltage is 280.0v. The figure shows that maximum efficiency of 87% can be achieved at $V_i=280v$ and $I_o=60A$. The maximum efficiency obtainable for the linearly controlled transformer regulator is about 82.5%. Generally speaking, a 4%-5% improvement can be achieved by PWM control scheme.

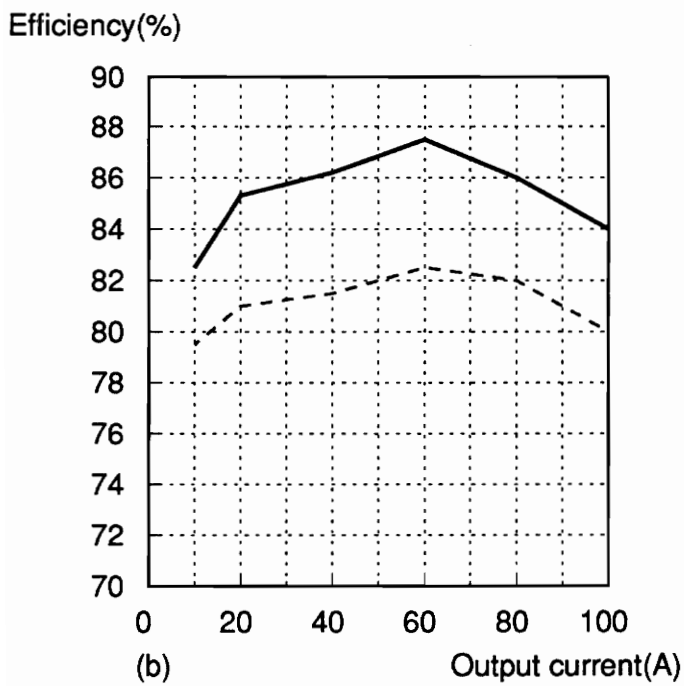
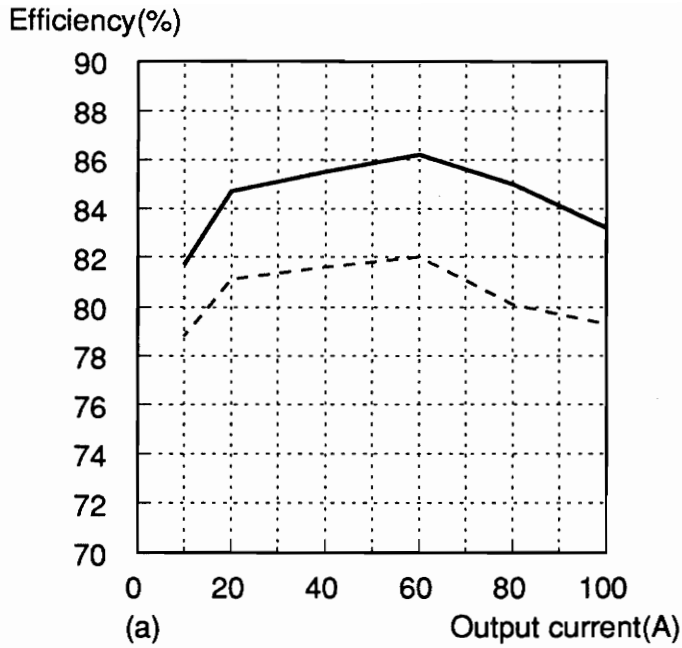


Fig. 4.7 Efficiency vs. output current for both PWM and linearly-controlled regulator.

— PWM control - - - Linear control

(a) $V_i=240v$, (b) $V_i=280v$

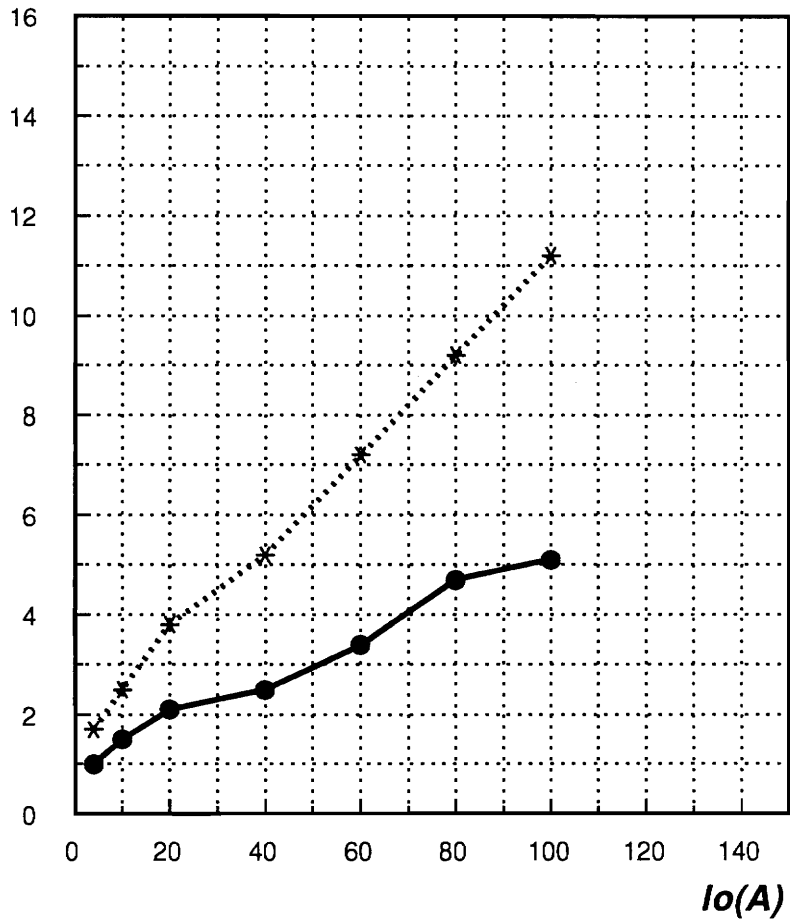
Control current and current gain

The converter efficiency depends, among other parameters, on the current gain of controlled transformer. Current gain in this case is defined as the ratio of the load current (I_o) to the rms current of the control current (I_{con}). Fig. 4.8 shows the magnitude of the control current vs. the load current and Fig. 4.9 shows the current gain vs. load current. It can be seen from Fig. 4.9 that the current gain of PWM scheme is much larger than that of a linear control scheme. This is the main reason PWM scheme is much more efficient than the linear control scheme. It can be seen from Fig. 4.9 that the current gain increases with load current. Appendix IV gives the theoretical calculated and experimental control current of PWM controlled transformer.

4.4 Estimation of Power Losses

In this subsection, power losses of various components in the circuit are estimated. Both the PWM control scheme and the linear control scheme are considered. The distribution of the power losses is estimated according to calculation based on the experimental waveforms. The total calculated power loss is then verified experimentally. For the case when output power is 200w and tables 4.4(a)(b) and (c) summarize the calculated and experimental results. For the case when output power is 500w and tables 4.5(a)(b) and (c) summarize the calculated and experimental results. The details of the theoretical estimation of power losses and efficiencies are given in the appendix II. It can be noted that the losses in the main power circuit are not much different for the two different control schemes, but the losses in the control circuit using PWM control are much less than the case of linear control. As can be

$I_{con}(A)$ (Peak value)

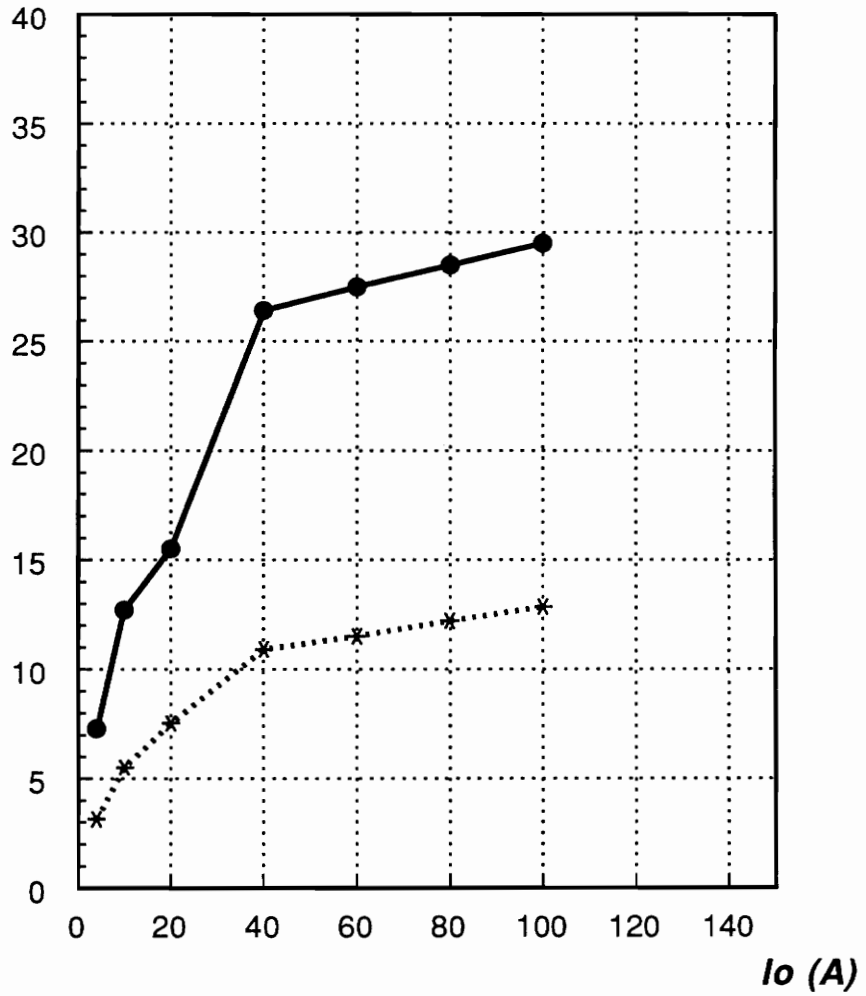


$V_i=240v$

- ***PWM control***
- * ***Linear control***

Fig. 4.8 Control current vs. output current

Current gain= $I_o/I_{con}(rms)$



$V_i=240v$

● *PWM control*

* *Linear control*

Fig. 4.9 Current gain vs. output current.

**Table 4.4(a) Comparison of efficiency and losses
for both control approaches**

Input voltage=240.0v Output power=200.0w

Eff. & losses	PWM control	Linear control
Measured overall eff. %	85.9	81.6
Loss in main circuit (w)	30.4	31.1
Loss in control circuit (w)	2.6	11.3
Calculated overall eff. %	85.8	82.5

**Table 4.4(b) Comparison of control circuit losses
for both control approaches**

Loss in devices	PWM control	Linear control	
S3	0.5	8.0	(w)
DR5	0.6	1.7	(w)
Snubber	0.1	0.1	(w)
Control core & winding	1.4	1.5	(w)
Loss in control circuit	2.6	11.30	(w)

**Table 4.4(c) Losses in main circuits
for both control approaches**

Loss in devices	PWM control	Linear control	
Q1 & Q2	5.8	6.0	(w)
DR3 & DR4 cond.	22.0	22.0	(w)
DR1 & DR2 cond.	0.2	0.3	(w)
Power transformer	2.1	2.5	(w)
Snubber	0.3	0.3	(w)
Loss in main circuit	30.4	31.1	(w)

Table 4.5(a) Comparison of efficiency and losses
for both control approaches

Input voltage=240.0v Output power=500.0w

Eff. & losses	PWM control	Linear control
Measured overall eff. %	83.0	79.3
Loss in main circuit (w)	94.4	97.3
Loss in control circuit (w)	4.6	26.4
Calculated overall eff. %	83.5	80.2

Table 4.5(b) Comparison of losses in control circuits
for both control approaches

Loss in devices	PWM control	Linear control	
S3	1.5	20.4	(w)
DR5	1.2	3.8	(w)
Snubber	0.2	0.2	(w)
Control core & winding	1.7	2.0	(w)
Loss in control circuit	4.6	26.4	(w)

Table 4.5(c) Losses in main circuits
for both control approaches

Loss in devices	PWM control	Linear control	
Q1 & Q2	22.2	24.8	(w)
DR3 & DR4 cond.	65.0	65.0	(w)
DR1 & DR2 cond.	0.3	0.3	(w)
Power transformer	6.5	6.7	(w)
Snubber	0.4	0.5	(w)
Loss in main circuit	94.4	97.3	(w)

Table 4.6

Comparison of voltage drops on MOSFETs S3
in linear and PWM-controlled transformer
 $V_i=280.0v$

$I_o(A)$	10.0	20.0	40.0	60.0	80.0	100.0
$V_o(v)$	5.0	5.0	5.0	5.0	5.0	5.0
$V_{ds(s3)}(v)$ linear	7.09	5.67	5.41	4.4	3.12	3.05
$V_{ds(s3)}(v)$ PWM	0.25	0.25	0.25	0.25	0.25	0.25

seen from Table 4.4, the control circuit dissipates 2.4w for PWM control and 11.3w for linear control to regulate 200w output. Power dissipation in the control circuit has two major components, which are conduction losses in MOSFET S3 and the conduction loss of the diode DR5, as expressed below:

$$P_c = I_{con(rms)}^2 \times R_{on(s3)} + I_{con(av)} \times V_d$$

Where $I_{con(rms)}$ is the rms current of the control current, $R_{on(s3)}$ is the on-resistance of MOSFET S3 and V_d is the forward voltage drop of diode DR5. From section 4.3, it is clear that control current of PWM-controlled transformer is much smaller than that of a linearly-controlled transformer. Another factor of control circuit loss is the effective on-resistance of MOSFET. MOSFET S3 of PWM-controlled approach is either completely turned on or off. Therefore, the on-resistance of MOSFET S3 of a PWM-controlled transformer is less than the linear control counterpart. MOSFET S3 by linear control is operated in the active region of the transistor. Table 4.6 summarizes the comparison of the voltage drop across S3 for both control approaches.

4.5 Control Performances

Loop gain

1) Control-to-output measurement

Figure 4.10 shows the measured control-to-output (V_o/V_{Fe}) Bode plot. It's different from control-to-output Bode plot of a conventional forward converter whose main switches are

```

A: T/R (dB) B:  $\theta$ 
A MAX 20.00 dB MKR
B MAX 180.0 deg GAIN
          -20.7564 dB
          -38.7523 deg
  
```

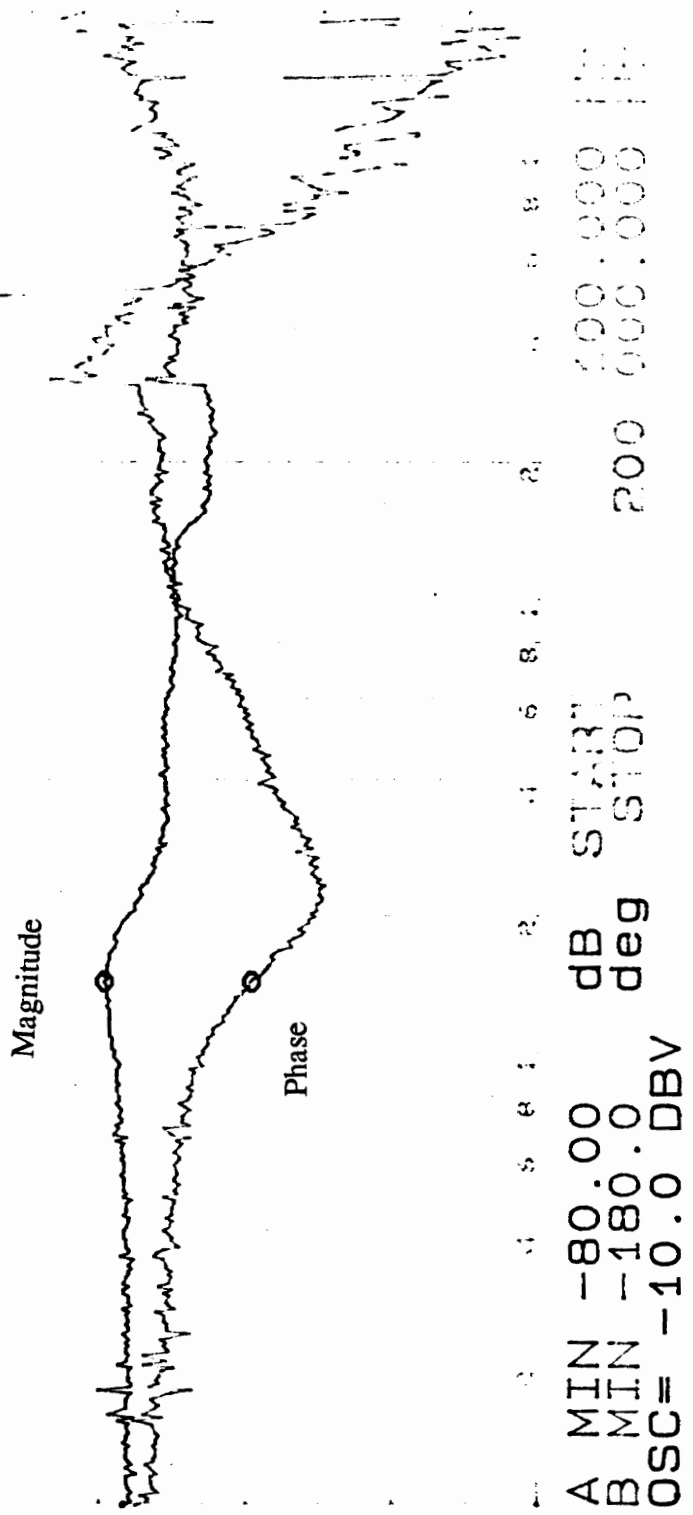


Fig. 4.10 Control-to-output transfer function Vo/Vfe.

controlled by feedback circuit. A small signal model is needed for the controlled transformer but this is beyond the scope of this thesis.

2) Loop gain measurement of dominant pole compensation (shown in Fig. 4.11)

$$V_i=254.0\text{v} \quad I_o=60.0\text{A} \quad V_o=5.0\text{v}$$

$$\text{Crossover frequency}=324.8\text{hz}$$

$$\text{Phase margin}=61.08\text{degree}$$

3) Loop gain measurement of three-pole and two-zero compensator (Fig. 4.12)

$$V_i=254.0\text{v} \quad I_o=60.0\text{A} \quad V_o=5.0\text{v}$$

$$\text{Crossover frequency}=2.34\text{KHZ}$$

$$\text{Phase margin}=80\text{degree}$$

Both compensators are implemented with two-transistor forward converter using controlled transformer. The same conversion efficiency and voltage regulation are obtained, however, transient responses are different, which is described in step responses of this section.

Output impedance

Fig. 4.13 shows the output impedance of the regulator. The maximum value of output impedance at crossover frequency (2.34KHZ) $Z_{\text{omax}}=-27.09\text{dB}$

Step responses

1) Step-load response of dominant pole compensator

Fig. 4.14(a) shows the step-load transient response when the dominant-pole compensation

A: T/R (dB) B: θ 0 MKR 324.833 Hz
 A MAX 15.00 dB GAIN 55.0523 mdB
 B MAX 250.0 deg PHASE 118.919 deg

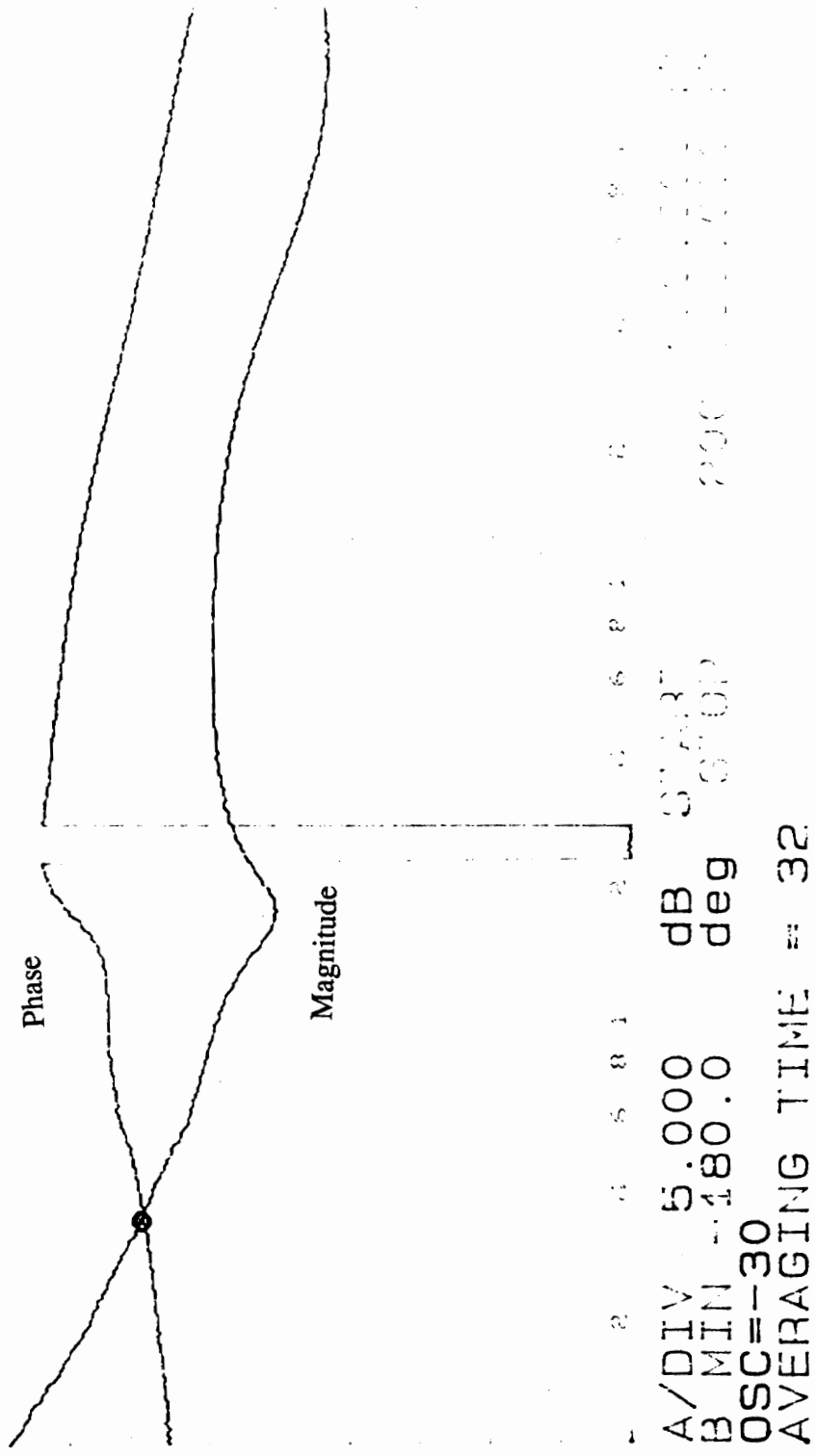
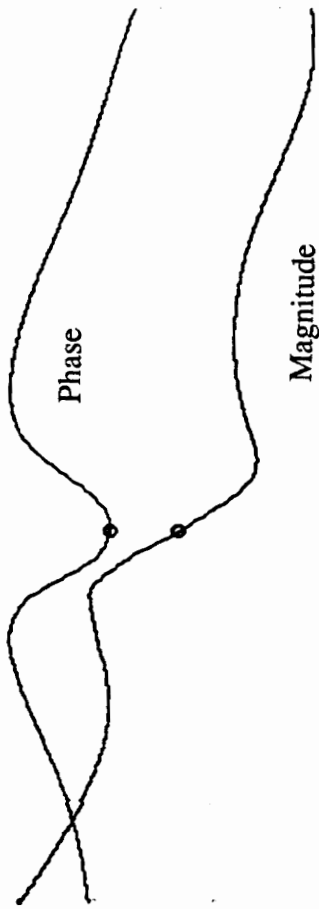


Fig. 4.11 Loop gain (Dominant pole compensation).

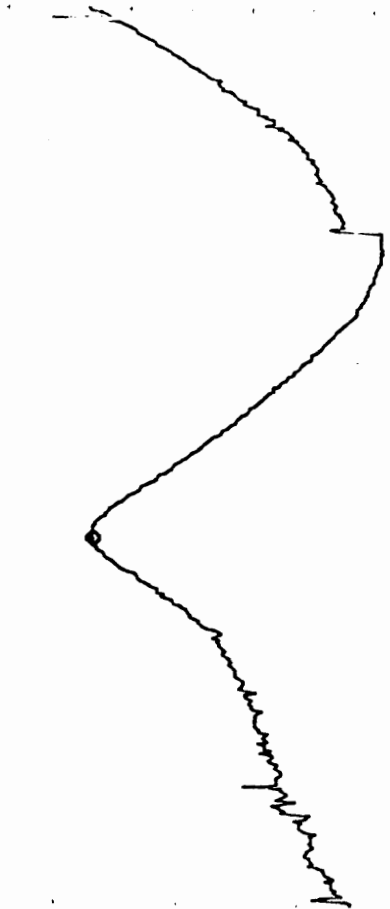
A: T/R (dB) B: θ
 A MAX 20.00 dB MKR 2 343.837 Hz
 B MAX 180.0 deg GAIN -353.854 mdB
 PHASE 99.8698 deg



.
 A MIN -40.00 dB S"Z" 100.000 Hz
 B MIN -180.0 deg S"Z" 200.000, 8000.0
 PSCALE= 2000, 2000, 8000, 8000

Fig. 4.12 Loop gain (Three-pole and two-zero compensation).

A: T/R (dB) B: θ
 A MAX -22.00 dB MKR
 B MAX 180.0 deg GAIN
 -27.0939 dB
 2 213.960 Hz
 -27.0939 dB
 deg



2 4 6 8 1 2 4 6 8 1
 A/DIV 2.000 dB START 200.000 Hz
 B MIN -180.0 deg STOP 200.000 Hz
 OSC= 15.0 DBM
 AVERAGING TIME = 64

Fig. 4.13 Output impedance (Three-pole and Two-zero compensation).

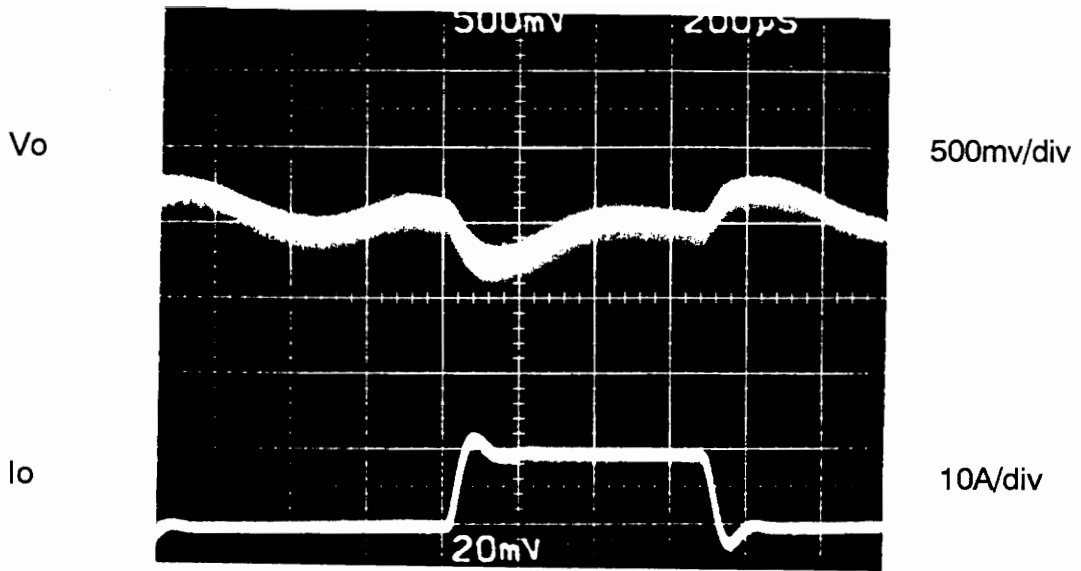
described in Appendix II. When the load current changes suddenly from 60.0A to 70.0A, the output voltage has a overshoot of 300mv and the setting time is approximately 380us.

2) Step-load response of three-pole and two-zero compensator

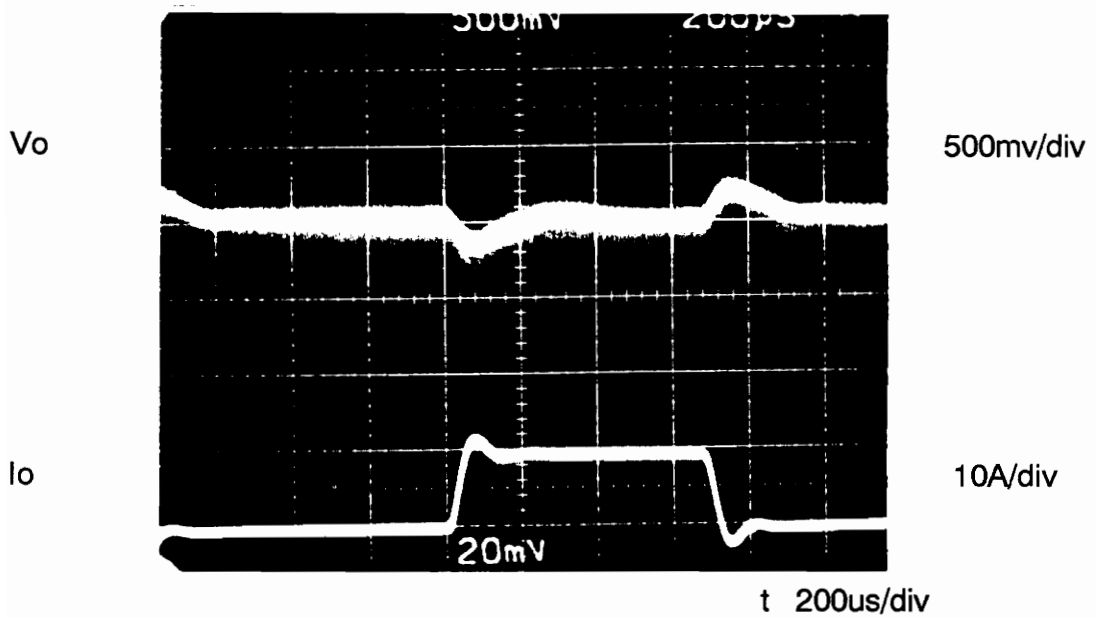
Fig. 4.14(b) shows the step-load transient response when the three-pole and two-zero compensation described in Appendix II. When the load current changes suddenly from 60.0A to 70.0A, the output voltage has a overshoot of 200mv and the setting time is approximately 200us.

Regulation

Fig. 4.15 shows the line and load regulation of two-transistor forward converter with PWM-controlled transformer. It can be seen that converter has good voltage regulation. When the input voltage is low line 240.0v and the output current is less than 80A, the output voltage is 5.0V; When the input voltage is low line and output current is larger than 80A, the output voltage is less than 5.0V since voltage drops on the main switches, rectifier diode, leakage inductance and parasitic resistance are substantial. When the input voltage is 280.0v, output voltage is tightly regulated at 5.0v for whole load range.



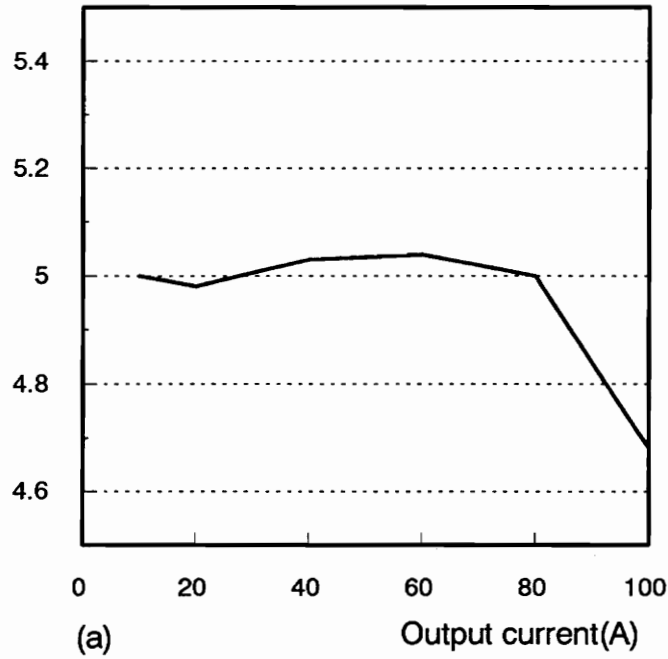
(a) Dominant pole compensator



(b) Three-pole and two-zero compensator

Fig. 4.14 Step load responses.

Output voltage(v)



Output voltage(v)

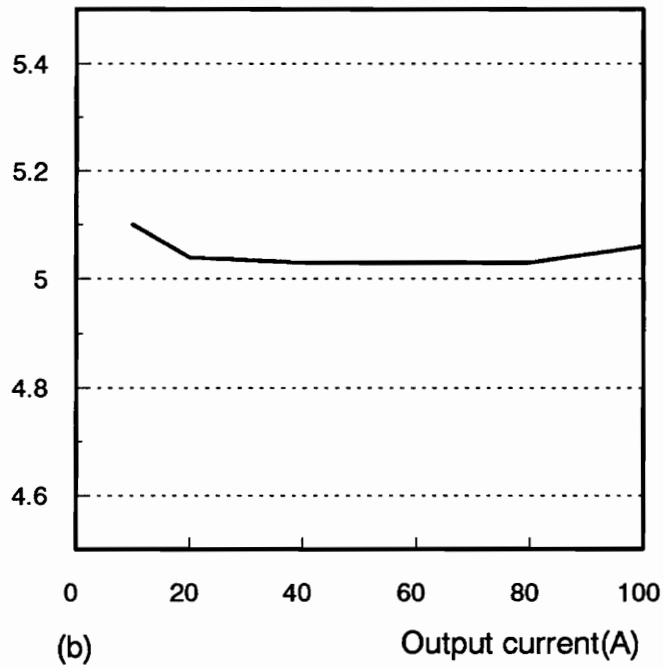


Fig. 4.15 Output voltage vs. output current.

(a) $V_i=240v$ (b) $V_i=280v$

Chapter 5

CONCLUSIONS AND SUGGESTIONS FOR FUTURE WORK

5.1 Conclusions

A PWM control scheme has been proposed and applied to a controlled-transformer regulator. When implemented in a 200-KHz, 500-w, 5-v output forward converter, a power conversion efficiency of 86% has been achieved. This is a significant improvement as compared to a linearly-controlled transformer regulator which is about 5% lower in efficiency. The theory of operation of the proposed circuit is described. The design considerations of the controlled transformer are presented. An numerical example is given to illustrate the design of the controlled transformer. An estimate of power losses for different parts of the circuit is also presented. The information should provide the designer with valuable insight into a PWM controlled-transformer regulator.

A comparison is made between the controlled-transformer regulators and magamp regulator. Soft ferrite cores can be used in a controlled transformer. Therefore, the core cost and the winding cost are much reduced, as compared to the conventional magamp. The design of the control portion of the controlled transformer can be standardized, which is a significant advantage. Furthermore, regulation of the controlled transformer can be done on the primary side rather than on the secondary side, which avoids problems, such as large reactor size, requirement of isolation in the feedback circuit, large power losses and EMI noise,

encountered in high-output-voltage and high-output-current applications. Therefore, a controlled-transformer regulator is particularly suited for high-output-voltage or/and high-output-current post regulation applications. Because of the ease of design standardization, it is also suited for custom-design applications, in which quick turn-around time is often much valued. With the advantages mentioned, it is concluded that a PWM controlled-transformer regulator fills in applications in which a magamp is not well suited.

5.2 Suggestions for Future Work

- 1) Development of small-signal control model for controlled transformer. In order to design the feedback compensator properly, a small signal model of controlled transformer should be developed.

- 2) Application of controlled transformer to mate with a half-bridge and full-bridge converter configuration. The discussion in the thesis is centered on forward converter configuration. An extension of the work is to investigate the possibility of applying the controlled transformer concept to both the half-bridge and the full-bridge configuration.

- 3) Operation of controlled transformer at very high frequency. Now that a ferrite core can be used satisfactorily in a controlled transformer regulator, the possibility of operation

beyond present 200KHz should be explored. The core losses of some ferrite can be extended into megahertz range, but the control characteristics at such frequency is not yet to be explored.

4) Investigation of using controlled-transformer for AC power distribution system. Issues such as regulation, dynamic interaction between local regulators should be investigated.

5) Mating of controlled transformer with soft-switching converters. Soft-switching converters are getting more popular. Using controlled transformer for post regulation for this type of converters is a possibility.

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Appendix I

Glossary of symbols

- A_c : Cross Sectional area of magnetic core
- A_w : Window area of magnetic core
- B_c : Reset flux density
- B_s : Saturation flux density
- B_r : Residual flux density
- B_m : The maximum flux density of power transformer
- B_{cmax} : The practical maximum flux density of control core
- B_{cmin} : The Practical minimum flux density of control core
- ΔB : Flux density swing
- ΔB_{hr} : Headroom of control core
- C1-C9: Capacitors
- D_1 : Duty cycle of primary switches
- D_2 : Duty cycle of secondary voltage
- DR1-DR5: Rectifier diodes
- f_s : Switching frequency
- f_c : The crossover frequency of control loop

f_o : The resonant frequency of converter output filter
 f_{ESR} : The frequency of ESR zero
 F_w : Window utilization factor of magnetic core
 H : Magnetic field force
 I_o : Output current
 I_{L1} : Current of output inductor
 I_{con} : Control current in the control secondary winding
 I_{pri} : Primary current of power transformer
 I_{sec} : Secondary current of power transformer
 I_{m1} : Magnetizing current of power transformer
 I_{m2} : Magnetizing current of control transformer
 I_{m2}' : Magnetizing current of control core reflected to secondary side
 I_{ds} : Drain-source current of MOSFET
 $I_{\text{ds(rms)}}$: RMS drain-source current of MOSFET
 I_{av} : Average current
 I_{in} : Input current
 L_c : Equivalent inductance of T_c
 L_s : Equivalent inductance of T_1
 l_{m1} : Average magnetic path of power core
 l_{m2} : Average magnetic path of control core
 N_m : The number of turns of control secondary winding
 N_{p1} : The number of turns of primary winding of power transformer
 N_{p2} : The number of turns of primary winding of control transformer

- N_2 : The number of turns of secondary winding of power transformer
- N_f : The total number of turns of Sampling winding
- P_{main} : Power dissipation in the main power circuit
- P_{ret} : Conduction loss of rectifiers
- P_{sw} : Switching loss of MOSFET
- P_{on} : Conduction loss of MOSFET
- P_{snub} : snubber loss
- P_{rd} : Reset diode loss
- Q1, Q2: MOSFET switch in the primary side
- R_c : Effective resistance of MOSFET S3
- R_{c1} : ESR resistance of output capacitor
- R_{p1} : Resistance of the primary winding
- R_{s1} : Resistance of the secondary winding
- R_{on} : On-resistance of MOSFET
- R_{on}' : On-resistance of MOSFET reflected to the secondary side
- R_w : Winding resistance of transformer, inductor and circuit wires
- R_L : Load resistance
- T_{on} : Period of the main switch turn-on
- T_{on}' : Power transfer period
- T_{off} : Period of the main switch turn-off
- T_1 : Power transformer
- T_c : Control transformer
- T_b : Blocking time

- T_r : Reset time
- T_{cl} : Clamp time
- $T_{d(on)}$: The turn-on time of MOSFET
- $T_{d(off)}$: The turn-off time of MOSFET
- T_s : Switching cycle
- S3: MOSFET in the control circuit
- V_r : Magnitude of the voltage on the control winding during reset time
- V_b : Magnitude of the voltage on the control winding during block time
- V_f : Secondary voltage amplitude during power transfer period
- V_{fl} : Primary voltage during main switches "on" period
- V_{rl} : Primary voltage during main switches "off" period
- V_{sy} : Synchronous voltage of the primary voltage on the sampling
- V_k : Triangular voltage of RC integrator
- V_e : The error voltage
- V_o : Output voltage
- V_d : Forward voltage drop of diode
- V_{ds} : Drain-to-source voltage of MOSFET
- V_i : Input voltage
- V_{Fe} : Sampled voltage of output voltage
- V_2 : Voltage on the output inductor
- V_{sec} : Voltage on the secondary winding
- V_{pq1} : Voltage from the source of MOSFET Q1 to ground
- V_{pri} : Voltage on the primary side of controlled transformer

V_{con} : Voltage on the control secondary winding

V_{gs} : The gate-to-source voltage of Q1 and Q2

V_{gs3} : The gate-to-source voltage of S3

V_{ref} : Reference voltage

V_{kp} : Voltage drop on the leakage inductance of transformer and parasitic resistance and inductance

μ_1 : Permeability of power core

μ_2 : Permeability of control core

Appendix II

Calculation of power losses

Various power losses of components are estimated for the controlled-transformer regulator, which are described in chapter 4. Both the PWM and the linearly controlled transformer are considered. The following calculation is based on the following specification of the converter operating condition when input voltage is 240.0V, output voltage is 5.0V and output current is 40.0A.

1. Losses and efficiency of the converter using PWM control

(1) Loss in the main circuit

1-1) Main power MOSFET Q1&Q2 (IRFP460)

a. Conduction loss

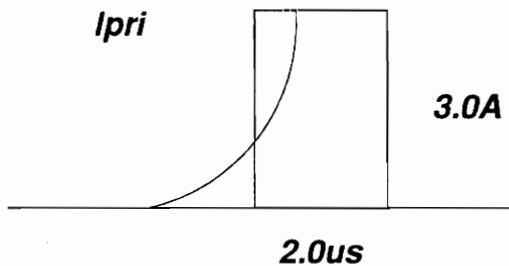


Fig. B.1 Primary current waveform

$$I_{ds(rms)} = \sqrt{\left(\frac{1}{5} \int_0^2 3 \times 3 dt\right)} = 1.897A$$

$$P_{on} = (I_{ds(rms)}^2 R_{on}) \times 2 = 1.897^2 \times 0.27 \times 2 = 1.944w$$

b. Switching loss

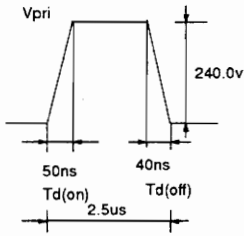


Fig. B.2 Primary voltage

$$T_{d(on)} = 50ns$$

$$T_{d(off)} = 40ns$$

$$P_{sw} = \frac{1}{T_s} \int_0^{T_s} V_{ds}(t) I_{ds}(t) dt$$

$$P_{sw} = \left(\frac{f_s}{2} \times V_{ds} I_{ds(min)} T_{d(off)} \right) \times 2 = \frac{2 \times 10^5}{2} \times 240 \times 2 \times 40 \times 10^{-9} \times 2 = 3.84w$$

Total MOSFET Q1&Q2 loss=1.944+3.84=5.78w

1-2) Reset diode DR1&DR2

$$P_{rd} = \left(I_{(pri)} \times \frac{T_r}{T_s} \times V_d \right) \times 2 = \left(2 \times \frac{0.5}{5} \times 0.5 \right) \times 2 = 0.2w$$

1-3) Rectifier diode DR3&DR4

$$P_{rect} = I_o \times V_d = 40 \times 0.55 = 22.0w$$

1-4) Snubber

$$P_{snub} = \left(\frac{1}{2} \times C_4 \times V_{sec}^2 f_s \right) \times 2 = 0.319w$$

1-5) Power transformer T1

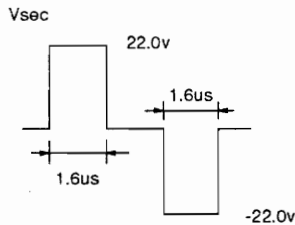


Fig. B.3 Secondary voltage waveform

$$V_{sec} = \frac{N_2 \Delta B A_e}{\Delta t} \times 10^8$$

$$\Delta B = \frac{V_{sec} \Delta t}{A_e N_2} \times 10^8 = 22 \times 1.6 \times \frac{10^{-6}}{0.78 \times 4} = 1128 \text{Gaus}$$

$$B_m = \frac{\Delta B}{2} = 564 \text{Gaus}$$

Magnetics J material core loss density = 220mw/cm³

$$vulume = 5.928 \text{cm}^3$$

a. Total Core loss = 0.22w/cm³ × 5.928cm³ = 1.30W

b. Winding copper loss = I_{pri(rms)}² × R_{pl} + I_{sec(rms)}² × R_{sl} = 0.018 + 0.797 = 0.815w

Total loss of power transformer=2.12w

Total loss of main power circuit P_{main}=30.42w

(2) Loss in the control circuit

2-1) MOSFET S3 (IRF540)

a. Conduction loss

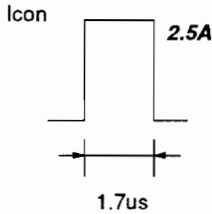


Fig. B.4 Control current waveform

$$I_{con(rms)} = \sqrt{\left(\frac{1.7}{5} \times 2.6\right)} = 1.516A$$

$$P_{on} = I_{con(rms)} \times R_{on} = 2.3 \times 0.077 = 0.177w$$

b. Switching loss

$$P_{sw} = \frac{f_s}{2} \times V_{ds} \times I_{con(min)} \times T_{d(off)} = 2 \times 10^5 \times 22 \times 2.5 \times 50 \times 10^{-9} = 0.275w$$

Total MOSFET loss=0.45w

2-2) Diode DR5

$$P_d = I_{av} \times V_d = \frac{1.7}{5} \times 2.6 \times 0.69 = 0.61w$$

2-3) Snubber

$$P_{snub} = \frac{1}{2} C_6 V^2 f_s = 0.0484w$$

2-4) Control Core and winding

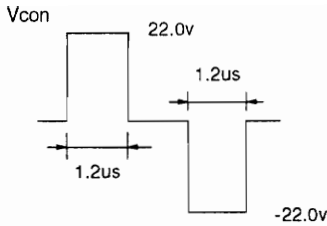


Fig. B.5 Control voltage waveform

$$\Delta B = \frac{V_{con} \Delta t}{A_e} N_m \times 10^8 = 22 \times 1.2 \times \frac{10^2}{(0.78 \times 6)} = 564.1 \text{ Gauss}$$

a. Control core loss = $225 \text{ mw/cm}^3 \times 5.93 \text{ cm}^3 = 1.31 \text{ w}$

b. Copper loss in the control winding

$$\text{Winding copper loss} = I_{pri(rms)}^2 \times R_{pl} + I_{con(rms)}^2 \times R_{sl} = 0.018 + 0.028 = 0.046 \text{ w}$$

$$P_{cu} = 0.018 + .028 = 0.046 \text{ w}$$

Total control core and winding loss = 1.36w

Total loss in the control circuit = 2.47w

(3) Total losses of the converter = $30.42 + 2.47 = 32.89 \text{ w}$

(4) Conversion efficiency

$$\text{Efficiency} = P_o / (P_o + \text{Total loss}) = 200.0 / (200.0 + 32.89) = 85.87\%$$

2. Losses and efficiency of the converter using linear control

(1) Loss in the main power circuit

1-1) Main power MOSFET Q1&Q2 (IRFP460)

a. Conduction loss

$$I_{ds(rms)} = 1.93A$$

$$P_{on} = (I_{ds(rms)}^2 R_{on}) \times 2 = 1.93^2 \times 0.27 \times 2 = 1.99w$$

b. Switching loss

$$P_{sw} = \left(\frac{f_s}{2} \times V_{ds} I_{ds(min)} T_{d(off)} \right) \times 2 = \frac{2 \times 10^5}{2} \times 240 \times 2.1 \times 40 \times 10^{-9} \times 2 = 4.03W$$

Total MOSFET Q1&Q2 loss=1.99+4.03=6.02w

1-2) Reset diode DR1&DR2

$$P_{rd} = \left(I_{(pri)} \times \frac{T_r}{T_s} \times V_d \right) \times 2 = \left(2.5 \times \frac{0.5}{5} \times 0.5 \right) \times 2 = 0.25W$$

1-3) Rectifier diode DR3&DR4

$$P_{rect} = I_o \times V_d = 40 \times 0.55 = 22.0w$$

1-4) Snubber

$$P_{snub} = \left(\frac{1}{2} \times C_4 \times V_{sec}^2 \right) \times 2 = 0.34w$$

1-5) Power transformer T₁

$$V_{sec} = \frac{N_2 \Delta B A_e}{\Delta t} \times 10^8$$

$$\Delta B = \frac{V_{sec} \Delta t}{A_c N_2} \times 10^8 = 22 \times 1.6 \times \frac{10^{-6}}{0.78 \times 4} = 1128Gaus$$

$$B_m = \frac{\Delta B}{2} = 564 \text{Gaus}$$

Magnetics J material core loss density = 230mw/cm^3

$$\text{vulume} = 5.928 \text{cm}^3$$

a. Total Core loss = $0.23 \text{w/cm}^3 \times 5.928 \text{cm}^3 = 1.36 \text{w}$

b. Winding copper loss = $I_{pri(rms)}^2 \times R_{pl} + I_{sec(rms)}^2 \times R_{sl} = 0.019 + 0.87 = 0.89 \text{w}$

$$P_{cu} = 0.018 + .028 = 0.046 \text{w}$$

Total loss of power transformer = 2.25w

Total loss of main power circuit $P_{\text{main}} = 30.86 \text{w}$

(2) Loss in the control circuit

2-1) MOSFET S3 (IRF540)

a. Conduction loss

$$P_{on} = I_{con(av)} \times V_{ds} = 5 \text{A} \times 0.5 \times 2.98 \text{V} = 7.45 \text{w}$$

b. Switching loss

$$P_{sw} = \frac{f_s}{2} \times V_{ds} \times I_{con(min)} \times T_{d(off)} = 2 \times 10^5 \times 22 \times 5.0 \times 50 \times 10^{-9} = 0.55 \text{W}$$

Total MOSFET S3 loss = $7.45 + 0.55 = 8.0 \text{w}$

2-2) Diode DR5

$$P_d = V_d \times I_{av} = 0.69 \times 0.5 \times 5 = 1.725 \text{w}$$

2-3) Snubber

$$P_{snub} = \frac{1}{2} C_6 V^2 f_s = 0.0514w$$

2-4) Control core and winding

$$\Delta B = \frac{V_{con} \Delta t}{A_e} N_m \times 10^8 = 22 \times 1.2 \times \frac{10^2}{(0.78 \times 6)} = 564.1 Gaus$$

a. Control core loss = $245mw/cm^3 \times 5.93cm^3 = 1.45w$

b. Copper loss in the control winding

$$\text{Winding copper loss} = I_{pri(rms)}^2 \times R_{pl} + I_{con(rms)}^2 \times R_{sl} = 0.065w$$

$$\text{Total control core and winding} = 1.45 + 0.065 = 1.52w$$

$$\text{Total loss in the control circuit} = 11.30w$$

(3) Total loss of converter = $30.86 + 11.30 = 42.16w$

(4) Efficiency = $P_o / (P_o + \text{Total loss}) = 200.0 / (200.0 + 42.16) = 82.59\%$

Appendix III

Analysis of DC control characteristics of PWM-controlled transformer and compensator design example

1. Analysis of DC control characteristics of PWM-controlled transformer

In this section, the DC control relationship between output voltage V_o and control error voltage V_e will be established. Symbols used in the derivation are referred to Fig. 3.2 and Fig. 3.9.

The volt-seconds of the control core has to be balanced. According to V_{con} waveform in Fig. 3.10, the block time T_b and reset time T_r have the following relationship:

$$V_b T_b = V_r T_r \quad (c.1)$$

Where V_r is magnitude of the voltage on the control winding during reset time

V_b is magnitude of the voltage on the control winding during block time

According to the waveform of secondary voltage V_{sec} , the output voltage can be expressed as:

$$V_o = V_f \frac{t_{on}}{T_s} \quad (c.2)$$

$$= V_f \frac{D_1 T_s - T_b}{T_s} \quad (c.3)$$

$$= D_1 V_f - V_f \frac{T_b}{T_s} \quad (c.4)$$

Where V_f is the secondary voltage amplitude during power transfer period

Substituting (c.1) into (c.4), then

$$V_o = D_1 V_f - V_f \frac{T_r V_r}{T_s V_b} \quad (c.5)$$

In general, $V_r = V_b$

$$V_o = D_1 V_f - V_f \frac{T_r}{T_s} \quad (c.6)$$

According to the waveform of primary winding voltage V_{pri} , the volt-seconds balance is:

$$T_{on} V_{fl} = T_{off} V_{rl} \quad (c.7)$$

$$T_{off} = \frac{V_{fl}}{V_{rl}} D_1 T_s \quad (c.8)$$

Where V_{fl} is the primary voltage during main switches "on" period

V_{rl} is the primary voltage during main switches "off" period

$$T_r = T_{off} - T_{cl} \quad (c.9)$$

$$V_{sy} = \frac{N_f}{N_2} V_{sec} \quad (c.10)$$

$$V_k = \frac{1}{RC} \int V_{sy} dt \quad (c.11)$$

$$= \frac{1}{\tau} \int V_{sy} dt \quad (c.12)$$

Where V_{sy} is the synchronous voltage of the primary voltage on the sampling winding and V_k is the triangular voltage of RC integrator. τ is time constant of RC integrator; Substitute (c.10) into (c.12), then (c.13) and (c.14).

$$\text{Positive slope} = \frac{1 N_f}{\tau N_2} V_f \quad (c.13)$$

$$\text{Negative slope} = \frac{1 N_f}{\tau N_2} V_c \quad (c.14)$$

The positive slope of ramp signal V_k is $V_c \frac{N_f}{\tau} N_2$, the negative slope is $-V_c \frac{N_f}{\tau N_2}$

Where $N_f = N_{f1} + N_{f2}$;

V_c is the secondary voltage amplitude during clamp period.

Clamp time is: $T_{cl} = V_c / \text{negative slope}$;

$$T_{cl} = \tau \frac{V_c N_2}{V_c N_f} \quad (c.15)$$

Substitute (c.15) into (c.9), then (c.16).

$$T_r = D_1 T_s \frac{V_{fl}}{V_{rl}} - \tau \frac{V_c N_2}{V_c N_f} \quad (c.16)$$

$$V_o = D_1 V_f - \frac{V_f}{T_s} \left(D_1 T_s \frac{V_{fl}}{V_{rl}} - \tau \frac{V_c N_2}{V_c N_f} \right) \quad (c.17)$$

Substitute (c.16) into (c.6), then (c.17).

$$= V_e \tau \frac{V_f N_2}{T_s V_c N_f} + D_1 \left(V_f - V_f \frac{V_{fl}}{V_{rl}} \right) \quad (c.18)$$

In general, $V_f = V_c$, $V_{fl} = V_{rl}$, thus:

$$V_o = V_e \tau \frac{N_2}{N_f T_s} \quad (c.19)$$

Equation (c.19) shows that the output voltage V_o is proportional to error voltage V_e , which is highly desirable to optimize the control loop design.

2. Compensator design examples

One of tasks in the design of the controlled-transformer regulator is to compensate the control loop to achieve system stability and good regulation performance both in DC and transient responses. Compensators of this circuit are designed partly by experience and partly by feedback control theory since the small signal model of forward converter with controlled transformer by PWM control is not available yet. Two different compensation schemes were used in controlled-transformer regulator. Design examples and comparison of two schemes are given. The measurement results of DC regulation and transient response of step load are presented in section 4.5.

1) Dominant pole compensation

Simplest dominant pole compensation can be used as shown in Fig. C.1. The compensation is to close the loop in such a way as to choose crossover frequency well below the resonant frequency of the output filter, before its phase drops very fast. The stable system can be achieved but the transient response may be slow.

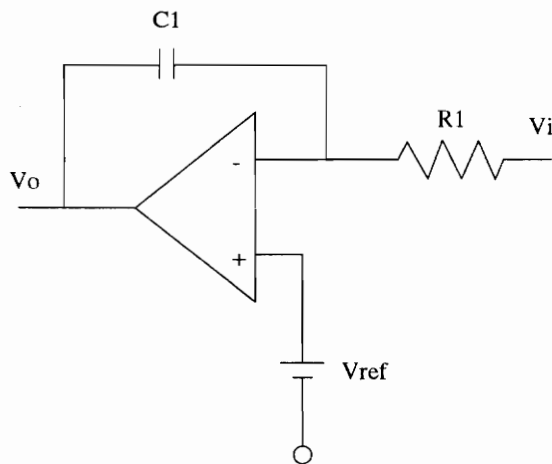


Fig. C.1 Dominant pole compensator

Resonant frequency $f_0 = \frac{1}{2\pi\sqrt{L_1 C_{out}}}$

$L_1 = 3.8\mu h$ $C_{out} = 4000\mu f$

So $f_0 = 1.29khz$

Let crossover frequency $f_c \ll f_0$ Stability of system is guaranteed (4-5)

Take $f_c = 0.8khz$ $f_c = \frac{1}{2\pi R_1 C_1}$

$C_1 = 0.033\mu f$ $R_1 = 4.7kohm$

$$Gain = \frac{1}{C_1 R_1} = 6447$$

2) Three-pole and two-zero compensator

The control-to-output transfer function of forward converter with controlled transformer using PWM control was measured and very popular three-pole and two-zero compensator for single control loop was designed. The circuit diagram and sketch of Bode plot of compensator are shown in Figure C.2.

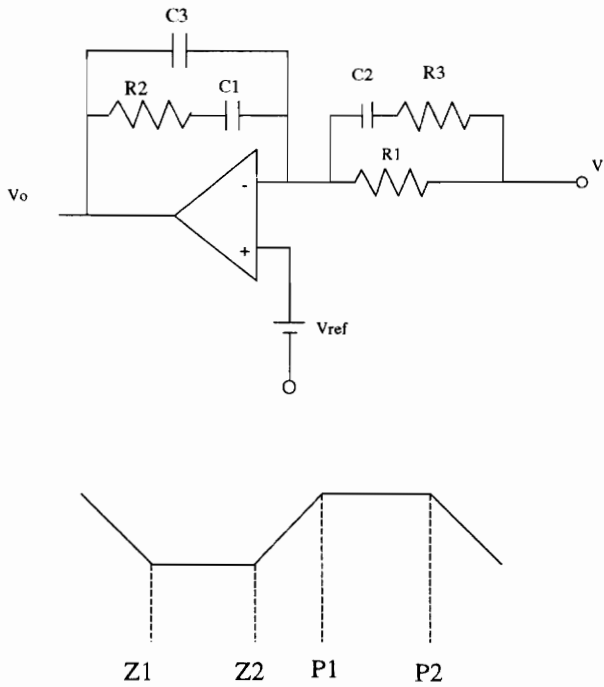


Fig. C.2 Three-pole and two-zero compensator

Three-pole and two-zero compensator transfer function is:

$$A(s) = \frac{\omega_z(1 + s/\omega_{z1})(1 + s/\omega_{z2})}{s\left(1 + \frac{s}{\omega_{p1}}\right)\left(1 + \frac{s}{\omega_{p2}}\right)}$$

Where Z1 is the first zero of compensator;

Z2 is the second zero of compensator;

P1 is the second pole of compensator;

P2 is the third pole of compensator;

The poles and zeros of compensator are placed as follows:

The first pole is placed at origin to tighten DC regulation.

The second pole is put at ESR zero to cancel ESR zero.

The third pole is at half switching frequency to reduce switching noise.

Choose crossover frequency $f_o < f_c < f_{ESR}$

To obtain enough phase boost to stabilize the system, the first zero is placed at half of resonant frequency and the second one is at twice of resonant frequency.

Resonant frequency $f_o = 1.29\text{kHz}$

$R_{c1} = 12\text{mohm}$ $C = 4000\text{uf}$

ESR zero $f_{ESR} = \frac{1}{2\pi R_{c1} C} = 3.31\text{KHZ}$

Choose crossover $f_c = 3.0\text{KHZ}$

First pole is at origin

Second pole : $f_{p1} = f_{ESR} = 3.31\text{KHZ}$ (cancel ESR zero)

Third pole : $f_{p2} = \frac{f_s}{2} = 100\text{KHZ}$ (reducing switching noise)

The first zero: $f_{z1} = \frac{f_o}{2} = 650\text{HZ}$

The second zero: $f_{z2} = 2f_o = 2.58\text{KHZ}$

Loop gain $\omega_l = 2 \times 6500 = 13000\text{rad/s}$

$$\omega_{z1} = 2\pi f_{z1} = 4084\text{rad/s}$$

$$\omega_{z2} = 2\pi f_{z2} = 16210\text{rad/s}$$

$$\omega_{ESR} = 2\pi f_{ESR} = 20800\text{rad/s}$$

$$\omega_{p1} = 2\pi f_{p1} = 20800\text{rad/s}$$

$$\omega_{p2} = 2\pi f_{p2} = 628000\text{rad/s}$$

If Choose $C1 \gg C3$, $R1 \gg R3$, then capacitances and resistances of compensator can be determined as follows:

$$\omega_l = \frac{1}{R_1 C_1};$$

$$\omega_{z1} = \frac{1}{R_2 C_1};$$

$$\omega_{p1} = \frac{1}{R_3 C_2};$$

$$\omega_{z2} = \frac{1}{R_1 C_2}$$

$$\omega_{p2} = \frac{1}{R_2 C_3}$$

Take $R1=10\text{Kohm}$, then:

$$R2=33\text{kohm}$$

$$R3=7.4\text{kohm}$$

$$C1 = 2 \times 0.0039 \mu f = 7800 \mu f \quad C2 = 2 \times 0.0033 \mu f = 6600 \mu f$$

$$C3 = 47 \mu f$$

The system using two designed compensators is very stable under whole load range and regulation is very good. Control performance is reported in section 4.5.

Appendix IV

Calculation of control current of PWM-controlled transformer

From section 3.5.1, the control current can be expressed as:

$$I_{con} = \frac{l_{m2}(B_s - \Delta B)}{N_{m1}\mu_2} \quad (D.1)$$

Where:

$$\Delta B = \frac{(D_1 - D_2)T_s}{A_e N_{p2}} V_i \quad (D.2)$$

$$D_2 = \frac{V_o}{\left(\frac{N_2}{N_{p1}} V_i - I_o (R_{on} + R_w) - V_{kp} \right)} \quad (D.3)$$

It is very difficult to calculate accurate D2 from eq. (D.3). Here, the measured value D₂ is used. The control current of PWM-controlled transformer is calculated under different load condition and listed as follows:

I _o	I _{con}	D ₂	Flux swing	Permeability
10.0A	1.7A	27%	650G	3500
20.0A	2.08A	29%	512G	3000
40.0A	2.75A	32%	486G	2300
60.0A	3.29A	35%	350G	2000
100.0A	4.87A	37%	250G	1500

The theoretical calculated and experimental results of control current are shown in Fig. D1. It can be seen that the calculated value of control current approximately agrees with the experimental results.

VITA

The author was born in Chongqing, Sichuan, P.R. China on December 21, 1962. He attended the 50th High School in Nanjing, P.R. China. He received his Bachelors Degree in 1984 and Masters Degree in 1987, both in Electrical Engineering at Southeast University, Nanjing, P.R. China.

From 1987 to 1988, the author worked as electronics designer engineer at Nanjing electronics Institute.

From 1989 to 1990, the author was a college teacher and researcher with Nanjing Institute of Posts and Telecommunications where he was involved in teaching power electronics course and the design of DC/DC converters.

From 1991 to 1993, the author joined Virginia Power Electronics Center in Bradley Electrical Engineering at Virginia Polytechnic Institute and State University and pursue Master Degree.