UNITRODE
APPLICATION NOTE

# THE UC1901 SIMPLIFIES THE PROBLEM OF ISOLATED FEEDBACK IN SWITCHING REGULATORS

1. Introduction

The UC1901 simplifies the task of closing the feedback loop in isolated, primary-side control, switching regulators by combining a precision reference and error amplifier with a complete amplitude modulation system. Using the IC's amplitude modulated output, loop error signals can be transformer coupled across high voltage isolation boundaries, providing stable and repeatable closed-loop characteristics. Coupling across an isolation boundary is nothing new in transformer technology, and the UC1901's ability to generate carrier frequencies of up to 5MHz keeps the transformer size and cost at a minimum. With a secondary reference and accurate coupling path for the feedback signal, isolated off-line supplies can reliably achieve the tolerances, regulation, and transient performance of their non-isolated counter parts and still take advantage of the benefits of primary-side control.

Closing a feedback loop in a simple or complex system requires a thorough understanding of all of the loop elements. Worse case variations of each element must be taken into account when loop stability, dynamic response, and operating point are determined. Unpredictability in any of the loop components will affect the overall design by making it, necessarily, more conservative. The transient response of a control loop, for example, will usually suffer if a loop must be heavily compensated to guarantee stability with component variations.

To obtain high levels of load and line regulation, the output voltage of a power supply must be sensed and compared to an accurate reference voltage. Any error voltage must be amplified and fed back to the supply's control circuitry where the sensed error can be corrected. In an isolated supply, the control circuitry is frequently located on the primary, or line, side of the supply. As shown in Figure 1, the feedback signal in this type of supply must cross the isolation boundary. Coupling this signal requires an element that will withstand the isolation potentials and still transfer the loop error signal. Though some significant drawbacks to their use exist, optical couplers are widely used for this function due to their ability to couple DC signals. Primarily, opto-couplers suffer from poor initial tolerance and sta-

bility. The gain, or current transfer ratio, through an opto-coupler is loosely specified and changes as a function of time and temperature. This variation will directly affect the overall loop gain of the system, making loop analysis more difficult and the resulting design more conservative. In addition, limited bandwidth capability prevents the use of optical couplers when an extended loop response is required.
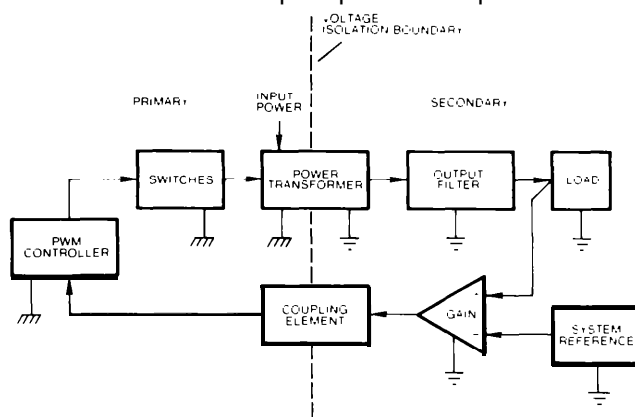
FIGURE 1: A Typical Closed-Loop Isolated Power Supply With Primary-Side Control.

With reliability firmly situated as an important aspect of electrical design, the benefits of primary-side control are increasingly attractive in off-line designs. The organization of an off-line switcher with primary-side control (See Figure 1) puts the control function on the same side of the isolation boundary as the switching elements. Not only does this simplify the interface between the controller and switches, it makes the protection of these switches much easier. Sensing of the switch currents and voltage can avoid failures and improve over-all supply performance. The argument for primary-side control has been further strengthened by the introduction of a new generation of control IC's. The controllers incorporate such features as low current start-up, high speed current sensing for pulse-by-pulse current limiting, and voltage feed-forward. Low current start-up alleviates the problem of efficiently supplying power to a line-side controller, while fast current limit circuitry and voltage feed-forward take advantage of the proximity of a primary-side controller to both the power switch(es) and the input supply voltage.

Combining all of the necessary functions to generate an AM feedback signal on the UC1901 make it the

first IC of its type. As will be seen, the UC1901 can be used in several modes to take full advantage of its functions. Recognizing the continuing evolution of power converter technology the UC1901 is intended to simplify the design of a new era of reliable and higher performance power converters.

2. The UC1901 Functions

The operation of the UC1901 is best understood by considering a typical application. In Figure 2, the UC1901 is shown providing the feedback signal to close the loop in an isolated switching power supply. With any feedback system it is desirable to compare the system output to the system reference with a minimum of intermediate circuitry. With the UC1901 situated on the secondary, or output side of the supply, the output voltage is simply divided down and compared to the 1.5V reference using the chip's high gain error amplifier. In this manner DC errors at the supply output are kept minimal even if significant non-linearities, or offsets, occur in the remainder of the power supply loop. Since the 1.5V output on the UC1901 is a trimmed, precision, reference, the need for a trim-pot to fine tune the output voltage is eliminated.

To make the UC1901 compatible with single output 5V power supplies it is designed to operate with input voltages as low as 4.5V This allows the part to be powered directly from a TTL compatible 5V output. A nominal supply current of only 5mA allows the part to be easily operated at its maximum input voltage rating of 40V without worry of excessive power dissipation.

The amplified error signal at the UC1901's compensation output is internally inverted and applied to the modulator. The other input to the modulator is the carrier signal from the oscillator. The modulator combines these two signals to produce a square wave output signal with an amplitude that is directly proportional to the error signal and whose frequency is that of the oscillator input. This output is buffered and applied to the coupling transformer. With the internal oscillator, carrier frequencies into the megahertz range can be generated. Operating at high frequencies can reduce both the size and cost of the coupling transformer. The secondary winding on the coupling transformer drives a diode-capacitor peak detector. With a simple resistive load to allow discharging of the holding capacitor an effective amplitude demodulator is formed. The small signal voltage gain from the error amplifier input to the detector output is a function of the feedback network around the error-amp, the modulator gain, the turns ratio of coupling transformer, and any loss in the demodulator.

In Figure 2 the relationship of the detector output to the sense supply voltage is non-inverting. This is necessary to guarantee start-up of the supply. Since the UC1901, as shown, is powered from the supply's output, the initial feedback signal back to the PWM controller will always be zero. The required 180° of DC phase shift is easily achieved by inverting the signal with the error amplifier that is present in most any PWM controller circuit.

In some applications it may be desirable to operate the carrier frequency of the UC1901 in synchroni-
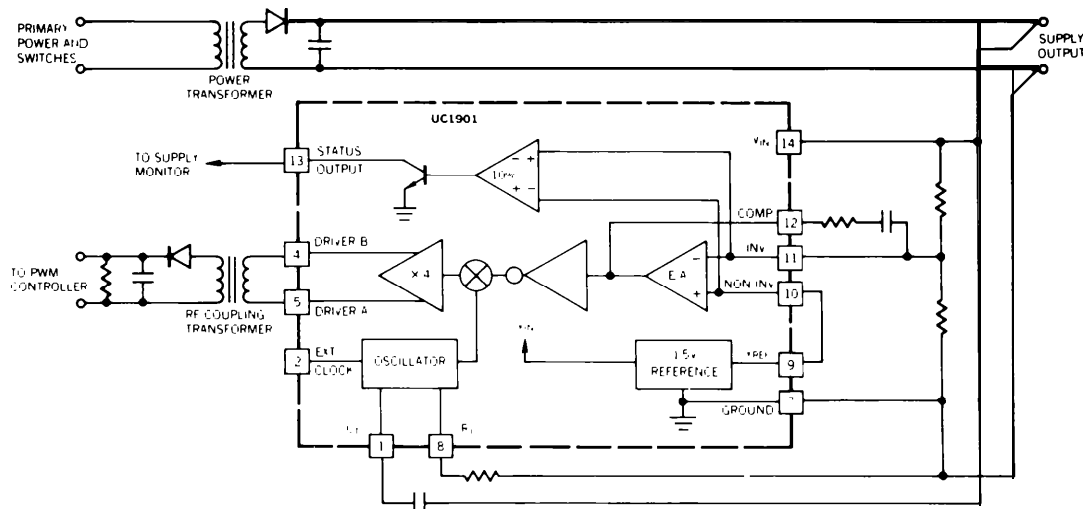
FIGURE 2: With a Precision Reference, and a Complete Amplitude Modulation System, the UC1901 Lets Isolated Feedback Loops be Closed Using a Small Signal Transformer:

3A
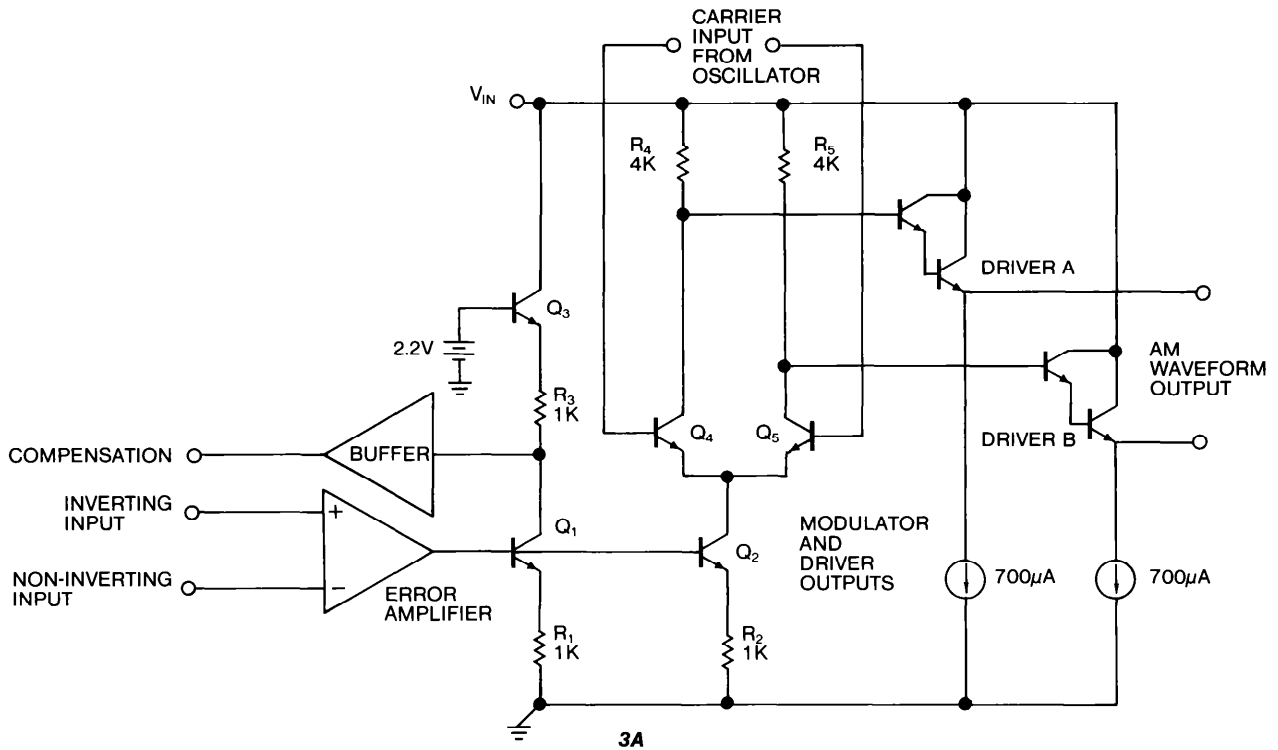
FIGURE 3: The Compensation Output on the UC7907 can be used to Accurately Control the AM Waveform Output.
A Simplified Schematic, (a) Shows the internal Signal Split into the Modulator: Voltage Waveforms, (b) Across
the Modulator Outputs, and at the Compensation Output show the Modulator Transfer Characteristic.

zation with a system clock, or reference frequency. In many situations, operation of the UC1901 at the switching frequency of the power supply can be beneficial. One such application is presented in this article. To accommodate this need the UC1901 has an external clock input.

One additional mode of operation is possible if the oscillator is left disabled and the external clock signal is kept low (or floated). In this condition the error amplifier can be used in a linear fashion with its output taken at the driver A output. The driver B output will be at a fixed DC voltage about 1.4V from the input supply voltage. If the external clock signal is tied high the roles of the two driver outputs are reversed. With 15mA of output current capacity, the two outputs can easily be combined to reference and drive an optical coupler. Although the instabilities of the coupler will still be present, the advantages of the UC1901's precision reference, high gain amplifier-driver, and 4.5V supply operation can be utilized.

3. A Controlled Feedback Response

There are many different topologies which can be used when implementing a switching power supply. For off-line supplies, fly-back and forward convert-
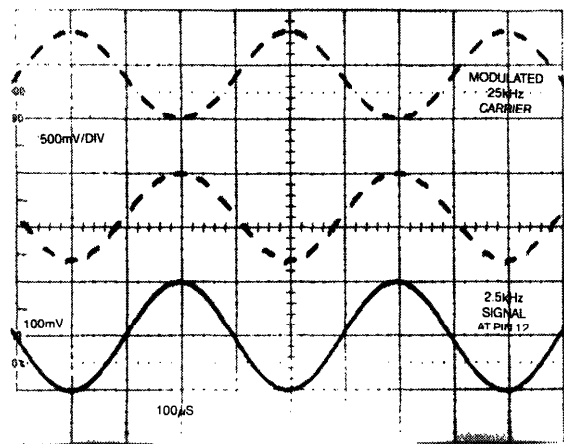

3B

ers are often designed. In the near future current-mode control versions of these may also be widely used. Each of these converter topologies has a different forward transfer characteristic and, within each type of converter, operating point, continuous or discontinuous inductor current, and voltage or current-mode duty cycle control are a few of the factors which can alter this characteristic. In short, the task of optimally designing a feedback network for one supply must usually be repeated when the next supply is designed.

Once the forward transfer function of a particular converter has been determined, various factors such as stability, line regulation, load regulation, and transient response will determine the overall loop response, and therefore feedback response, required. One of the objectives of the UC1901, in addition to allowing a controlled isolated feedback response, is to make the task of implementing a given response as easy as possible. With the compensation node on the UC1901, local R-C feedback networks can be used to shape the small signal gain and phase frequency response of the overall feed back network.

The error amplifier on the chip has a typical open loop gain of 60dB and is internally compensated to have a unity gain bandwidth of just above 1MHz. Both of these characteristics are measured with respect to the compensation node (Pin 12). As shown in Figure 3a, the amplified error signal is internally split, at the collectors of Q_1 and Q_2, and fed to both the modulator and the compensation output. Applying feedback from the compensation output to the error amplifier's inverting input controls the small signal collector current through Q_1. Since Q_2 sees the same base voltage, and its emitter resistance is the same, its collector current will track that of Q_1. The collector current of Q_2 feeds the modulator and determines the amplitude of its output signal. The 4-to-1 ratio of resistors R_4 (or R_5) and R_2 results in a fixed 12dB of small signal gain measured as the ratio of the amplitude of the differential signal at the modulator outputs to the compensation mode signal. This relationship, as well as the function of the modulator, is shown in Figure 3b. The scope traces show a 200mV peak to peak sinusoid at 2.5kHz, measured at the compensation output, and the resulting 800mV variations in the peak amplitude of a 25kHz square wave carrier as measured across the modulator's differential output.

The remaining factors influencing the response of the feedback path are the signal gain through the transformer, the detector circuit, and the circuitry between the detector output and the supply's PWM. The signal gain through the transformer is simply the turns ratio of transformer. The small signal detector gain can usually be assumed to be unity as long as the AC load presented to the detector is kept small. Some load on the detector is necessary to allow its output to slew in a negative direction. Figure 4 summarizes the transfer and output characteristics of a typical transformer and detector.
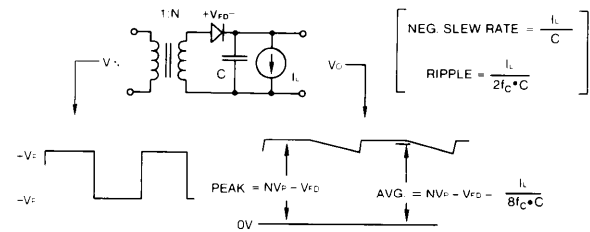
FIGURE 4: A Typical Detector Model and its Output Characteristics.

Here the load on the detector is modeled as a current source, simplifying the equations. In actual practice the operating point of the detector output will be determined by the circuitry which interfaces it with the PWM input. Since the minimum recovery from the detector is zero volts a nominal positive operating level which provides adequate dynamic range for DC and transient conditions should be chosen.

The UC1901 is specified to generate maximum carrier levels equal to or in excess of 1.6V peak. This indicates that a turns ratio of greater than one-to-one will be required for the coupling transformer if the detector output must exceed approximately 1V, (allowing for a detector diode drop of 0.6V). It should be noted that many switching power supplies now being designed include an integrated PWM control IC. A typical PWM IC includes a dedicated error amplifier which amplifies and buffers the input error voltage and applies it to the PWM ramp comparator. This amplifier can be readily used to fix a nominal detector operating point that is compatible with a one-to-one transformer. Additionally, the error amplifier on the UC1901 and the PWM's amplifier can be combined to achieve both large DC loop gains for improved load and line regulation, and the optimization of the loop gain and phase frequency response for improved transient and stability performance.

4. Transformer Requirements

The coupling transformer used with the UC1901 has two primary requirements. First, it must provide DC isolation. Secondly, it should transfer voltage information across the isolation boundary. Meeting the first requirement of DC isolation will depend on specific applications. In general, though, small signal transformers can be readily built to meet the isolation requirements of today's line-operated systems.

For the most stringent applications, E-type cores with bobbin carried windings are inexpensively available or built. Where small size is most important, a simple toroid core can be used.

The second requirement of the transformer primarily determines the amount of magnetizing inductance it must have. The magnetizing inductance of a transformer refers to the actual inductance formed by the windings around the core material. In many classical transformer examples, the magnetizing inductance is ignored. This is a valid approximation since, in these examples, the magnetizing current required is much less than the reflected load currents. In this case, the load currents are small and, as the transformer inductance is reduced, the magnetizing currents become dominant.

The driver outputs on the UC1901 are emitter followers which are biased at 700μA. Therefore, if the drivers are operated without additional bias current the peak current through the transformer's primary winding cannot exceed this value. Figure 5a illustrates the relationship of the magnetizing current to the voltage across the transformer's input. If the reflected load currents are neglected, it can be seen that the minimum magnetizing inductance required for linear transfer of the modulator squarewave is given by:

(1) $$L_M \geq \frac{V_P}{4f_c I_P}$$

Where: L_M = the magnetizing inductance,

V_P = the peak carrier voltage across transformer inputs,

f_c = the UC1 operating frequency,

I_P = the bias current of the UC1901 drivers.

As an example, consider the case where Vp is equal to 2V, f_c is 100kHz, and the drivers are operating at their internal bias levels. Using equation 1, the inductance looking into the primary winding with no secondary load must be greater than 7.1 mH. Alternatively, if the carrier frequency is raised to 1 MHz and the bias levels of the UC1901 drivers are increased to 3.5mA, then L_M can be as low as 150μH. Using high permeability ferrite material, this level of magnetizing inductance can be realized with as little as 10 turns on a small toroid core.

Equation 1 sets a minimum limit on the magnetizing inductance for linear transfer of the carrier wave-

5A

5B
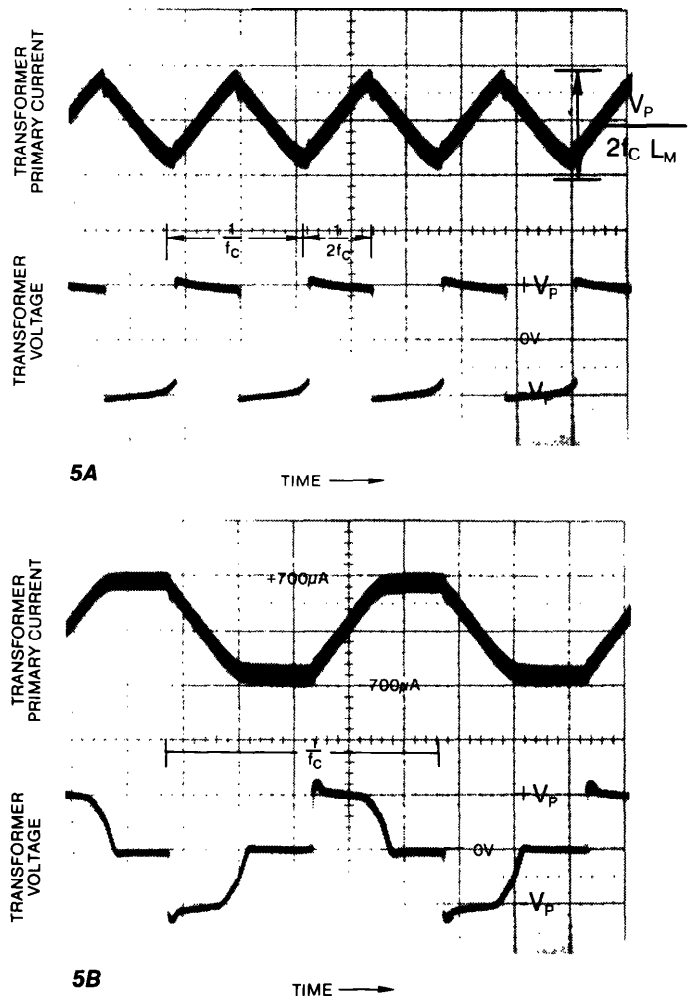
FIGURE 5: *The UC1901 Driver Outputs Follow the Modulator Output Square Wave, (a.), Sourcing and Sinking Current Levels Dependent on Transformer Inductance, Carrier Frequency, and Voltage Level. When the Bias Level of the Driver Outputs, I_P, is Reached, (b.), a Tri-state Waveform is Coupled Across the Transformer; the Peak Voltage Level Though, Remains Approximately the Same. The Reflected Load Currents are Assumed Negligible.*

form. Actually, the amplitude information is still coupled even when the inductance is less than this minimum. In this case, the UC1901 drivers will support the voltage across the coil until the peak current is reached. The result, illustrated in Figure 5b, is a tri-state waveform at the transformer's input and output. Peak detection of this waveform yields the same amplitude information as the linear transfer case, although detection ripple will increase. Another situation which results in a tri-state waveform exists when the carrier duty cycle is not 50%. In this case, the volt-seconds across the transformer will be balanced by an "imbalancing" of the driver

bias levels. The imbalance will be sufficient to cause the peak current to be reached during the > 50% portion of the carrier waveform.

5. The High Frequency Oscillator

The oscillator circuit on the UC1901 is designed to operate at frequencies of up to 5MHz. To achieve this operating range the circuit shown in Figure 6 uses only NPN transistors in those parts of circuit which are dynamically involved in the actual oscillation. The standard bipolar process used to produce the UC1901 characteristically yields high f_T, typically 250MHz, NPN devices. Conversely, the same process has PNP structures with f_T's of only 1 to 2MHz. In the oscillator, PNP's are used only in determining quiescent operating points of the circuit.

The latched comparator formed by Q_1-Q_4, diodes D_1 and D_2, and resistors R_1 and R_2 has a controlled input hysteresis which determines the peak to peak voltage swing on the timing capacitor C_T. The timing capacitor C_T is referenced to V_{IN} since this is the reference point for the latched comparator's thresholds. The comparator's outputs at D_1 and D_2 switch the 2X current source through Q_{10} changing the net current into the timing capacitor from positive to negative, reversing the capacitor voltage's dv/dt.

When the resulting ramp reaches the comparator's lower threshold, the current is switched back to Q_{11} and the ramp reverses until the upper threshold is reached and the process begins again. This results in a triangle waveform at C_T and a squarewave signal at D_1 and D_2.

The magnitude of the charging current is controlled by the external resistor, R_T and the internally generated voltage across it. This voltage is compensated to track variations in the comparator hysteresis. The tracking characteristics of this voltage stabilize the oscillation frequency over temperature and enhance the initial frequency tolerance. Typically, repeatability and temperature stability of the operating frequency are both better than 5%.

The oscillator circuit has been optimized for a nominal R_T of 1 A desired operating frequency is obtained by choosing the correct value for C_T. As shown in Figure 7, the oscillator frequency is give by the relation:

$$(2) \qquad \qquad f_{OSC.} \quad = \quad \frac{1.24}{R_T \, C_T},$$

for frequencies below 500kHz. Above 500kHz, the solid line indicates appropriate C_T values. There is
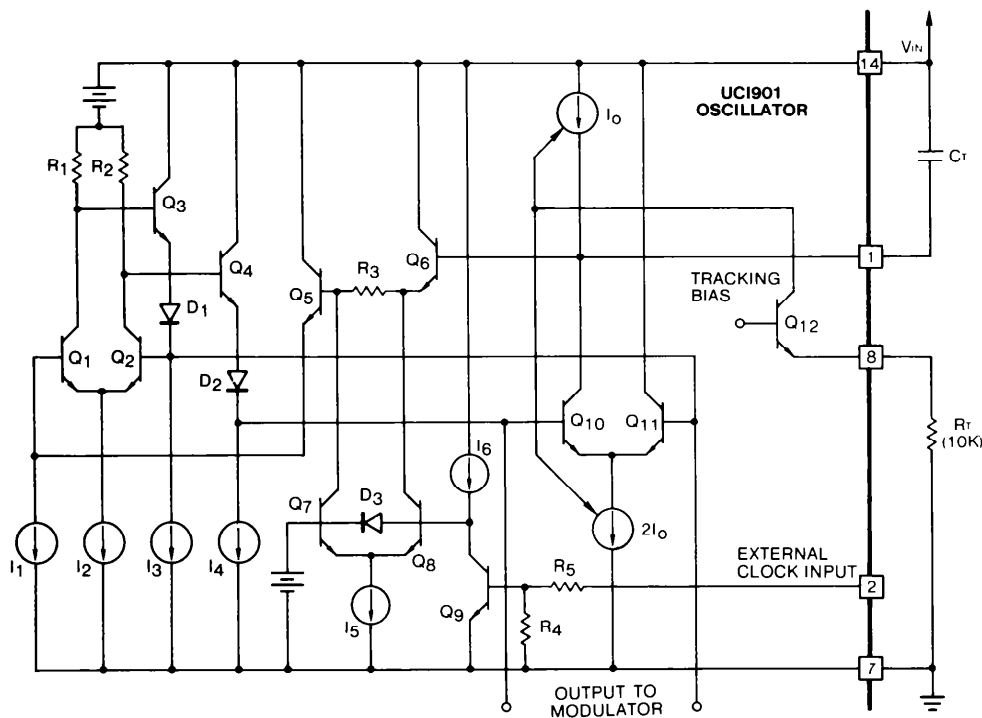

FIGURE 6: UC1901 High Frequency Oscillator Simplified Schematic.

no upper limit on the size of-the capacitor used, thus allowing the oscillator to have an arbitrarily long period if desired.

$$f_{osc} = \frac{1.24}{R_T C_T}$$

$V_{IN} = 10V$
$R_T = 10K\Omega$
$T_j = 25°C$

OSCILLATOR FREQUENCY – Hz
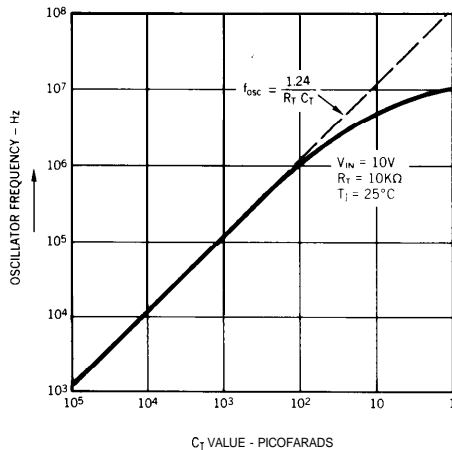
C_T VALUE - PICOFARADS

FIGURE 7: UC1901 Oscillator Frequency.

To allow operation of the modulator with a carrier frequency that is driven from a system operating frequency or clock, the oscillator can be over-ridden. Tying C_T to the input supply voltage disables the oscillator. The modulator circuit can now be switched in synchronization with a signal at the external clock input. Internally, the clock signal is applied to the latched comparator via the input device Q_9, and the differential pair Q_7 and Q_8. As the clock input goes high, Q_9 turns Q_8 off and Q_7 on, creating an offset across R_3 that is sufficient to switch the comparator. The comparator then, as before, drives the modulator. When the clock input returns low, the process is reversed. Using the external clock input, both the frequency and duty cycle of the modulator outputs are controlled.

6. A Status Output is More Than Just a Green Light

Many systems today require a monitoring function on the supply output. The status output on the UC1901 can fill this need, a green light function, and can also be used to fill some more "sophisticated" needs. The circuit in Figure 8 takes advantage of the status output in the start-up of an off-line forward converter. The UC1901 is being used in an application where the switching supply must be synchronized to a system clock. The clock signal is generated on the secondary or output side of the supply. To allow start-up, the PWM oscillator is free-running when the line voltage is applied. As the supply voltage rises, the UC1901's external clock input is driven at the switching frequency rate through resistors R_1 and R_2. When the supply output
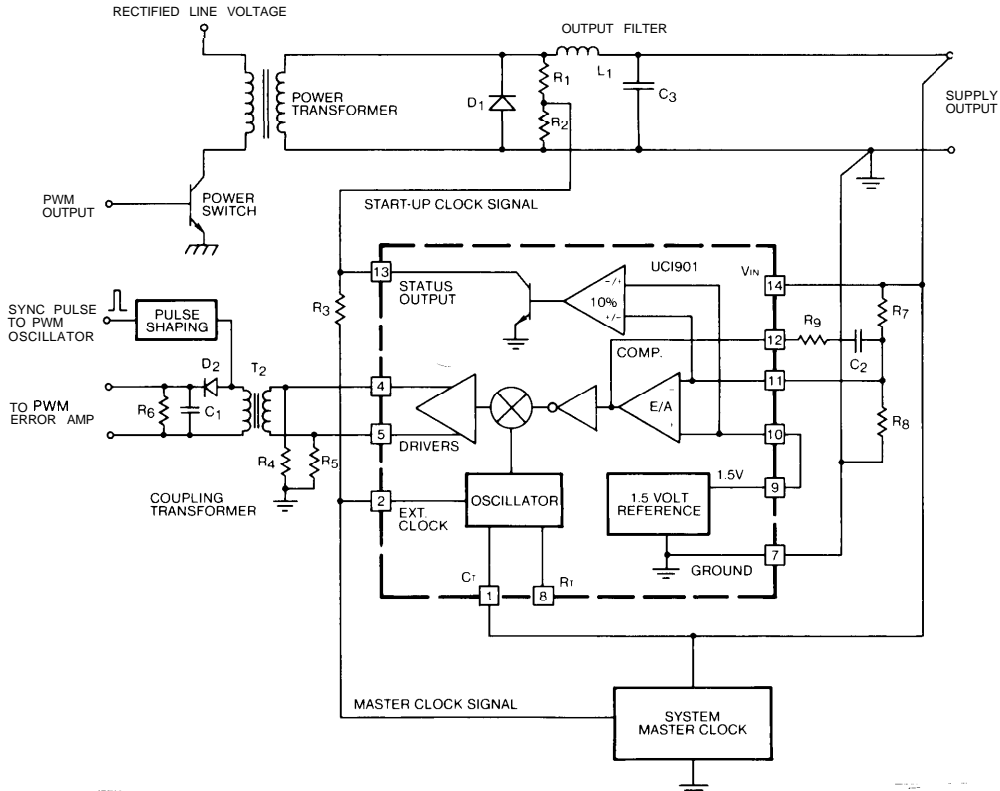

FIGURE 8: The Status Output on the UC1901 is used in the Start-Up of a Power Supply Synchronized to a Secondary Referenced Master Clock The Coupling Transformer Carries the Feedback and Clock Signals. The Status Output is used to Sequence Clock Signals to the UC1901 External Clock Input During Start-Up.

reaches 90% of its operating level, the status output decouples the external clock input from the switcher and enables the UC1901's clock input to be driven from the now operational system clock.

On the primary side, the output of the coupling transformer is used before demodulation to provide a synchronization pulse to the PWM control oscillator. Under normal operation, the entire power supply, including the feedback system, will be synchronized to the system clock.

7. The UC1901 in an Off Line Flyback Converter

As alluded to previously, flyback converters see wide use in off-line applications. The flyback topology has some general cost benefits which have spurred its use in low cost, low power (< 150W), off-line systems. Perhaps the two most significant of which are the need for only a single power magnetic element in the supply (no output filter inductor is required), and the ability to easily obtain multi-output systems by adding one additional winding to the coupling power inductor for each extra output. Also, the flyback topology, especially when used in the discontinuous mode, lends itself very well to the benefits of voltage feed-forward.

7a. 60 Watt Dual Output Converter

Shown in Figure 9 is a flyback converter designed with the UC1901 and a primary side control IC, the UC1840. The converter has two 30W outputs, one at 5V/6A, and another at 12V/2.5A. Minimum loads of 1A are specified at each output. The UC1901 is used to sense and regulate the 5V output. This output is specified at ±2 percent (untrimmed), with load and line regulation of better than 0.2 percent. Respectively, the 12V output is specified at ±5 percent with ±6 percent load and line regulation. Regulation of the 12V output relies on close coupling between the 5V and 12V output circuits.

The UC1840 controller has all of the features discussed previously for an off-line controller. In addition, it has some advanced fault protection features. Only parts of the UC1840's capabilities are discussed here. For those desiring a more complete description, it can be found in the second reference mentioned at the end of this article. In the supply, the UC1840 sequences itself through start-up using the energy stored in C_4 by the trickle resistor R_{11}. Once the supply is up and running W_4, the auxiliary winding on L_1, provides power to the controller and the switch drive circuitry. The primary

winding on the coupled inductor, W_1 is applied across the rectified and filtered line voltage at a 60kHz rate via the FET switching device. L_1 is referred to as a coupled inductor, rather than as a transformer, since the primary and secondary windings do not conduct at the same time. Energy is stored in the inductor core as the switching device conducts, and then "dumped" to the secondary outputs when the device is turned off.

The converter operates in the discontinuous mode. Operating in this mode, the total current in the coupled inductor goes to zero during each cycle of operation. In other words, the energy stored in the core during the beginning of a cycle is entirely expended to the load before the end of the cycle. This allows the inductor size to be minimized since its average energy level is kept low. The price paid for discontinuous operation is higher peak currents in the switching and rectifying devices. Also, high ripple currents at the supply's output(s) make ESR, (equivalent series resistance), requirements on the output filter capacitors more stringent.

7b. Discontinuous Flyback's Forward Transfer Function

The process of designing a feedback network for the supply begins with determining the small signal transfer function of the converter's forward control path. This path can be defined as the small signal dependency of the output voltage, V_{OUT}, to, V_C, the control voltage at the input to the PWM comparator. As defined, the control voltage on the UC1840 appears at the compensation output of its internal error amplifier. The transfer function of this path for the discontinuous converter is given by equation (3).

$$(3) \quad \frac{v_{OUT}}{v_C}(s) = \frac{V_{IN}}{V_R} \sqrt{\frac{T_P R_L}{2L_M}} \cdot \frac{1 + sC_F R_S}{1 + \dfrac{sC_F R_L}{2}}$$

Where:

V_{IN} = level of the rectified line voltage,

V_R = The equivalent peak PWM ramp voltage-equal to the extrapolated control voltage input which would result in a 100% switch duty cycle,

T_P = One period of the switching frequency,

L_M = Magnetizing inductance of the primary winding,

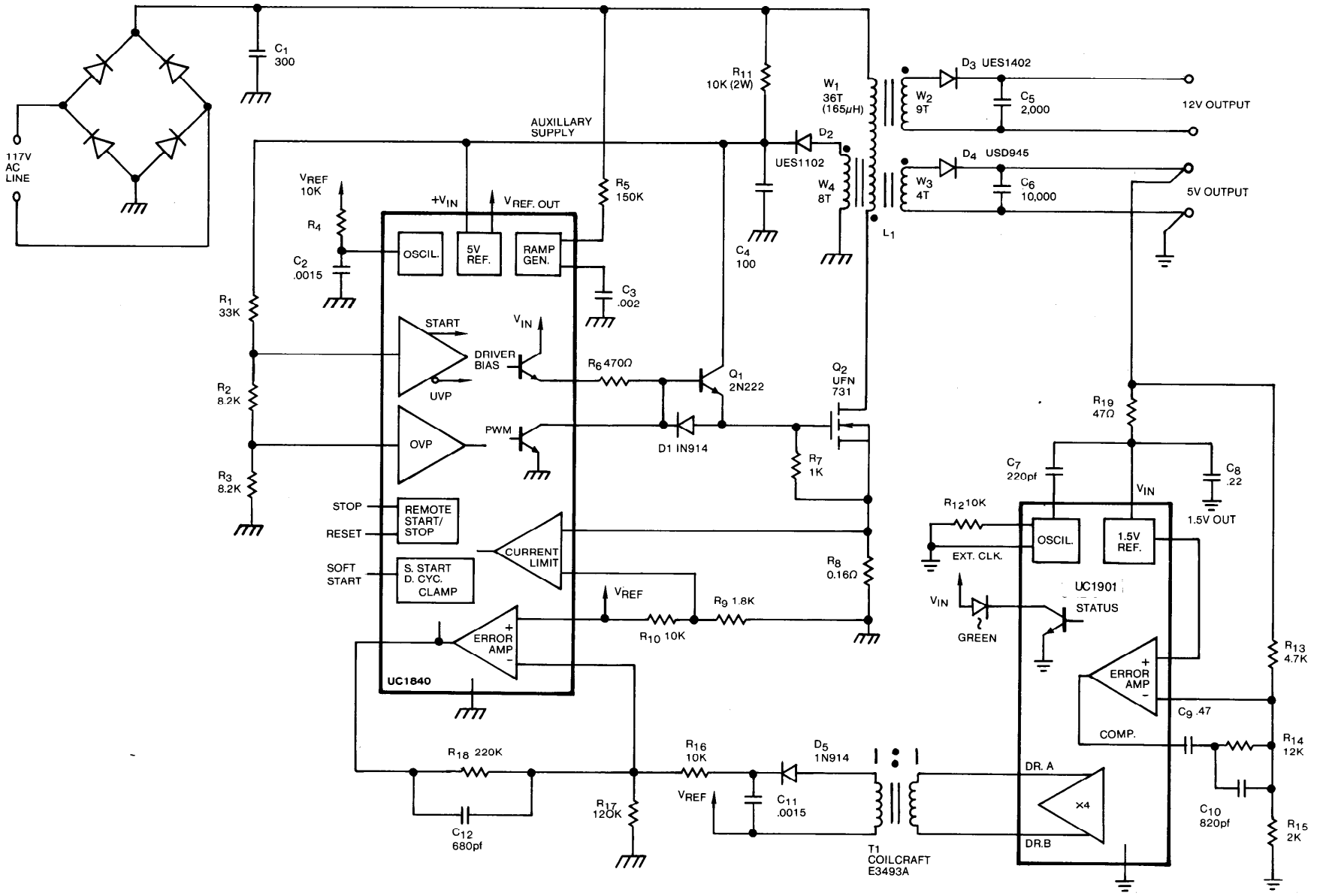C_F = A total effective output filter capacitor,

FIGURE 9: The UC1901 Combines With an Advanced PWM Controller in a 60W Off-Line Converter.

R_L = The total effective load, (assumed resistive),

R_S = ESR of the filter capacitor,

s = $2\pi j f$, f is frequency in hertz.

The word effective is used in describing R_L and C_F since, although we are interested in calculating the response to the 5V output, the loads at the 12V and auxiliary outputs must be accounted for. This is easily done by reflecting these loads to the 5V output using the corresponding turns ratio on the inductor.

7c. Voltage Feedforward Steadies Response

Equation 3 indicates a substantial dependency of the control response to both the load R_L, and the input voltage, V_{IN}. This can slightly complicate the design of the feedback network since both the gain and phase response of the loop will vary with operating conditions.

The benefits of feed-forward are easily illustrated at this point by examining its effect in this circuit. The UC1840 controller uses resistor R_5 to sense the input voltage and proportionately scale the charging current into the PWM ramp capacitor, C_3. Scaling the ramp slope is the same as scaling V_R, the equivalent peak ramp voltage. The result is a modeled ramp voltage given by:

$$(4) \qquad V_R = \frac{V_{IN} \, T_P}{R_5 \, C_3}$$

When this expression for V_R is substituted into equation 3, the result is a forward transfer function that is independent of the input voltage. Not only does this simplify the feedback analysis, it also vastly improves the supply's inherent rejection of line voltage variations.

The forward response of the converter, plotted in Figure 10, has a single pole roll-off occurring between 11Hz and 38Hz depending on the load. The single pole roll-off allows the feedback network a bit of latitude since, from a stability standpoint, the loop bandwidth can be extended by simply adding broadband gain with an appropriate roll-off frequen-
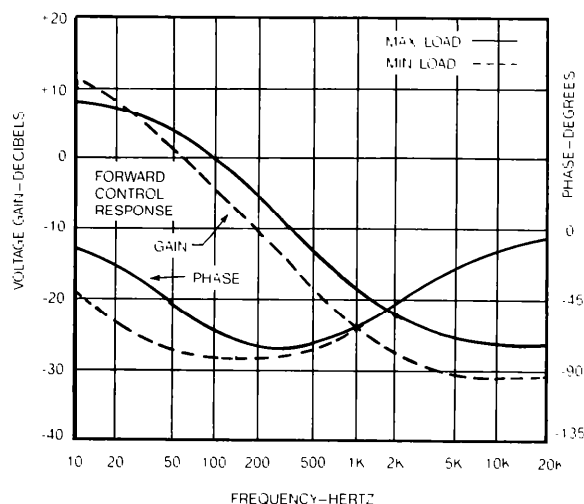
FIGURE 10: Closing the Feedback Loop is Preceeded by the Characterization of the Converter's Forward Small Signal Transfer Function.

cy. No mid-band zeros or led-lag networks are necessary, as might be for converters with double pole responses. Although, the zero resulting from the ESR of the filter capacitors can, if not taken into account, appreciably extend the loop bandwidth beyond its intended value.

7d. Wide Bandwidth Gives Fast Transient Response At 5V Output

This supply was designed to have a unity gain loop bandwidth of between 5 and 10kHz. With this bandwidth the supply's control response to step load and line changes occurs in fractions of a millisecond. This is only true with regard to the 5V output. There is no feedback from the 12V output therefore the output impedence of the 12V supply will be determined by IR losses, the dynamic impedance of the rectifying diodes, and the coupling efficiency between the inductor windings. This impedance is not reduced by the loop gain, as it is at the 5V output. As a result, the time constant of the response at this output will be considerably longer.

The fast response of the 5V output and the relatively slow response of the 12V output are illustrated in Figure 11 which shows three oscilliscope traces in response to a 3.0A load change at the 5V output. The upper trace is the response of the 5V output

which has been expanded and lowpass (< 15kHz) filtered slightly so the small signal loop characteristics can be seen. The trace below this is the 12V output's deviation due to cross-regulation limitations, the longer time constants involved are obvious. Both the fast response of the 5V loop, and the longer settling time of the 12V output are apparent in the third trace. This trace is the fed back correction signal at the UC1840's error amplifier output. From the middle trace the output impedance of the 12V supply can be estimated by noting the approximate 1ms time constant and dividing it by the $2000\mu F$ value of the 12V output filter capacitor. This gives a value of 0.5Ω for the output impedance. This agrees well with actual measurements of the 12V output's load regulation.

50 mV DV

0 5 V DIV

0.5 V/DIV

1 ms DIV
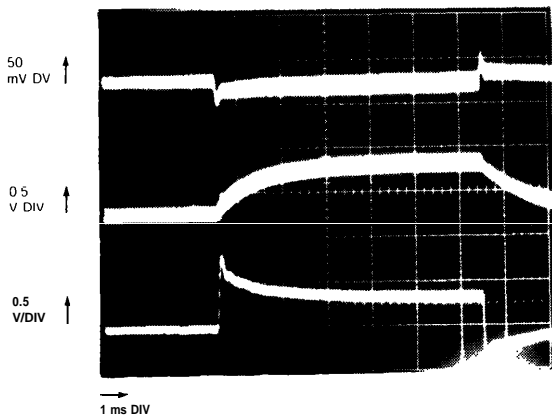
FIGURE 11: The Transient Response of the 5V Output (Top Trace), to a 3.0A Step Load Change Reflects the Extended Bandwidth of the 5V Loop. The Open-Loop 12V Output (Middle), Responds to the Effects of Cross Regulation. The Feedback Error Signal (Lower) Coupled Through the UC1901 is Measured at the UC1840 Error Amp. Output.
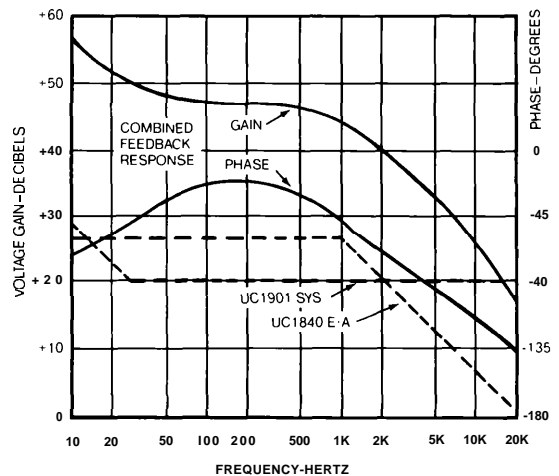
7e. The Feedback Response

Plotted in Figure 12 is the response of the feedback network. Also plotted are the asymptotic gain lines of the two contributing gain blocks, the UC1901 response (from 5V output to detector output) and the UC1840 error amp response (detector output to the PWM control voltage). The UC1901's error amplifier is run open loop at DC but is quickly rolled off to 8dB. With the 12dB of modulator gain, the UC1901 feedback system has a broadband gain of 20dB. A pole at 16kHz is added to reduce the gain through the UC1901 error amplifier at the 60kHz switching frequency. As mentioned earlier, excessive gain at the switching frequency can "use up" the dynamic range of the UC1901's AM output.

The UC1901 is operated with a carrier frequency of 500kHz. The coupling transformer, a Coilcraft E3493A, (double E core, bobbin wound construction), has a magnetizing inductance of 2.1mH. At 500kHz the peak current required to drive the primary winding is only $475\mu A$ per peak volt. The reflected load current is kept much smaller. This allows the transformer to be easily driven from the UC1901 driver outputs. The E3493A is widely used as a common mode line choke, and is rated for V.D.E. and U.L. isolation requirements. The transformer has a current rating of 2A, greatly exceeding the requirements of this application. Even though the device is larger than some alternatives, its availability and high volume pricing, as well as its isolation capability, make it a very suitable choice.

At the output of the transformer the diode-capacitor detector is referenced, along with the inverting input of the UC1840 error amplifier, to the UC1840's 5V reference. The operating point of the detector is fixed at 0.5V by the divider formed by R_{16} and R_{17} in Figure 9. This in turn sets the operating point of the carrier, with a detector diode drop of 0.5V, at about 1V peak. This level is reflected back through the one-to-one transformer to the UC1901 outputs. A 1V operating point is approximately at the center of the devices dynamic range.

The load current at the detector output is $50\mu A$, set by the 0.5V operating level and R_{16}. The peak to peak detector ripple, at 500kHz, across the $.0015\mu F$ holding capacitor is about 35mV. The gain through the UC1840 error amplifier at 500kHz is -26dB,


FIGURE 12: Local Feedback Around the UC1901 and 1840 Error Amplifiers is Used to Obtain the Desired Feedback Response.

attenuating the ripple to less than 2mV at the error amplifier output.

The response of the UC1840 error amplifier is flat out to 1kHz where the gain is rolled off to set the loop's 0db frequency. The DC gain is kept as high as possible, to fix the detector operating point, without actually having a series integrating capacitor in the feedback. If both the UC1901 and the UC1840 error amplifiers are run open loop at DC, with series R-C networks to set the AC gain, the total phase margin at low frequencies can become small or nonexistent. The result can be instability or, more likely, a peaked closed loop response that can increase the low frequency noise level of the supply.

The distribution of gain between the UC1901 and UC1840 error amplifiers is somewhat, although not entirely, arbitrary. Keeping the 500kHz ripple at the PWM comparator input below a certain level puts restrictions on the AC gain of the PWM's error amplifier To much AC gain through the UC1901's amplifier can degrade the supply's transient response under large signal conditions. A suitable distribution for any application will, more than likely, be an iterative procedure. A simple computer or programmable calculator program can be a great tool when massaging these aspects of a design.
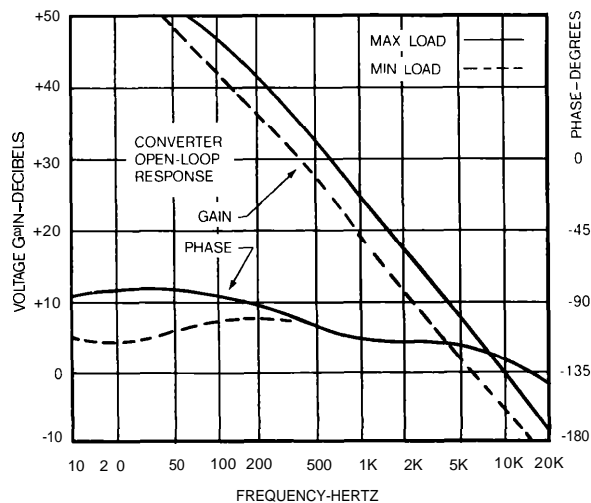
FIGURE 13: The Over-All Open-Loop Response of the Supply Will Determine the Supply's Over-All Stability and Small Signal Transient Response.

The overall open-loop responses, plotted in Figure 13, will not vary significantly except as indicated with load. The desired loop bandwidth has been achieved with an adequate phase margin of > 50°.

The result is a supply with very repeatable, as well as stable, operating characteristics. The same type of analysis for determining the required feedback response can be used in applying the UC1901 to any type of isolated closed loop supply. The choice of coupling transformer and carrier frequency used with the UC1901 should be based on individual system requirements.

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