A Fourth Generation of LCD Backlight Technology
Component and Measurement Improvements Refine Performance

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PREFACE

Current generation portable computers and instruments utilize backlit LCDs (liquid crystal displays). These displays have also appeared in applications ranging from medical equipment to automobiles, gas pumps and retail terminals. Cold cathode fluorescent lamps (CCFLs) provide the highest available efficiency for backlighting the display. These lamps require high voltage AC to operate, mandating an efficient high voltage DC/AC converter. In addition to good efficiency, the converter should deliver the lamp drive in sine wave form. This is desirable to minimize RF emissions. Such emissions can cause interference with other devices, as well as degrading overall operating efficiency. The sine wave excitation also provides optimal current-to-light conversion in the lamp. The circuit should permit lamp control from zero to full brightness with no hysteresis or "pop-on," and must also regulate lamp intensity vs power supply variations.

The small size and battery-powered operation associated with LCD equipped apparatus mandate low component count and high efficiency for these circuits. Size constraints place severe limitations on circuit architecture and long battery life is usually a priority. Laptop and handheld portable computers offer an excellent example. The CCFL and its power supply are responsible for almost 50% of the battery drain. Additionally, these components, including PC board and all hardware, usually must fit within the LCD enclosure with a height restriction of 0.25 inches.

A practical, efficient LCD backlight design is a classic study of compromise in a transduced electronic system. Every aspect of the design is interrelated, and the physical embodiment is an integral part of the electrical circuit. The choice and location of the lamp, wires, display housing and other items have a major effect on electrical characteristics. The greatest care in every detail is required to achieve a practical high efficiency LCD backlight. Getting the lamp to light is just the beginning!

First generation backlights were crude, with poor performance in almost all areas. LTC (Linear Technology Corporation) has introduced feedback stabilization and optimized lamp driving configurations in three successive generations of technology. The effort has culminated in dedicated ICs for backlight driving.

This fourth publication reviews our recent work in components and measurement techniques applicable to LCD backlighting. Theoretical considerations are presented with practical suggestions, remedies and circuits. As always, we welcome reader comments, questions and requests for consultation.

> **UPDATE**
> While LED backlights have mostly replaced CCFLs, this Application Note shows proper design and layout for high voltage inverters

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INTRODUCTION

This scribing marks the fourth LTC publication in as many years concerning LCD illumination.[1] The extraordinary user response to previous efforts has resulted in a continuing LCD backlight development effort by our company. This level of interest, along with significant performance advances since the last publication, justifies further discussion of LCD backlighting.

Development of attractive solutions for LCD illumination has necessitated the longest sustained LTC application engineering effort to date. A single circuit in a 1991 publication (*Measurement and Control Circuit Collection*, LTC Application Note 45, June 1991) has resulted in four years of continuous investigation, summarized in three successive, dedicated publications.

The impetus for all this bustle has been an overwhelming and continuously ascending reader response. Practical, high performance LCD backlighting solutions are needed in a wide range of applications. The optical, transductive and electronic aspects combine (conspire?) to present an extraordinarily challenging problem. The LCD backlight problem's interdisciplinary nature, along with highly interactive effects, provides an exquisitely subtle engineering exercise. Backlights present the most complex set of interdependencies the author has ever encountered. Our academic interest in this challenge is, of course, well patinaed with capitalistic intent. Substantial comfort arrives with the certainty that the audience is similarly acculturated.

This publication includes pertinent information from previous efforts in addition to updated sections and a large body of new material. The partial repetition is a small penalty compared to the benefits of text flow, completeness and time efficient communication. Older material has been altered, abridged or augmented as appropriate, while simultaneously introducing new findings. Previous work has emphasized obtaining and verifying high efficiency. This characteristic is still quite desirable, but other backlight requirements have become evident. These include low voltage operation, improved system interface, minimization of display-induced losses, circuitry compaction and better measurement/optimization techniques. These advances have been enabled by development of new ICs and instrumentation.

Finally, this preamble must appreciate the text's arrangement and review by various LTC personnel and customers. They transmuted a psychotic uproar of a manuscript into this finessed presentation. Hopefully, readers will join the author in applause.

Note 1: Previous publications are annotated in References 1, 18 and 25.

PERSPECTIVES ON DISPLAY EFFICIENCY

The LCD displays currently available require two power sources, a backlight supply and a contrast supply. The display backlight is the single largest power consumer in a typical portable apparatus, accounting for almost 50% of battery drain with the display at maximum intensity. As such, every effort must be expended to maximize backlight efficiency.

Study of LCD energy management should consider the problem from an interdisciplinary viewpoint. The backlight presents a cascaded energy attenuator to the battery (Figure 1). Battery energy is lost in the electrical-to-electrical conversion to high voltage AC to drive the CCFL. This section of the energy attenuator is the most efficient; conversion efficiencies exceeding 90% are possible. The CCFL, although the most efficient electrical-to-light converter available today, has losses exceeding 80%. Additionally, the optical transmission efficiency of present displays is under 10% for monochrome with color types much lower.
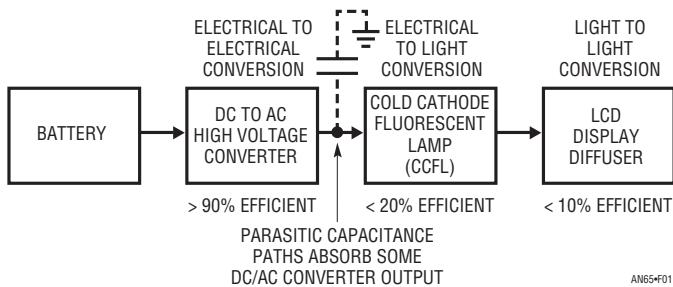

Figure 1. Backlit LCD Display Presents a Cascaded Energy Attenuator to the Battery. DC/AC Conversion Is Significantly More Efficient Than Energy Conversions in Lamp and Display

The very high DC/AC conversion efficiency highlights some significant issues. Anything that improves energy transfer in the other "attenuator" areas will have greater impact than further electrical efficiency improvements. Additional improvements in electrical efficiency, while certainly desirable, are reaching the point of diminishing returns. Clearly, overall backlight efficiency gains must come from lamp and display improvements.

There is very little electrical workers can do to improve lamp and display efficiency besides call attention to the problems (see the following sections on lamps and displays).[2] Improvements are, however, possible in related areas. In particular, the *form* of drive applied to the lamp is quite critical. The waveshape supplied to the lamp influences its current-to-light conversion efficiency. Thus, dissimilar waveforms containing equivalent power can produce different amounts of lamp light output. This implies that a more *electrically* efficient inverter with a nonoptimal output waveshape could produce less light than a "less efficient" inverter with a more appropriate waveform. Experiment reveals this to be true. As such, distinction between electrical and photometric efficiency is necessary and requires attention.

Another practical area where improvement is possible is transmission of inverter drive to the lamp. The high frequency AC waveform is subject to losses due to parasitic capacitances in the wiring and display. Controlling the parasitic capacitances and the manner in which lamp drive is applied can yield significant efficiency improvement.

Practical methods addressing both aforementioned areas are contained in subsequent sections of this publication.

Cold Cathode Fluorescent Lamps (CCFLs)

Any discussion of CCFL power supplies must consider lamp characteristics. These lamps are complex transducers, with many variables affecting their ability to convert electrical current to light. Factors influencing conversion efficiency include the lamp's current, temperature, drive waveform characteristics, length, width, gas constituents and the proximity to nearby conductors.

These and other factors are interdependent, resulting in a complex overall response. Figures 2 through 8 show some typical characteristics. A review of these curves hints at the difficulty in predicting lamp behavior as operating conditions vary. The lamp's current, temperature and warm-up time are clearly critical to emission, although electrical efficiency may not necessarily correspond to the best optical efficiency point. Because of this, both electrical and photometric evaluation of a circuit is often required. It is possible, for example, to construct a CCFL circuit with 94% electrical efficiency which produces less light output

Note 2: "Call attention to the problems" constitutes a pleasant euphemism for complaining. This publication's section on displays presents such complaints in visual form along with suggested remedies.

than an approach with 80% electrical efficiency. (See Appendix L, "A Lot of Cut Off Ears and No Van Goghs—Some Not-So-Good Ideas.") Similarly, the performance of a very well matched lamp/circuit combination can be severely degraded by a lossy display enclosure or excessive high voltage wire lengths. Display enclosures with too much conducting material near the lamp have huge losses due to capacitive coupling. A poorly designed display enclosure can easily degrade efficiency by 20%. High voltage wire runs typically cause 1% loss per inch of wire.
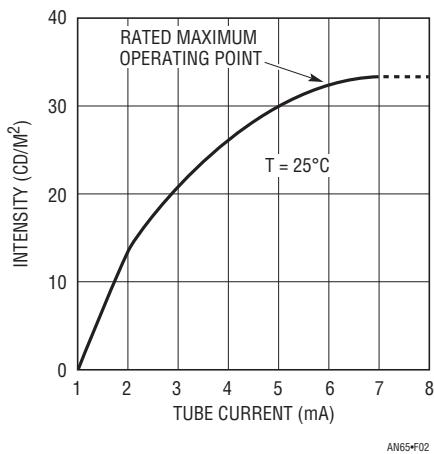
Figure 2. Emissivity for a Typical 5mA Lamp. Curve Flattens Badly Above 6mA
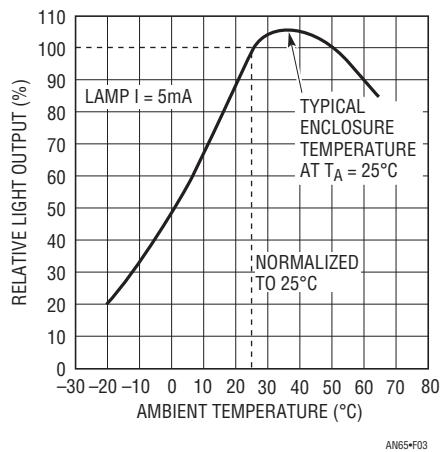
Figure 3. Ambient Temperature Effects on Emissivity of a Typical 5mA Lamp. Lamp and Enclosure Must Come to Thermal Steady State Before Measurements Are Made
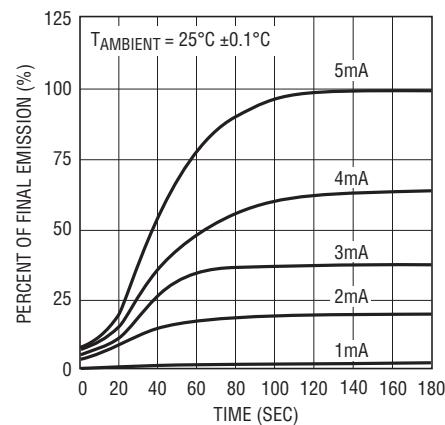
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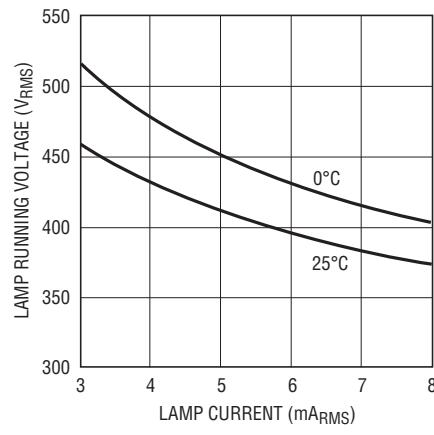
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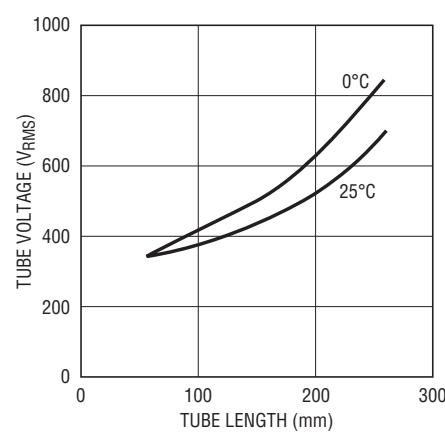
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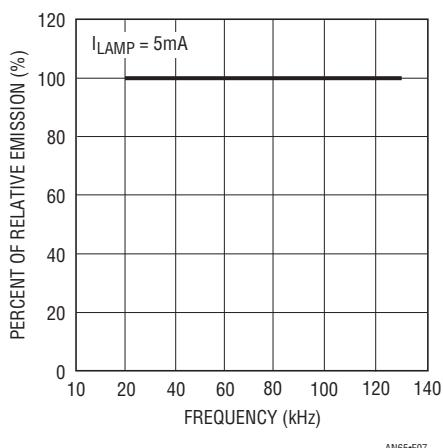
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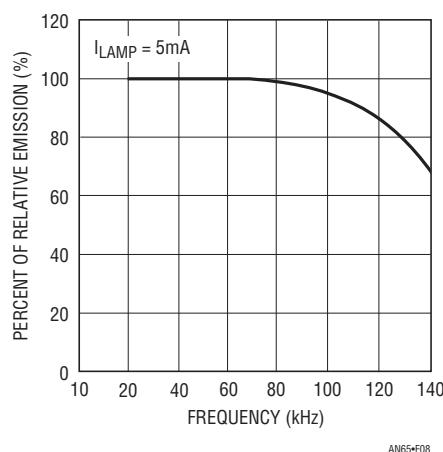
Figure 8. Figure 7's Lamp Shows Significant Emission vs Drive Frequency Degradation When Mounted in a Display. Cause Is Frequency-Dependent Loss Due to Display's Parasitic Capacitance Paths

The optimum drive frequency is determined by display and wiring losses, not lamp characteristics. Figure 7 shows lamp emissivity is essentially flat over a wide frequency range. Figure 8 shows results with the same lamp mounted in a typical display.

The apparent emissivity fall-off at high frequencies is caused by reduced lamp current due to parasitic capacitance-induced losses. As frequency increases, the display's parasitic capacitance diverts progressively more energy, lowering lamp current and emission. This effect is sometimes misinterpreted, leading to the mistaken conclusion that lamp emissivity degrades with increasing frequency.

CCFL Load Characteristics

These lamps are a difficult load to drive, particularly for a switching regulator. They have a "negative resistance" characteristic; the starting voltage is significantly higher than the operating voltage. Typically, the start voltage is about 1000V, although higher and lower voltage lamps are common. Operating voltage is usually 300V to 500V, although other lamps may require different potentials. The lamps will operate from DC, but migration effects within the lamp will quickly damage it. As such, the waveform must be AC. No DC content should be present.

Figure 9a shows an AC driven lamp's characteristics on a curve tracer. The negative resistance-induced "snap-back" is apparent. In Figure 9b another lamp, acting against the

(9a)

(9b)
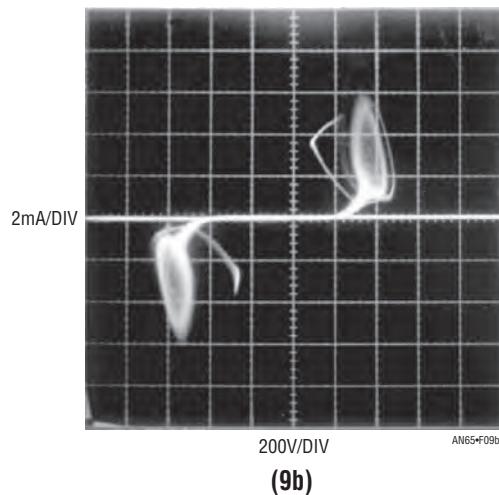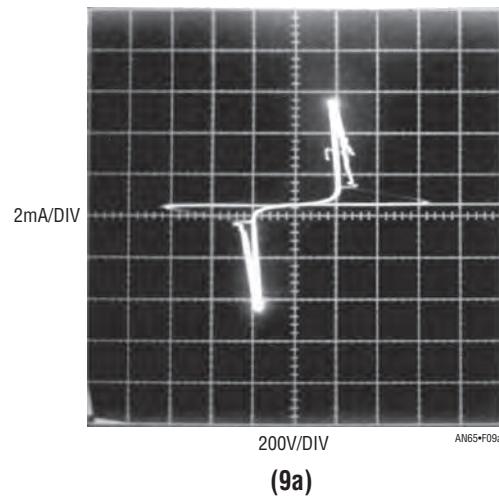
Figure 9. Negative Resistance Characteristic for Two CCFL Lamps. "Snap-Back" Is Readily Apparent, Causing Oscillation in 9b. These Characteristics Complicate Power Supply Design

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curve tracer's drive, produces oscillation. These tendencies, combined with the frequency compensation problems associated with switching regulators, can cause severe loop instabilities, particularly on start-up. Once the lamp is in its operating region it assumes a linear load characteristic, easing stability criteria. Lamp operating frequencies are typically 20kHz to 100kHz and a sine-like waveform is preferred. The sine drive's low harmonic content minimizes RF emissions, which could cause interference and efficiency degradation.[3] A further benefit to the continuous sine drive is its low crest factor and controlled rise times, which are easily handled by the CCFL. CCFL's RMS current-to-light output efficiency and lifetime degrades with fast rise, high crest factor drive waveforms.[4]

Display and Layout Losses

The physical layout of the lamp, its leads, the display housing and other high voltage components are integral parts of the circuit. Placing the lamp into a display introduces pronounced electrical loading effects which must be considered. Poor layout can easily degrade efficiency by 25% and higher layout-induced losses have been observed. Producing an optimal layout requires attention to how losses occur. Figure 10 begins our study by examining potential parasitic paths between the transformer's output and the lamp. Parasitic capacitance to AC ground from any point between the power supply output and the lamp creates a path for undesired current flow. Similarly, stray coupling from any point along the lamp's length to AC ground induces parasitic current flow. All parasitic current flow

is wasted, causing the circuit to produce more energy to maintain desired current flow in the lamp. The high voltage path from the transformer to the display housing should be as short as possible to minimize losses. A good rule of thumb is to assume 1% efficiency loss per inch of high voltage lead. Any PC board traces, ground or power planes should be relieved by at least 1/4" in the high voltage area. This not only prevents losses but eliminates arcing paths.

Parasitic losses associated with lamp placement within the display housing require attention. High voltage wire length within the housing must be minimized, particularly for displays using metal construction. Ensure that the high voltage is applied to the shortest wire(s) in the display. This may require disassembling the display to verify wire length and layout. Another loss source is the reflective foil commonly used around lamps to direct light into the actual LCD. Some foil materials absorb considerably more field energy than others, creating loss. Finally, displays supplied in metal enclosures tend to be lossy. The metal absorbs significant energy and an AC path to ground is unavoidable. Direct grounding of metal enclosed displays further increases losses. Some display manufacturers have addressed this issue by relieving the metal in the lamp area with other materials. Losses introduced by the display are substantial and vary widely with different displays. These

Note 3: Many of the characteristics of CCFLs are shared by so-called "Hot" cathode fluorescent lamps. See Appendix A, "Hot" Cathode Fluorescent Lamps.
Note 4: See Appendix L, "A Lot of Cut Off Ears and No Van Goghs—Some Not-So-Great Ideas."
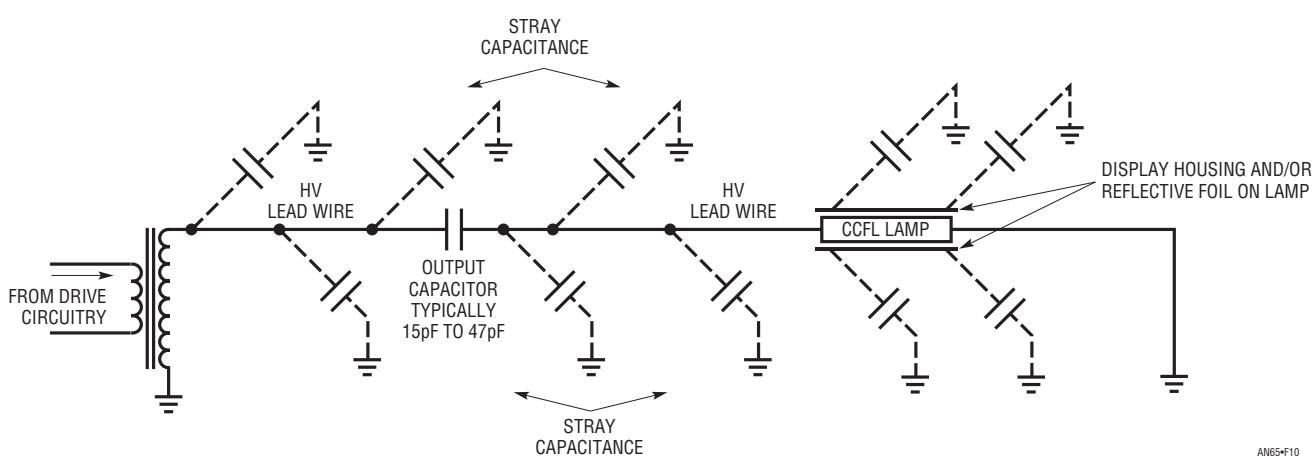
Figure 10. Loss Paths Due to Stray Capacitance in a Practical LCD Installation. Minimizing These Paths Is Essential for Good Efficiency

losses not only degrade overall efficiency, but complicate meaningful determination of the lamp current. Figure 11 shows effects of distributed parasitic capacitance loss paths on lamp current. The display housing and reflective foil-induced loss paths provide a continuous conduit for loss current flow. This results in a continuously varying value of "lamp current" along the lamp's length. In cases where one end of the lamp is at or near ground, the current fall-off is greatest in the lamp's high voltage regions. Although parasitic capacitance is usually uniformly distributed, its effect becomes far greater as voltage scales up.

These effects illustrate why designing around lamp specifications is such a frustrating exercise. Display vendors typically call out lamp operating parameters based on information received from the lamp manufacturer. Lamp vendors often determine operating characteristics in a completely different enclosure, or none at all. This set of uncertainties complicates design effort. The only viable solution is to determine lamp performance with the display of interest. This is the only practical way to maximize performance and ensure against overdriving the lamp, which wastes power and shortens lamp life.

In general, the display introduces parasitics which degrade performance. Latter portions of this text discuss some compensatory techniques, but the deleterious effects of display parasitics dominate practical backlight design.

There are some benefits to lossy displays. One advantage of display parasitics is that they effectively lower lamp breakdown voltage. The parasitic shunt capacitance along the tube's length forms a distributed electrode, effectively shortening the breakdown path, lowering the lamp's turn-on voltage. This accounts for the fact that many display mounted lamps start at lower voltages than the "naked" lamp breakdown voltage specification suggests. This effect aids low temperature start-up (see Figures 5 and 6).

A second potential advantage of distributed parasitic lamp capacitance is enhancement of low current operation. In some cases extended dimming range is possible because the parasitics provide a more evenly distributed field along the lamp's length. This tends to maintain illumination along the lamp's entire length at low operating currents, allowing low luminosity operation.
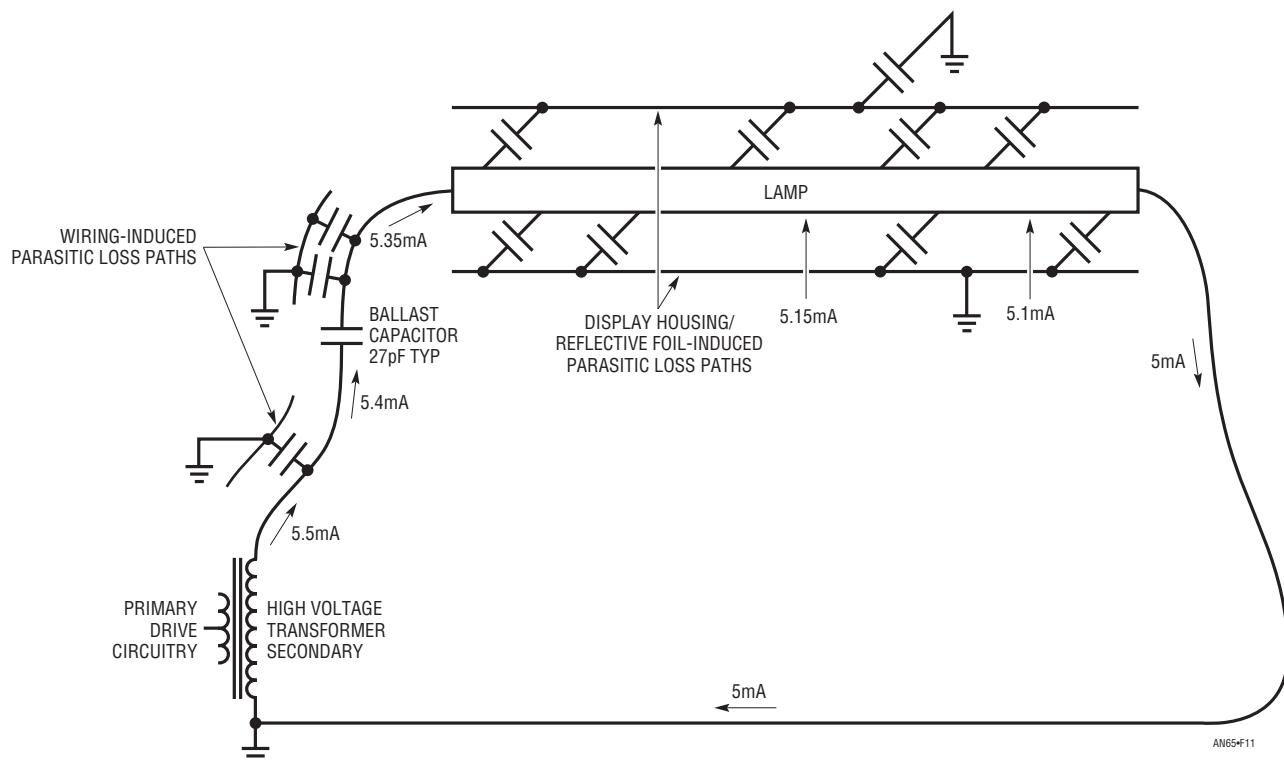
Figure 11. Distributed Parasitic Capacitances in a Practical Situation Cause Continuous Downward Shift in Measured "Lamp Current." In This Case 0.5mA Is Lost to Parasitic Paths. Most Loss Occurs in High Voltage Regions

The lessons here are clear. A thorough characterization of lamp/display losses is crucial to understanding trade-offs and obtaining the best possible performance. The highest efficiency "in system" backlights have been produced by careful attention to these issues. In some cases the entire display enclosure was re-engineered for lower losses.

The display loss issue, central to backlight design, merits detailed attention. The following briefly commented photographs (Figures 12 through 32) illustrate a variety of display situations. Hopefully, this visual tour will alert display users and manufacturers to the problems involved, promoting appropriate action by both.
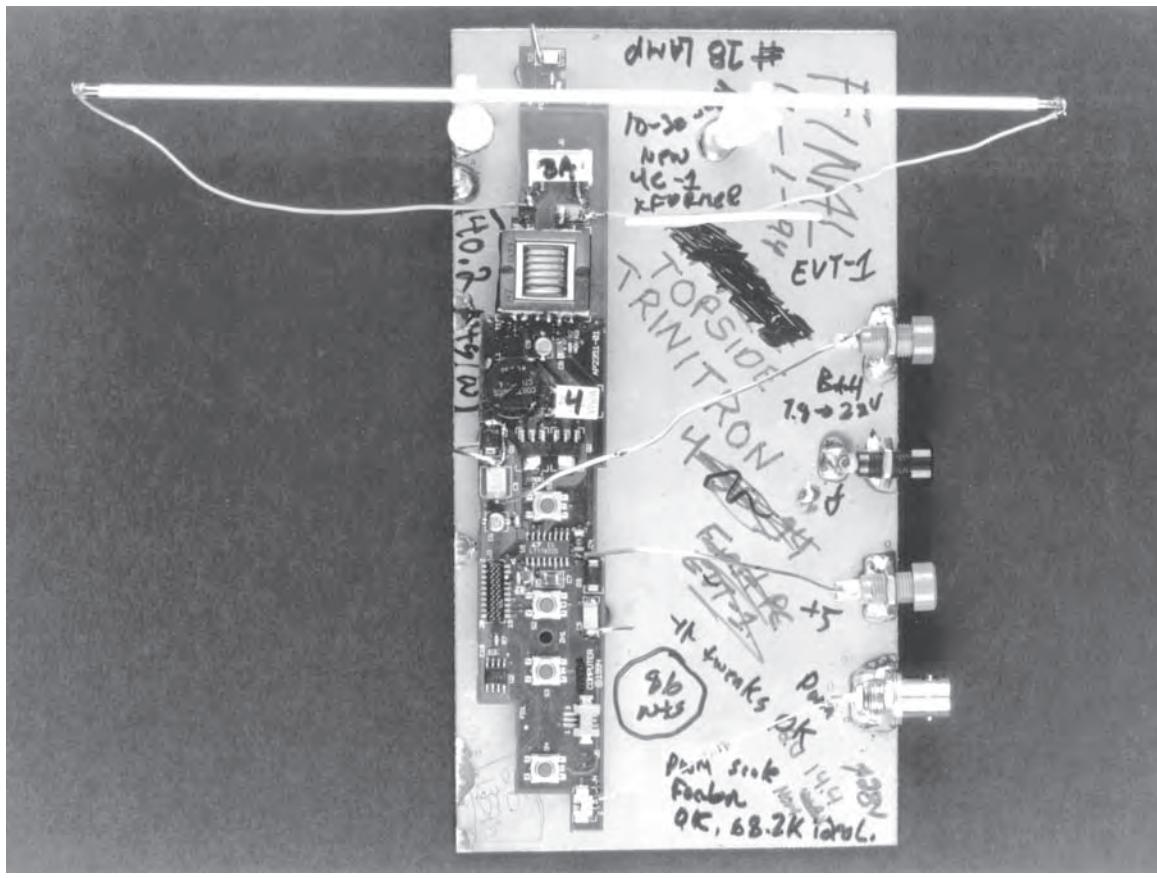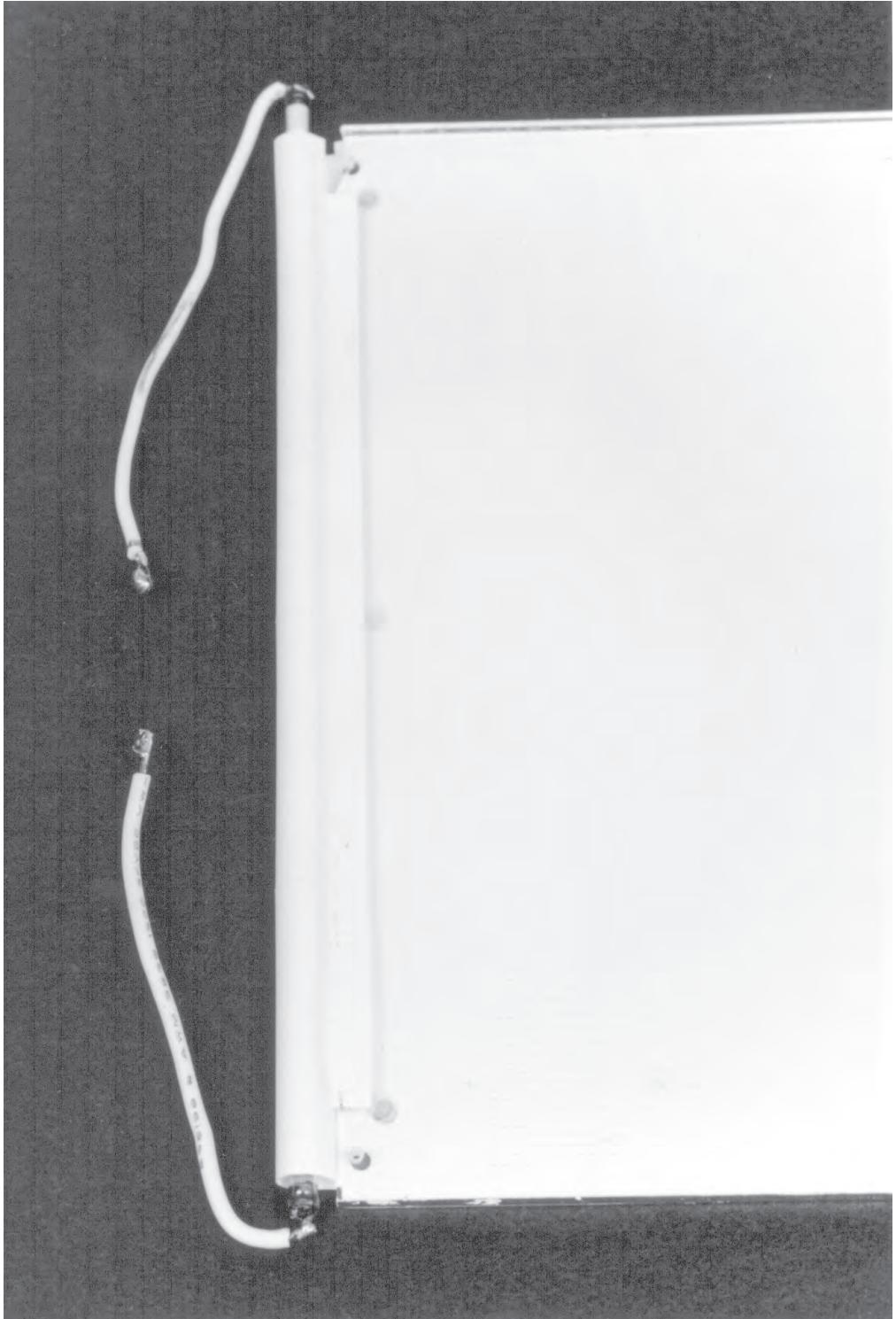
Figure 12. The Ideal Display Is No Display. Drive Electronics Connected to a "Naked" Lamp Simulates a Zero Loss Display. Note Nylon Stand-Offs. Results Obtained Have No Relationship to Practical Display Driving

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Figure 13. Measuring Lamp Wire to Display Frame Capacitance. Technique Gives Lead Wire-to-Frame Loss Information but Not Lamp-to-Foil-or-Frame Loss Data. Lamp Must Be Energized Before Its Parasitics Are Measurable
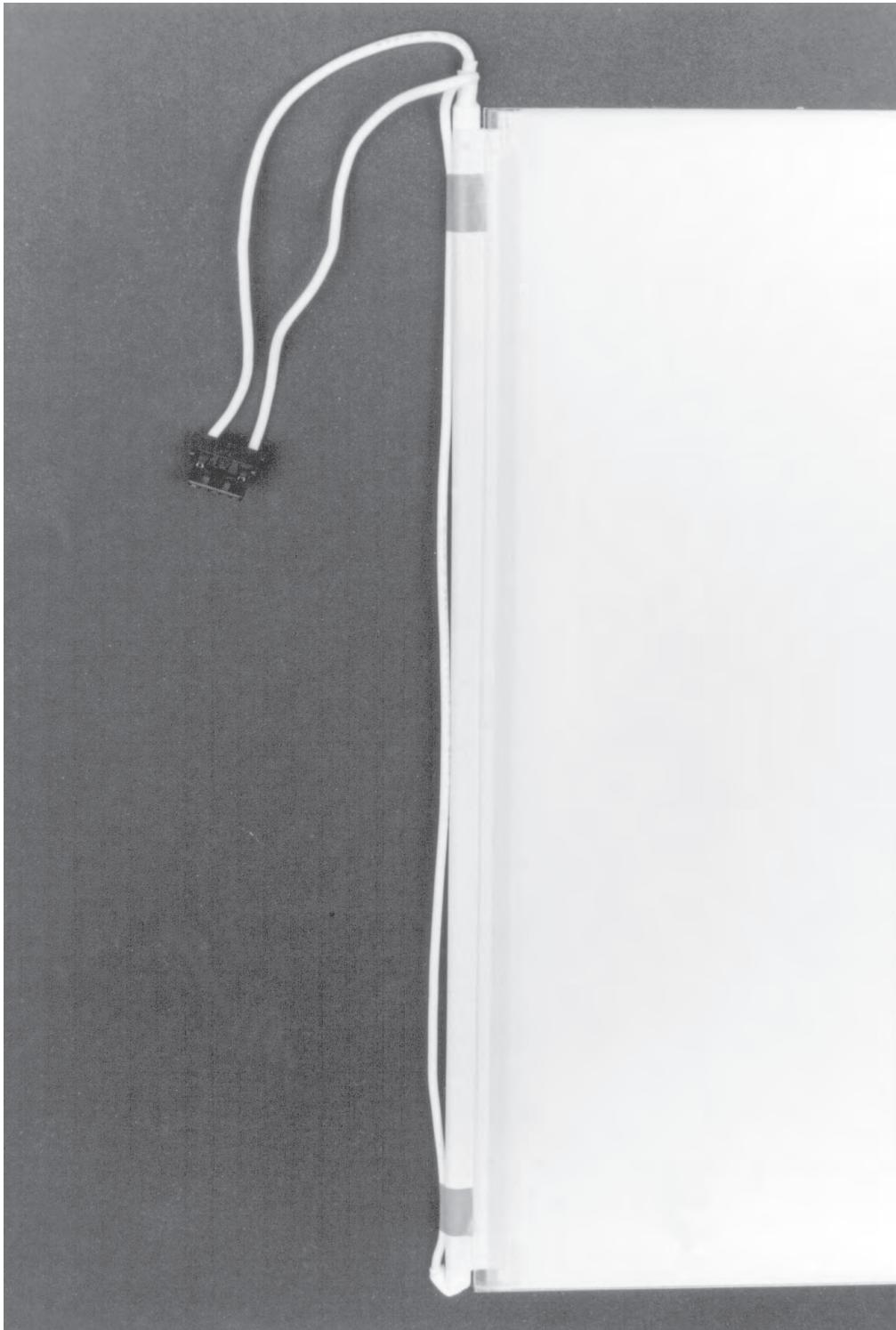
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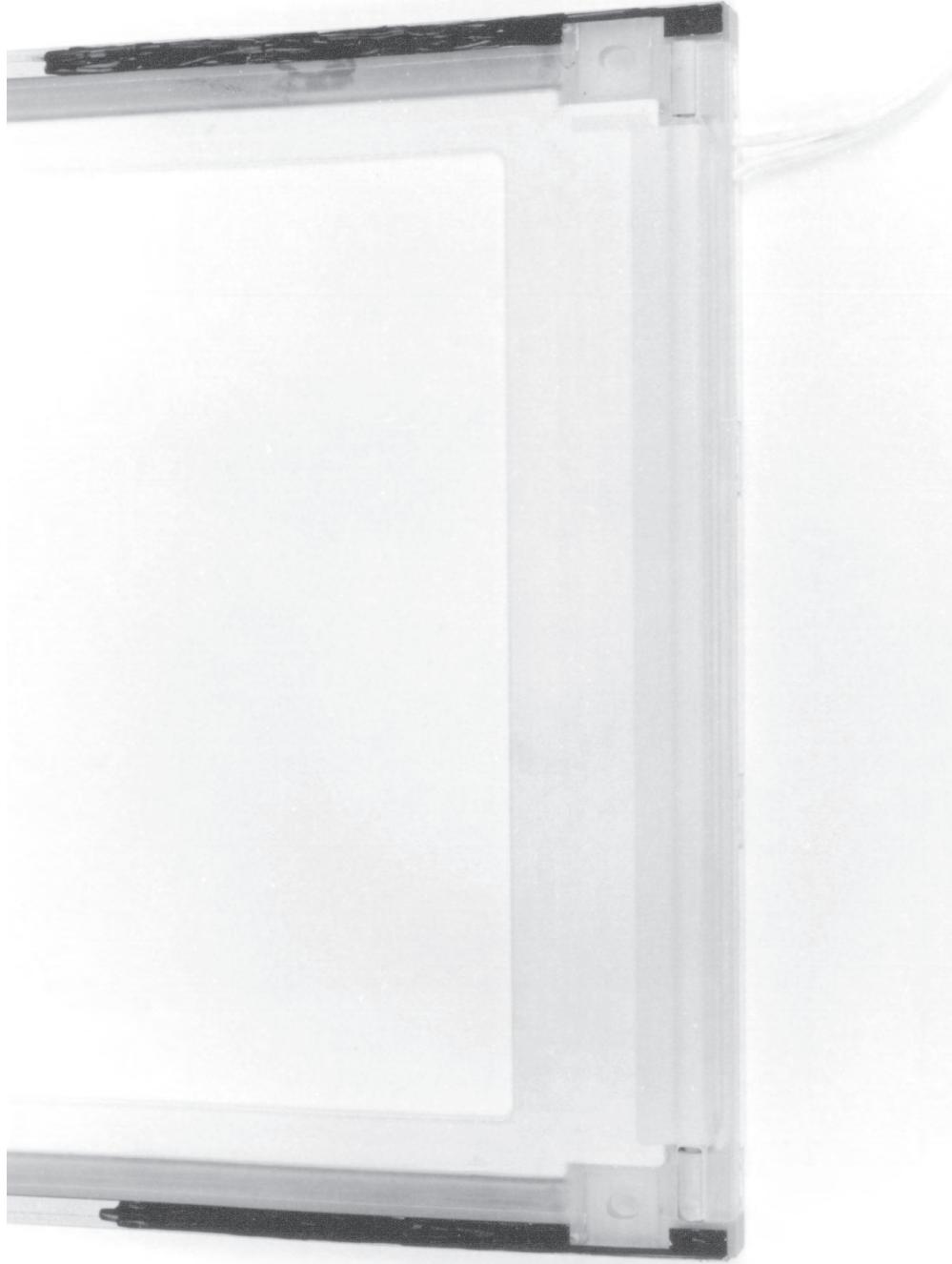
Figure 14. Low Loss Display Has No Metal in Lamp Region. Reflective Foil Floats from Ground and Has Low Absorption. Display Loss About 1.5%

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Figure 15. Another Low Loss Display Has Similar Characteristics to Figure 14. Running Long Wire Return Across Lamp Length Increases Loss to about 4%. Spacing Wire Away from Lamp Would Cut Loss by Half

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Figure 16. A Custom Designed, Extremely Low Loss Display. All Metal Is Eliminated in Lamp Area (Lower Portion of Photo). A Good Compromise Between Mechanical Strength and Loss Control

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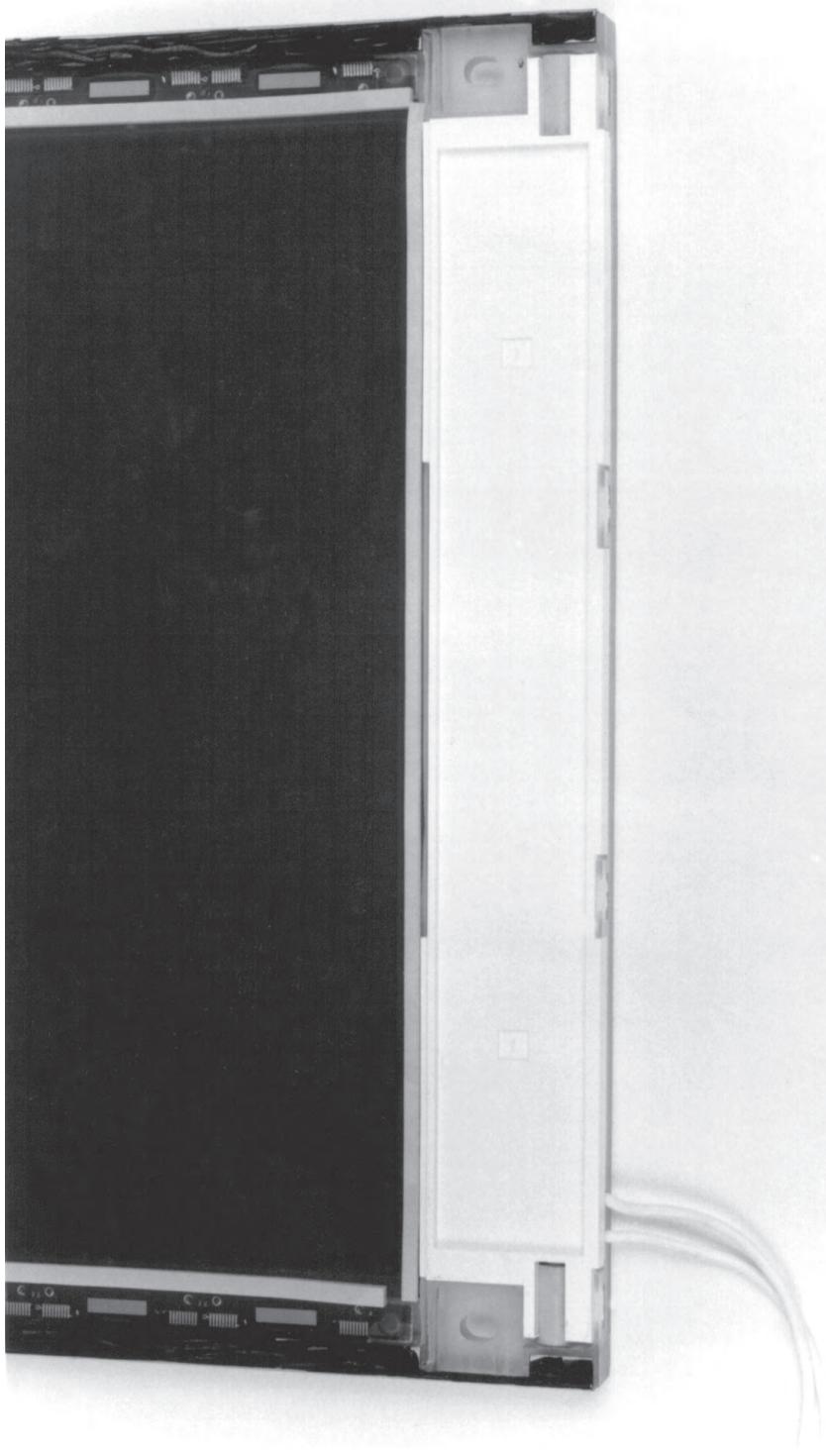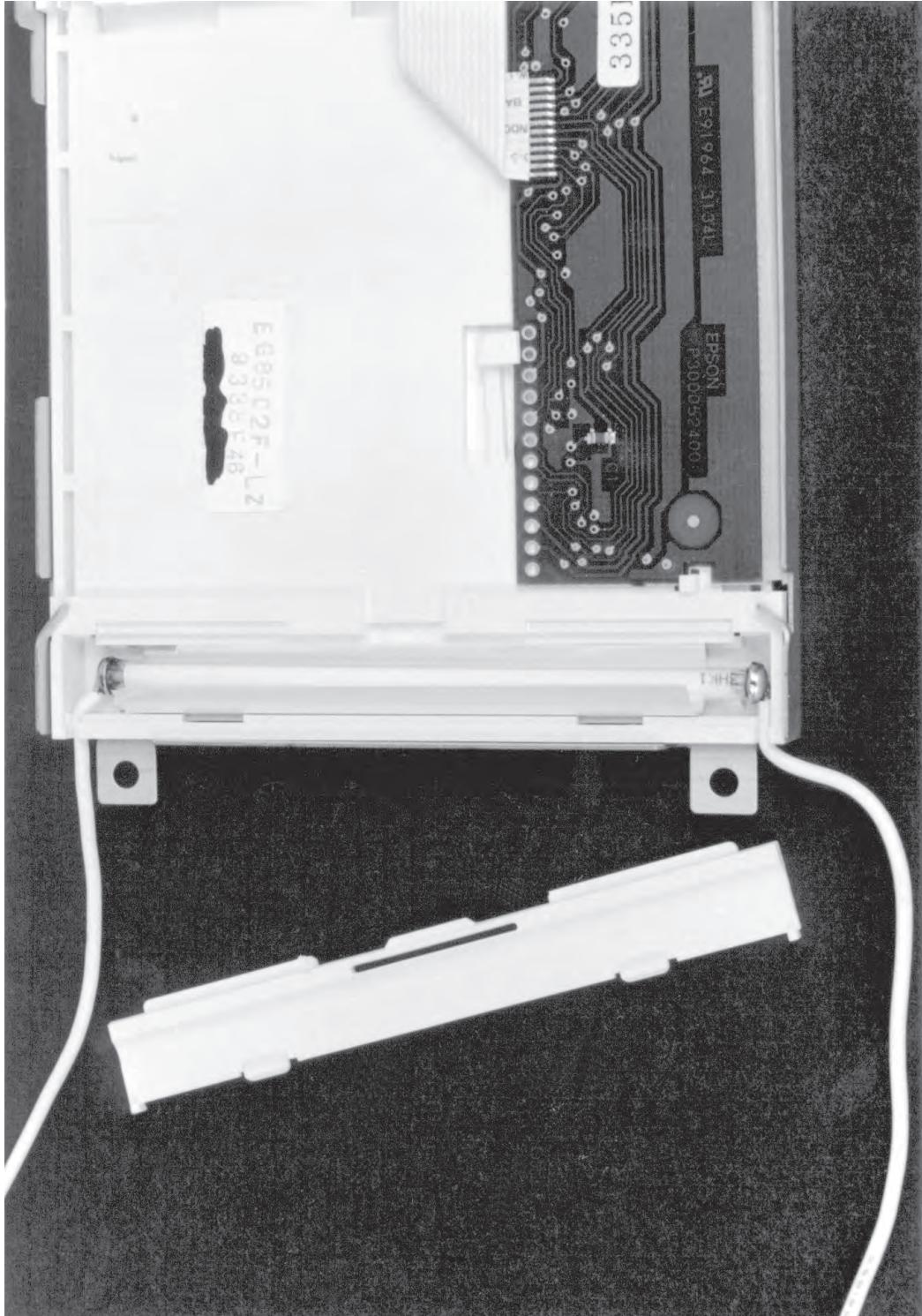
Figure 17. Figure 16's Reverse Side. All Metal Is Relieved in Lamp Area, Maintaining Low Losses. An Excellent, Practical Display

Figure 18. Plastic "Cocoon" Cuts Losses. Metallic Foil Is Absorptive but Floats from Grounded Display Frame. A Good Compromise with about 4% Loss

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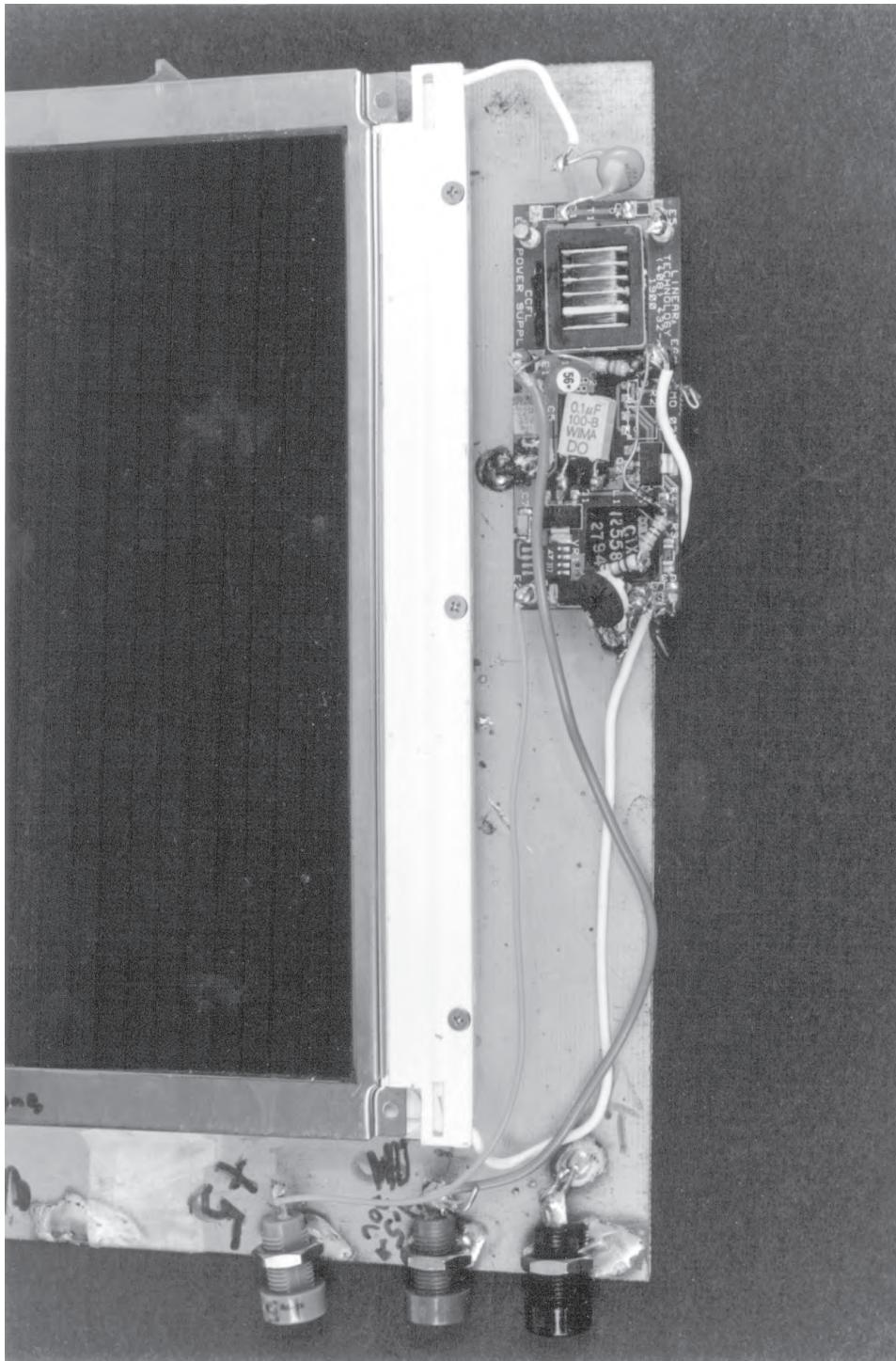
Figure 19. Plastic "Outrigger" Isolates Lamp from Metal Display Frame Loss Path
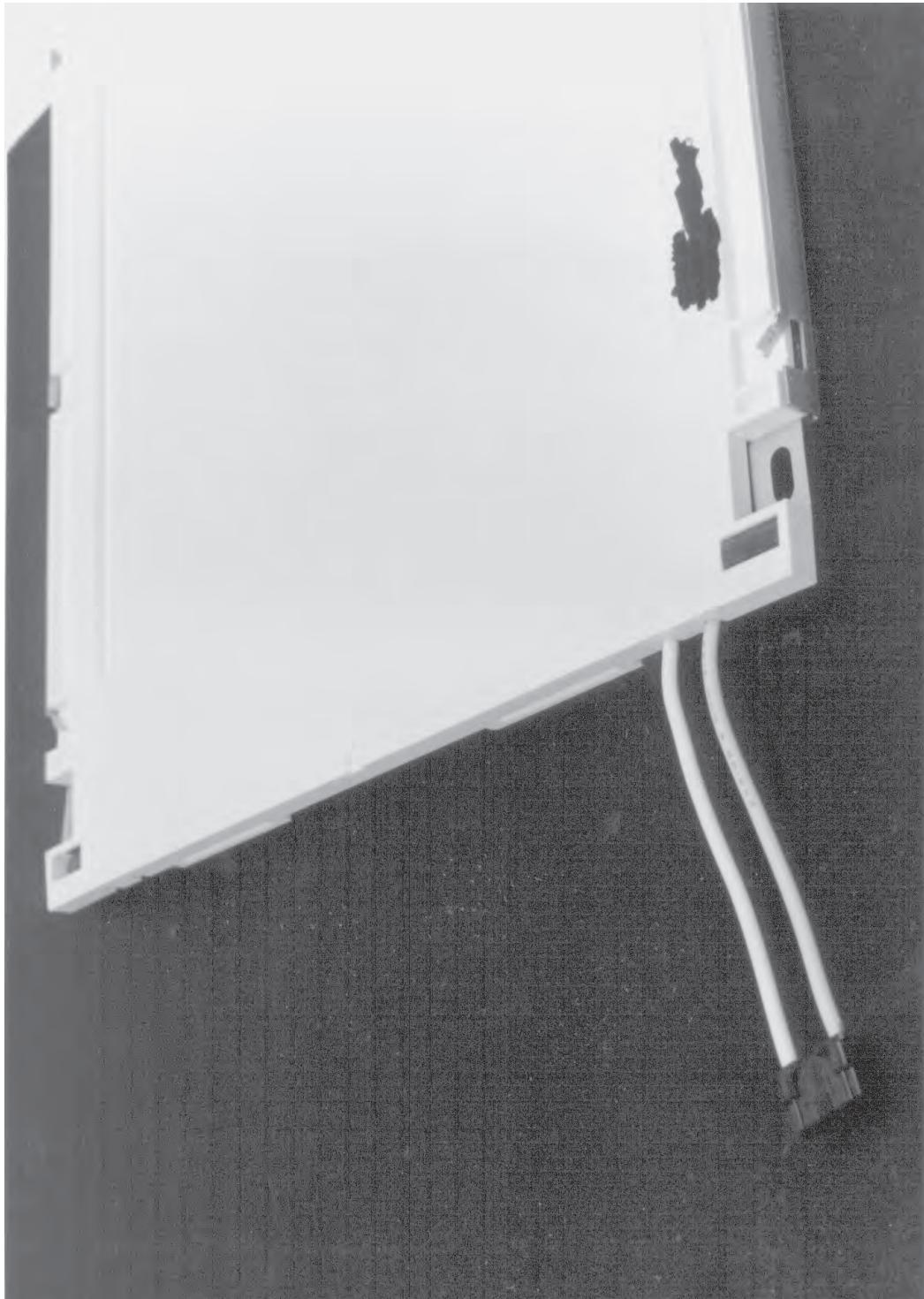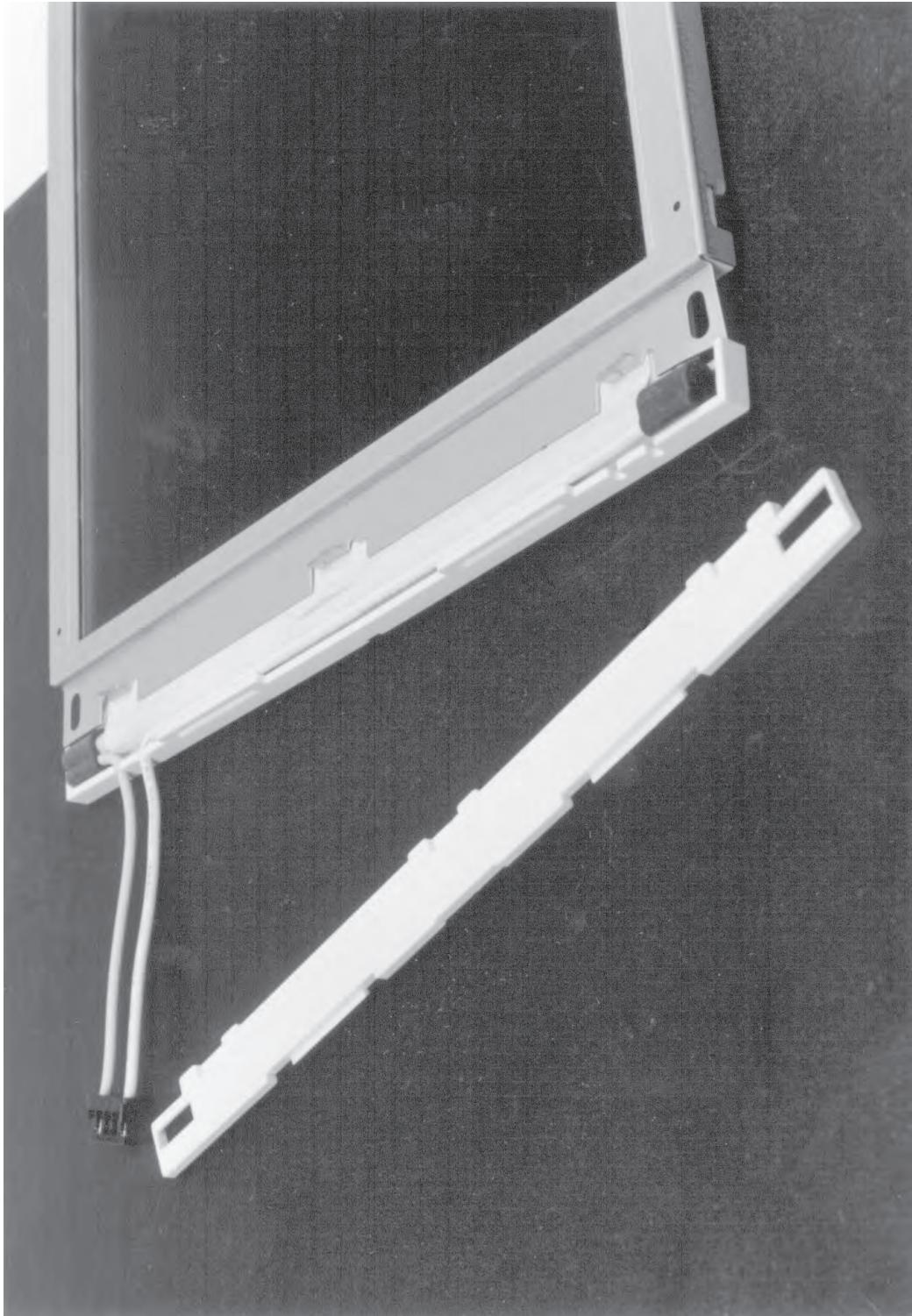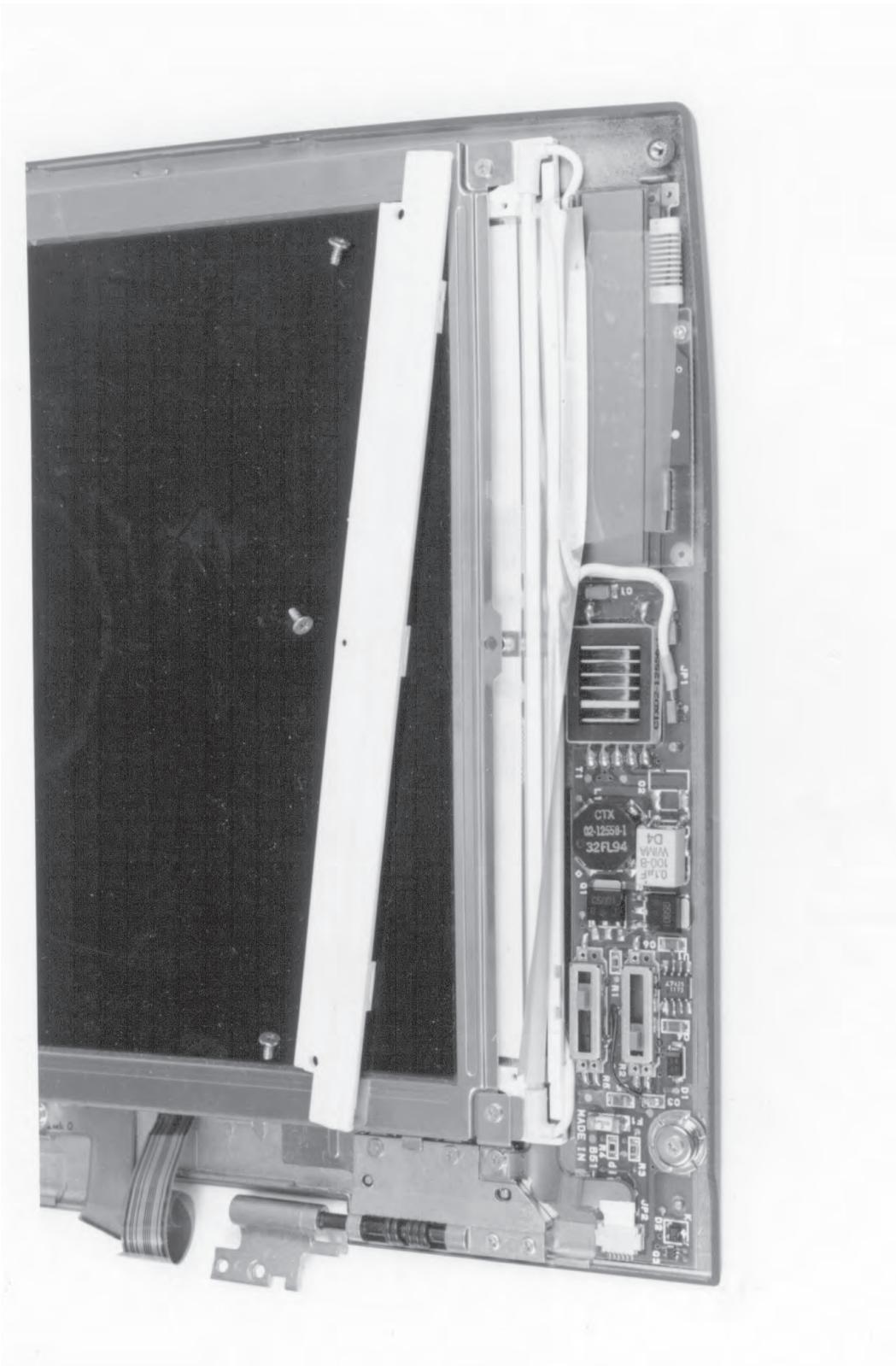
Figure 20. Plastic Isolates Lamp from Metal Frame in This Display's Rear View
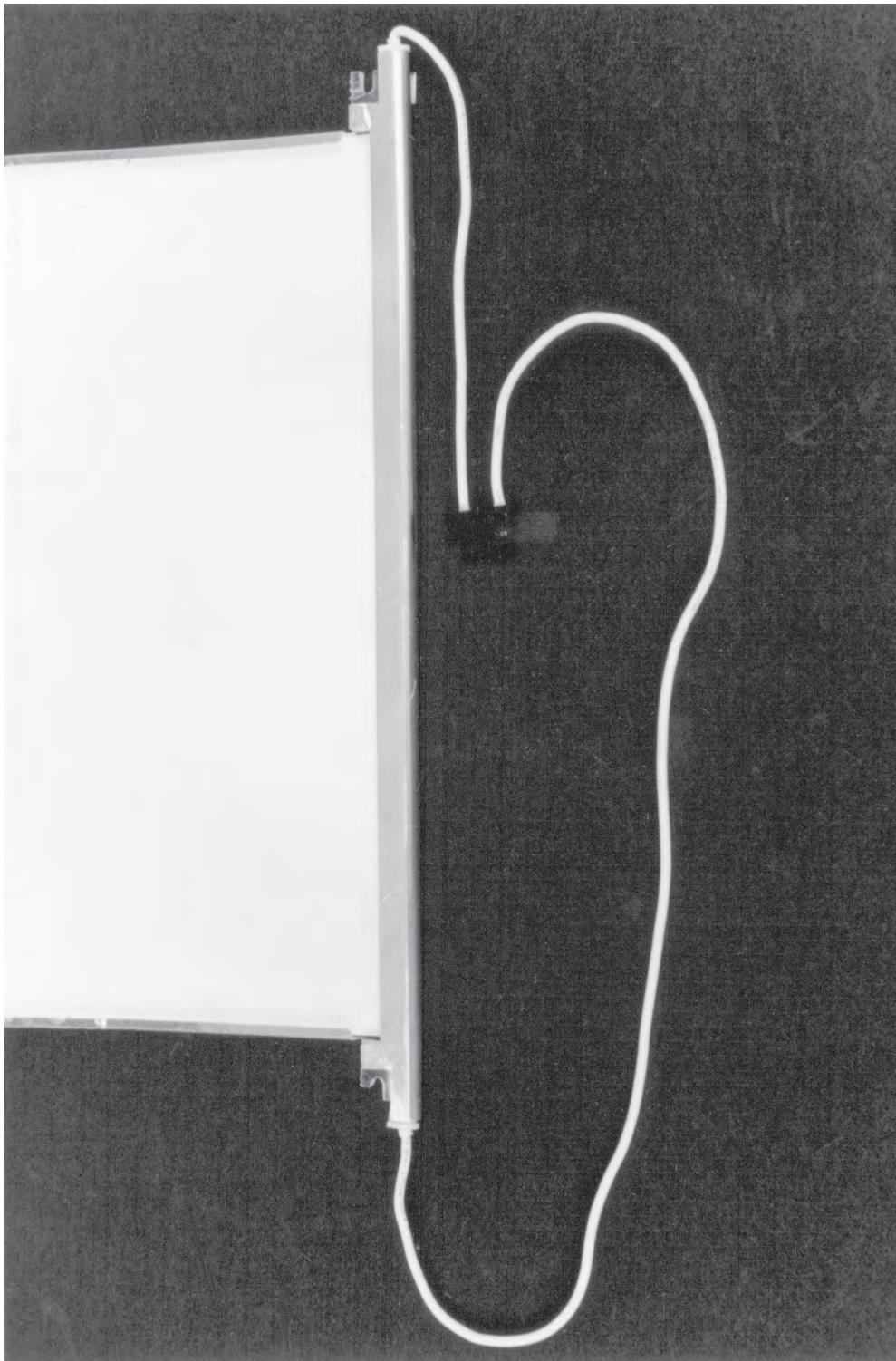
Figure 21. Figure 20's Display Front View Continues Plastic Isolation Treatment but Reflective Foil (over Lamp) Contacts Metal Frame. Massive Losses via This Path Cause Overall 12% Loss. Trimming Foil from Metal Cuts Loss to 4%
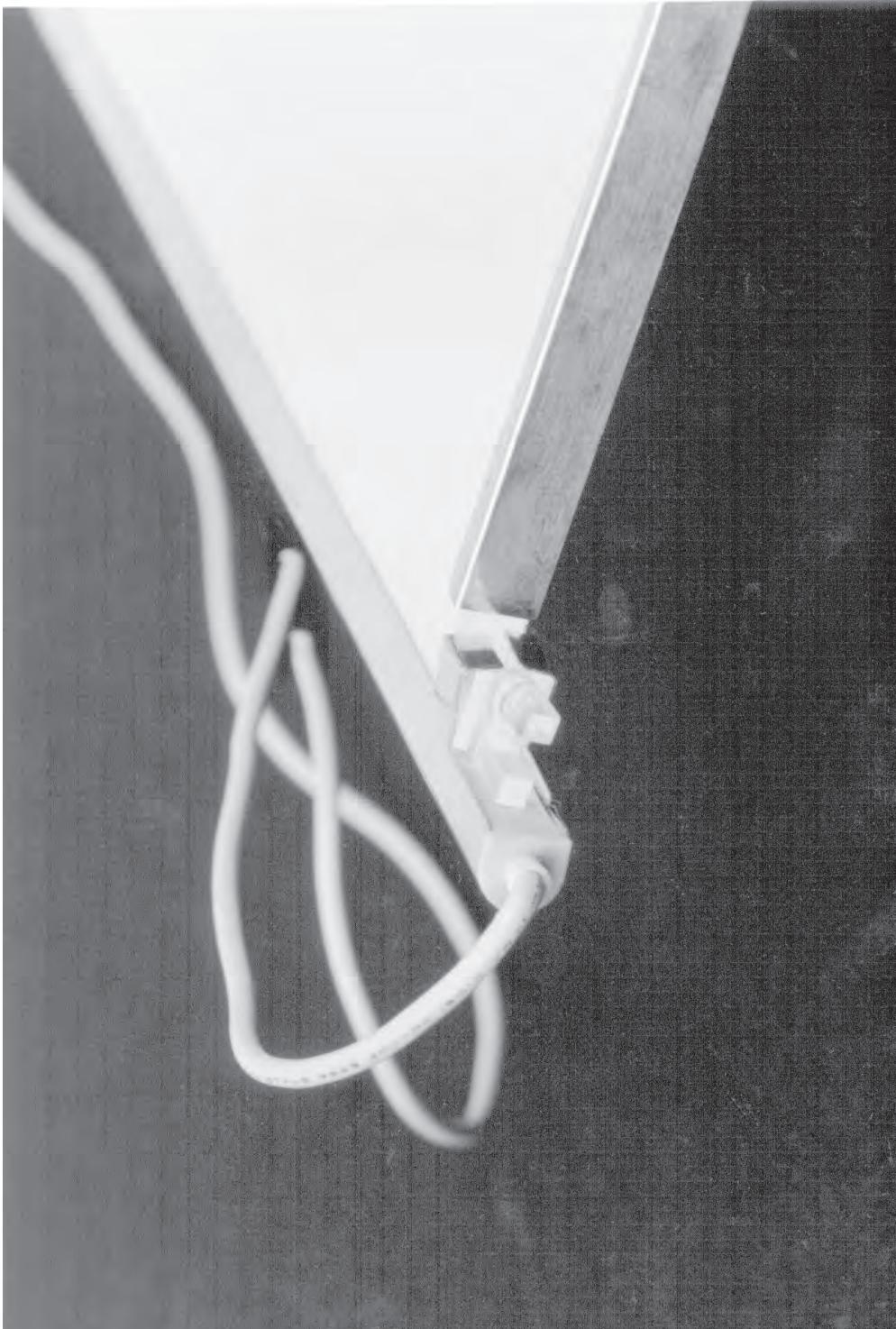
Figure 22. Another "Outrigged" Plastic Enclosure Suffers Foil Contacting Display's Frame Metal. Relieving Foil from Metal Cuts Losses from 13% to 6%. Poor Wire Routing (Lower Right) Causes 3% Loss
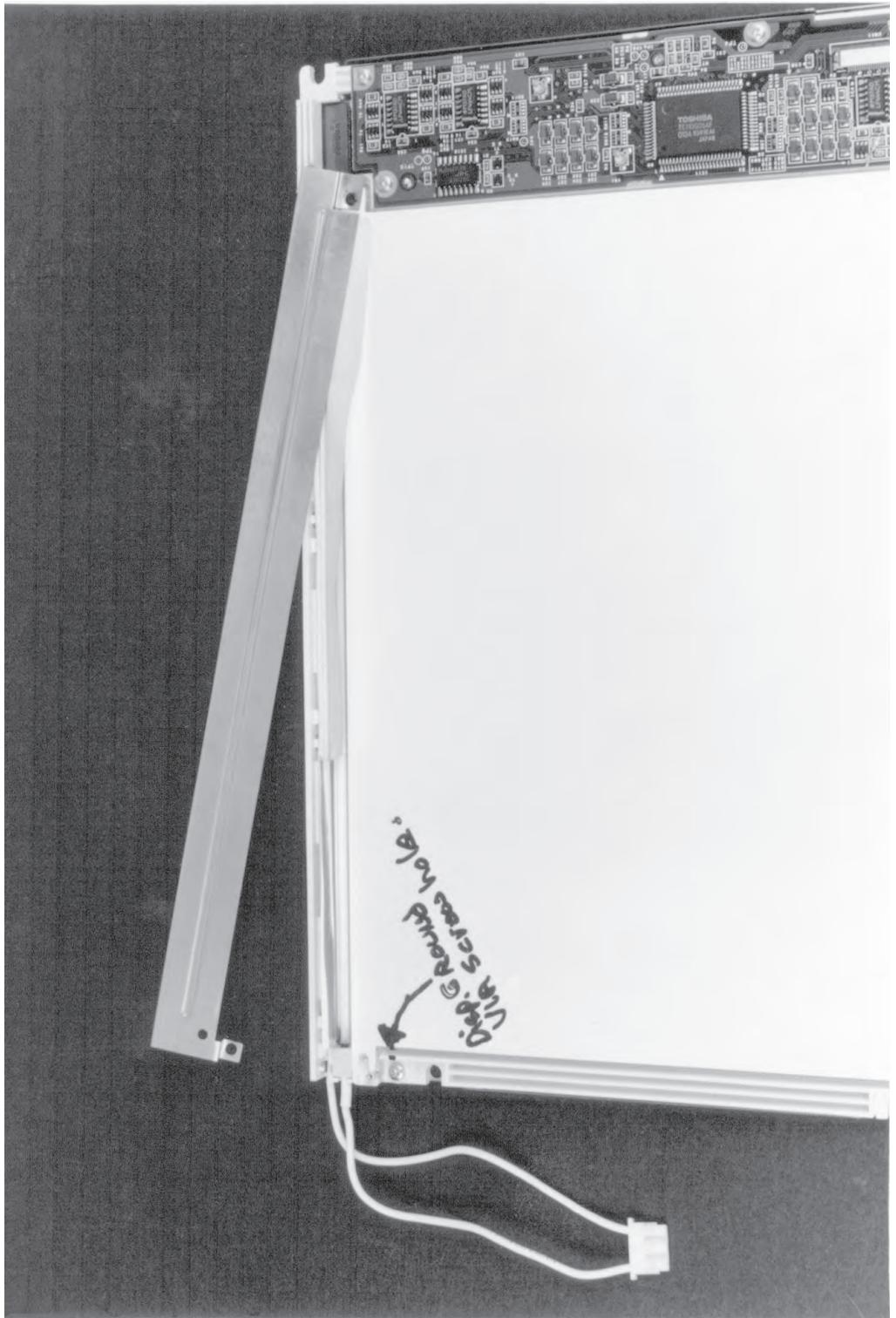
Figure 23. Isolation Slits (Center Right and Left) in Metal Reflector Prevent Losses to Grounded Metal Frame (Upper Right and Left). Overall Losses About 6%

Figure 24. Close-Up of Figure 23's Isolation Slit Construction. Secondary Benefit Is Control of Reflector-to-Lamp Distance, Minimizing Capacitance

Figure 25. Metal Cover over Lamp Causes 15% Loss. Replacing Cover Securing Screws with Nylon Types Floats Cover from Ground, Dropping Loss to 8%. Replacing Cover with Plastic Improves Loss to Only 3% . . . a 5X Improvement!
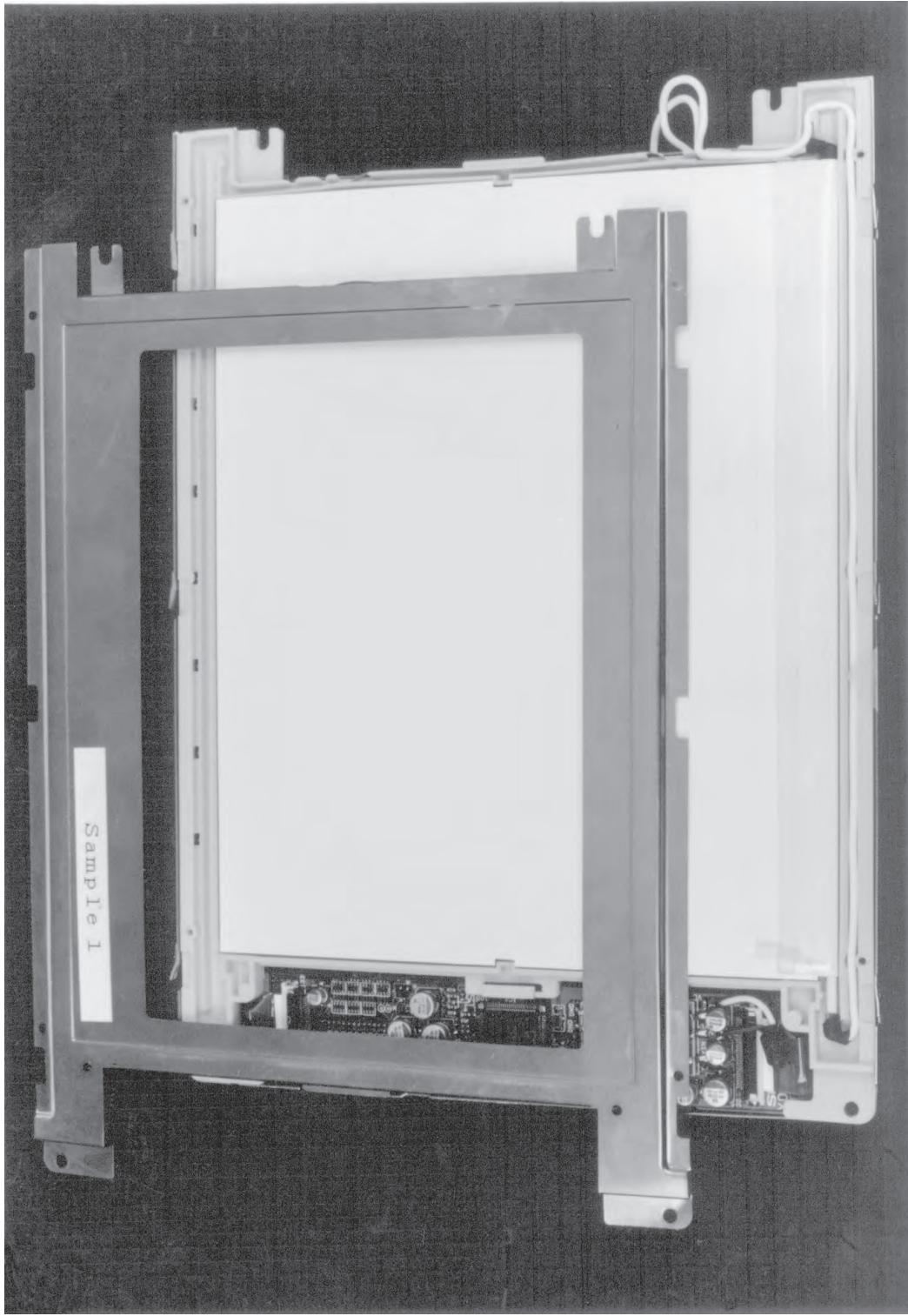
Figure 26. Huge Metal Area over Lamp Causes 14% Loss. Replacing Metal in Lamp Area with Plastic Cuts Loss to 6%
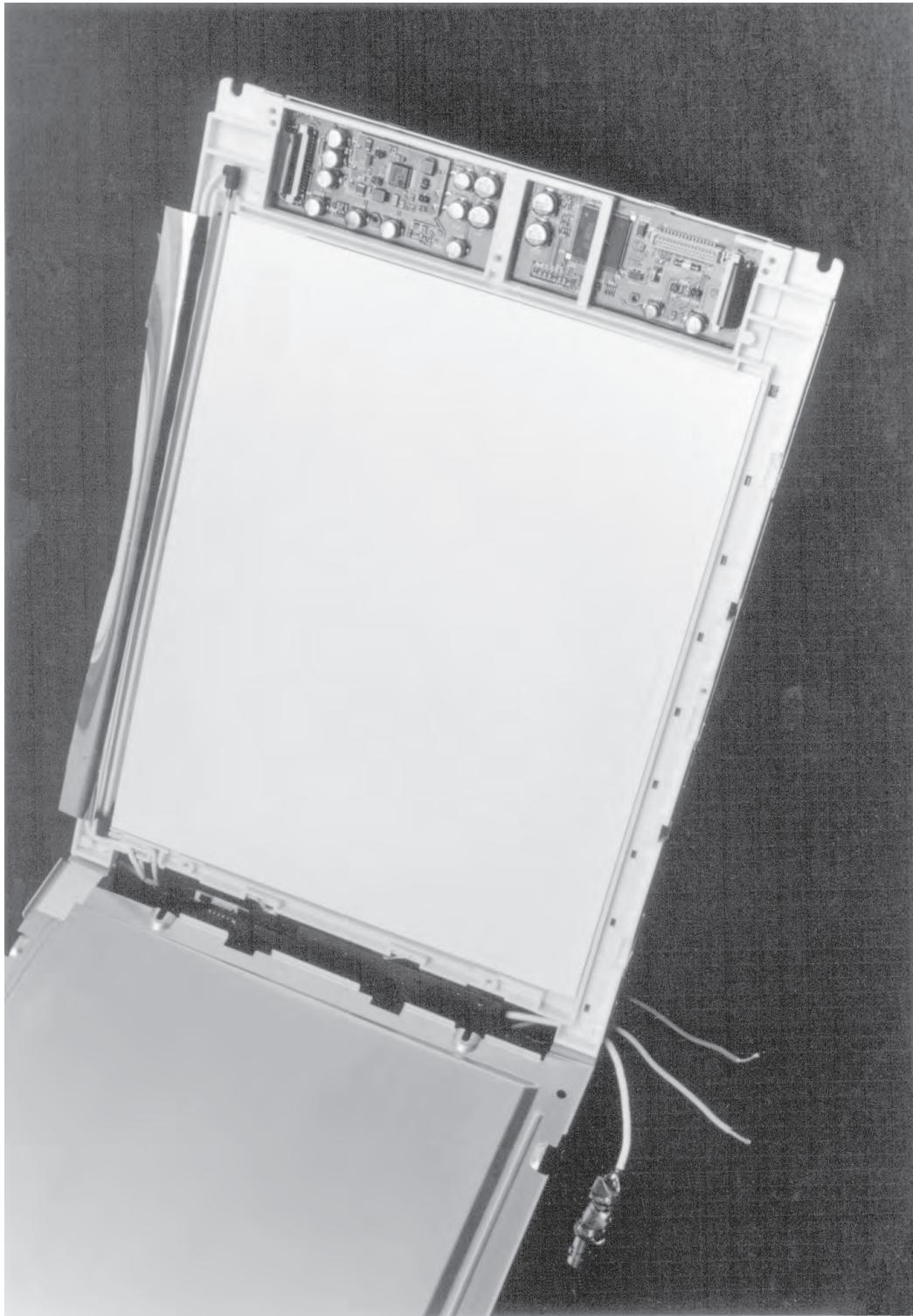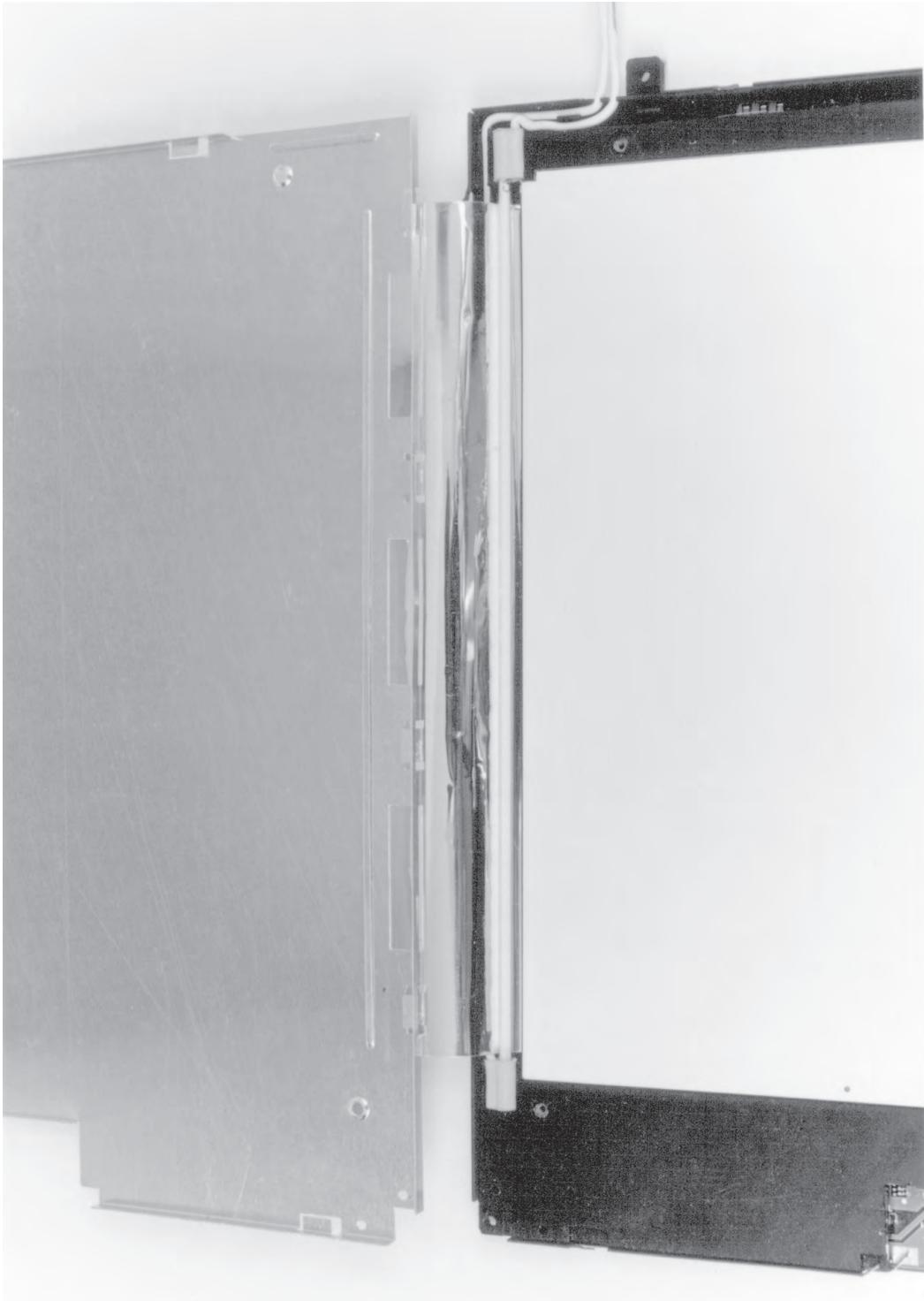
Figure 27. Metallic Foil over Lamp (Upper Center) Dumps Absorbed Energy to Metal Rear Cover. 16% Loss Results
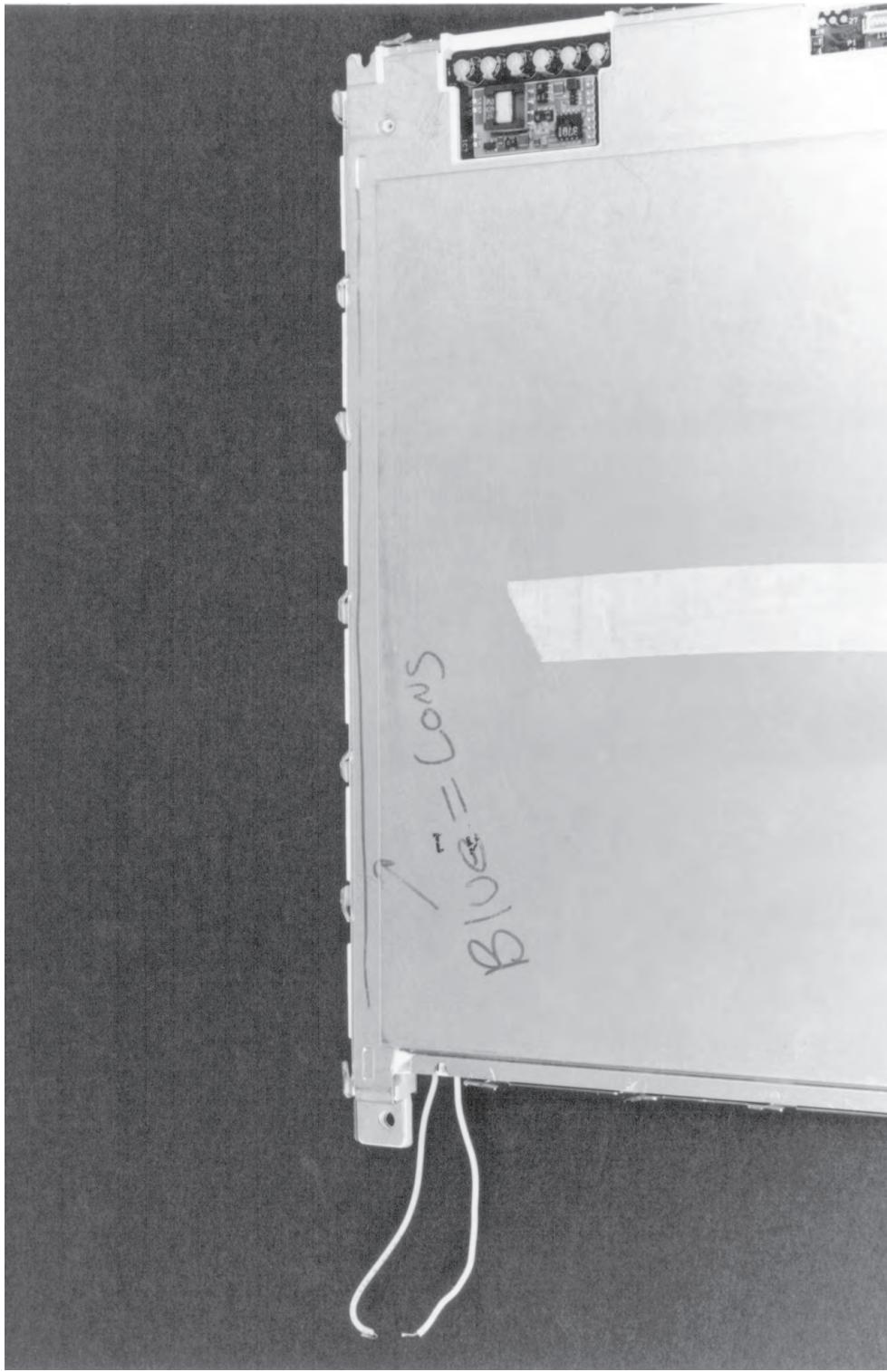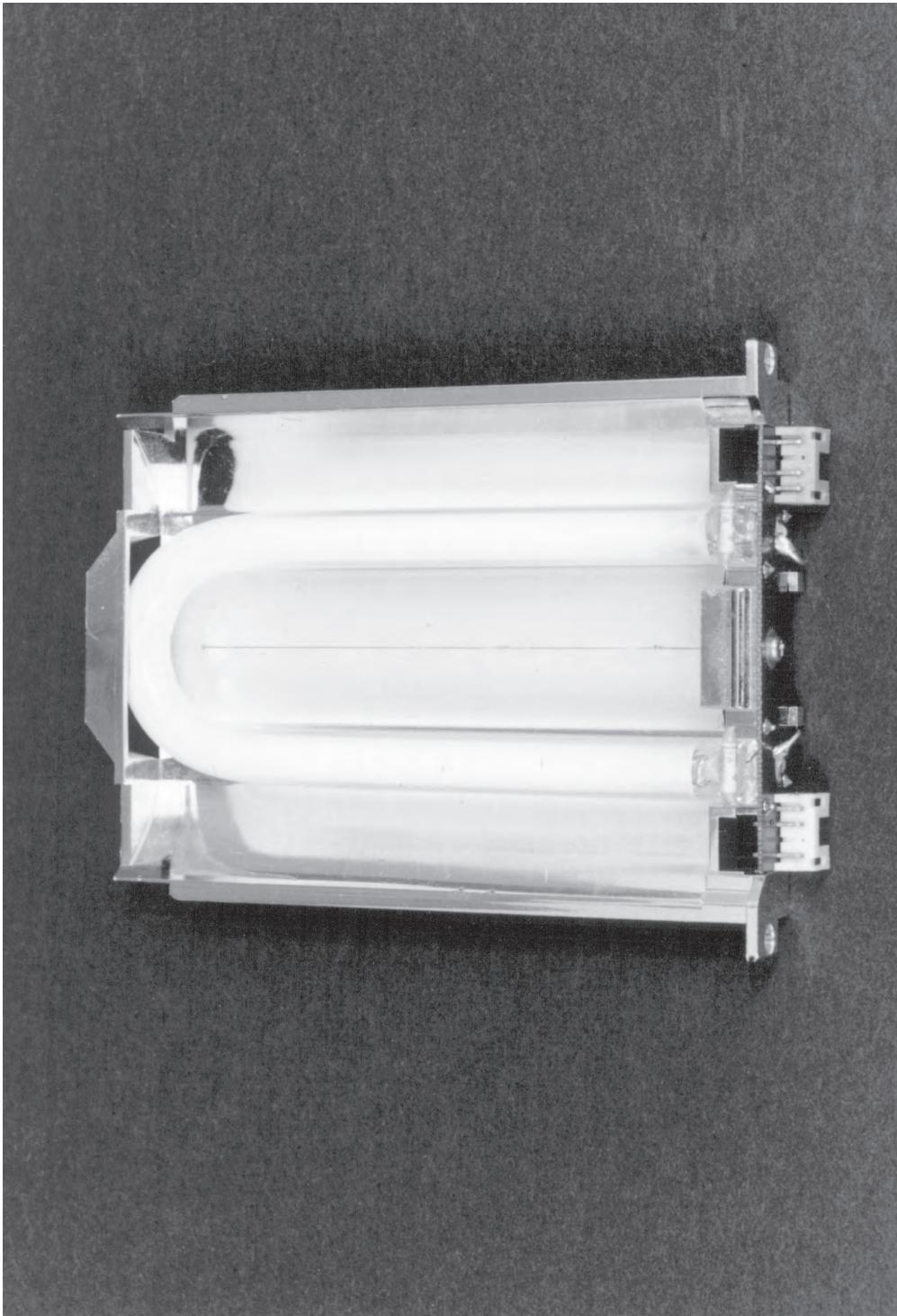
Figure 28. Low Losses of the Display's Nonconductive Frame (Black Plastic) Are Thrown Away by Lossy Reflective Foil Contacting Massive Metal Rear Cover. 15% Loss Results
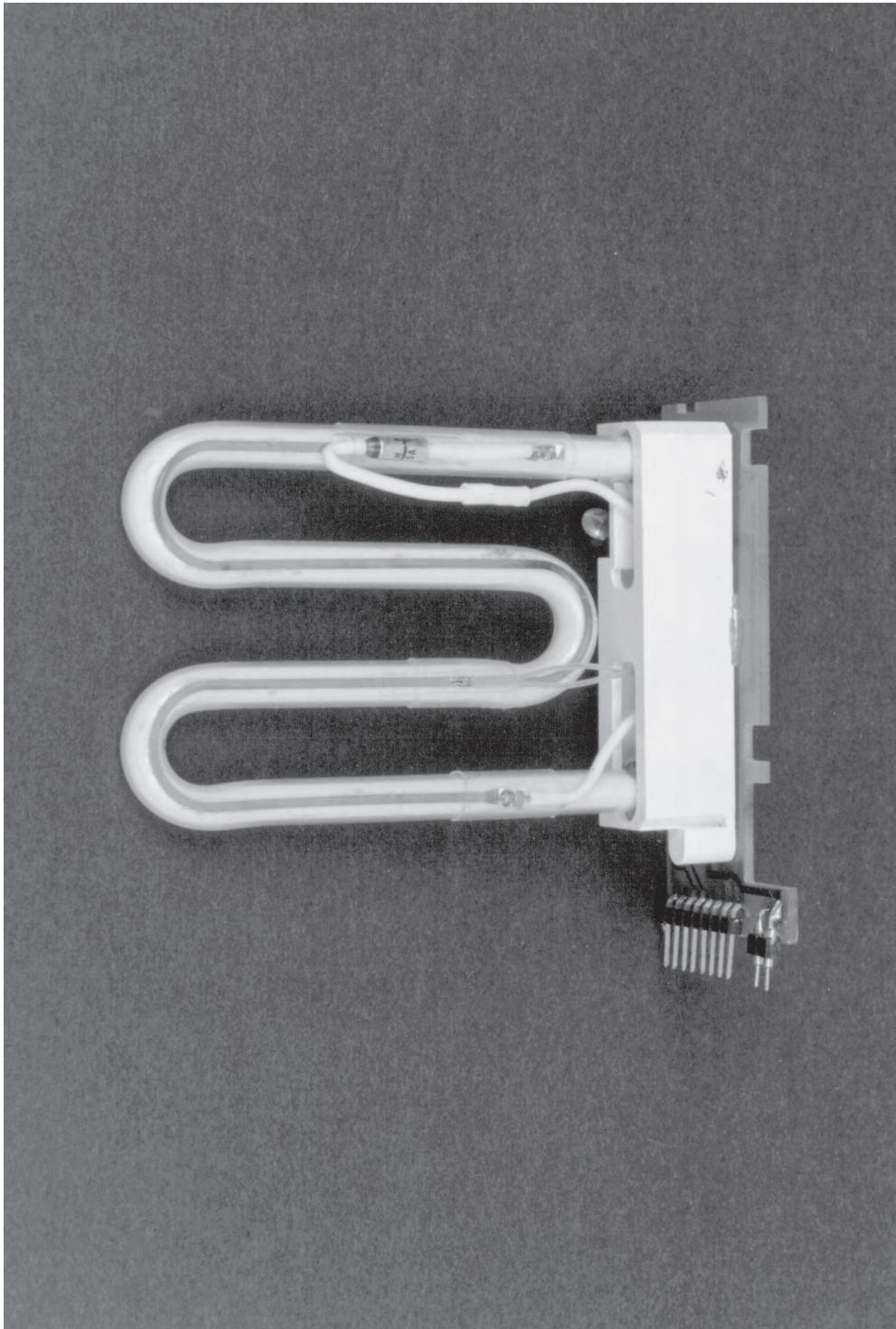
Figure 29. A Similar Situation to Figure 28. Large Metal Rear Cover Contacts Lossy Foil (Not Visible), Causing Huge Losses

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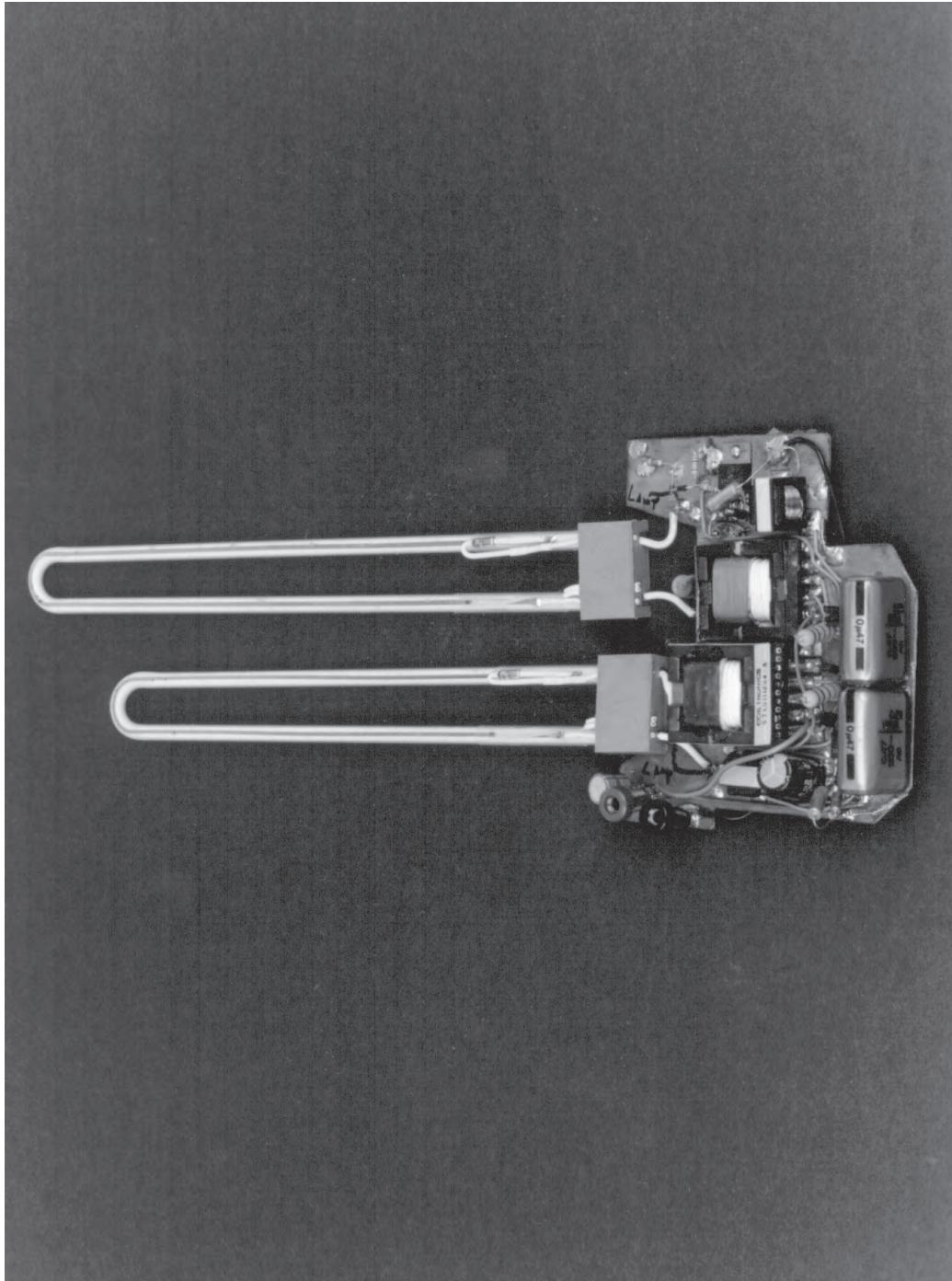
Figure 30. Grounded Metallic Optical Reflector in Automotive Lamp Introduces 18% Loss. Optical Gain Over Nonmetallic Reflector May Justify Large Electrical Loss

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Figure 31. Metallic Heater on Lamp in This Automotive Application Eases Low Temperature Starting but Causes 31% Loss

Figure 32. Similar to Figure 31. Metallic Cold Start Heaters in Automotive Application Induce 23% Loss

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Considerations for Multilamp Designs

Multiple-lamp designs are not recommended if lamp intensity matching is important. Maintaining emission matching over time, temperature and production variations is quite difficult. In some restricted cases multilamp displays may be a viable option, but a single lamp with good diffuser optics is almost always the better approach. Information on dual-lamp displays is presented here for reference purposes only.[5]

Systems using two lamps have some unique layout problems. Almost all dual-lamp displays are color units. The lower light transmission characteristics of color displays necessitates more light. As such, display manufacturers sometimes use two lamps to produce more light. The wiring layout of these dual-lamp color displays affects efficiency and illumination balance in the lamps. Figure 33 shows an "x-ray" view of a typical display. This symmetrical arrangement presents equal parasitic losses. If C1 and C2 and the lamps are well-matched, the circuit's current output splits evenly and equal illumination occurs.

Figure 34's display arrangement is less friendly. The asymmetrical wiring forces unequal losses and the lamps receive imbalanced current. Even with identical lamps, illumination may not be balanced. This condition is partially correctable by skewing values of C1 and C2 . C1, because it drives greater parasitic capacitance, should be larger than C2. This tends to equalize the currents, promoting equal lamp drive. It is important to realize that this compensation does nothing to recapture the lost energy—efficiency is still compromised. There is no substitute for minimizing loss paths. Similarly, any change in lamp characteristics (e.g., aging) can cause imbalanced illumination to recur.

In general, imbalanced illumination causes fewer problems than might be supposed at high intensity levels. Unequal illumination is much more noticeable at lower levels. In the worst case the dimmer lamp may only partially illuminate. This phenomenon, sometimes called "Thermometering," is discussed in detail in the text section, "Floating Drive Circuits."

Note 5: The text's tone is intended to convey our distaste for multilamp displays. They are the very soul of heartache.
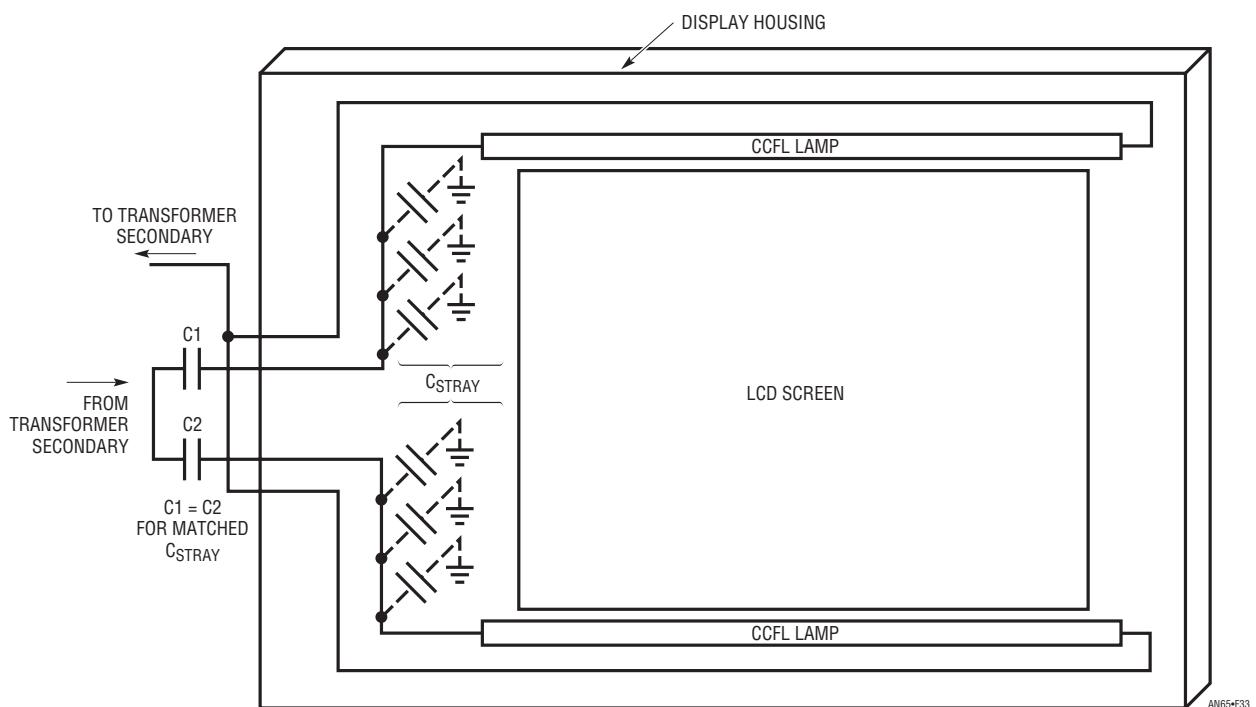
Figure 33. Loss Paths for "Best Case" Dual-Lamp Display. Symmetry Promotes Balanced Illumination, but Lamp Limitations Dominate Achievable Results
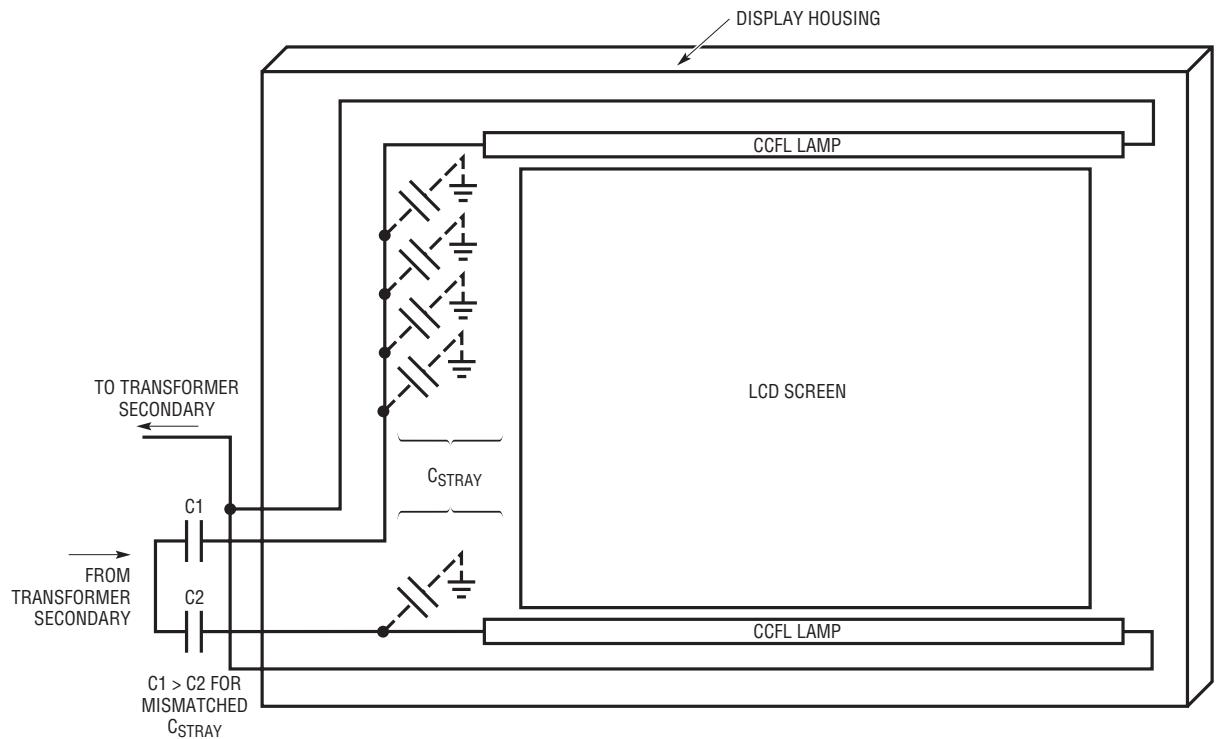
Figure 34. Asymmetric Losses in a Dual-Lamp Display. Skewing C1 and C2 Values Compensates Imbalanced Loss Paths but Not Wasted Energy

CCFL Power Supply Circuits
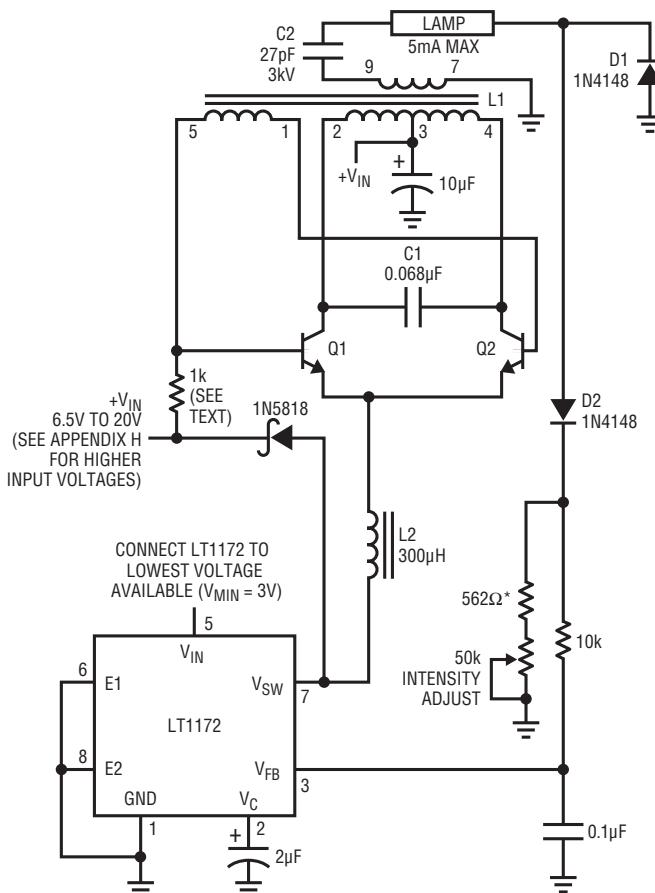
Choosing an approach for a general purpose CCFL power supply is difficult. A variety of disparate considerations make determining the "best" approach a thoughtful exercise. Above all, the architecture must be extraordinarily *flexible*. The sheer number and diversity of applications demands this. The considerations take many degrees of freedom. Power supply voltages range from 2V to 30V with output power from minuscule to 50W. The load is highly nonlinear and varies over operating conditions. The backlight is often located some distance from the primary power source, meaning the supply must tolerate substantial supply bus impedances. Similarly, it must not corrupt the supply bus with noise, or introduce appreciable RFI into the system or environment. Component count should be low and the supply must be physically quite small as space is usually extremely limited. Additionally, the circuit must be relatively layout-insensitive because of varying board shape requirements. Interface for shutdown and dimming control should accommodate either digital or analog

inputs, including voltage, current, resistive, PWM or serial bit-stream addressing. Finally, lamp current should be predictable and stable with changes in time, temperature and supply voltage.

A current-fed, feedback-controlled resonant Royer converter meets these requirements.[6] This approach, because of its extreme flexibility, is a favorable compromise. It operates over wide supply ranges and scales well over a broad output power range. Current is taken from the supply bus almost continuously, making the circuit tolerate supply bus impedance. This characteristic also means that circuit operation does not corrupt power supply lines. There is no RFI problem and component count is low. It is small, relatively insensitive to layout and easy to interface to. Lastly, lamp current is stable and predictable over operating conditions.

Note 6: See Appendices K and L for detailed discussion on architecture selection and the Royer configuration.

Figure 35 is a practical CCFL power supply circuit based on the above discussion. Efficiency is 88% with an input voltage range of 6.5V to 20V. This efficiency figure can be degraded by about 3% if the LT172 V_{IN} pin is powered from the same supply as the main circuit V_{IN} terminal. Lamp intensity is continuously and smoothly variable from zero to full intensity. When power is applied the LT1172 switching regulator's Feedback pin is below the device's internal 1.2V reference, causing full duty cycle modulation at the V_{SW} pin (Trace A, Figure 36). V_{SW} conducts current (Trace B) which flows from L1's center tap, through the transistors, into L2. L2's current is deposited in switched fashion to ground by the regulator's action.

A = 20V/DIV
B = 0.4A/DIV
C = 20V/DIV
D = 20V/DIV
E = 1000V/DIV
F = 5V/DIV

A, B = 4µs/DIV
C THRU F = 20µs/DIV
TRIGGERS FULLY INDEPENDENT

AN65•F36

Figure 36. Waveforms for the Cold Cathode Fluorescent Lamp Power Supply. Note Independent Triggering on Traces A and B, and C through F

L1 and the transistors comprise a current driven Royer class converter[7] which oscillates at a frequency primarily set by L1's characteristics (including its load) and the 0.068µF capacitor. LT1172 driven L2 sets the magnitude of the Q1/Q2 tail current, hence L1's drive level. The 1N5818 diode maintains L2's current flow when the LT1172 is off. The LT1172's 100kHz clock rate is asynchronous with respect to the push/pull converter's (60kHz) rate, accounting for Trace B's waveform thickening.

The 0.068µF capacitor combines with L1's characteristics to produce sine wave voltage drive at the Q1 and Q2 collectors (Traces C and D respectively). L1 furnishes voltage step-up and about $1400V_{P-P}$ appears at its secondary (Trace E). Current flows through the 27pF capacitor into the lamp. On negative waveform cycles, the lamp's current is steered to ground via D1. Positive waveform cycles are directed via D2 to the ground referred 562Ω/50k potentiometer chain. The positive half-sine appearing across the resistors (Trace F) represents 1/2 the lamp current. This signal is filtered by the 10k/0.1µF pair and presented to the LT1172's Feedback pin. This connection closes a control loop which regulates lamp current. The 2µF capacitor at the LT1172's V_C pin provides stable loop compensation. The loop forces the LT1172 to switch mode modulate L2's

Note 7: See Appendix K, "Who Was Royer and What Did He Design?" See also Reference 2.

C1 = MUST BE A LOW LOSS CAPACITOR. METALIZED POLYCARB WIMA
    MKP-20 (GERMAN) OR PANASONIC ECH-U RECOMMENDED
L1 = SUMIDA 6345-020 OR COILTRONICS CTX110092-1
    (PIN NUMBERS SHOWN FOR COILTRONICS UNIT)
L2 = COILTRONICS CTX300-4
Q1, Q2 = ZETEX ZTX849, ZDT1048 OR ROHM 2SC5001
    * = 1% FILM RESISTOR
DO NOT SUBSTITUTE COMPONENTS
COILTRONICS (407) 241-7876, SUMIDA (708) 956-0666

AN65•F35

Figure 35. An 88% Efficiency Cold Cathode Fluorescent Lamp Power Supply

average current to whatever value is required to maintain constant current in the lamp. The constant current's value, and hence lamp intensity, may be varied with the potentiometer. The constant current drive allows full 0% to 100% intensity control with no lamp dead zones or "pop-on" at low intensities.[8] Additionally, lamp life is enhanced because current cannot increase as the lamp ages.

The circuit's 0.1% line regulation is notably better than some other approaches. This tight regulation prevents lamp intensity variation when abrupt line changes occur. This typically happens when battery-powered apparatus is connected to an AC-powered charger. The circuit's excellent line regulation derives from the fact that L1's drive waveform never changes shape as input voltage varies. This characteristic permits the simple $10k\Omega/0.1\mu F$ RC to produce a consistent response. The RC averaging characteristic has serious error compared to a true RMS conversion, but the error is constant and "disappears" in the 562Ω shunt's value.

This circuit is similar to one previously described[9] but its 88% efficiency is 6% higher. The efficiency improvement is primarily due to the transistor's higher gain and lower saturation voltage. The base drive resistor's value (nominally 1k) should be selected to provide full V_{CE} saturation without inducing base overdrive or beta starvation. A procedure for doing this is described in a following section, "General Optimization and Measurement Considerations."

Figure 37's circuit is similar, but uses a transformer with lower copper and core losses to increase efficiency to 91%. The trade-off is slightly larger transformer size. Additionally, a higher frequency switching regulator offers slightly lower V_{IN} current, aiding efficiency. L1's smaller value, a result of the higher frequency operation, permits slightly reduced copper loss. The transformer options listed allow efficiency optimization over the supply range of interest. Value shifts in C1, L2 and the base drive resistor reflect different transformer characteristics. This circuit also features shutdown and a DC or pulse width controlled dimming input. Appendix F, "Intensity Control and Shutdown Methods," details operation of these features. Figure 38, directly derived from Figure 37, produces 10mA output to drive color LCDs at 92% efficiency. The

C1 = WIMA MKP-20, PANASONIC ECH-U
L1 = COILCRAFT DT3316-333
Q1, Q2 = ZETEX ZTX849, ZDT1048
  OR ROHM 2SC5001
T1 = COILTRONICS CTX 02-12614-1 OR
  CTX110600-1 (SEE TEXT)
* = 1% FILM RESISTOR

DO NOT SUBSTITUTE COMPONENTS
COILTRONICS (407) 241-7876
COILCRAFT (708) 639-6400

AN65•F37
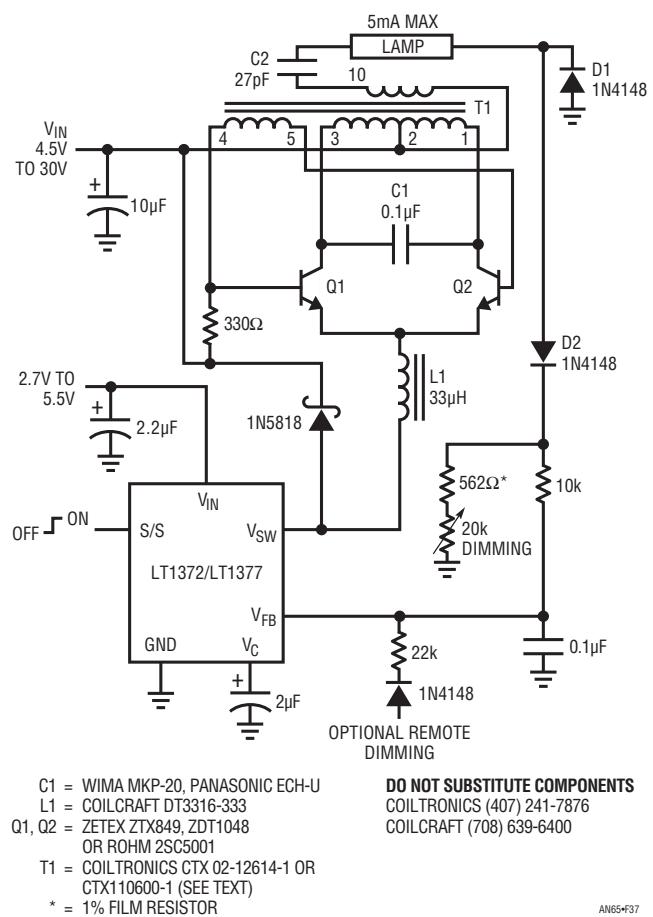
Figure 37. A 91% Efficient CCFL Supply for 5mA Loads Features Shutdown and Dimming Inputs. Higher Frequency Switching Regulator Reduces L1's Size While Requiring Less V_{IN} Current

slight efficiency improvement comes from a reduction in regulator "housekeeping" current as a percentage of total current drain. Value changes in components are the result of higher power operation. The most significant change involves driving two lamps. Accommodating two lamps involves separate ballast capacitors but circuit operation is similar. Dual-lamp designs reflect slightly different loading back through the transformer's primary. C2 usually ends
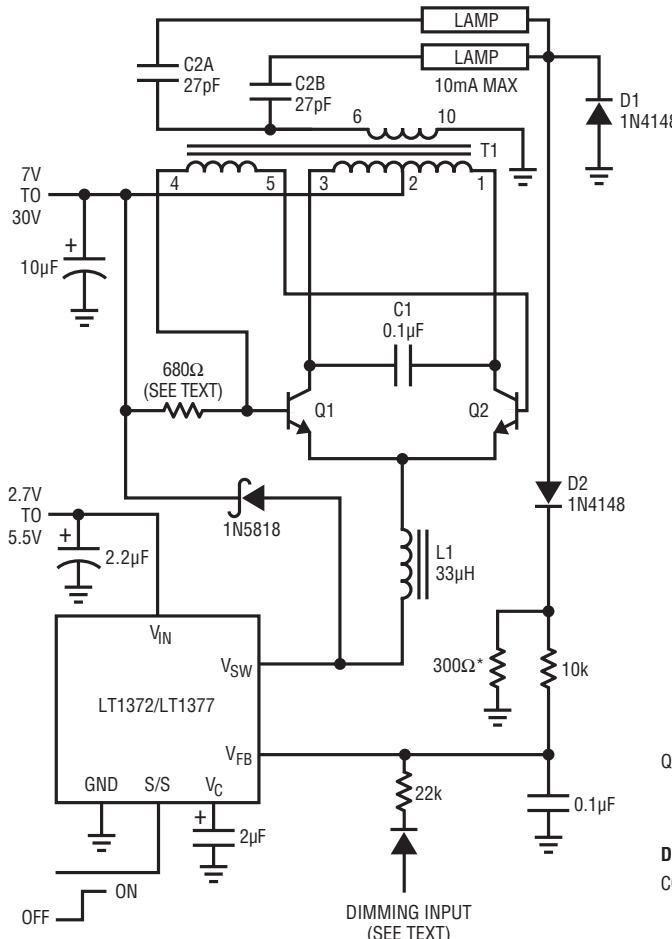
Note 8: Controlling a nonlinear load's current, instead of its voltage, permits applying this circuit technique to a wide variety of nominally evil loads. See Appendix I, "Additional Circuits."
Note 9: See "Illumination Circuity for Liquid Crystal Displays," Linear Technology Corporation, Application Note 49, August 1992 and "Techniques for 92% Efficient LCD Illumination," Linear Technology Corporation, Application Note 55, August 1993.

up in the 10pF to 47pF range. Note that C2A and B appear with their lamp loads in parallel across the transformer's secondary. As such, C2's value is often smaller than in a single-lamp circuit using the same type lamp. Ideally, the transformer's secondary current splits evenly between the C2-lamp branches, with the total load current being regulated. In practice, differences between C2A and B, and differences in lamps and lamp wiring layout preclude a perfect current split. Practically, these differences are small and the lamps appear to emit an equal amount of light at high intensity. Layout and lamp matching can influence C2's value. Some techniques for dealing with these issues appear in the text section, "Considerations for Multilamp Designs." As previously stated, dual-lamp designs are distinctly not recommended, particularly if balanced illumination over wide dimming ranges is required.

Figure 39 uses a dedicated CCFL IC, the LT1183, to enhance circuit performance. The Royer-based high voltage converter portion is recognizable from previous circuits, with the 200kHz LT1183 performing the switching regulator/feedback function. This IC also features open lamp protection circuitry, simplified frequency compensation, a separate regulator providing LCD contrast and other features.[10] The contrast supply is driven by the LT1183 with L3 and associated discrete components completing the function. The CCFL and contrast outputs may be adjusted with DC, PWM or potentiometers.

Note 10: Open lamp protection is often desirable and may be added to the previous circuits at the cost of some discrete components. See Appendix E, "Open Lamp/Overload Protection." Frequency compensation issues are covered in the text section "Feedback Loop Stability Issues." See Appendix J for discussion of LCD contrast supplies.

C1 = WIMA MKP2 OR PANASONIC ECH-U
L1 = COILTRONICS CTX150-4
Q1, Q2 = ZETEX ZTX849, ZDT1048 OR ROHM 2SC5001
T1 = COILTRONICS CTX210605 OR SUMIDA EPS-207
  (PIN NUMBERS SHOWN FOR COILTRONICS UNIT)
* = 1% FILM RESISTOR

DO NOT SUBSTITUTE COMPONENTS

COILTRONICS (407) 241-7876, SUMIDA (708) 956-0666

AN65•F38

Figure 38. A 92% Efficient CCFL Supply for 10mA Loads Features Shutdown and Dimming Inputs. Dual-Lamp Designs, Typical of Early Color Displays, Are Not Recommended
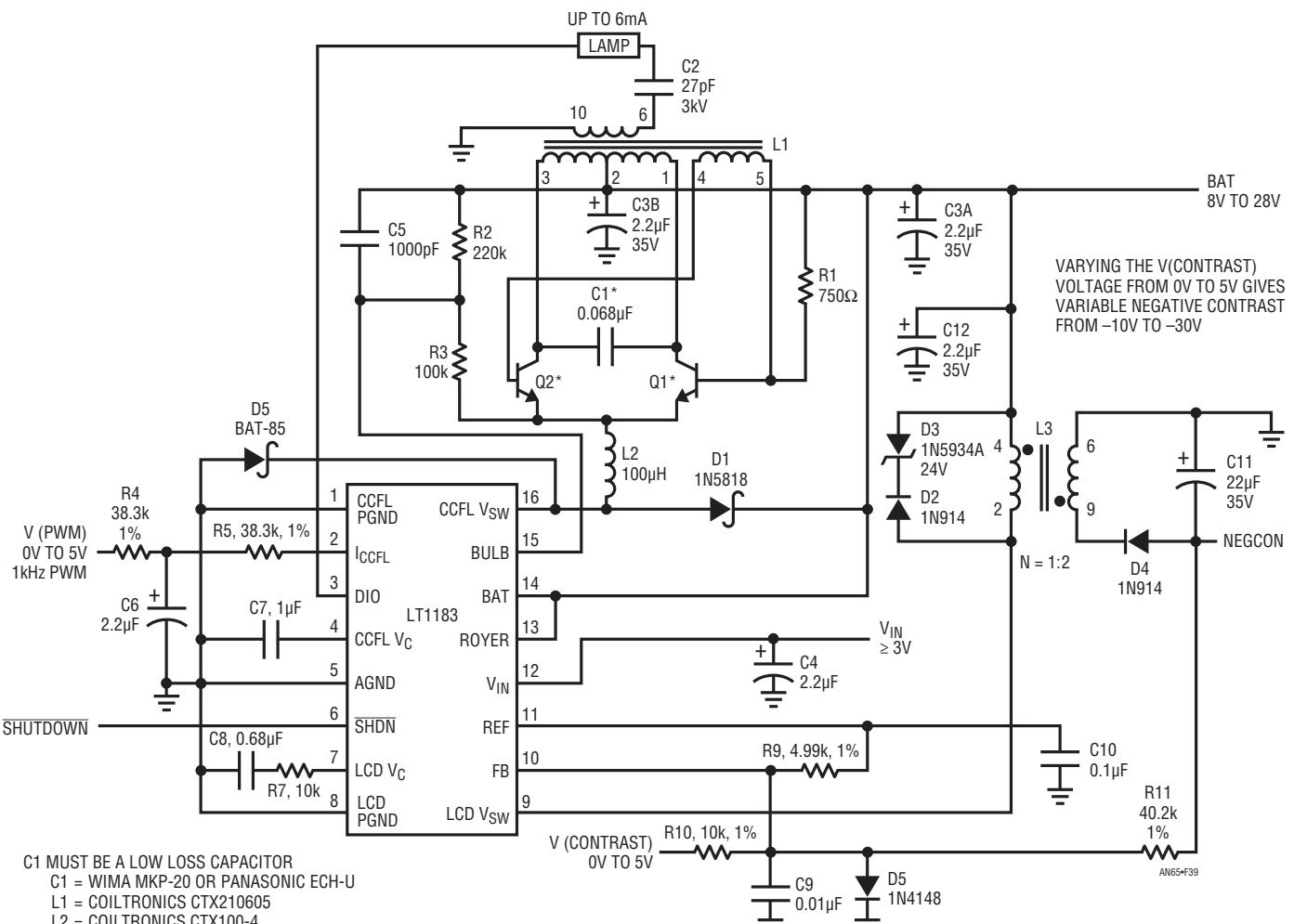
Figure 39. Dedicated Backlight IC Includes Switching Regulator, Open Lamp Protection and LCD Contrast Supply. 200kHz Operation Minimizes L2 Size. Shutdown and Control Inputs Are Simplified

C1 MUST BE A LOW LOSS CAPACITOR
    C1 = WIMA MKP-20 OR PANASONIC ECH-U
    L1 = COILTRONICS CTX210605
    L2 = COILTRONICS CTX100-4
    L3 = COILTRONICS CTX02-12403
Q1, Q2 = ZETEX ZTX849, ZDT1048 OR ROHM 2SC5001
*DO NOT SUBSTITUTE COMPONENTS
COILTRONICS (407) 241-7876
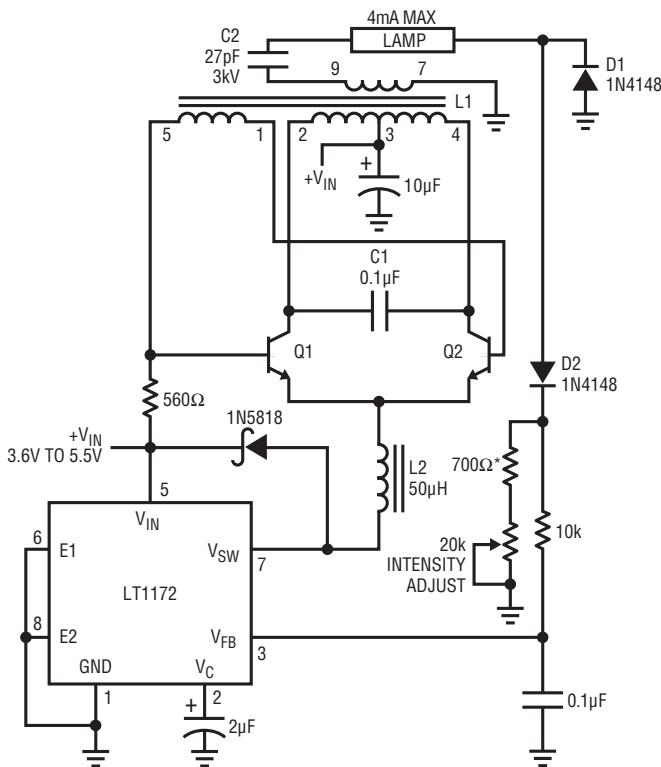
Low Power CCFL Power Supplies

Many applications require relatively low power CCFL backlighting. Figure 40's variation, optimized for low voltage inputs, produces 4mA output. Circuit operation is similar to the previous examples. The fundamental difference is L1's higher turns ratio, which accommodates the reduced available drive voltage. The circuit values given are typical, although some variation occurs with various lamps and layouts.

Figure 41's design, the so-called "dim backlight," is optimized for very low current lamp operation. The circuit is meant for use at low input voltages, typically 2V to 6V with a 1mA maximum lamp current. This circuit maintains control down to lamp currents of 1µA, a very dim light! It is intended for applications where the longest possible battery life is desired. Primary supply drain ranges from hundreds of microamperes to 100mA with lamp currents of microamps to 1mA. In shutdown the circuit pulls only 100µA. Maintaining high efficiency at low lamp currents requires modifying the basic design.

Achieving high efficiency at low operating current requires lowering quiescent power drain. To do this the previously employed pulse width modulator-based devices are replaced with an LT1173. The LT1173 is a Burst Mode® operation regulator. When this device's Feedback pin is too low it delivers a burst of output current pulses, putting energy into the transformer and restoring the feedback point. The regulator maintains control by appropriately modulating the burst duty cycle. The ground referred diode at the V_{SW} pin prevents substrate turn-on due to excessive L2 ring-off.
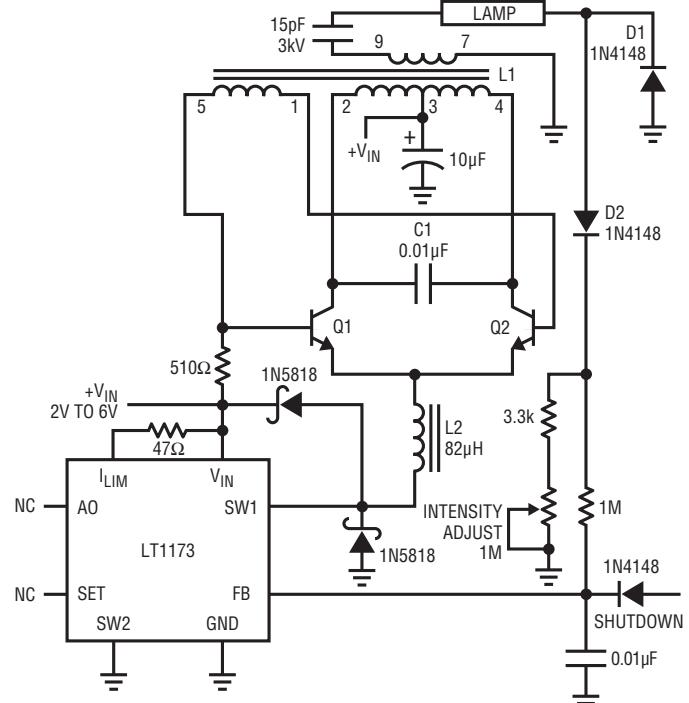
C1 = MUST BE A LOW LOSS CAPACITOR.
      METALIZED POLYCARB
      WIMA MKP-20 (GERMAN) OR PANASONIC ECH-U RECOMMENDED
L1 = COILTRONICS CTX110654-1
L2 = COILTRONICS CTX50-4
Q1, Q2 = ZETEX ZTX849, ZDT1048 OR ROHM 2SC5001
      * = 1% FILM RESISTOR
      DO NOT SUBSTITUTE COMPONENTS
COILTRONICS (407) 241-7876, SUMIDA (708) 956-0666   AN65•F40

Figure 40. A 4mA Design Intended for Low Voltage Operation. L1's Modified Turns Ratio Allows Operation Down to 3.6V

C1 = MUST BE A LOW LOSS CAPACITOR.
      METALIZED POLYCARB
      WIMA FKP2, MKP-20 (GERMAN) OR PANASONIC ECH-U RECOMMENDED
L1 = SUMIDA 6345-020 OR COILTRONICS CTX110092-1
      PIN NUMBERS SHOWN FOR COILTRONICS UNIT
L2 = TOKO 262LYF-0091K (408) 432-8251
Q1, Q2 = ZETEX ZTX849, ZDT1048 OR ROHM 2SC5001
      DO NOT SUBSTITUTE COMPONENTS   AN65•F41

Figure 41. Low Power CCFL Power Supply. Circuit Controls Lamp Current Over a 1µA to 1mA Range

During the off periods the regulator is essentially shut down. This type of operation limits available output power, but cuts quiescent current losses. In contrast, the other circuit's pulse width modulator type regulators maintain "housekeeping" current between cycles. This results in more available output power but higher quiescent currents.

Figure 42 shows operating waveforms. When the regulator comes on (Trace A, Figure 42) it delivers bursts of output current to the L1/Q1/Q2 high voltage converter. The converter responds with bursts of ringing at its resonant frequency.[11] The circuit's loop operation is similar to the previous designs except that T1's drive waveform varies with supply. Because of this, line regulation suffers and the circuit is not recommended for wide ranging inputs.

Some lamps may display nonuniform light emission at very low excitation currents. See the text section, "Floating Lamp Circuits."

A CCFL power supply that addresses the previous circuit's line regulation problems and operates from 2V to 6V is detailed in Figure 43. This circuit, contributed by Steve Pietkiewicz of LTC, can drive a small CCFL over a 100μA to 2mA range.

A = 5V/DIV

B = 5V/DIV

50μs/DIV AN65•F42

Figure 42. Waveforms for the Low Power CCFL Power Supply. LT1173 Burst Type Regulator (Trace A) Periodically Excites the Resonant High Voltage Converter (Q1 Collector Is Trace B)

Note 11: The discontinous energy delivery to the loop causes substantial jitter in the burst repetition rate, although the high voltage section maintains resonance. Unfortunately, circuit operation is in the "chop" mode region of most oscilloscopes, precluding a detailed display. "Alternate" mode operation causes waveform phasing errors, producing an inaccurate display. As such, waveform observation requires special techniques. Figure 42 was taken with a dual-beam instrument (Tektronix 556) with both beams slaved to one time base. Single sweep triggering eliminated jitter artifacts. Most oscilloscopes, whether analog or digital, will have trouble reproducing this display.

L1 = COILCRAFT D03316-473
Q1, Q2 = ZETEX ZTX849, ZDT1048 OR ROHM 2SC5001
T1 = COILTRONICS CTX110654-1
0.068μF = WIMA MKP-20 OR PANASONIC ECH-U

AN65•F43

Figure 43. Low Power Cold Cathode Fluorescent Lamp Supply Is Optimized for Low Voltage Inputs and Small Lamps

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The circuit uses an LT1301 micropower DC/DC converter IC in conjunction with a current driven Royer class converter comprised of T1, Q1 and Q2. When power and intensity adjust voltage are applied, the LT1301's I_{LIM} pin is driven slightly positive, causing maximum switching current through the IC's internal Switch pin (SW). Current flows from T1's center tap, through the transistors, into L1. L1's current is deposited in switched fashion to ground by the regulator's action.

Circuit efficiency ranges from 80% to 88% at full load, depending on line voltage. Current mode operation combined with the Royer's consistent waveshape vs input results in excellent line rejection. The circuit has none of the line rejection problems attributable to the hysteretic voltage control loops typically found in low voltage micropower DC/DC converters. This is an especially desirable characteristic for CCFL control, where lamp intensity must remain constant with shifts in line voltage.
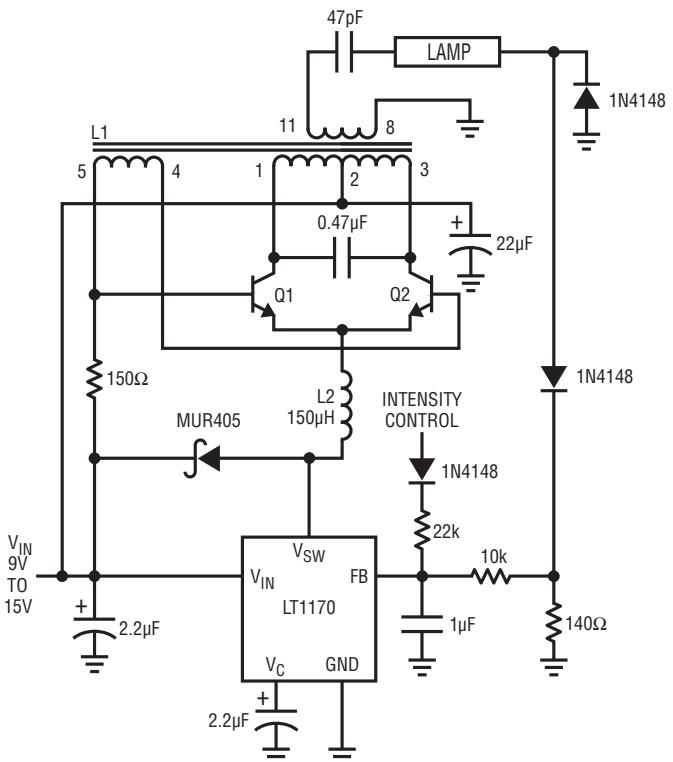
The Royer converter oscillates at a frequency primarily set by T1's characteristics (including its load) and the 0.068µF capacitor. LT1301 driven L1 sets the magnitude of the Q1/Q2 tail current, hence T1's drive level. The 1N5817 diode maintains L1's current flow when the LT1301's switch is off. The 0.068µF capacitor combines with T1's characteristics to produce sine wave voltage drive at the Q1 and Q2 collectors. T1 furnishes voltage step-up and about 1400V_{P-P} appears at its secondary. Alternating current flows through the 22pF capacitor into the lamp. On positive half-cycles the lamp's current is steered to ground via D1. On negative half-cycles the lamp's current flows through Q3's collector and is filtered by C1. The LT1301's I_{LIM} pin acts as a 0V summing point with about 25µA bias current flowing out of the pin into C1. The LT1301 regulates L1's current to equalize Q3's average collector current, representing 1/2 the lamp current, and R1's current, represented by V_A/R1. C1 smooths all current flow to DC. When V_A is set to zero, the I_{LIM} pin's bias current forces about 100µA bulb current.

High Power CCFL Power Supply

As mentioned, the CCFL circuit approach presented here scales quite nicely over a wide range of output power. Most circuits are in the 0.5W to 3W region due to the application's small size and battery-driven nature. Automotive, aircraft, desktop computer and other displays often require much higher power.

Figure 44's arrangement is a scaled-up version of the text's CCFL circuits. This design, similar to ones employed for automotive use, drives a 25W CCFL. There are virtually no configuration changes, although most component power ratings have increased. The transistors can handle the higher currents, but all other power components are higher capacity. Efficiency is about 80%.

Additional high power circuits appear in Appendix I, "Additional Circuits."

L1 = COILTRONICS CTX02-11128
L2 = COILTRONICS CTX150-3-52
Q1, Q2 = ZETEX ZTX849, ZDT1048 OR ROHM 2SC5001
0.47µF = WIMA 3X 0.15mF TYPE MKP-20
COILTRONICS (407) 241-7876

AN65•F44

Figure 44. A 25W CCFL Supply Is a Scaled Version of Lower Power Circuits

"Floating" Lamp Circuits

All circuits presented to this point drive the lamp in single-ended fashion. Similarly, Figure 45 shows one lamp electrode receiving drive with the other terminal essentially at ground. This causes significant loss via parasitic paths associated with the lamp's driven end. This is so because of the large voltage swing in this region. The parasitic paths near the lamp's grounded end undergo relatively little swing, contributing small energy loss. Unfortunately, the lost energy is heavily voltage-dependent ($E = 1/2\ CV^2$) and net energy loss is excessive if driven end parasitics are large. Figure 46 minimizes the losses by altering the drive scheme. In this case the lamp is driven from both ends instead of grounding one end. This "floating" lamp arrangement requires only half the voltage swing at each lamp end instead of full swing at one end. This introduces more loss in the parasitic paths previously associated with

the grounded end. In most cases these increased losses are favorably offset by the reduced swing because of the V^2 loss term associated with voltage amplitude.

The advantage gained varies considerably with display type, although a 10% to 20% reduction in lost energy is common. In some displays loss reduction is not as good, and occasionally improvement is negligible. Heavily asymmetric wiring to or within the display can sometimes make floating drive more lossy than grounded drive. In such cases testing in both modes is necessary to determine which type drive is most efficient.

A second advantage of floating operation is extended illumination range. "Grounded" lamps operating at relatively low currents may display the "thermometer effect," that is, light intensity may be nonuniformly distributed along lamp length.
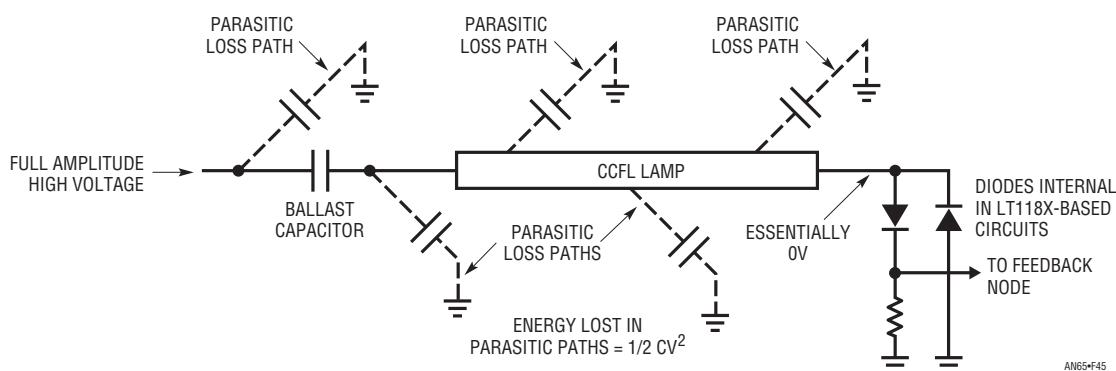
Figure 45. Ground Referred Lamp Drive Has Large Energy Loss in High Voltage Regions Due to Full Amplitude Swing
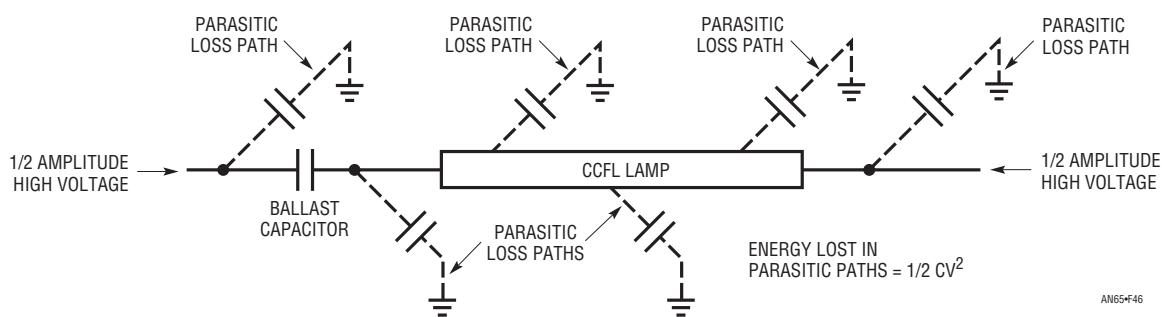
Figure 46. "Floating" Lamp Allows Reduced, Bipolar Drive, Cutting Losses Due to Parasitic Capacitance Paths. Formerly Grounded Lamp End's Paths Absorb More Energy Than Before, but Overall Loss Is Lower Due to Equation's V^2 Term

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Figure 47 shows that although lamp current density is uniform, the associated field is imbalanced. The field's low intensity, combined with its imbalance, means that there is not enough energy to maintain uniform phosphor glow beyond some point. Lamps displaying the thermometer effect emit most of their light near the driven electrode, with rapid emission fall-off as distance from the electrode
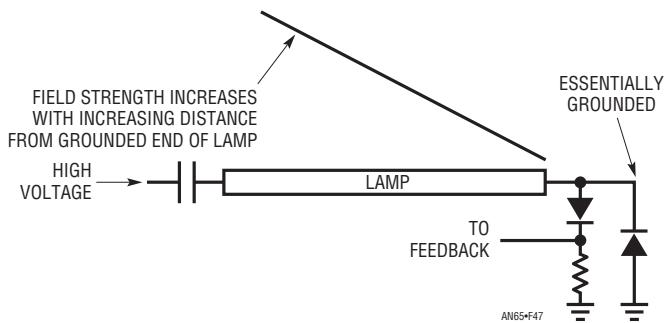
Figure 47. Field Strength vs Distance for a Ground Referred Lamp. Field Imbalance Promotes Uneven Illumination at Low Drive Levels

increases. Placing a conductor along the lamp's length largely alleviates "thermometering." The trade-off is decreased efficiency due to energy leakage.[12] It is worth noting that various lamp types have different degrees of susceptibility to the thermometer effect.

Some displays require extended illumination range. "Thermometering" usually limits the lowest practical illumination level. One acceptable way to minimize "thermometering" is to eliminate the large field imbalance. The floating drive used to reduce energy loss also provides a way to minimize "thermometering." Figure 48 reviews a circuit originally

Note 12: A very simple experiment quite nicely demonstrates the effects of energy leakage. Grasping the lamp at its low voltage end (low field intensity) with thumb and forefinger produces almost no change in circuit input current. Sliding the thumb/forefinger combination towards the high voltage (higher field intensity) lamp end produces progressively greater input currents. Don't touch the high voltage lead or you may receive an electrical shock. Repeat: Do not touch the high voltage lead or you may receive an electrical shock.
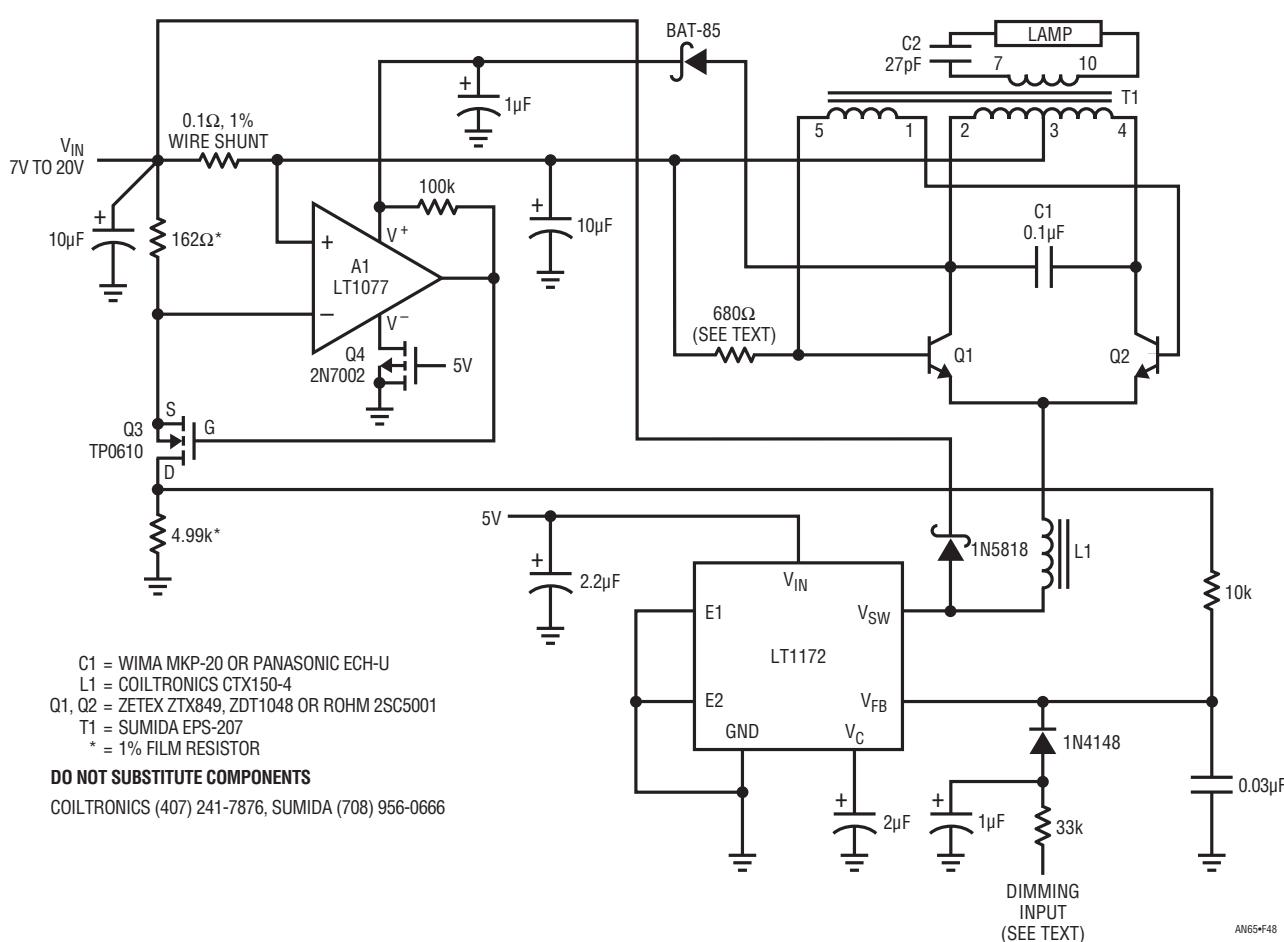
Figure 48. Practical "Floating" Lamp Drive Circuit. A1 Senses Royer Input Current with Q3 Providing Resultant Feedback Information to Switching Regulator. Circuit Reduces Lost Energy Due to Parasitics by 10% to 20%

introduced in a previous publication.[13] The circuit's most significant aspect is that the lamp is fully floating—there is no galvanic connection to ground as in the previous designs. This allows T1 to deliver symmetric, differential drive to the lamp. Such balanced drive eliminates field imbalance, reducing thermometering at low lamp currents. This approach precludes any feedback connection to the now floating output. Maintaining closed-loop control necessitates deriving a feedback signal from some other point. In theory, lamp current proportions to T1's or L1's drive level and some form of sensing this can be used to provide feedback. In practice, parasitics make a practical implementation difficult.[14]

Figure 48 derives the feedback signal by measuring Royer converter current and feeding this information back to the LT1172. The Royer's drive requirement closely proportions to lamp current under all conditions. A1 senses this current across the 0.1Ω shunt and biases Q3, closing a local

feedback loop. Q3's drain voltage presents an amplified, single-ended version of the shunt voltage to the feedback point, closing the main loop. A1's power supply pin is bootstrapped to T1's boosted swing via the BAT-85 diode, permitting it to sense across the supply-fed shunt resistor. Internal A1 characteristics ensure start-up and substitution of this device is not recommended.[15]

The lamp current is not as tightly controlled as before but 0.5% regulation over wide supply ranges is possible. The dimming in this circuit is controlled by a 1kHz PWM signal. Note the heavy filtering (33k/1μF) *outside* the feedback loop. This allows a fast time constant, minimizing turn-on overshoot.[16]

Note 13: See Reference 1.
Note 14: See Appendix L, "A Lot of Cut Off Ears and No Van Goghs—Some Not-So-Great Ideas," for details.
Note 15: See Reference 1, then don't say we didn't warn you.
Note 16: See text section, "Feedback Loop Stability Issues."
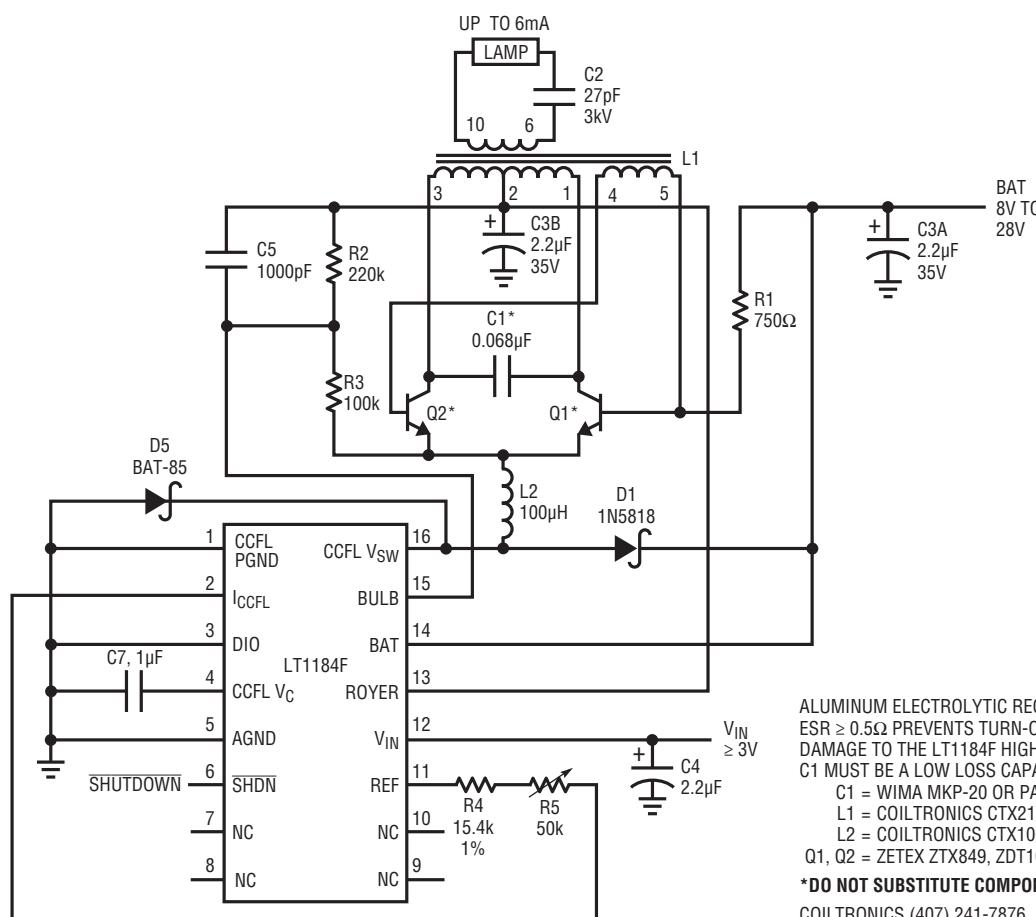
Figure 49. LT1184F IC Version of Figure 48's Floating Lamp Circuit Offers Similar Performance with Fewer Components. Open Bulb Protection and Shutdown Are Included

In all other respects operation is similar to the previous circuits. This circuit typically permits the lamp to operate with less energy loss and over a 40:1 intensity range without "thermometering." The normal feedback connection is usually limited to a 10:1 range.

IC-Based Floating Drive Circuits

Figure 49 compacts Figure 48 into a low component count, floating drive circuit. The LT1184F IC contains all functions except the Royer-based high voltage converter. The circuit also has "open lamp" protection and a 1.23V reference for biasing the dimming potentiometer.

Figure 50 adds a bipolar LCD contrast supply output to Figure 49. The LT1182 allows setting contrast supply polarity by simply grounding the appropriate output terminal. The CCFL portion is similar to the previous circuit, although intensity is controlled with a varying PWM or 0V to 5V input.

ALUMINUM ELECTROLYTIC RECOMMENDED FOR C3B.
ESR ≥ 0.5Ω PREVENTS TURN-ON SURGE CURRENT
DAMAGE TO THE LT1182 HIGH SIDE SENSE RESISTOR.
C1 MUST BE A LOW LOSS CAPACITOR.
    C1 = WIMA MKP-20 OR PANASONIC ECH-U
    L1 = COILTRONICS CTX210605
    L2 = COILTRONICS CTX100-4
    L3 = COILTRONICS CTX02-12403
    Q1, Q2 = ZETEX ZTX849, ZDT1048 OR ROHM 2SC5001

*DO NOT SUBSTITUTE COMPONENTS

COILTRONICS (407) 241-7876

AN65•F50

Figure 50. LT1182 Has Bipolar Output Contrast Supply in Addition to Floating Lamp Drive

Figure 51's circuit is similar, although no contrast supply is included. The LT1186 implements a floating lamp drive similar to Figure 49. This IC contains an internal D/A converter which may be addressed by accumulating a bit stream or serial protocol. Figure 52 shows a typical arrangement using an 80C31 type microcontroller. Figure 53 gives the complete software listing which was written by Tommy Wu of LTC.

ALUMINUM ELECTROLYTIC RECOMMENDED FOR C3B.
ESR ≥ 0.5Ω PREVENTS TURN-ON SURGE CURRENT
DAMAGE TO THE LT1186 HIGH SIDE SENSE RESISTOR.
C1 MUST BE A LOW LOSS CAPACITOR.

C1 = WIMA MKP-20 OR PANASONIC ECH-U
L1 = COILTRONICS CTX210605
L2 = COILTRONICS CTX100-4
Q1, Q2 = ZETEX ZTX849, ZDT1048 OR ROHM 2SC5001

***DO NOT SUBSTITUTE COMPONENTS**

COILTRONICS (407) 241-7876

AN65•F51

Figure 51. LT1186 Permits Serial or Bit Stream Data Addressing to Set Floating Lamp Current

AN65•F52
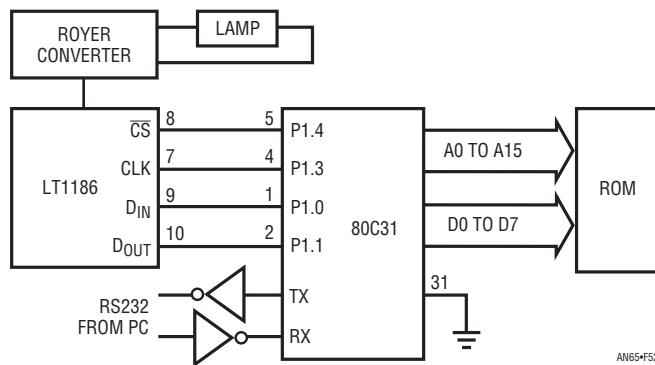
Figure 52. Typical Processor Interface for Figure 51

The LT1186 DAC algorithm is written in assembly code in a file named LT1186A.ASM as a function call from the MAIN fuction below.

Note: A user inputs an integer from 0 to 255 on a keyboard and the LT1186 adjusts the IOUT programming current to control the operating lamp current and the brightness of the LCD display.

```c
#include <stdio.h>
#include <reg51.h>
#include <absacc.h>

extern char lt1186(char);  /* external assembly function in lt1186a.asm*/
sbit Clock = 0x93;

main()
{
    int number = 0;
    int LstCode;

    Clock = 0;

    TMOD = 0x20;   /* Establish serial communication 1200 baud */
    TH1 = 0xE8;
    SCON = 0x52;
    TCON = 0x69;

    while(1)     /*  Endless loop  */
    {
        printf("\nEnter any code from 0 - 255:");
        scanf("%d",&number);
        if((0>number)l(number>255))
          {
             number = 0;
             printf("The number exceeds its range. Try again!");
          }
           else
           {
              LstCode = lt1186(number);
              printf("Previous # %u",(LstCode&0xFF)); /* AND the previous number with 0xFF to turn off sign
                                    extension */
           }
        number = 0;
    }
}
```

```
; The following assembly program named LT1186A.ASM receives the Din word from the main C program,
; lt1186 lt1186(). Assembly to C interface headers, declarations and memory allocations are listed before the
; actual assembly code.
;
; Port p1.4 = CS
; Port p1.3 = CLK
; Port p1.1 = Dout
; Port p1.0 = Din
;
```

Figure 53. Complete Software Listing for Figure 52's Processor Interface

```
NAME  LT1186_ CCFL
PUBLIC lt1186, ?lt1186?BYTE

?PR?ADC_INTERFACE?LT1186_CCFL SEGMENT CODE
?DT?ADC_INTERFACE?LT1186_CCFL SEGMENT  DATA

            RSEG  ?DT?ADC_INTERFACE?LT1186_CCFL
?lt1186?BYTE: DS  2

            RSEG  ?PR?ADC_INTERFACE?LT1186_CCFL

CS      EQU   p1.4
CLK     EQU   p1.3
DOUT    EQU   p1.1
DIN     EQU   P1.0

lt1186: setb   CS              ;set CS high to initialize the LT1186
        mov    r7,?lt1186?BYTE ;move input number(Din) from keyboard to R7
        mov    p1, #01h        ;setup port p1.0 becomes input
        clr    CS              ;CS goes low, enable the DAC
        mov    a, r7           ;move the Din to accumulator
        mov    r4, #08h        ;load counter 8 counts
        clr    c               ;clear carry before rotating
        rlc    a               ;rotate left Din bit(MSB) into carry
loop:   mov    DIN, c          ;move carry bit to Din port
        setb   CLK             ;Clk goes high for LT1186 to latch Din bit
        mov    c, DOUT         ;read Dout bit into carry
        rlc    a               ;rotate left Dout bit into accumulator
        clr    CLK             ;clear clock to shift the next Dout bit
        djnz   r4, loop        ;next data bit loop
        mov    r7, a           ;move previous code to R7 as character return
        setb   CS              ;bring CS high to disable DAC
        ret
        END

Note: When CS goes low, the MSB of the previous code appears at Dout.
```

Figure 53 (continued). Complete Software Listing for Figure 52's Processor Interface

High Power Floating Lamp Circuit

High power floating lamp circuits require more current than the LT118X series can deliver. In such cases the function can be built from discrete components and ICs. Figure 54 shows a 30W CCFL circuit used in an automotive application. This 4-lamp circuit uses an LT1269 current-fed Royer converter to provide high power. Lamp current is sensed in current transformer T2. A1 and associated components form a synchronous rectifier for T2's low level output. A2 provides gain and closes a loop back at the LT1269's feedback terminal. T2's isolated sensing permits the advantages of floating operation with the LT1269 providing high power capability. This circuit has about 83% efficiency at 30W output, a wide dimming range and 0.1% line regulation.

Selection Criteria for CCFL Circuits

Selecting which CCFL circuit to use for a specific application involves numerous trade-offs. A variety of issues determine which circuit is the "best" approach. At a minimum,

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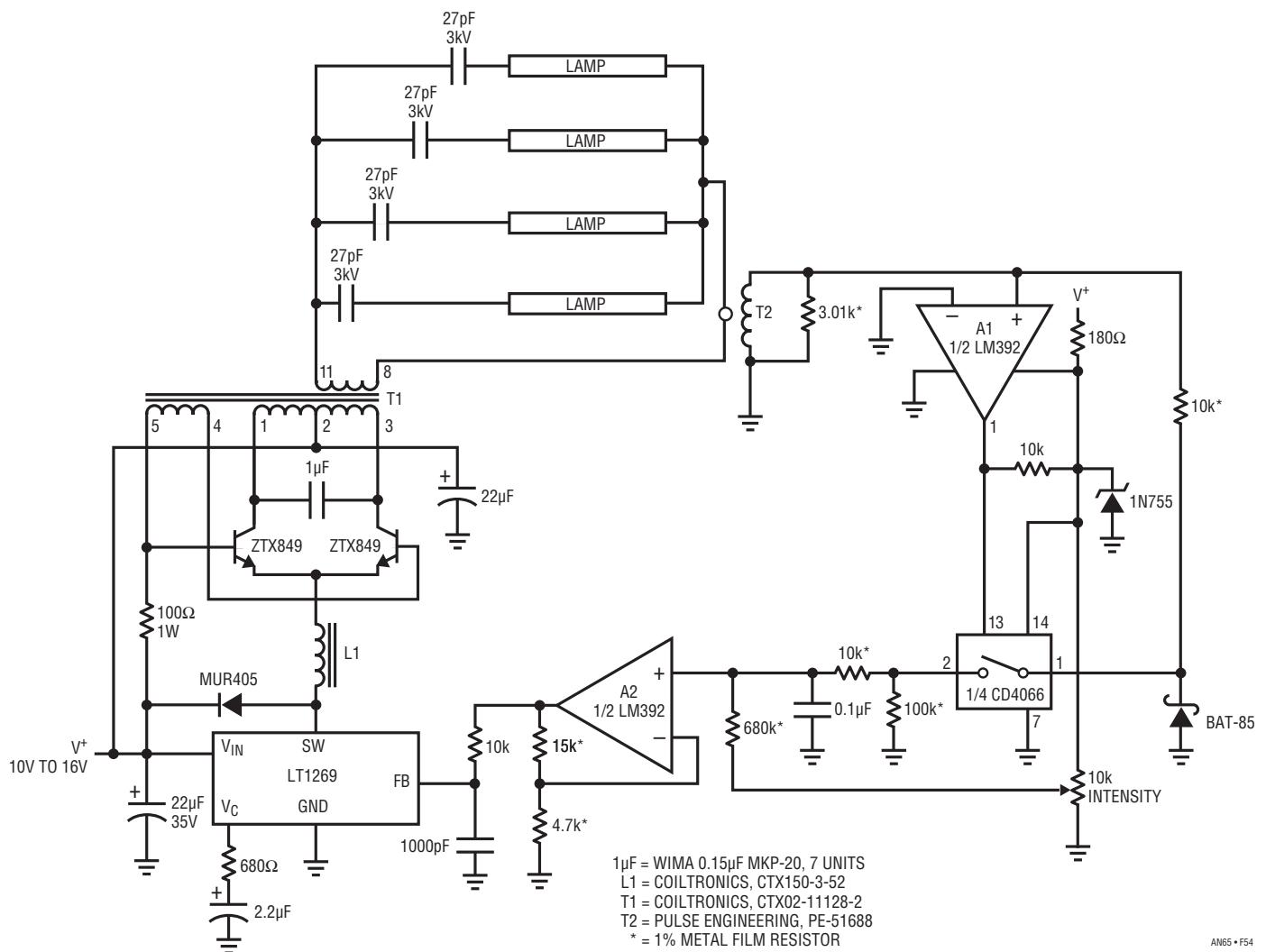
Figure 54. A High Power, Multilamp Display Using the Floating Drive Approach. Power Requirement Necessitates LT1269 Regulator and Discrete Component Approach. Floating Feedback Path Is Via Current Transformer

the user should consider the following guidelines before committing to any approach. Related discussion to all of the following topics is covered in appropriate text sections.

Display Characteristics

The display characteristics (including wiring losses) should be well-understood. Typically, display manufacturers list *lamp* requirements. These specifications are often obtained from the lamp vendor, who usually tests in free air, with no significant parasitic loss paths. This means that actual required power, start and running voltages may significantly differ from data sheet specifications. The only way to be certain of display characteristics is to measure them. The

measured display energy loss can determine if a floating or grounded circuit is applicable. Low loss displays (relatively rare) usually provide better overall efficiency with grounded drive. As losses become worse (unfortunately, relatively common) floating drive becomes a better choice. Efficiency measurements may be required in both modes to determine the best choice. (See "General Optimization and Measurement Considerations.")

Operating Voltage Range

The operating voltage range includes the minimum to maximum voltages the circuit must operate from. In battery-driven apparatus supply range can easily be

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3:1, and sometimes greater. Best backlight performance is usually obtained in the 8V to 28V range. In general, potentials below 7V require some efficiency trade-offs at moderate (1.5W to 3W) power levels. Some systems reduce backlight power when running from the battery, and this can have a pronounced effect on the design. Even seemingly small (e.g., 20%) reductions in power may make painful trade-offs unnecessary. In particular, high turns ratio transformers are required to support low voltage operation at full lamp output. They work well but somewhat less efficiently than lower ratio types due to the higher peak currents characteristic of their operation. Current trends in battery technology encourage system operation at low voltages, necessitating extreme care in transformer selection and Royer circuit design.

Auxiliary Operating Voltages

Auxiliary, logic supply voltages should be used (if available) to run CCFL "housekeeping" currents, such as IC "V_{IN}" pins. This saves power. Always run switching regulators from the lowest potential available, usually 3.3V or 5V. Many systems provide these voltages in switched form, making separate shutdown lines unnecessary. Simply turning off the switching regulator's supply shuts the entire backlight circuit down.

Line Regulation

Grounded lamp circuits, by virtue of their true global feedback, provide the best line regulation. For abrupt changes, a user may notice anything beyond 1% regulation. A grounded circuit easily meets this requirement; a floating circuit usually will. Slowly changing line inputs causing excursions outside 1% are not normally a problem because they are not detectable. Rapid line changes, such as plugging in a systems AC line adapter, require good regulation to avoid annoying display flicker.

Power Requirements

The CCFL's power requirement, *including display and wiring losses*, should be well-defined over all conditions, including temperature and lamp specification variations. Usually, IC versions of floating lamp circuits are restricted to 3W to 4W output power while grounded circuit power is easily scaled.

Supply Current Profile

The backlight is often physically located far "forward" in the system. Impedances in cables, switches, traces and connectors can build up to significant levels. This means that a CCFL circuit should draw operating power continuously, rather than requiring discrete, high current "chunks" from a lossy supply line. Royer-based architectures are nearly ideal in this regard, pulling current smoothly over time and requiring no special bypassing, supply impedance or layout treatment. Similarly, Royer type circuits do not cause significant disturbances to the supply line, preventing noise injection back into the supply.

Lamp Current Certainty

The ability to predict lamp current at full intensity is important to maintain lamp life. Excessive overcurrent greatly shortens lamp life, while yielding little luminosity benefit (see Figure 2). Grounded circuits are excellent in this category with 1% usually achieved. Floating circuits are typically in the 2% to 5% range. Tight current tolerances do not benefit unit/unit display luminosity because lamp emission and display attenuation variations approach ±20% and vary over life.

Efficiency

CCFL backlight efficiency should be considered from two perspectives. The electrical efficiency is the ability of the circuit to convert DC power to high voltage AC and deliver it to the load (lamp and parasitics) with minimum loss. The optical efficiency is perhaps more meaningful to the user. It is simply the ratio of display luminosity to DC power into the CCFL circuit. The electrical and optical losses are lumped together in this measurement to produce a luminosity vs power specification. It is quite significant that *the electrical and optical peak efficiency operating points do not necessarily coincide.* This is primarily due to the lamp's emissivity dependence on waveshape. The optimum waveshape for emissivity may or may not coincide with the circuit's electrical operating peak. In fact, it is quite possible for "inefficient" circuits to produce more light than "more efficient" versions. The only way to ensure peak efficiency in a given situation is to optimize the circuit to the display.

Shutdown

System shutdown almost always requires turning off the backlight. In many cases the low voltage supply is already available in switched form. If this is so, the CCFL circuits shown go off, absorbing very little power. If switched low voltage power is not available the shutdown inputs may be used, requiring an extra control line.

Transient Response

The CCFL circuit should turn on the lamp without attendant overshoot or poor control loop settling characteristics. This can cause objectionable display flicker, and in the worst case result in transformer overstress and failure. Properly prepared floating and grounded CCFL circuits have good transient response, with LT118X-based types inherently easier to optimize.

Dimming Control

The method of dimming should be considered early in the design. All of the circuits shown can be controlled by potentiometers, DC voltages and currents, pulse width modulation or serial data protocol. A dimming scheme with high accuracy at maximum current prevents excessive lamp drive and should be employed.

Open Lamp Protection

The CCFL circuits deliver a current source output. If the lamp is broken or disconnected, compliance voltage is limited by transformer turns ratio and DC input voltage. Excessive voltages can cause arcing and resultant damage. Typically, the transformers withstand this condition but open lamp protection ensures against failures. This feature is built into the LT118X series; it must be added to other circuits.

Size

Backlight circuits usually have severe size and component count limitations. The board must fit within tightly defined dimensions. LT118X series-based circuits offer lowest component count, although board space is usually dominated by the Royer transformer. In extremely tight spaces it may be necessary to physically segment the circuit but this should be considered as a last resort.[17]

Contrast Supply Capability

Some LT118X parts provide contrast supply outputs. The other circuits do not. The LT118X's onboard contrast supply is usually an advantage but space is sometimes so restricted that it cannot be used. In such cases the contrast supply must be remotely located.

Emissions

Backlight circuits rarely cause emission problems and shielding is usually not required. Higher power versions (e.g., >5W) may require attention to meet emission requirements. The fast rise switching regulator output sometimes causes more RFI than the high voltage AC waveform. If shielding is used, its parasitic effects are part of the inverter load and optimization must be carried out with the shield in place.

Summary of Circuits

The interdependence of backlight parameters makes summarizing or rating various approaches a hazardous exercise. There is simply no intellectually responsible way to streamline the selection and design process if optimum results are desired. A meaningful choice *must* be the outcome of laboratory-based experimentation. There are just too many interdependent variables and surprises for a systematic, theoretically based selection. Pure analytics are pretty; working circuits come from the bench. Some generalizations having limited usefulness are, however, possible. Figures 55 and 56 attempt to summarize salient characteristics vs part type and may (however cautiously) be considered a beginning point.[18]

Figure 55 summarizes characteristics of all the circuits. Figure 56 focuses on the features of the LT118X series parts.

Note 17. See Appendix G, "Layout, Component and Emissions Considerations."
Note 18. Readers detecting author ambivalence about inclusion of Figures 55 and 56 are not hallucinating.

ISSUES	LT118X SERIES	LT117X SERIES	LT137X SERIES
Optical Efficiency	Grounded output versions display dependent. Floating versions usually 5% to 20% better.	Display dependent	Display dependent
Electrical Efficiency	Grounded output versions— 75% to 90%, depending on supply voltage and display. Floating output versions slightly lower.	75% to 90%, depending on supply voltage and display	75% to 92%, depending on supply voltage and display
Lamp Current Certainty	1% to 2% for grounded versions, 1% to 4% for floating output types	2% maximum	2% maximum
Line Regulation	0.1% to 0.3% for grounded types, 0.5% to 6% for floating versions	0.1% to 0.3%	0.1% to 3%
Operating Voltage Range	5.3V to 30V, depending on output power, temperature range, display, etc.	4.0V to 30V, depending on output power, temperature range, display, etc.	4.0V to 30V, depending on output power, temperature range, display, etc.
Power Range	0.75W to 6W typical	0.75W to 20W typical	0.5W to 6W typical
Supply Current Profile	Continuous—no high current peaks	Continuous—no high current peaks	Continuous—no high current peaks
Shutdown Control	Yes—logic compatible	Requires small FET or bipolar transistor	Yes—logic compatible
Transient Response— Overshoot	Excellent—no optimization required	Excellent—requires optimization in some cases	Excellent—requires optimization in some cases
Dimming Control	Pot., PWM, variable DC voltage or current. LT1186 has serial digital input with data storage.	Pot., PWM, variable DC voltage or current	Pot., PWM, variable DC voltage or current
Emissions	Low	Low	Low, although high power versions may require attention to layout and shielding
Open Lamp Protection	Internal to IC	Requires external small-signal transistor and some discretes at high supply voltages	Requires external small-signal transistor and some discretes at high supply voltages
Size	Low component count, small overall board footprint. 200kHz magnetics.	Small—100kHz magnetics	Small—1MHz magnetics for fastest versions
Contrast Supply Capability	Various contrast supply options available, including bipolar output	No	No

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Figure 55. Design Issues vs Typical Part Choice. Chart Makes Simplistic Assumptions and Is Intended As a Guide Only

LT1269/LT1270	LT1301	LT1173
Display dependent	Display dependent	Display dependent
75% to 90%, depending on supply voltage and display	70% to 88%, depending on supply voltage and display	65% to 75%, depending on supply voltage and display
2% maximum	2% typical	5%
0.1% to 0.3%	0.1% to 0.3%	8% to 10%
4.5V to 30V, depending on output power, temperature range, display, etc.	2V to 10V practical	2V to 6V practical
5W to 35W typical	0.02W to 1W practical	Essentially 0W to about 0.6W
Continuous—no high current peaks	Continuous—no high current peaks	Irregular—relatively high current peaking requires attention to supply rail impedance
Requires small FET or bipolar transistor	Yes—logic compatible	Logic compatible shutdown practical
Excellent—requires optimization in some cases	Excellent—no optimization required	Excellent—no optimization required
Pot., PWM, variable DC voltage or current	Pot., PWM, variable DC voltage or current	Pot., PWM, variable DC voltage or current
High power mandates attention to layout and shielding	Real low	Itsy-bitsy
Requires external small-signal transistor and some discretes at high supply voltages	Requires external small-signal tansistor and some discretes, but low supply voltages usually eliminate this consideration	None, but low supply, low power operation usually eliminates this issue
Relatively large due to high power 100kHz magnetics	Very small—low power magnetics cut size	Small—low power magnetics cut size
No	No	No

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	LT1182	LT1183	LT1184	LT1184F	LT1186
Floating Lamp Operation	Yes	Yes	No	Yes	Yes
Grounded Lamp Operation	Yes	Yes	Yes	Yes	Yes
Contrast Supply	Bipolar Contrast Outputs	Unipolar Contrast Outputs	No	No	No
Voltage Reference Available	No	Yes	Yes	Yes	No
Internal Control DAC	No	No	No	No	Yes

Figure 56. Features of Various LT118X IC Backlight Controllers

General Optimization and Measurement Considerations

Once a display/lamp combination has been picked, the appropriate circuit can be selected and optimized. "Optimization" implies maximizing performance in those areas most important in a particular application. This may involve trading off characteristics in one area to gain advantage in another. The circuit types described impose mild penalty in this regard because they are quite flexible.

A desirable characteristic is something often loosely referred to as "efficiency." There are really two types of efficiency in a backlight circuit. The *optical* efficiency measures the circuit/display combination as a transducer. It is the ratio of light output to electrical power input. This ratio lumps the converter's electrical loss with lamp and display losses. A backlight's *electrical* efficiency measures the converter's electrical input vs output power without regard to optical performance. Obviously, high electrical efficiency is required and a reliable way to measure it is desirable. More subtly, the ability to measure and manipulate purely electrical terms offers a way to influence optical efficiency. This is so because the lamp is sensitive to the drive waveform's shape. Best emissivity and lifetime are usually obtained with low crest factor, sinusoidal waveforms. The Royer circuit's transformer and capacitors can be selected to provide this characteristic for any given display/lamp combination. Doing this optimizes lamp drive but also effects the converter's electrical efficiency. This interaction between the optimum electrical and optical operating points must be accounted for to obtain best optical efficiency. The relationship is quite complex with a number of variables determining just where peak optical efficiency occurs.

Typically, optical output peaking occurs with a fairly clean, low harmonic waveform at the Royer collectors (Figure 57). This is usually the result of a relatively large resonating capacitor and a small ballast capacitor. Conversely, the converter's peak electrical efficiency point usually comes just as appreciable second harmonic appears in the Royer collector waveform (Figure 58). The peak electrical and optical efficiency points almost never coincide and optical efficiency often occurs 5% or more off the electrical efficiency peak. Happily, this very messy situation can be resolved by a relatively simple functional trim. The trimming procedure assumes transformer turns ratio and ballast capacitor values commensurate with the lowest


5V/DIV

5µs/DIV   AN65•F57
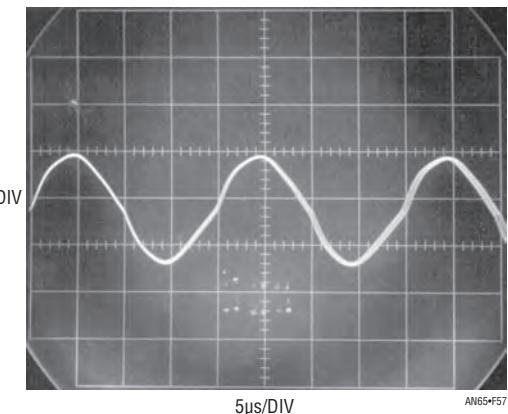
Figure 57. Typical Royer Collector Waveform at the Peak Optical Output Point. Relatively Large Resonating Capacitor May Degrade Electrical Efficiency


5V/DIV

5µs/DIV   AN65•F57
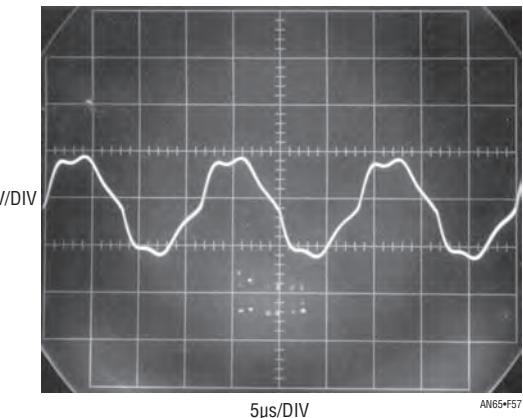
Figure 58. Typical Royer Collector Waveform at the Peak Electrical Efficiency Point. Relatively High Harmonic Content May Degrade Optical Efficiency

required circuit operating voltage have been chosen. If this factor is not considered, the optical efficiency peak will be realized but the design may not regulate at low supply voltages. Low supply voltage operation mandates high turns ratio and larger ballast capacitor values for a given display loss. If display loss is high, ballast capacitor value generally must rise to offset voltage dividing effects between it and the display's parasitic loss paths. Establish the lowest values of turns ratio and ballast capacitor that maintain regulation at minimum supply voltage before performing the trim.

Achieving peak optical efficiency involves comparing display luminosity to input power for different resonating capacitor values. For a given lamp/transformer/ ballast capacitor combination different resonating capacitors produce varying amounts of light. Large values tend to smooth harmonics, peaking optical output but increasing converter circulating losses. Smaller values promote lower circulating currents but less light output. Figure 59 shows typical results for five capacitor values at a forced 10V main supply and 5mA lamp current. Large values produce more light but require more supply current. The raw data is expressed as the ratio of light output-per-watt of input power in the right-most column. This Nits-per-Watt ratio peaks at 0.1µF, indicating the best optical efficiency.[19]

This test must be performed in a stable thermal environment because of the lamp's emission sensitivity to temperature (see Figure 3). Additionally, some arrangement for rapidly switching the capacitor values is desirable. This avoids power interruptions and the resultant long display warm-up times.

CAPACITOR (µF)	10V MAIN SUPPLY CURRENT	5V SUPPLY CURRENT	TOTAL SUPPLY WATTS	INTENSITY (NITS)	NITS/WATT
0.15	0.304	0.014	3.11	118	37.9
0.1	0.269	0.013	2.75	112	40.7
0.068	0.259	0.013	2.65	101	38.1
0.047	0.251	0.013	2.57	95	37.3
0.033	0.240	0.013	2.46	88	35.7

Note: Maintain I_{MAIN} Supply = 10.0V and I_{LAMP} = 5mA$_{RMS}$ under all conditions.

Figure 59. Typical Data Taken for Optical Efficiency Optimization. Note Emissivity Peak (Nits/Watt) for 0.1µF Resonating Value, Indicating Best Trade-Off Point for Electrical vs Optical Efficiency. Data Should Be Retaken for Several Ballast Capacitor Values to Ensure Maximum Optical Efficiency

Electrical Efficiency Optimization and Measurement

Several points should be kept in mind when observing operation of these circuits. The high voltage secondary can only be monitored with a wideband, high voltage probe fully specified for this type of measurement. *The vast majority of oscilloscope probes will break down and fail if used for this measurement.*[20] Tektronix probe types P-6007 and P-6009 (acceptable in some cases) or types P6013A and P6015 (preferred) probes must be used to read L1's output.

Another consideration involves observing waveforms. The switching regulator frequency is completely asynchronous from the Royer converter's switching. As such, most oscilloscopes cannot simultaneously trigger and display all the circuit's waveforms. Figure 36 was obtained using a dual beam oscilloscope (Tektronix 556).Traces A and B are triggered on one beam while the remaining traces are triggered on the other beam. Single beam instruments with alternate sweep and trigger switching (e.g., Tektronix 547) can also be used but are less versatile and restricted to four traces.

Obtaining and verifying high electrical efficiency[21] requires some amount of diligence. The optimum efficiency values given for C1 and C2 (C1 is the resonating capacitor and C2 is the ballast capacitor) are typical and will vary for specific types of lamps. An important realization is that the term "lamp" includes the *total* load seen by the transformer's secondary. This load, reflected back to the primary, sets transformer input impedance. The transformer's input impedance forms an integral part of the LC tank that produces the high voltage drive. Because of this, circuit efficiency must be optimized with the wiring,

Note 19: Optical measurement units are beyond arcane; a monument to obscuration. Candela/Meter^2 is a basic unit, and 1 Nit = 1 Candela/Meter^2. "Nit" is a contracted form of the Latin word "Nitere," meaning "to emit light . . . to sparkle."
Note 20: Don't say we didn't warn you!
Note 21: The term "efficiency" as used here applies to electrical efficiency. In fact, the ultimate concern centers around the efficient conversion of power supply energy into light. Unfortunately, lamp types show considerable deviation in their current-to-light conversion efficiency. Similarly, the emitted light for a given current varies over the life and history of any particular lamp. As such, this text portion treats "efficiency" on an electrical basis; the ratio of power removed from the primary supply to the power delivered to the lamp. When a lamp/display combination has been selected, the ratio of primary supply power to lamp emitted light energy may be measured with the aid of a photometer. This is covered in the immediately preceding text and Appendix D.

display housing and physical layout arranged *exactly* the same way they will be built in production. Deviations from this procedure will result in lower efficiency than might otherwise be possible. In practice, a "first cut" efficiency optimization with "best guess" lead lengths and the intended lamp in its display housing usually produces results within 5% of the achievable figure. Final values for C1 and C2 may be established when the physical layout to be used in production has been decided on. C1 sets the circuit's resonance point, which varies to some extent with the lamp's characteristic. C2 ballasts the lamp, effectively buffering its negative resistance characteristic. Small C2 values provide the most load isolation but require relatively large transformer output voltage for loop closure. Large C2 values minimize transformer output voltage but degrade load buffering. C2 values also affect waveform distortion, influencing lamp emissivity and optical efficiency (see previous text discussion). Also, C1's "best" value is somewhat dependent on the lamp type used. Both C1 and C2 must be selected for given lamp types. Some interaction occurs, but generalized guidelines are possible. Typical values for C1 are 0.01µF to 0.15µF. C2 usually ends up in the 10pF to 47pF range. C1 *must* be a low loss capacitor and substitution of the recommended devices is not recommended. A poor quality dielectric for C1 can easily degrade efficiency by 10%. Before capacitor selection the Q1/Q2 base drive resistor should be set to a value which ensures saturation, e.g., 470Ω. Next, C1 and C2 are selected by trying different values for each and iterating towards best efficiency. During this procedure ensure that loop closure is maintained. Several trials usually produce the optimum C1 and C2 values. Note that the highest efficiencies are not necessarily associated with the most esthetically pleasing waveshapes, particularly at Q1, Q2 and output. Finally, the base drive resistor's value should be optimized.

The base drive resistor's value (nominally 1k) should be selected to provide full V_{CE} saturation without inducing base overdrive or beta starvation. This point may be established for any lamp type by determining the peak collector current at full lamp power.

The base resistor should be set at the largest value that ensures saturation for worst-case transistor beta. This condition may be verified by varying the base drive

resistor about the ideal value and noting small variations in input supply current. The minimum obtainable current corresponds to the best beta vs saturation trade-off. In practice, supply current rises slightly on either side of this point. This "double value" behavior is due to efficiency degradation caused by either excessive base drive or saturation losses.

Other issues influencing efficiency include lamp wire length and energy leakage from the lamp. The high voltage side(s) of the lamp should have the smallest practical lead length. Excessive length results in radiative losses which can easily reach 3% for a 3-inch wire. Similarly, no metal should contact or be in close proximity to the lamp. This prevents energy leakage which can exceed 10%.[22]

It is worth noting that a custom designed lamp affords the best possible results. A jointly tailored lamp/circuit combination permits precise optimization of circuit operation, yielding highest efficiency.

These considerations should be made with knowledge of other LCD issues. See Appendix B, "Mechanical Design Considerations for Liquid Crystal Displays." This section was guest-written by Charles L. Guthrie of Sharp Electronics Corporation.

Special attention should be given to the circuit board layout since high voltage is generated at the output. The output coupling capacitor must be carefully located to minimize leakage paths on the circuit board. A slot in the board will further minimize leakage. Such leakage can permit current flow outside the feedback loop, wasting power. In the worst case, long term contamination buildup can increase leakage inside the loop, resulting in starved lamp drive or destructive arcing. It is good practice for minimization of leakage to break the silk screen line which outlines the transformer. This prevents leakage from the high voltage

Note 22: This footnote annotates similar issues raised in Footnote 12 and associated text. The repetition is based on the necessity for emphasis. A very simple experiment quite nicely demonstrates the effects of energy leakage. Grasping the lamp at its low voltage end (low field intensity) with thumb and forefinger produces almost no change in circuit input current. Sliding the thumb/forefinger combination towards the high voltage (higher field intensity) lamp end produces progressively greater input currents. Don't touch the high voltage lead or your may receive an electrical shock. Repeat: Do not touch the high voltage lead or you may receive an electrical shock.

secondary to the primary. Another technique for minimizing leakage is to evaluate and specify the silk screen ink for its ability to withstand high voltages. Appendix G, "Layout, Component and Emissions Considerations," details high voltage layout practice.

Electrical Efficiency Measurement

Once these procedures have been followed efficiency can be measured. Efficiency may be measured by determining lamp current and voltage. Measuring current involves utilization of a wideband, high accuracy clip-on current probe having a true (thermally based) RMS readout. No commercially manufactured current probe will meet the accuracy and bandwidth requirements and the probe must be constructed.[23]

Lamp RMS voltage is measured at the lamp with a wideband, properly compensated high voltage probe.[24] Multiplying these two results gives power in watts, which may be compared to the DC input supply (E)(I) product. In practice, the lamp's current and voltage contain small out-of-phase components but their error contribution is negligible.

Both the current and voltage measurements require a wideband true RMS voltmeter. The meter must employ a thermal type RMS converter—the more common logarithmic computing type-based instruments are inappropriate because their bandwidth is too low.

The previously recommended high voltage probes are designed to see a 1MΩ/10pF-22pF oscilloscope input. The RMS voltmeters have a 10MΩ input. This difference necessitates an impedance matching network between the probe and the voltmeter. Floating lamp circuits require this matching and differential measurement, severely complicating instrumentation design. See Footnote 24.

Feedback Loop Stability Issues

The circuits shown to this point rely on closed-loop feedback to maintain the operating point. All linear closed-loop systems require some form of frequency compensation to achieve dynamic stability. Circuits operating with relatively low power lamps may be frequency compensated by simply overdamping the loop. Text Figures 35, 37 and 38 use this approach. The higher power operation

associated with color displays requires more attention to loop response. The transformer produces much higher output voltages, particularly at start-up. Poor loop damping can allow transformer voltage ratings to be exceeded, causing arcing and failure. As such, higher power designs may require optimization of transient response characteristics. LT118X series parts almost never require optimization because their error amplifier's gain/phase characteristics are specially tailored to CCFL load characteristics. The LT1172, LT1372 and other general purpose switching regulators require more attention to ensure proper behavior. The following discussion, applicable to general purpose LTC switching regulators in CCFL applications, uses the LT1172 as an example.

Figure 60 shows the significant contributors to loop transmission in these circuits. The resonant Royer converter delivers information at about 50kHz to the lamp. This information is smoothed by the RC averaging time constant and delivered to the LT1172's feedback terminal
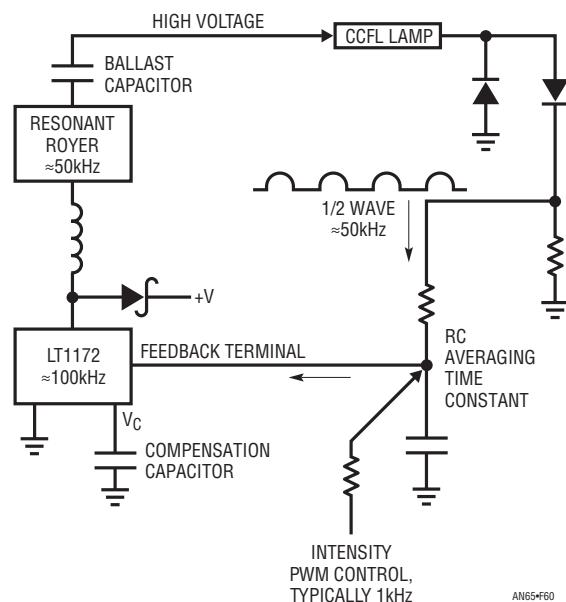

Figure 60. Delay Terms in the Feedback Path. The RC Time Constant Dominates Loop Transmission Delay and Must Be Compensated for Stable Operation

Note 23: Justification for this requirement and construction details appear in Appendix C, "Achieving Meaningful Electrical Measurements."
Note 24: Measuring floating lamp circuit voltages is a particularly demanding exercise requiring a wideband differential high voltage probe. Probe construction details appear in Appendix C.

as DC. The LT1172 controls the Royer converter at a 100kHz rate, closing the control loop. The capacitor at the LT1172 rolls off gain, nominally stabilizing the loop. This compensation capacitor must roll off the gain-bandwidth at a low enough value to prevent the various loop delays from causing oscillation.

Which of these delays is the most significant? From a stability viewpoint the LT1172's output repetition rate and the Royer's oscillation frequency are sampled data systems. Their information delivery rate is far above the RC averaging time constants delay and is not significant. The RC time constant is the major contributor to loop delay. This time constant must be large enough to turn the half wave rectified waveform into DC. It also must be large enough to average any intensity control PWM signal to DC. Typically, these PWM intensity control signals come in at a 1kHz rate (see Appendix F, "Intensity Control and Shutdown Methods"). The RC's resultant delay dominates loop transmission. It must be compensated by the capacitor at the LT1172. A large enough value for this capacitor rolls off loop gain at low enough frequency to provide stability. The loop simply does not have enough gain to oscillate at a frequency commensurate with the RC delay.[25]

This form of compensation is simple and effective. It ensures stability over a wide range of operating conditions. It does, however, have poorly damped response at system turn-on. At turn-on the RC lag delays feedback, allowing output excursions well above the normal operating point. When the RC acquires the feedback value the loop stabilizes properly. This turn-on overshoot is not a concern if it is well within transformer breakdown ratings. Color displays, running at higher power, usually require large initial voltages. If loop damping is poor, the overshoot may be dangerously high. Figure 61 shows such a loop responding to turn-on. In this case the RC values are 10k and 4.7µF, with a 2µF compensation capacitor. Turn-on overshoot exceeds 3500V for over 10ms! Ring-off takes over 100ms before settling occurs. Additionally, an inadequate (too small) ballast capacitor and excessively lossy layout force a 2000V output once loop settling occurs. This photo was taken with a transformer rated well below this figure. The resultant arcing caused transformer destruction, resulting in field failures. A typical destroyed transfomer appears in Figure 62.

A = 1000V/DIV

20ms/DIV   AN65•F61
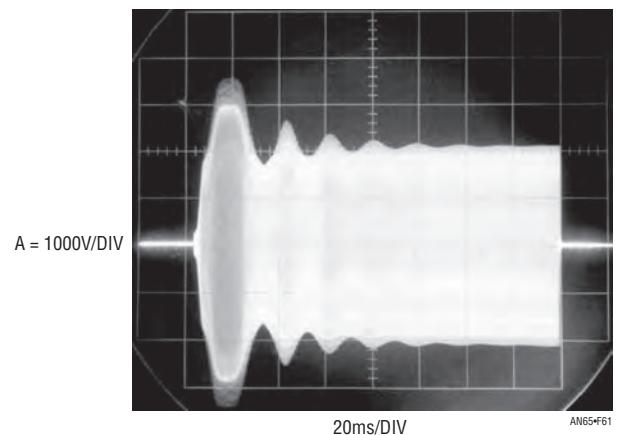
Figure 61. Destructive High Voltage Overshoot and Ring-Off Due to Poor Loop Compensation. Transformer Failure and Field Recall Are Nearly Certain. Job Loss May also Occur


AN65•F62

Figure 62. Poor Loop Compensation Caused This Transformer Failure. Arc Occured in High Voltage Secondary (Lower Right). Resultant Shorted Turns Caused Overheating

Figure 63 shows the same circuit with the RC values reduced to 10k and 1µF. The ballast capacitor and layout have also been optimized. Figure 63 shows peak voltage reduced to 2.2kV with duration down to about 2ms (note horizontal scale change). Ring-off is also much quicker with lower amplitude excursion. Increased ballast capacitor value and wiring layout optimization reduce running voltage to 1300V. Figure 64's results are even better. Changing the compensation capacitor to a 3kΩ/2µF network introduces a leading response into the loop, allowing faster acquisition. Now, turn-on excursion is slightly lower, but greatly reduced in duration (again, note horizontal scale change). The running voltage remains the same.

Note 25. The high priests of feedback refer to this as "Dominant Pole Compensation." The rest of us are reduced to more pedestrian descriptives.

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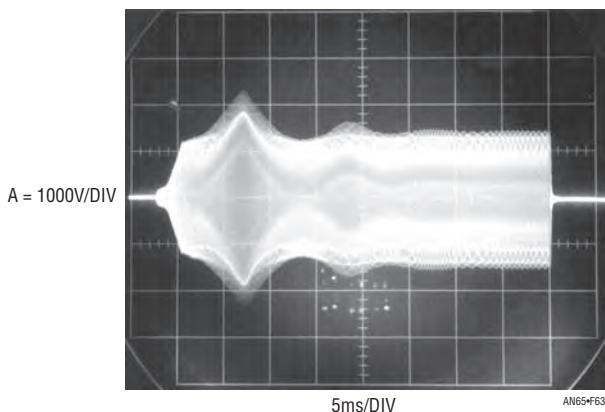
Figure 63. Reducing RC Time Constant Improves Transient Response, Although Peaking, Ring-Off and Run Voltage Are Still Excessive

A = 1000V/DIV

5ms/DIV   AN65•F63


A = 1000V/DIV

2ms/DIV   AN65•F64
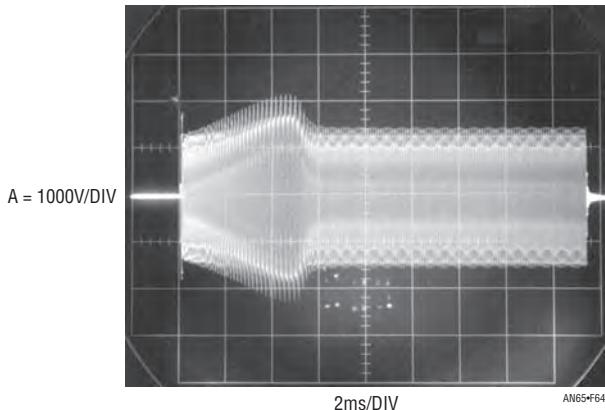
Figure 64. Additional Optimization of RC Time Constant and Compensation Capacitor Reduces Turn-On Transient. Run Voltage Is Large, Indicating Possible Lossy Layout and Display

The photos show that changes in compensation, ballast value and layout result in dramatic reductions in overshoot amplitude and duration. Figure 62's performance almost guarantees field failures while Figures 63 and 64 do not overstress the transformer. Even with the improvements, more margin is possible if display losses can be controlled. Figures 62, 63 and 64 were taken with an exceptionally lossy display. The metal enclosure was very close to the metallic foil wrapped lamp, causing large losses with subsequent high turn-on and running voltages. If the display is selected for lower losses, performance can be greatly improved.

Figure 65 shows a low loss display responding to turn-on with a 2µF compensation capacitor and 10k/1µF RC values. Trace A is the transformer's output while Traces B and C are the LT1172's $V_{COMPENSATION}$ and Feedback pins, respectively. The output overshoots and rings badly, peaking to about 3000V. This activity is reflected by overshoots at the $V_{COMPENSATION}$ pin (the LT1172's error amplifier output) and the Feedback pin. In Figure 66 the RC is reduced to 10kΩ/0.1µF. This substantially reduces loop delay. Overshoot goes down to only 800V—a reduction of almost a factor of four. Duration is also much shorter.

A = 2000V/DIV

B = 0.5V/DIV

C = 1V/DIV

10ms/DIV   AN65•F65
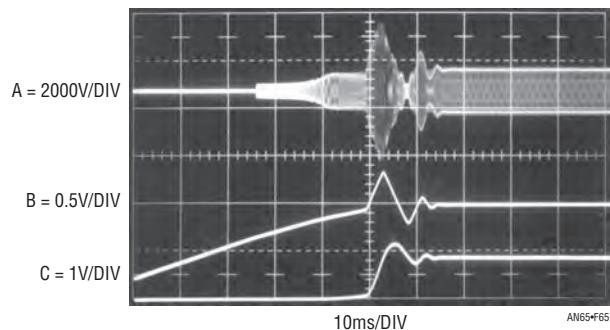
Figure 65. Waveform for a Lower Loss Layout and Display. High Voltage Overshoot (Trace A) Is Reflected at Compensation Node (Trace B) and Feedback Pin (Trace C)


A = 2000V/DIV

B = 0.5V/DIV

C = 1V/DIV

10ms/DIV   AN65•F66
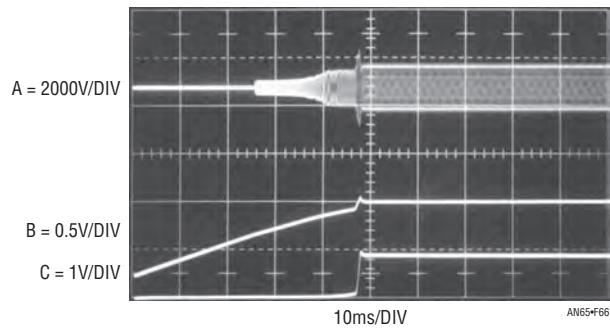
Figure 66. Reducing RC Time Constant Produces Quick, Clean Loop Behavior. Low Loss Layout and Display Result in 650V$_{RMS}$ Running Voltage

The V$_{COMPENSATION}$ and Feedback pins reflect this tighter control. Damping is much better, with slight overshoot induced at turn-on. Further reduction of the RC to 10kΩ/0.01µF (Figure 67) results in even faster loop capture but a new problem appears. In Trace A lamp turn-on is so fast the overshoot does not register in the photo. The V$_{COMPENSATION}$ (Trace B) and feedback nodes (Trace C) reflect this with exceptionally fast response. Unfortunately, the RC's light filtering causes ripple to appear when the feedback node settles. As such, Figure 66's RC values are probably more realistic for this situation.

The lesson from this exercise is clear. The higher voltages involved in color displays mandate attention to transformer outputs. Under running conditions layout and display losses can cause higher loop compliance voltages, degrading efficiency and stressing the transformer. At turn-on improper compensation causes huge overshoots, resulting in possible transformer destruction. Isn't a day of loop and layout optimization worth a field recall?


A = 2000V/DIV

B = 0.5V/DIV

C = 1V/DIV

10ms/DIV   AN65•F67
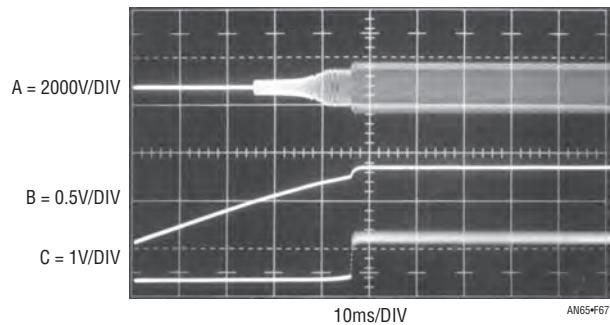
Figure 67. Very Low RC Value Provides Even Faster Response, but Ripple at Feedback Pin (Trace C) Is Too High. Figure 66 Is the Best Compromise

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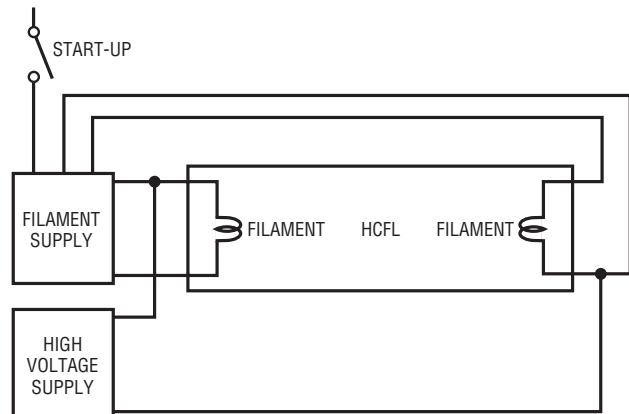
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APPENDIX A

"HOT" CATHODE FLUORESCENT LAMPS

Many CCFL characteristics are shared by so-called "hot" cathode fluorescent lamps (HCFLs). The most significant difference is that HCFLs contain filaments at each end of the lamp (see Figure A1). When the filaments are powered they emit electrons, lowering the lamp's ionization potential. This means a significantly lower voltage will start the lamp. Typically, the filaments are turned on, a relatively modest voltage impressed across the lamp and start-up occurs. Once the lamp starts, filament power is removed. Although HCFLs reduce the high voltage requirement they require a filament supply and sequencing circuitry. The CCFL circuits shown in the text will start and run HCFLs without using the filaments. In practice, this involves simply driving the filament connections at the HCFL ends as if they were CCFL electrodes.


Figure A1. A Conceptual Hot Cathode Fluorescent Lamp Power Supply. Heated Filaments Liberate Electrons, Lowering the Lamp's Start-Up Voltage Requirement. CCFL Supply Discussed in Text Eliminates Filament Supply

APPENDIX B

MECHANICAL DESIGN CONSIDERATIONS FOR LIQUID CRYSTAL DISPLAYS

Charles L. Guthrie, Sharp Electronics Corporation

Introduction

As more companies begin the manufacturing of their next generation of computers, there is a need to reduce the overall size and weight of the units to improve their portability. This has sparked the need for more compact designs where the various components are placed in closer proximity, thus making them more susceptible to interaction from signal noise and heat dissipation. The following is a summary of guidelines for the placement of the display components and suggestions for overcoming difficult design constraints associated with component placement.

In notebook computers the thickness of the display housing is important. The design usually requires the display to be in a pivotal structure so that the display may be folded down over the keyboard for transportation. Also, the outline dimensions must be minimal so that the package will remain as compact as possible. These two constraints drive the display housing design and placement

of the display components. This discussion surveys each of the problems facing the designer in detail and offers suggestions for overcoming the difficulties to provide a reliable assembly.

The problems facing the pen-based computer designer are similar to those realized in notebook designs. In addition, however, pen-based designs require protection for the face of the display. In pen-based applications, as the pen is moved across the surface of the display, the pen has the potential for scratching the front polarizer. For this reason the front of the display must be protected. Methods for protecting the display face while minimizing effects on the display image are given.

Additionally, the need to specify the flatness of the bezel is discussed. Suggestions for acceptable construction techniques for sound design are included. Further, display components likely to cause problems due to heat buildup are identified and methods for minimization of the heat's effects are presented.

The ideas expressed here are not the only solutions to the various problems and have not been assessed as to whether they may infringe on any patents issued or applied for.

Flatness and Rigidity of the Bezel

In the notebook computer the bezel has several distinct functions. It houses the display, the inverter for the backlight, and in some instances, the controls for contrast and brightness of the display. The bezel is usually designed to tilt to set the optimum viewing angle for the display.

It is important to understand that the bezel must provide a mechanism to keep the display flat, particularly at the mounting holds. Subtle changes in flatness place uneven stress on the glass which can cause variations in contrast across the display. Slight changes in pressure may cause significant variation in the display contrast. Also, at the extreme, significantly uneven pressures can cause the display glass to fail.

Because the bezel must be functional in maintaining the flatness of the display, consideration must be made for the strength of the bezel. Care must be taken to provide structural members while minimizing the weight of the unit. This may be executed using a parallel grid, normal to the edges of the bezel, or angled about 45° off of the edges of the bezel. The angled structure may be more desirable in that it provides resistance to torquing the unit while lifting the cover with one hand. Again, the display is sensitive to stresses from uneven pressure on the display housing.

Another structure which will provide excellent rigidity, but adds more weight to the computer, is a "honeycomb" structure. This "honeycomb" structure resists torquing from all directions and tends to provide the best protection for the display.

With each of these structures it is easy to provide mounting assemblies for the display. "Blind nuts" can be molded into the housing. The mounting may be done to either the front or rear of the bezel. Attachment to the rear may provide better rigidity for placement of the mounting hardware.

One last caution is worth noting in the development of a bezel. The bezel should be engineered to absorb most of the shock and vibration experienced in a portable computer. Even though the display has been carefully designed, the notebook computer presents extraordinary shock and abuse problems.

Avoiding Heat Buildup in the Display

Several of the display components are sources for heat problems. Thermal management must be taken into account in the design of the display bezel. A heated display may be adversely affected; a loss of contrast uniformity usually results. The cold cathode fluorescent tube (CCFT) itself gives off a small amount of heat relative to the amount of power dissipated in its glow discharge. Likewise, even though the inverters are designed to be extremely efficient, there is some heat generated. The buildup of heat in these components will be aggravated by the typically "tight" designs currently being introduced. There is little ventilation designed into most display bezels. To compound the problem, the plastics used are poor thermal conductors, thus causing the heat to build up which may affect the display.

Some current designs suffer from poor placement of the inverter and/or poor thermal management techniques. These designs can be improved even where redesign of the display housing with improved thermal management is impractical.

One of the most common mistakes in current designs is that there has been no consideration for the buildup of heat from the CCFT. Typically, the displays for notebook applications have only one CCFT to minimize display power requirements. The lamp is usually placed along the right edge of the display. Since the lamp is placed very close to the display glass, it can cause a temperature rise in the liquid crystal. It is important to note that variations in temperature of as little as 5°C can cause an apparent non-uniformity in display contrast. Variations caused by slightly higher temperature variations will cause objectionable variations in the contrast and display appearance.

To further aggravate the situation, some designs have the inverter placed in the bottom of the bezel. This has a tendency to cause the same variations in contrast, particularly when the housing does not have any heat sinking for the inverter. This problem manifests itself as a "blooming" of the display, just above the inverter. This "blooming" looks like a washed out area where, in the worst case, the characters on the display fade completely.

The following section discusses the recommended methods for overcoming these design problems.

Placement of the Display Components

One of the things that can be done is to design the inverter into the base of the computer with the motherboard. In some applications this is impractical because this requires the high voltage leads to be mounted within the hinges connecting the display bezel to the main body. This causes a problem with strain relief of the high voltage leads and thus with UL certification.

One mistake made most often is placing the inverter at the bottom of the bezel next to the lower edge of the display. It is a fact that heat rises, yet this is one of the most overlooked problems in new notebook designs. Even though the inverters are very efficient, some energy is lost in the inverter in the form of heat. Because of the insulating properties of the plastic materials used in the bezel construction, heat builds up and affects the display contrast.

Designs with the inverter at the bottom can be improved in one of three ways. The inverter can be relocated away from the display, heat sinking materials can be placed between the display and the inverter, or ventilation can be provided to remove the heat.

In mature designs it may be impractical to do what is obvious and move the inverter up to the side of the display towards the top of the housing. In these cases the inverter may be insulated from the display with a "heat dam." One method of accomplishing this would be to use a piece of mica insulator die cut to fit tightly between the inverter and the display. This heat dam would divert the heat around the end of the display bezel to rise harmlessly to the top of the housing. Mica is recommended in this application because of its thermal and electrical insulation properties.

The last suggestion for removing heat is to provide some ventilation to the inverter area. This has to be done very carefully to prevent exposing the high voltage. Ventilation may not be a practical solution because resistance to liquids and dust is compromised.

The best solution for the designer of new hardware is to consider the placement of the inverter to the side of the display and at the top of the bezel. In existing designs of this type the effects of heat from the inverter, even in tight housings, has been minimal or nonexistent.

One problem which is aggravated by the placement of the inverter at the bezel is heat dissipated by the CCFT. In designs where the inverter is placed up and to the side of the display, fading of the display contrast due to CCFT heat is not a problem. However, when the inverter is placed at the bezel bottom, some designs experience a loss of contrast aggravated by the heat from the CCFT and inverter.

In cases where the inverter must be left at the bottom, and the CCFT is causing a loss of contrast, the problem can be minimized by using an aluminum foil heat sink. This does not remove the heat from the display but dissipates it over the entire display area, thus normalizing the display contrast. The aluminum foil is easy to install and in some present designs has successfully improved the display contrast.

Remember that the objection to the contrast variation stems more from nonuniformity than from a total loss of contrast.

Protecting the Face of the Display

One of the last considerations in the design of notebook and pen-based computers is protection of the display face. The front polarizer is made of a mylar base and thus is susceptible to scratching. The front protection for the display, along with providing scratch protection, may also provide an antiglare surface.

There are several ways that scratch resistance and antiglare surfaces can be incorporated. A glass or plastic cover may be placed over the display, thus providing protection. The material should be placed as close to the display as possible to minimize possible parallax problems due to reflections off of the cover material. With antiglare materials the further the material is from the front of the display the greater the distortion.

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In pen applications, the front antiscratch material is best placed in contact with the front glass of the display. The cover glass material normally needs to be slightly thicker to protect the display from distortion when pressure is being exerted on the front.

There are several methods for making the pen input devices. Some use the front surface of the cover glass to provide input data and some use a field effect to a printed wiring board on the back of the display. When the pen input is on the front of the display, the input device is usually on a glass surface.

To limit specular reflection in this application, the front cover glass should be bonded to the display. Care must be taken to ensure that the coefficient of thermal expansion is matched for all of the materials used in the system. Because of the difficulties encountered with the bonding of the cover glass, and the potential to destroy the display through improper workmanship, consulting an expert is strongly recommended.

APPENDIX C

ACHIEVING MEANINGFUL ELECTRICAL MEASUREMENTS

Obtaining reliable efficiency data for the CCFL circuits presents a high order difficulty measurement problem. The accuracy required in the high frequency AC measurements is uncomfortably close to the state-of-the-art. Establishing and maintaining accurate wideband AC measurements is a textbook example of attention to measurement technique. The combination of high frequency, harmonic laden waveforms and high voltage makes meaningful results difficult to obtain. The choice, understanding and use of test instrumentation is crucial. Clear thinking is needed to avoid unpleasant surprises![1]

The lamp's current and voltage waveforms contain energy content over a wide frequency range. Most of this energy is concentrated at the inverter's fundamental frequency and immediate harmonics. However, if 1% measurement uncertainty is desirable, then energy content out to 10MHz must be accurately captured. Figure C1, a spectrum analysis of lamp current, shows significant energy out to 500kHz. Diminished, but still significant, content shows

Note 1: It is worth considering that various constructors of text Figure 35 have reported efficiencies ranging from 8% to 115%.
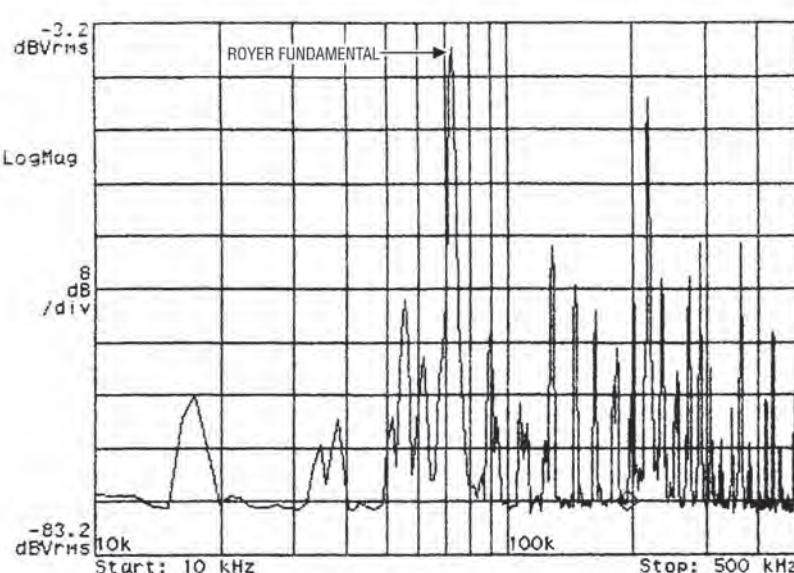

Figure C1. Hewlett-Packard HP89410A Spectral Plot of Lamp Current Shows Significant Energy Out to 500kHz

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up in Figure C2's 6MHz wide plot. This data suggests that monitoring instrumentation must maintain high accuracy over wide bandwidth.

Accurate determination of RMS operating current is important for electrical and emissivity efficiency computations and to ensure long lamp life. Additionally, it is desirable to be able to perform current measurements in the presence of high common mode voltage (> 1000V$_{RMS}$). This capability allows investigation and quantification of display and wiring-induced losses, regardless of their origins in the lamp drive circuitry.

Current Probe Circuitry

Figure C3's circuitry meets the discussed requirements. It signal-conditions a commercially available "clip-on" current probe with a precision amplifier to provide 1% measurement accuracy to 10MHz. The "clip-on" probe provides convenience, even in the presence of the high common voltages noted. The current probe biases A1, operating at a gain of about 3.75. No impedance matching is required due to the probe's low output impedance termination. Additional amplifiers provide distributed gain,
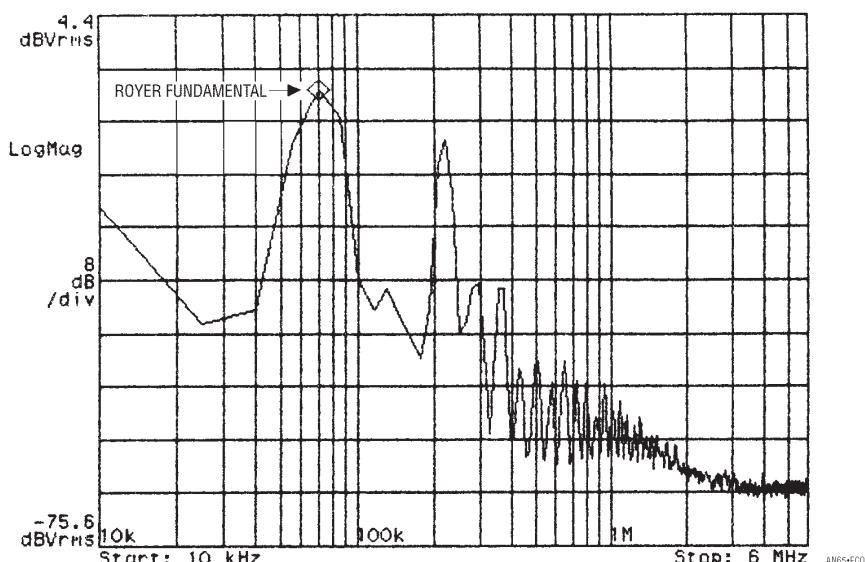

Figure C2. Extended HP89410A Spectral Plot Shows Lamp Current Has Measurable Energy Well into MHz Range. Data Indicates that Lamp Voltage and Current Instrumentation Must Have Precision, Wideband Response
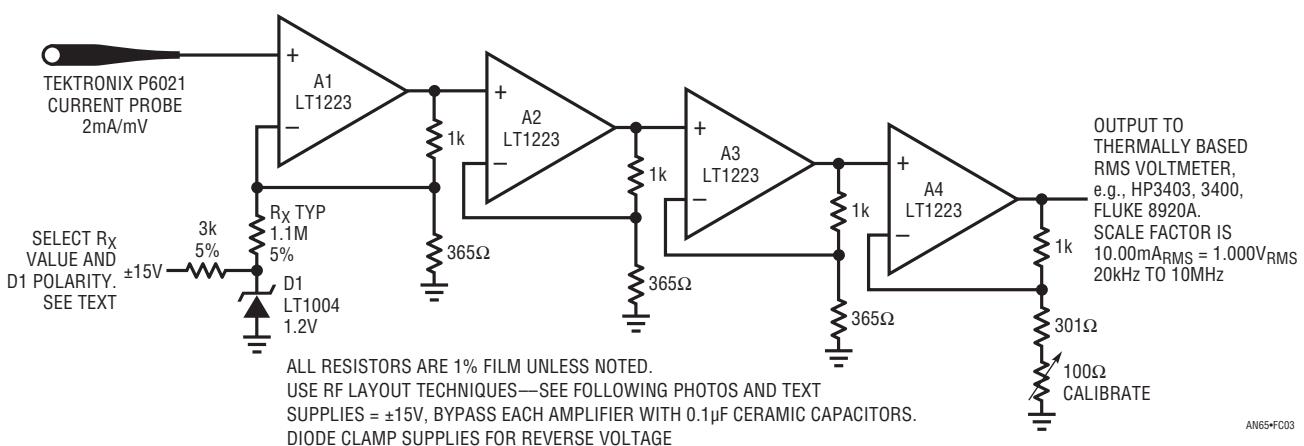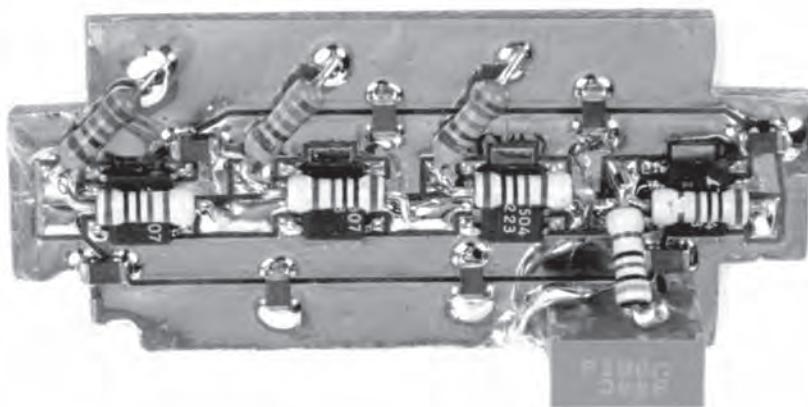

Figure C3. Precision "Clip-On" Current Probe for CCFL Measurements Maintains 1% Accuracy Over 20kHz to 10MHz Bandwidth

maintaining wide bandwidth with an overall gain of about 200. The individual amplifiers avoid any possible crosstalk-based error that could be introduced by a monolithic quad amplifier. D1 and R_X are selected for polarity and value to trim overall amplifier offset. The 100Ω trimmer sets gain, fixing the scale factor. The output drives a thermally based, wideband RMS voltmeter. In practice, the circuit is built into a 2.25" × 1" × 1" enclosure which is *directly* connected via BNC hardware to the voltmeter. No cable is used. Figure C4 shows the probe/amplifier combination. Figure C5 details RF layout techniques used in the amplifier's construction. Figure C6 shows a version of the
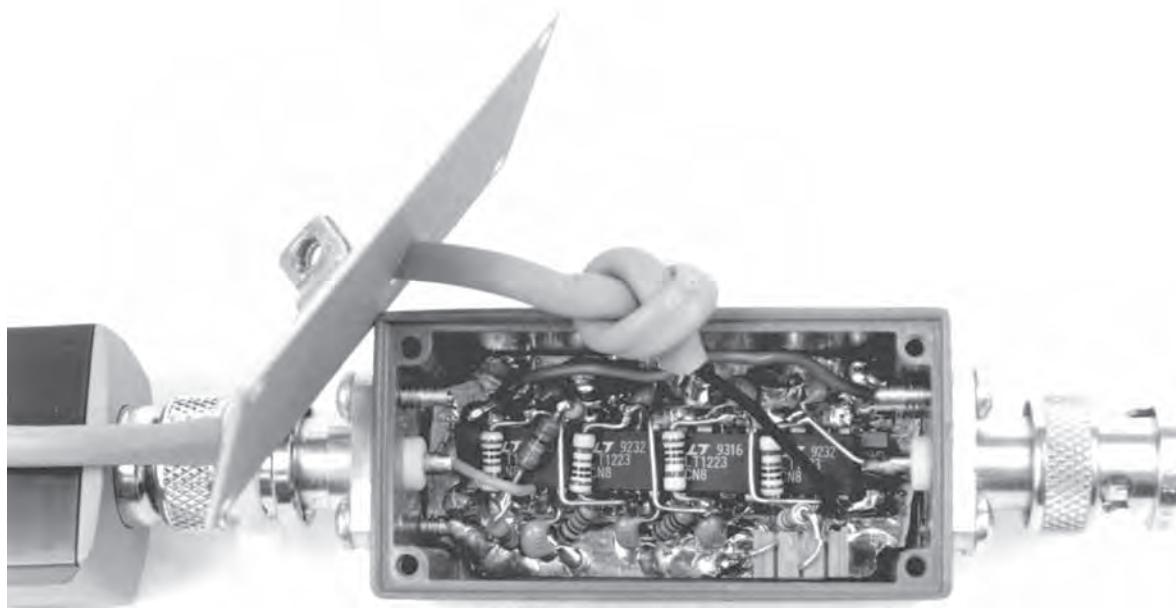

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Figure C4. Current Probe Amplifier Mated to the Current Probe Termination Box

AN65•FC05

**Figure C5. RF Layout Technique for the Current Probe Amplifier Is Required for Performance Levels Quoted in Text**
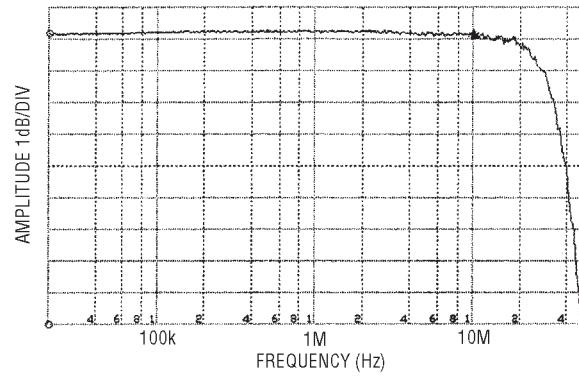
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Figure C6. A Version of the Current Probe Amplifier in Its Housing. Current Probe Terminator Is at Left

amplifier, detailing enclosure layout and construction. The result is a "clip-on" current probe with 1% accuracy over a 20kHz to 10MHz bandwidth. This tool has proven to be indispensable to any rigorously conducted backlight work. Figure C7 shows response for the probe/amplifier as measured on a Hewlett-Packard HP4195A network analyzer.

Figure C7. Amplitude vs Frequency Output of HP4195A Network Analyzer. Current Probe/Amplifier Maintains 1% (0.1dB) Error Bandwidth from 20kHz to 10MHz. Small Aberrations Between 10MHz and 20MHz Are Test Fixture Related

Current Calibrator

Figure C8's circuit, a current calibrator, permits calibration of the probe/amplifier and can be used to periodically check probe accuracy. A1 and A2 form a Wein bridge oscillator. Oscillator output is rectified by A4 and A5 and compared to a DC reference at A3. A3's output controls Q1, closing an amplitude stabilization loop. The stabilized amplitude is terminated into a 100Ω, 0.1% resistor to provide a precise 10.00mA, 60kHz current through the series current loop. Trimming is performed by altering the nominal 15k resistor for exactly 1.000V$_{RMS}$ across the 100Ω unit.

In use, this current probe has shown 0.2% baseline stability with 1% absolute accuracy over one year's time. The sole maintenance requirement for preserving accuracy is to keep the current probe jaws clean and avoid rough or abrupt handling of the probe.[2] Figure C9a shows the probe/calibrator used with an RMS voltmeter. Figure C9b shows the current probe in use, in this case determining display frame parasitic loss.

Note 2: Private communication, Tektronix, Inc.
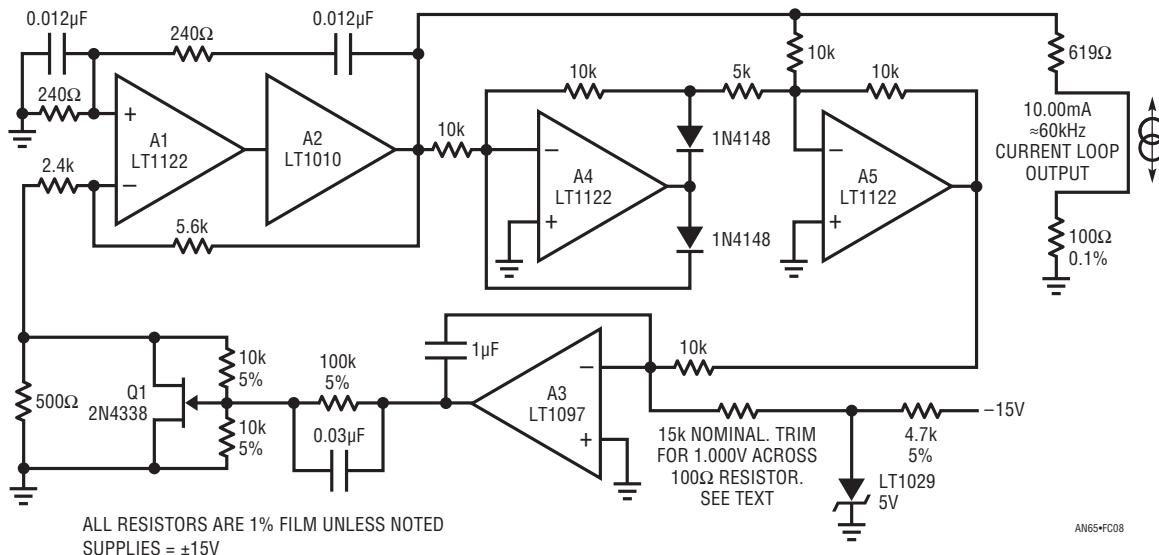
**Figure C8. Current Calibrator for Probe Trimming and Accuracy Checks. Stabilized Oscillator Forces 10.00mA Through Output Current Loop at 60kHz**

Figure C9a. Complete Current Probe Test Set Includes Probe, Amplifier, Calibrator and Thermally Based RMS Voltmeter. Accuracy Is 1% to 10MHz

Figure C9b. Current Probe Measuring Display Frame Parasitic Current. "Clip-On" Capability Allows Measurement at Any Point in Lamp Circuit

Voltage Probes for Grounded Lamp Circuits

The high voltage measurement across the lamp is quite demanding on the probe. The simplest case is measuring grounded lamp circuits. The waveform fundamental is at 20kHz to 100kHz, with harmonics into the MHz region. This activity occurs at peak voltages in the kilovolt range. The probe must have a high fidelity response under these conditions. Additionally, the probe should have low input capacitance to avoid loading effects which would corrupt the measurement. The design and construction of such a probe requires significant attention. Figure C10 lists some recommended probes along with their characteristics. As stated in the text, almost all standard oscilloscope probes *will fail* [3] if used for this measurement. Attempting to circumvent the probe requirement by resistively dividing the lamp voltage also creates problems. Large value resistors often have significant voltage coefficients and their shunt capacitance is high and uncertain. As such, simple voltage dividing is not recommended. Similarly, common high voltage probes intended for DC measurement will have large errors because of AC effects. The P6013A and P6015 are the favored probes; their 100MΩ input and small capacitance introduces low loading error.

The penalty for their 1000X attenuation is reduced output, but the recommended voltmeters (discussion to follow) can accommodate this.

All of the recommended probes are designed to work into an oscilloscope input. Such inputs are almost always 1MΩ paralleled by (typically) 10pF to 22pF. The recommended voltmeters, which will be discussed, have significantly different input characteristics. Figure C11's table shows higher input resistances and a range of capacitances. Because of this the probe must be compensated for the voltmeter's input characteristics. Normally, the optimum compensation point is easily determined and adjusted by observing probe output on an oscilloscope. A known amplitude square wave is fed in (usually from the oscilloscope calibrator) and the probe adjusted for correct response. Using the probe with the voltmeter presents an unknown impedance mismatch and raises the problem of determining when compensation is correct.

The impedance mismatch occurs at low and high frequency. The low frequency term is corrected by placing an appropriate value resistor in shunt with the probe's output. For

Note 3: That's twice we've warned you nicely.

TEKTRONIX PROBE TYPE	ATTENUATION FACTOR	ACCURACY	INPUT RESISTANCE	INPUT CAPACITANCE	RISE TIME	BAND-WIDTH	MAXIMUM VOLTAGE	DERATED ABOVE	DERATED TO AT FREQUENCY	COMPENSATION RANGE	ASSUMED TERMINATION RESISTANCE
P6007	100X	3%	10M	2.2pF	14ns	25MHz	1.5kV	200kHz	700V_{RMS} at 10MHz	15pF to 55pF	1M
P6009	100X	3%	10M	2.5pF	2.9ns	120MHz	1.5kV	200kHz	450V_{RMS} at 40MHz	15pF to 47pF	1M
P6013A	1000X	Adjustable	100M	3pF	7ns	50MHz	12kV	100kHz	800V_{RMS} at 20MHz	12pF to 60pF	1M
P6015	1000X	Adjustable	100M	3pF	4.7ns	75MHz	20kV	100kHz	2000V_{RMS} at 20MHz	12pF to 47pF	1M

Figure C10. Characteristics of Some Wideband High Voltage Probes. Output Impedances Are Designed for Oscilloscope Inputs

MANUFACTURER AND MODEL	FULL-SCALE RANGES	ACCURACY AT 1MHz	ACCURACY AT 100kHz	INPUT RESISTANCE AND CAPACITANCE	MAXIMUM BANDWIDTH	CREST FACTOR
Hewlett-Packard 3400 Meter Display	1mV to 300V, 12 Ranges	1%	1%	0.001V to 0.3V Range = 10M and < 50pF, 1V to 300V Range = 10M and < 20pF	10MHz	10:1 At Full Scale, 100:1 At 0.1 Scale
Hewlett-Packard 3403C Digital Display	10mV to 1000V, 6 Ranges	0.5%	0.2%	10mV and 100mV Range = 20M and 20pF ±10%, 1V to 1000V Range = 10M and 24pF ±10%	100MHz	10:1 At Full Scale, 100:1 At 0.1 Scale
Fluke 8920A Digital Display	2mV to 700V, 7 Ranges	0.7%	0.5%	10M and < 30pF	20MHz	7:1 At Full Scale, 70:1 At 0.1 Scale

Figure C11. Pertinent Characteristics of Some Thermally Based RMS Voltmeters. Input Impedances Necessitate Matching Network and Compensation for High Voltage Probes

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a 10MΩ voltmeter input a 1.1M resistor is suitable. This resistor should be built into the smallest possible BNC equipped enclosure to maintain a coaxial environment. No cable connections should be employed; the enclosure should be placed *directly* between the probe output and the voltmeter input to minimize stray capacitance. This arrangement compensates the low frequency impedance mismatch. Figure C12 shows the impedance matching box attached to the high voltage probe.

Correcting the high frequency mismatch term is more involved. The wide range of voltmeter input capacitances combined with the added shunt resistor's effects presents problems. How is the experimenter to know where to set the high frequency probe compensation adjustment? One solution is to feed a known value RMS signal to the probe/voltmeter combination and adjust compensation for a proper reading. Figure C13 shows a way to generate a known RMS voltage. This scheme is simply a standard backlight circuit reconfigured for a constant voltage output. The op amp permits low RC loading of the 5.6kΩ feedback termination without introducing bias current error. The 5.6kΩ value may be series or parallel trimmed for a 300V output. Stray parasitic capacitance in the feedback network affects output voltage. Because of this, all feedback associated nodes and components should be rigidly fixed and the entire circuit built into a small metal box. This prevents any significant change in the parasitic terms. The result is a known $300V_{RMS}$ output.
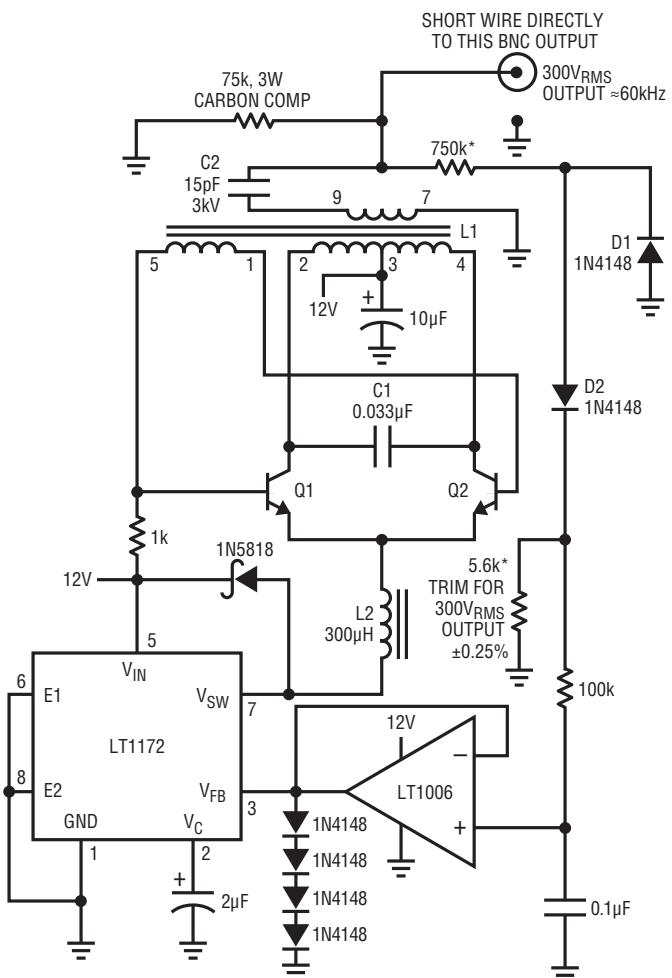
Now, the probe's compensation is adjusted for a 300V voltmeter indication using the shortest possible connection (e.g., BNC-to-probe adapter) to the calibrator box. This procedure, combined with the added resistor, completes the probe-to-voltmeter impedance match. If the probe compensation is altered (e.g., for proper response on an oscilloscope) the voltmeter's reading will be erroneous.[4] It is good practice to verify the calibrator box output before and after every set of efficiency measurements. This is done by *directly* connecting, via BNC adapters, the calibrator box to the RMS voltmeter on the 1000V range.

Note 4: The translation of this statement is to hide the probe when you are not using it. If anyone wants to borrow it, look straight at them, shrug your shoulders and say you don't know where it is. This is decidedly dishonest, but eminently practical. Those finding this morally questionable may wish to re-examine their attitude after producing a day's worth of worthless data with a probe that was unknowingly readjusted.

Figure C12. The Impedance Matching Box (Extreme Left) Mated to the High Voltage Probe. Note Direct Connection. No Cable Is Used

C1 = MUST BE A LOW LOSS CAPACITOR.
    METALIZED POLYCARB
    WIMA FKP2, MKP-20 (GERMAN) OR PANASONIC ECH-U RECOMMENDED
L1 = SUMIDA 6345-020 OR COILTRONICS CTX110092-1
    PIN NUMBERS SHOWN FOR COILTRONICS UNIT
L2 = COILTRONICS CTX300-4
Q1, Q2 = ZETEX ZTX849 OR ZDT1048
   * = 1% FILM RESISTOR (TEN 75k RESISTORS IN SERIES)

DO NOT SUBSTITUTE COMPONENTS

COILTRONICS (407) 241-7876, SUMIDA (708) 956-0666

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Figure C13. High Voltage RMS Calibrator Is Voltage Output Version of CCFL Circuit

Voltage Probes for Floating Lamp Circuits

Measuring voltage of floating lamp circuits requires a nearly heroic effort. Floating lamp measurement involves all the difficulties of the grounded case but also needs a fully differential input. This is so because the lamp is freely floating from ground. The two probes must not only be properly compensated but matched and calibrated within 1%. Additionally, a fully floating source is required to check calibration instead of Figure C13's simple single-ended approach.

Figure C14's differential amplifier converts the differential output of the high voltage probes to a single-ended signal for driving an RMS voltmeter. It introduces less than 1% error in 10MHz bandwidth if probe compensation and calibration are correct (discussion to follow). Both probe inputs feed source followers (Q1-Q4) via RC networks that provide proper probe termination. Q2 and Q4 bias differential amplifier A2, running at a gain of ≈2. FET DC and low frequency differential drift is controlled by A1. A1 measures a band limited version of A2's inputs and biases Q4's gate termination resistor. This forces Q4 and Q2 to equal source voltages. This control loop eliminates

DC and low frequency error due to FET mismatches.[5] Q1 and Q3 also follow the probe output and feed a small, frequency-dependent, summed signal to A2's auxiliary input. This term is used to correct high frequency common mode rejection limitations of A2's main inputs. A2's output drives the RMS voltmeter via a 20:1 divider. The divider combines with A2's gain-bandwidth characteristics to give 1% accuracy out to 10MHz at the voltmeter input.

To calibrate the amplifier, tie both inputs together and select R_X (shown at Q4) so A1's output is near 0V. It may be necessary to place R_X at Q2 to make this trim. Next, drive the shorted inputs with a 1V, 10MHz sine wave. Adjust the "10MHz CMRR trim" for a minimum RMS voltmeter reading, which should be below 1mV. Finally, lift the "+" input from ground, apply $1V_{RMS}$ at 60kHz and set A2's gain trim for a 100mV voltmeter reading. As a check, grounding the "+" input and driving the "−" input with the 60kHz signal should produce an identical meter reading. Further,

Note 5: A more obvious and less complex way to control FET mismatch-induced offset would utilize a matched dual monolithic FET. Readers are invited to speculate on why this approach has unacceptable high frequency error.
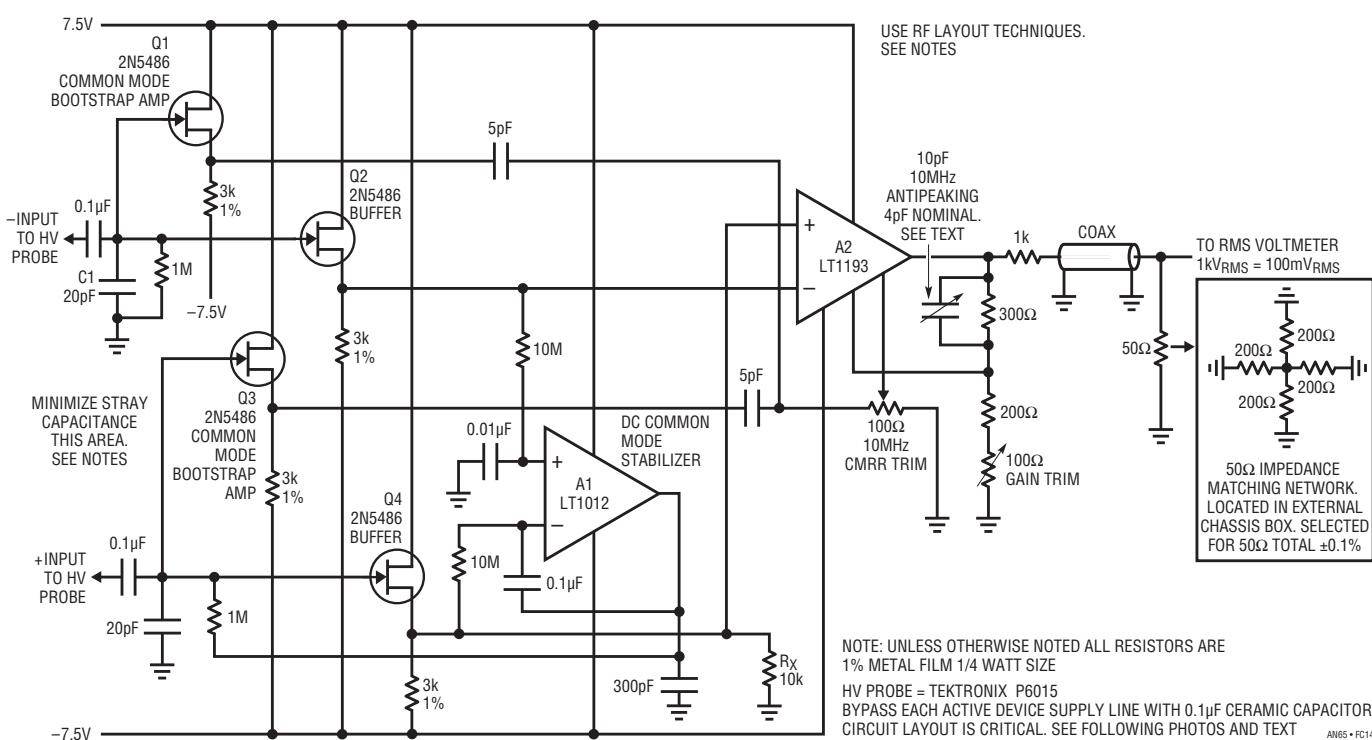

Figure C14. Precison Wideband Differential Probe Amplifier Permits Floating Lamp Voltage Measurement. Source Followers Combine with Impedance Matching Networks to Unload Probes. A2 Provides Differential-to-Single-Ended Transition

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known differential inputs at any frequency from 10kHz to 10MHz should produce corresponding calibrated and stable RMS voltmeter readings within 1%. Errors outside this figure at the highest frequencies are correctable by adjusting the "10MHz antipeaking" trim. This completes amplifier calibration.

The high voltage probes must be properly frequency-compensated to give calibrated results with the amplifier. The RC values at the amplifier inputs approximate the termination impedance the probe is designed for. Individual probes must, however, be precisely frequency-compensated to achieve required accuracy. This is quite a demanding exercise because of probe characteristics.
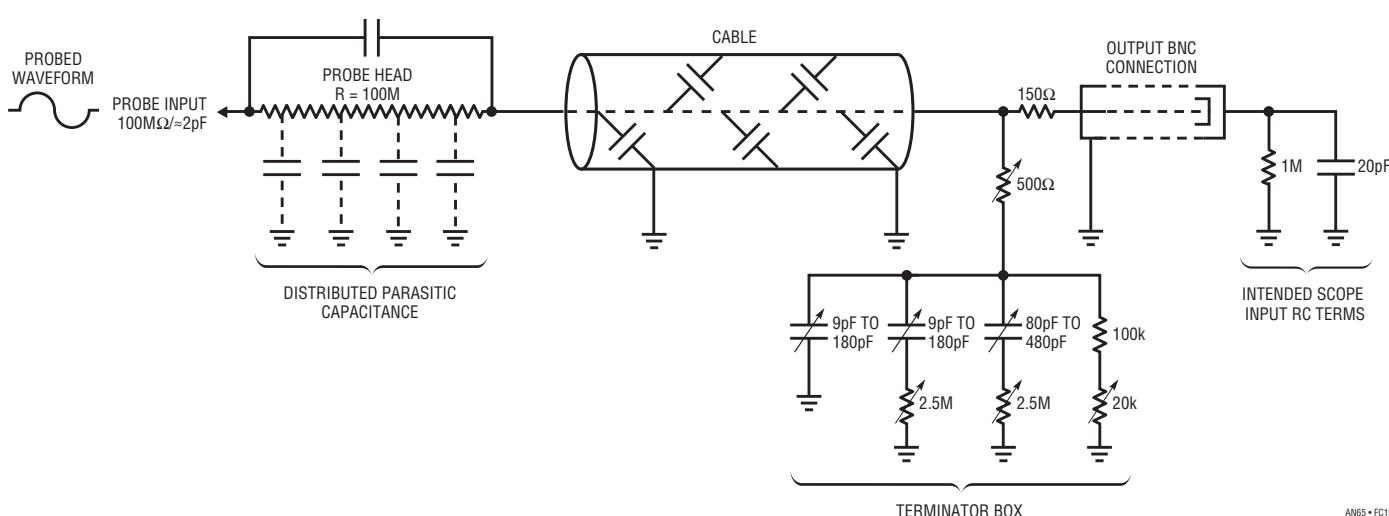
Figure C15 is an approximate schematic of the Tektronix P6015 high voltage probe. A physically large, 100M resistor occupies the probe head. Although the resistor has repeatable wideband characteristics, it suffers distributed parasitic capacitances. These distributed capacitances combine with similar cable losses, presenting a distorted version of the probed waveform to the terminator box. The terminator box impedance-frequency characteristic, when properly adjusted, corrects the distorted information, presenting the proper waveform at the output. The probe's 1000X attenuation factor, combined with its high impedance, provides a safe, minimally invasive measure of the input waveform.

The large number of parasitic terms associated with the probe head and cable result in a complex, multitime-constant response characteristic. Faithful wideband response requires the terminator box components to separately compensate each of these time constants. As such, no less than seven user adjustments are required to compensate the probe to any individual instrument input. These trims are interactive, requiring a repetitive sequence before the probe is fully compensated. The probe manual describes the trimming sequence, using the intended oscilloscope display as the output. In the present case the ultimate output is an RMS voltmeter via the differential amplifier just described. This complicates determining the probe's proper compensation point but can be accommodated.

To compensate the probes, connect them *directly* to the calibrated differential amplifier (see Figures C16, C17, C18) and ground the probe associated with the "–" input. Drive the "+" input probe with a 100V, 100kHz square wave that has a clean 10ns edge with minimal aberrations following the transition.[6] The absolute amplitude of the waveform is unimportant. Monitor this waveform[7] on an oscilloscope. Additionally, monitor A2's output in the differential

Note 6: Suitable instruments include the Hewlett-Packard 214A and the Tektronix type 106 pulse generators.
**Note 7:** Use a properly compensated probe, please!

Figure C15. Approximate Schematic of Tektronix P6015 High Voltage Probe. Distributed Parasitic Capacitances Necessitate Numerous Interactive Trims, Complicating Probe Matching to Voltmeter

Figure C16. Complete Differential Probe and Calibrator. BNC Outputs Provide Precision, Floating 500V$_{RMS}$ Calibration Source to Check Probe/Amplifier Section

amplifier (see Figure C14) with the oscilloscope.[8] Perform the compensation procedure described in the Tektronix P6015 manual until both waveforms displayed on the oscilloscope have identical shapes. When this state is reached, repeat this procedure with the "–" input probe driven and the "+" input probe grounded. This sequence brings the probe's interactive adjustments reasonably close to the optimum points.

To complete the calibration, connect the 50Ω precision termination (see Figures C14 and C16) and the RMS voltmeter to the differential amplifier's output (see Figure C16). Ground the "–" input probe and drive the "+" probe with a known amplitude high voltage waveform of about 60kHz.[9]

Perform *very slight* readjustments of this probe's compensation trims to get the voltmeter's reading to agree with the calibrated input (account for scaling differences—e.g., ignore the voltmeter's range and decimal point indications). The trim(s) having the greatest influence should be utilized for this adjustment—only a slight adjustment should be required. Upon completing this step repeat the procedure using the 100V, 100kHz square wave, verifying input/output waveform edge fidelity. If waveform fidelity has been lost retrim and try again. Several iterations may be necessary until both conditions are met.

Repeat the above procedure for the "–" probe adjustment with the "+" probe grounded.

Note 8: See Footnote 7.
**Note 9:** Figure C13's calibrator is appropriate.
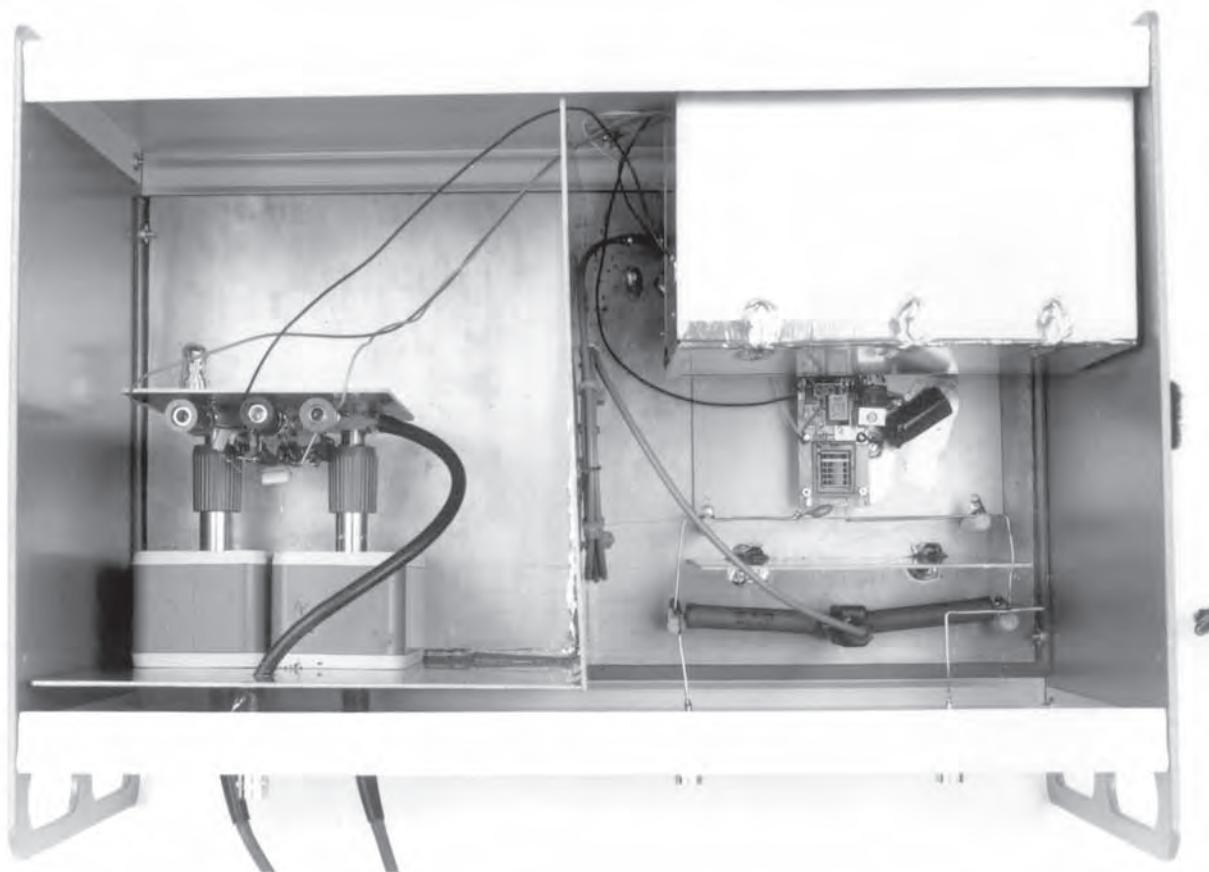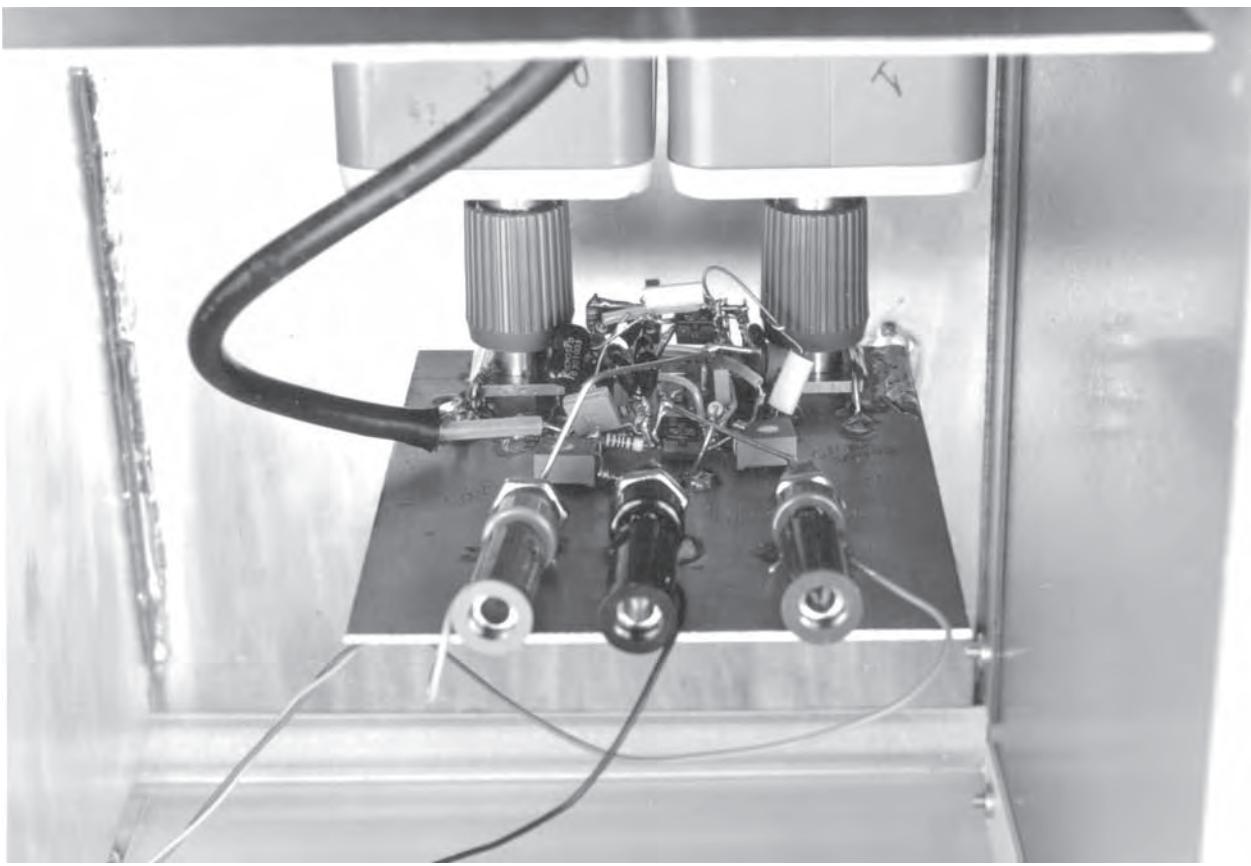
Figure C17. Top View of Differential Probe/Calibrator. Probes Are Directly Mated to Differential Amplifier (Left). Calibrator Is to Right. Current Transformer Is Located Between Load Resistors

Short both probes together and drive them with the 100V, 100kHz square wave. The RMS voltmeter should read (ideally) zero. Typically, it should indicate well below 1% of input. The differential amplifier's "10MHz CMRR trim" (Figure C14) can be adjusted to minimize the voltmeter reading.

Next, with the probes still shorted, apply a swept 20kHz to 10MHz sine wave with the highest amplitude available. Monitor A2's output with an RMS voltmeter, ensuring that it never rises above 1% of input amplitude. Finally,[10] apply the highest available known amplitude, swept 20kHz to

10MHz signal to each probe with the other probe grounded. Verify that the RMS voltmeter indicates correct and flat gain over the entire swept frequency range for each case. If any condition described in this paragraph is not met, the entire calibration sequence must be repeated. This completes the calibration.

Note 10: "Finally" is more than an appropriate descriptive. Achieving a wideband, matched probe response involving 14 interactive adjustments takes time, patience and utter determination. Allow at least six hours for the entire session. You'll need it.

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Figure C18. Detail of Probe/Amplifier Connection Shows Direct, Low Loss BNC Coupling

Differential Probe Calibrator

A calibrator with a fully floating, differential output allows periodic operational checking of the differential probe's accuracy. This calibrator is built into the same enclosure as the differential probe (Figure C16). Figure C19 is a schematic of the calibrator.

The circuit is a highly modified form of the basic backlight power supply. Here, T1's output drives two precision resistors which are well-specified for high frequency, high voltage operation. The resistor's current is monitored by L2, a wideband current transformer. L2's placement between the resistors combines with T1's floating drive to minimize the effects of L2's parasitic capacitance. Although L2 has parasitic capacitance, it is bootstrapped to essentially 0V, negating its effect.

L2's secondary output is amplified by A1 and A2, with A3 and A4 serving as a precision rectifier. A4's output is smoothed by the 10kΩ/0.1µF filter, closing a loop at the LT1172's Feedback pin. Similar to previously described CCFL circuits, the LT1172 controls Royer drive, setting T1's output.

To calibrate this circuit, ground the LT1172's V_C pin, open T1's secondary and select the LT1004's polarity and associated resistor value for 0V at A4's output. Next, put a 5.00mA, 60kHz current through L2.[11] Measure A4's smoothed output (the LT1172's Feedback pin) and adjust the "output trim" for 1.23V. Next, reconnect T2's secondary, remove the current calibrator connection and

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Note 11: Figure C8's output, rescaled for 5.00mA, is a source of calibrated current.

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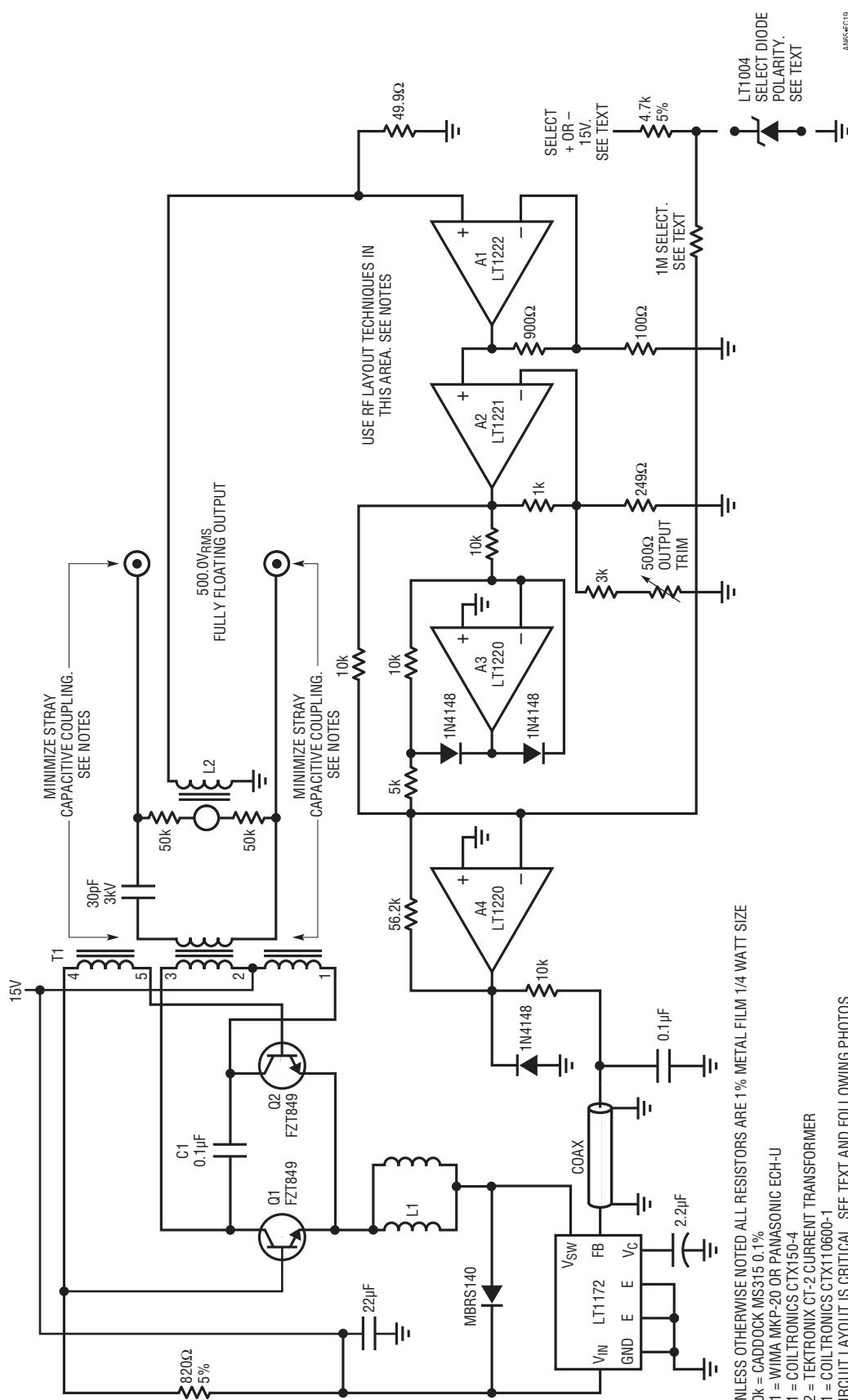
Figure C19. The Floating Output Calibrator. Current Transformer Permits Floating Output While Maintaining Tight Loop Control. Amplifiers Provide Gain to Inverter Circuit's Feedback Node

UNLESS OTHERWISE NOTED ALL RESISTORS ARE 1% METAL FILM 1/4 WATT SIZE
50k = CADDOCK MS315 0.1%
C1 = WIMA MKP-20 OR PANASONIC ECH-U
L1 = COILTRONICS CTX150-4
L2 = TEKTRONIX CT-2 CURRENT TRANSFORMER
T1 = COILTRONICS CTX110600-1
CIRCUIT LAYOUT IS CRITICAL. SEE TEXT AND FOLLOWING PHOTOS
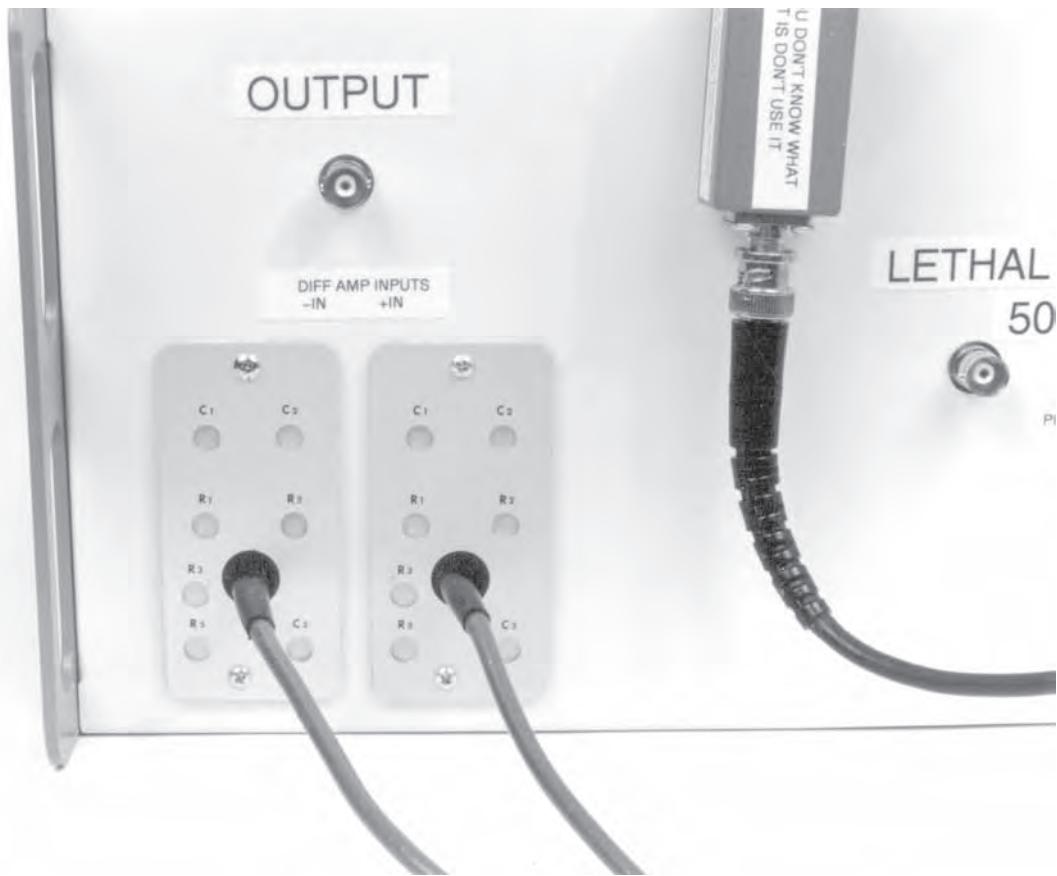
unground the LT1172 V_C pin. The result is 500V_{RMS} at the calibrator's differential output. This may be checked with the differential probe. Reversing the probe connections should have no effect, with readings well within 1%.[12]

The differential probe and floating output calibrator require almost fanatical attention to layout to achieve the performance levels noted. The wideband amplifier sections utilize RF layout techniques which are reasonably well-documented.[13] Practical construction considerations for parasitic capacitance related issues are photographically detailed in Figures C17 through C23.
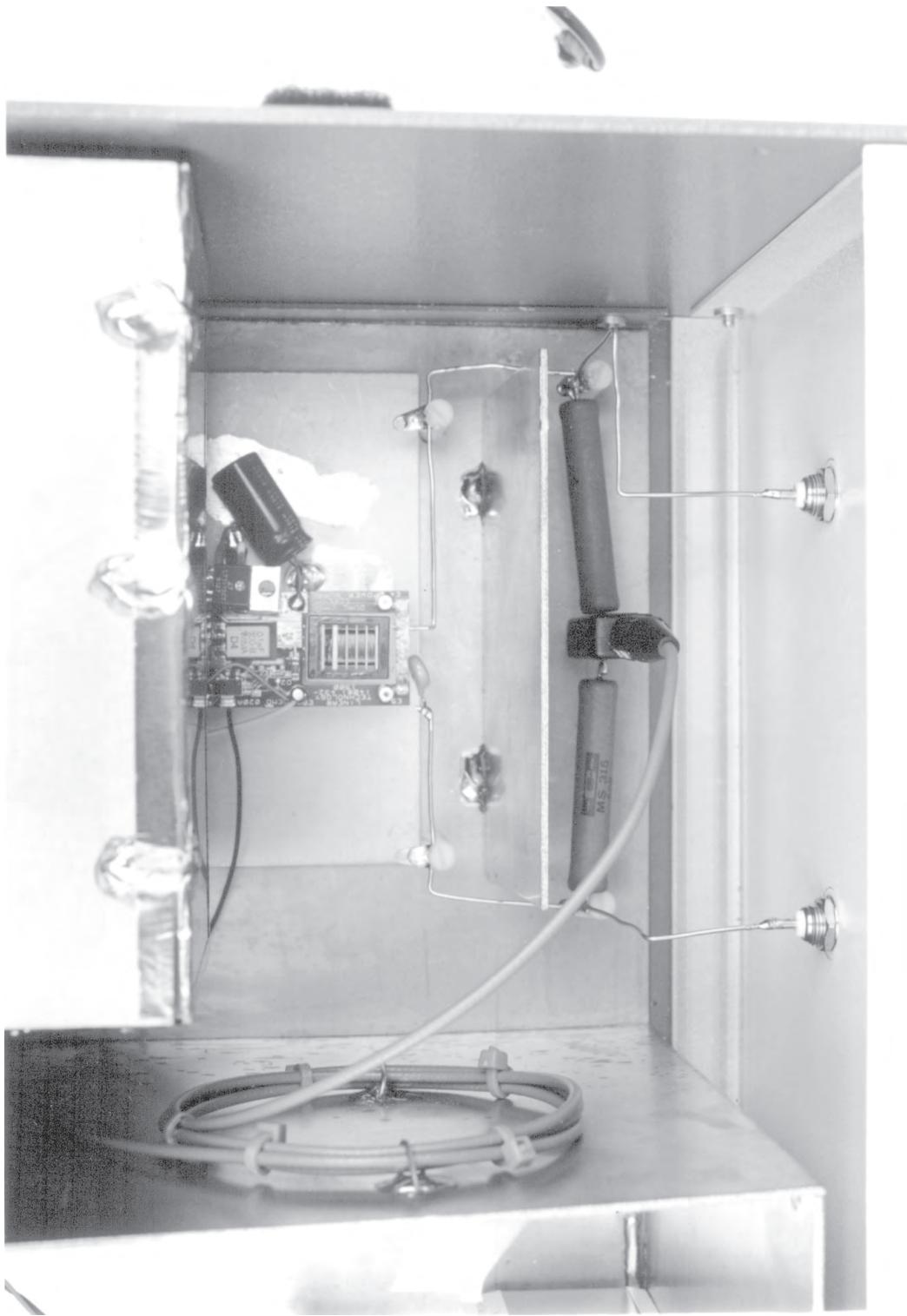
Note 12: Those who construct and trim the differential probe and calibrator will experience the unmitigated joy that breaks loose when they agree within 1%.
**Note 13:** See Reference 26.
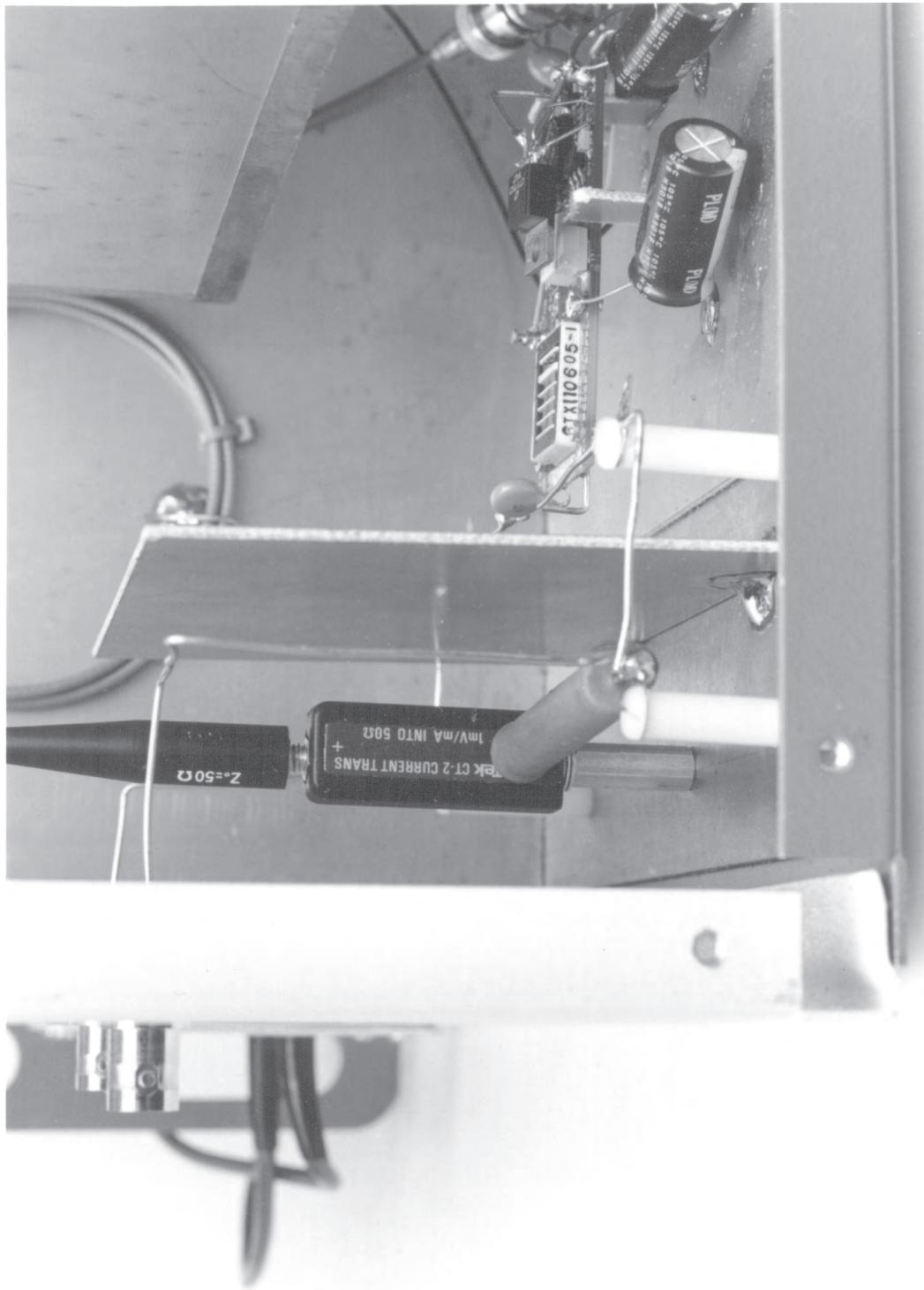
Figure C20. Differential Probes Are Mechanically Secured to Chassis, Discouraging Unauthorized Removal. All Compensation Access Holes Are Sealed, Preventing Unwanted Adjustment

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Figure C21. Calibrator Section Detail. Inverter in Center with Load Resistors and Current Transformer in Foreground. Note Shield Between Inverter and Load Resistors, and Low Capacitance Layout

Figure C22. Calibrator Output Detail. Current Transformer Is Bootstrapped to Load Resistor's 0V Midpoint. Shield (Center) Prevents Interaction Between Transformer Field and Load Resistors or Current Transformer. Bus Wire and Nylon Stand-Offs Minimize Stray Capacitance
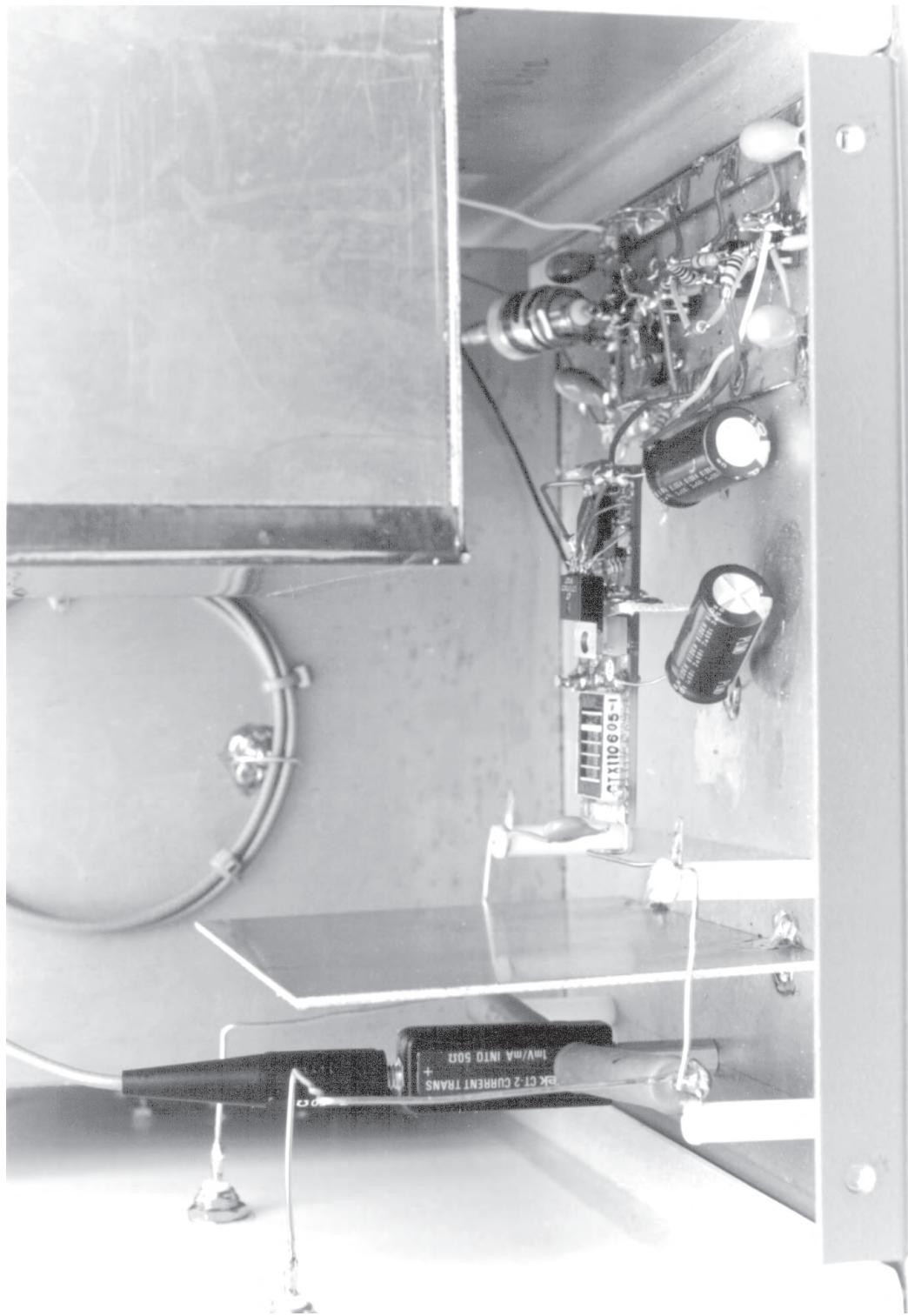
Figure C23. Calibrator Section Showing Bus Wire/Nylon Post Construction Used to Minimize Stray Capacitance (Left). Inverter in Center; Control Electronics Located at Lower Right. Power Supply Is Enclosed in Shielded Box at Upper Right

RMS Voltmeters

The efficiency measurements require an RMS responding voltmeter. This instrument must respond accurately at high frequency to irregular and harmonically loaded waveforms. These considerations eliminate almost all AC voltmeters, including DVMs with AC ranges.

There are a number of ways to measure RMS AC voltage. Three of the most common include *average, logarithmic* and *thermally* responding. Averaging instruments are calibrated to respond to the average value of the input waveform, which is almost always assumed to be a sine wave. Deviation from an ideal sine wave input produces errors. Logarithmically based voltmeters attempt to overcome this limitation by continuously computing the input's true RMS value. Although these instruments are "real time" analog computers, their 1% error bandwidth is well below 300kHz and crest factor capability is limited. Almost all general purpose DVMs use such a logarithmically based approach and, as such, are not suitable for CCFL efficiency measurements. Thermally based RMS voltmeters are direct acting thermoelectronic analog computers. They respond to the input's RMS heating value. This technique is explicit, relying on the very definition of RMS (e.g., the heating power of the waveform). By turning the input into heat, thermally based instruments achieve vastly higher bandwidth than other techniques.[14] Additionally, they are insensitive to waveform shape and easily accommodate large crest factors. These characteristics are necessary for the CCFL efficiency measurements.

Figure C24 shows a conceptual thermal RMS/DC converter. The input waveform warms a heater, resulting in increased output from its associated temperature sensor. A DC amplifier forces a second, identical, heater/sensor pair to the same thermal conditions as the input driven pair. This differentially sensed, feedback enforced loop makes ambient temperature shifts a common mode term, eliminating their effect. Also, although the voltage and thermal interaction is nonlinear, the input/output RMS voltage relationship is linear with unity gain.

The ability of this arrangement to reject ambient temperature shifts depends on the heater/sensor pairs being isothermal. This is achievable by thermally insulating them with a time constant well below that of ambient shifts. If the time constants to the heater/sensor pairs are matched,
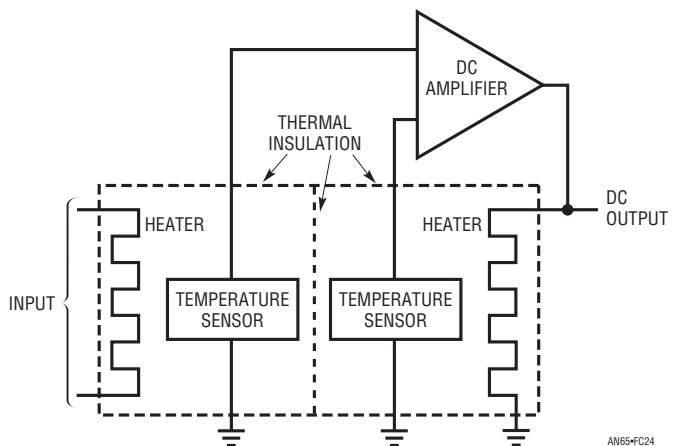

Figure C24. Conceptual Thermal RMS/DC Converter

ambient temperature terms will affect the pairs equally in phase and amplitude. The DC amplifier rejects this common mode term. Note that although the pairs are isothermal, they are insulated from each other. Any thermal interaction between the pairs reduces the system's thermally based gain terms. This would cause unfavorable signal-to-noise performance, limiting dynamic operating range.

Figure C24's output is linear because the matched thermal pair's nonlinear voltage/temperature relationships cancel each other.

The advantages of this approach have made its use popular in thermally based RMS/DC measurements.

The instruments listed in Figure C11, while considerably more expensive than other options, are typical of what is required for meaningful results. The HP3400A and the Fluke 8920A are currently available from their manufacturers. The HP3403C, an exotic and highly desirable instrument, is no longer produced but readily available on the secondary market.

Figure C25 shows an RMS voltmeter which can be constructed instead of purchased.[15] Its small size permits it to be built into bench and production test equipment. As shown, it is designed to be used with Figure C14's differential probe, although the configuration is adaptable to any CCFL-related measurement. It provides a true

Note 14: Those finding these descriptions intolerably brief are commended to References 4, 5, 6, 9, 10, 11 and 12.
Note 15: This circuit derives from Reference 27.
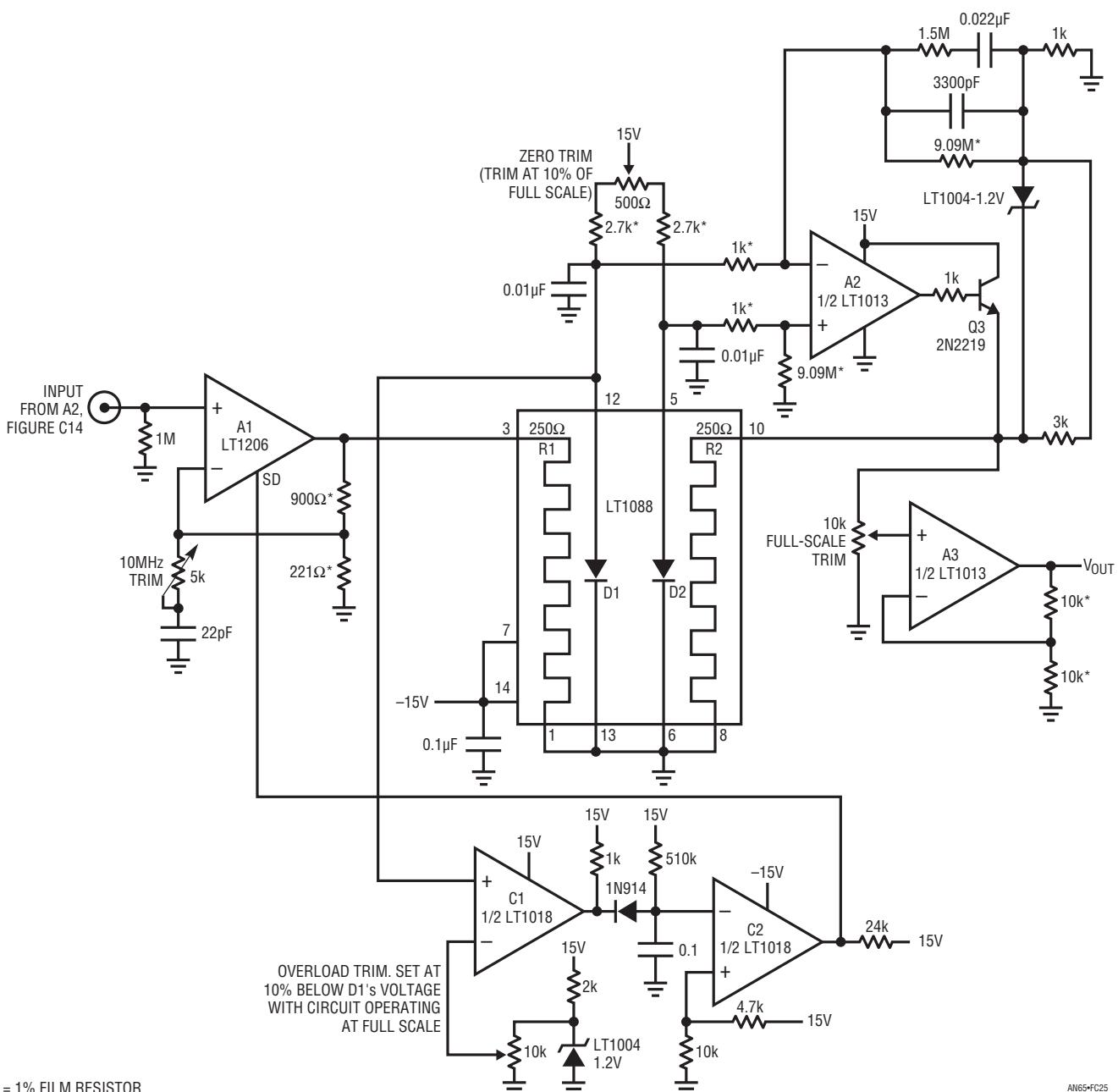
Figure C25. Wideband RMS/DC Converter for Use with Differential Probe/Amplifier. Circuit Is Also Usable with Current Probe/Amplifier with Appropriate Gain Adjustments

RMS/DC conversion from DC to 10MHz with less than 1% error, regardless of input signal waveshape. It also features high input impedance and overload protection.

The circuit consists of three blocks; a wideband amplifier, the RMS/DC converter and overload protection. The amplifier provides high input impedance and gain, and drives the RMS/DC converter's input heater. Input resistance is defined by the 1M resistor with input capacitance

about 10pF. The LT1206 provides a flat 10MHz bandwidth gain of 5. The 5kΩ/22pF network gives A1 a slight peaking characteristic at the highest frequencies, allowing 1% flatness to 10MHz. A1's output drives the RMS/DC converter.

The LT1088-based RMS/DC converter is made up of matched pairs of heaters and diodes and a control amplifier. The LT1206 drives R1, producing heat which lowers D1's

voltage. Differentially connected A2 responds by driving R2 via Q3 to heat D2, closing a loop around the amplifier. Because the diodes and heater resistors are matched, A2's DC output is related to the RMS value of the input, regardless of input frequency or waveshape. In practice, residual LT1088 mismatches necessitate a gain trim, which is implemented at A3. A3's output is the circuit output. The LT1004 and associated components provide loop compensation and good settling time over wide ranges of operating conditions (see Footnote 14).

Start-up or input overdrive can cause A1 to deliver excessive current to the LT1088 with resultant damage. C1 and C2 prevent this. Overdrive forces D1's voltage to an abnormally low potential. C1 triggers low under these conditions, pulling C2's input low. This causes C2's output to go high, putting A1 into shutdown and terminating the overload. After a time determined by the RC at C2's input, A1 will be enabled. If the overload condition still exists the loop will almost immediately shut A1 down again. This oscillatory action will continue, protecting the LT1088 until the overload condition is removed.

Performance for the circuit is quite impressive. Figure C26 plots error from DC to 11MHz. The graph shows 1% error bandwidth of 11MHz. The slight peaking out to 5MHz is due to the gain boost network at A1's negative input. The peaking is minimal compared to the total error envelope, and a small price to pay to get the 1% accuracy to 10MHz.
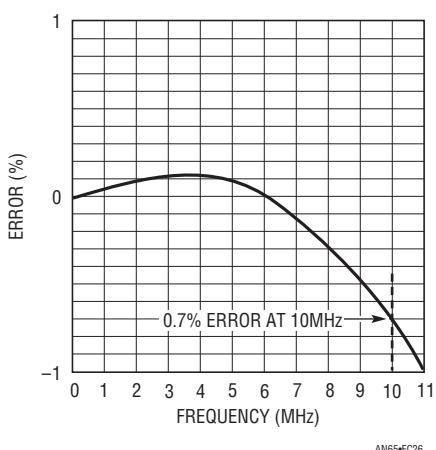

Figure C26. Error Plot for the RMS/DC Converter. Frequency-Dependent Gain Boost at A1 Preserves 1% Accuracy but Causes Slight Peaking Before Roll-Off

To trim this circuit put the 5kΩ potentiometer at its maximum resistance position and apply a 100mV, 5MHz signal. Trim the 500Ω adjustment for exactly $1V_{OUT}$. Next, apply a 5MHz, 1V input and trim the 10kΩ potentiometer for $10.00V_{OUT}$. Finally, put in 1V at 10MHz and adjust the 5kΩ trimmer for $10.00V_{OUT}$. Repeat this sequence until circuit output is within 1% accuracy for DC-10MHz inputs. Two passes should be sufficient.

Calorimetric Correlation of Electrical Efficiency Measurements

Careful measurement technique permits a high degree of confidence in the efficiency measurement's accuracy. It is, however, a good idea to check the method's integrity by measuring in a completely different domain. Figure C27 does this by calorimetric techniques. This arrangement, identical to the thermal RMS voltmeter's operation (Figure C24), determines power delivered by the CCFL circuit by measuring its load temperature rise. As in the thermal RMS voltmeter, a differential approach eliminates ambient temperature as an error term. The differential amplifier's output, assuming a high degree of matching in the two thermal enclosures, proportions to load power. The ratio of the enclosure's E • I products yields efficiency information. In a 100% efficient system the amplifier's output energy would equal the power supply's output. Practically it is always less as the CCFL circuit has losses. This term represents the desired efficiency information.

Figure C28 is similar except that the CCFL circuit board is placed within the calorimeter. This arrangement nominally yields the same information, but is a much more demanding measurement because far less heat is generated. The signal-to-noise (heat rise above ambient) ratio is unfavorable, requiring almost fanatical attention to thermal and instrumentation considerations.[16] It is significant that the *total* uncertainty between electrical and both calorimetric efficiency determinations was 3.3%. The two thermal approaches differed by about 2%. Figure C29 shows the calorimeter and its electronic instrumentation. Descriptions of this instrumentation and thermal measurements can be found in the References section following the main text.

Note 16: Calorimetric measurements are not recommended for readers who are short on time or sanity.
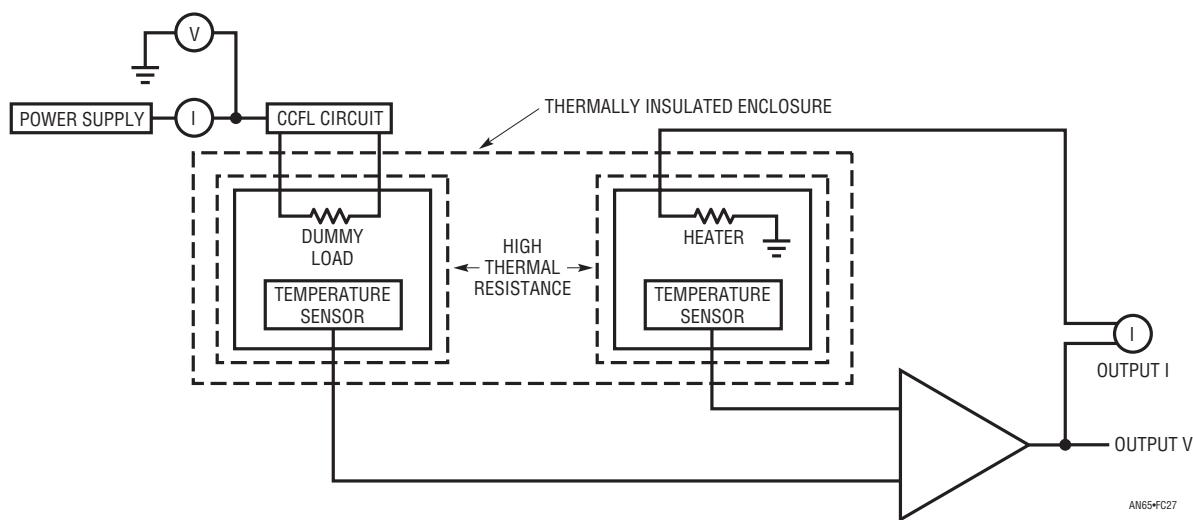
Figure C27. Efficiency Determination via Calorimetric Measurement. Ratio of Power Supply to Output Energy Gives Efficiency Information
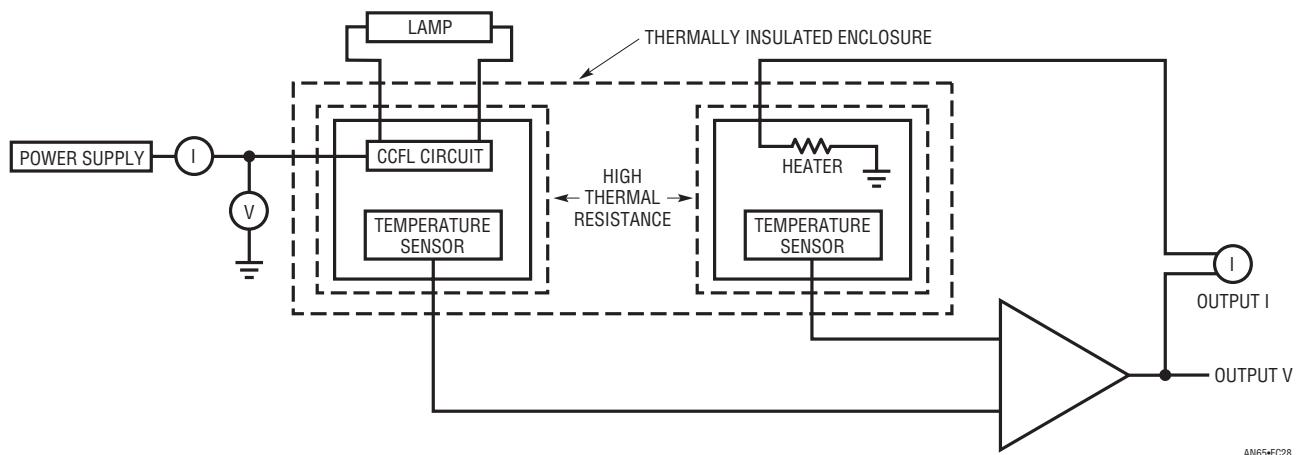

Figure C28. The Calorimeter Measures Efficiency by Determining Circuit Heating Losses
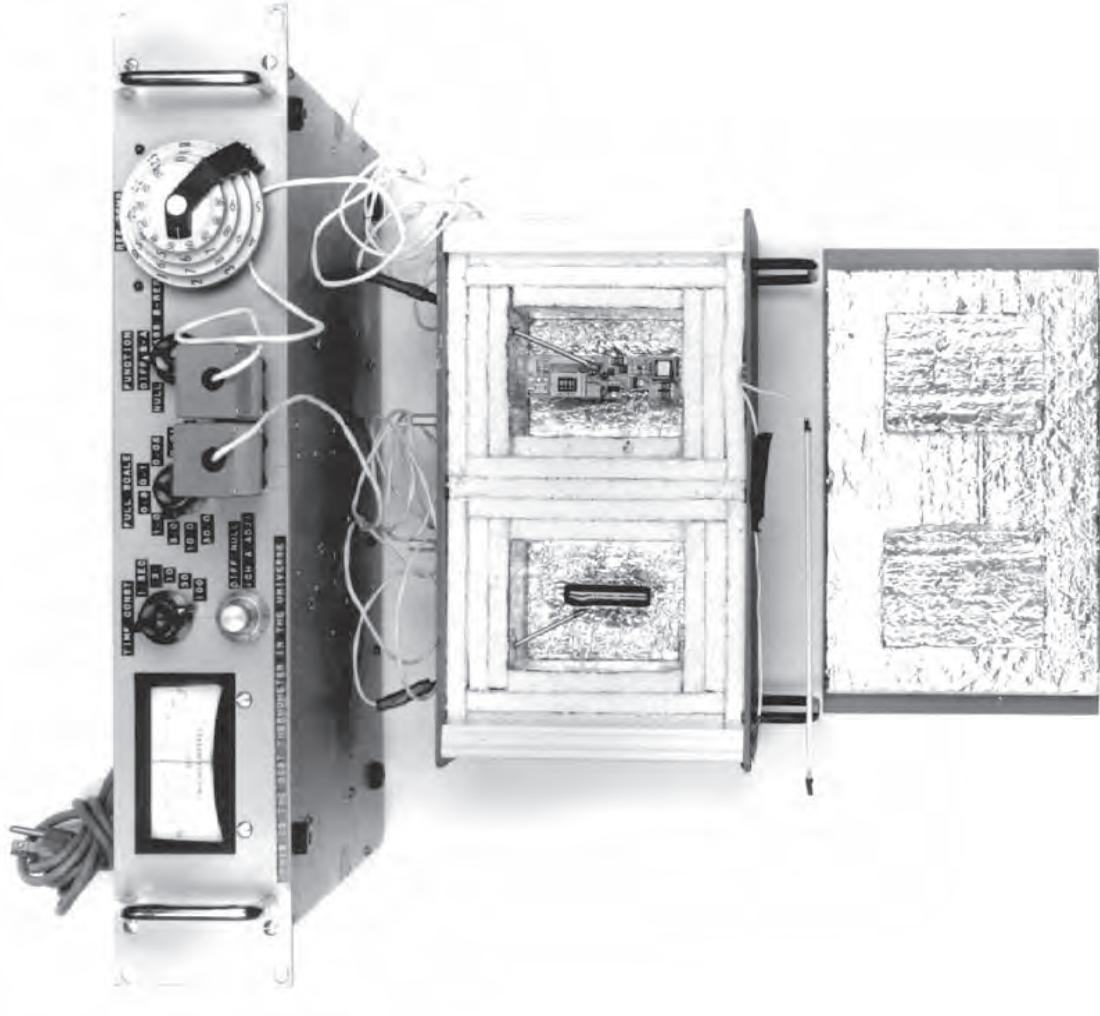
Figure C29. The Calorimeter (Center) and Its Instrumentation (Top). Calorimeter's High Degree of Thermal Symmetry Combined with Sensitive Servo Instrumentation Produces Accurate Efficiency Measurements. Lower Portion of Photo Is Calorimeter's Top Cover

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APPENDIX D

PHOTOMETRIC MEASUREMENTS

In the final analysis, the ultimate concern centers around the efficient conversion of power supply energy to light. Emitted light varies monotonically with power supply energy,[1] but certainly not linearly. In particular, lamp luminosity may be highly nonlinear, particularly at high power vs drive power. There are complex trade-offs involving the amount of emitted light vs power consumption, drive waveform shape and battery life. Evaluating these trade-offs requires some form of photometer. The relative luminosity of lamps may be evaluated by placing the lamp in a light tight tube and sampling its output with photodiodes. The photodiodes are placed along the lamp's length and their outputs electrically summed. This sampling technique is an uncalibrated measurement, providing relative data only. It is, however, quite useful in determining relative lamp emittance under various drive conditions. Additionally, because the enclosure has essentially no parasitic capacitance, lamp performance may be evaluated under "zero loss" conditions. Figure D1 shows this "glometer," with its uncalibrated output appropriately scaled in "brights." The switches allow various sampling diodes along the lamp's length to be disabled. The photodiode signal-conditioning electronics are mounted behind the switch panel with the drive electronics located to the left.

Figure D2 details the drive electronics. A1 and A2 form a stabilized output Wein bridge sine wave oscillator. A1 is the oscillator and A2 provides gain stabilization in concert with Q1. The stabilizing loop's operating point is derived from the LT1021 voltage reference. A3 and A4 constitute a voltage-controlled amplifier which feeds power stage A5. A5 drives T1, a high ratio step-up transformer. T1's output sources current to the lamp. Lamp current is rectified and its positive portion terminated into the 1k resistor. The voltage appearing across this resistor, indicative of lamp current, biases A6. Band-limited A6 compares the lamp current-derived signal against the LT1021 reference and closes a loop back to A3. This loop's operating point, and hence lamp current, is set by the "current amplitude" adjustment over a 0mA to 6mA range. A1's "Frequency Adjust" control permits a 20kHz to 130kHz frequency operating range. The switch located at A1's output permits external sources of various waveforms and frequencies to drive the amplifier.

Note 1: But not always! It is possible to build highly electrically efficient circuits that emit less light than "less efficient" designs. See previous text and Appendix L, "A Lot of Cut Off Ears and No Van Goghs—Some Not-So-Great Ideas."

AN65•FD01

Figure D1. "Glometer" Measures Relative Lamp Emissivity Under Various Drive Conditions. Test Lamp Is Inside Cylindrical Housing. Photodiodes on Housing Convert Light to Electrical Output (Center) Via Amplifiers (Not Visible in Photo). Electronics (Left) Permit Varying Drive Waveforms and Frequency
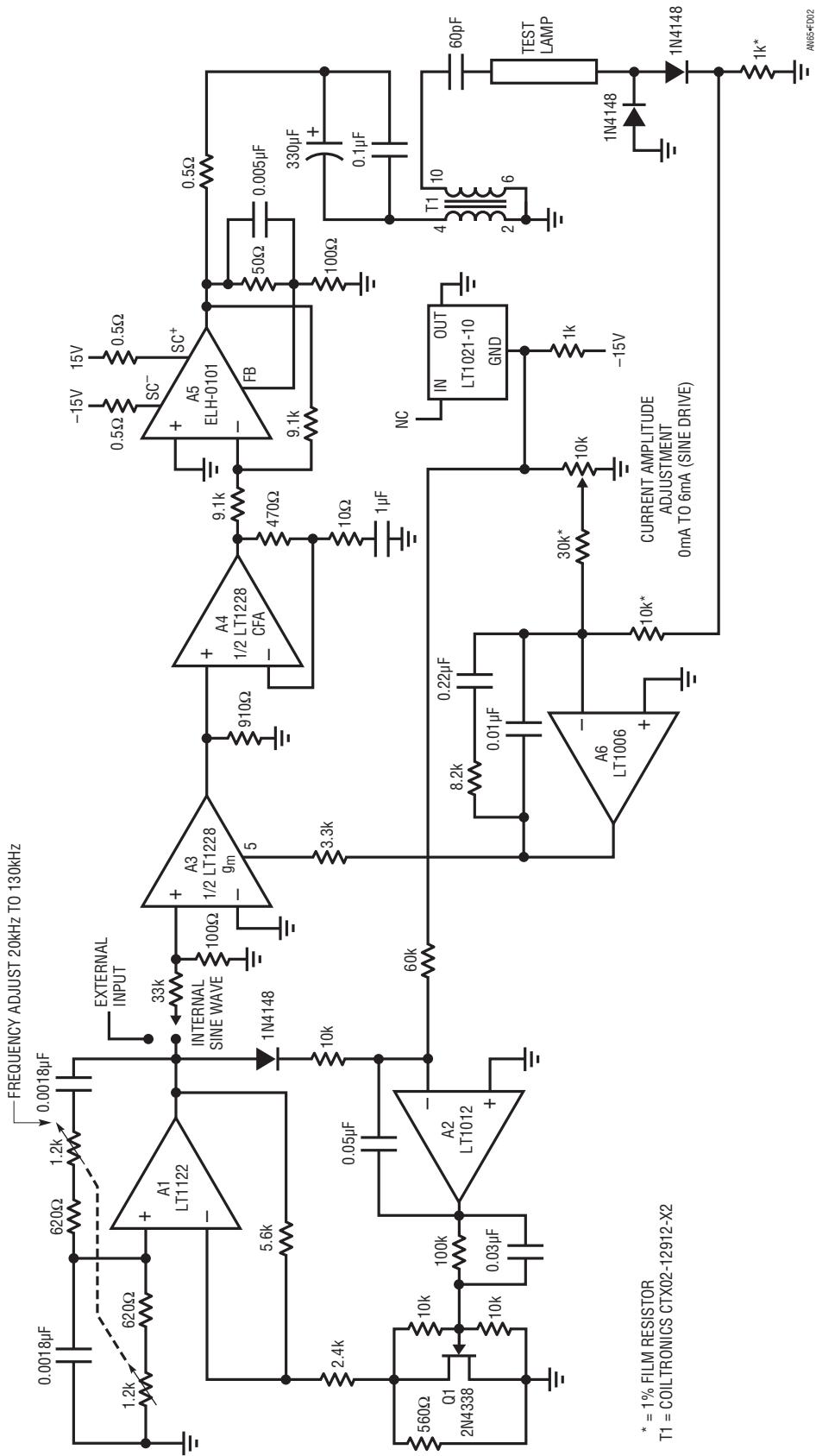
Figure D2. Glometer Drive Electronics Permits Varying Frequency and Waveshape Applied to Test Lamp. Resultant Data Shows Lamp Sensitivity to These Parameters

* = 1% FILM RESISTOR
T1 = COILTRONICS CTX02-12912-X2

AN65-FD02

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The drive scheme and wideband transformer provide extremely faithful response. Figure D3 shows waveform fidelity at 100kHz with a 5mA lamp load. Trace A is T1's primary drive and Trace B is the high voltage output. Figure D4, a horizontal and vertical expansion of D3, indicates well-controlled phase shift. Residual effects cause slight primary impedance variations (note primary drive nonlinearity at the sixth vertical division), although the output remains singularly clean.


A = 20V/DIV

B = 200V/DIV

2μs/DIV

AN65•FD03
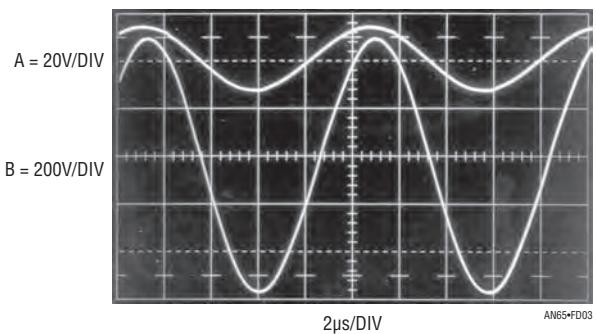
Figure D3. Wideband Transformers Input (Trace A) and Output (Trace B) Waveforms Indicate Clean Response at 100kHz

A = 5V/DIV

B = 200V/DIV
(UNCALIBRATED)

1μs/DIV

AN65•FD04
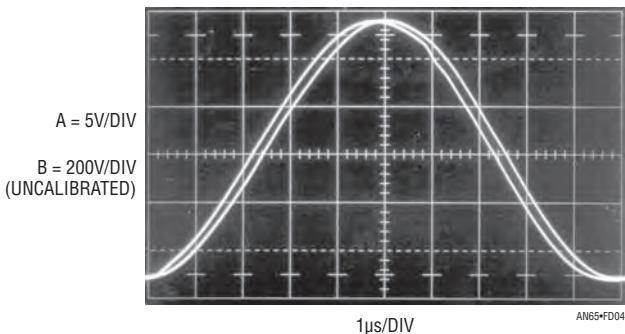
Figure D4. Magnified Versions of Figure D3's Waveforms. Output (Trace B) Is Undistorted Despite Slight Drive (Trace A) Deformity at Sixth Vertical Division. Transformer Roll-Off Dictates This Desirable Behavior

Figure D5 shows the photodiode signal conditioning. Groups of various photodiodes bias amplifiers A1 through A6. Each amplifier's output is fed via a switch to summing amplifier A7. The switches permit establishment of "dead zones" along the test lamp's length, enhancing ability to study emissivity vs location. A7's output represents the summation of all sensed lamp emission.

The glometer's ability to measure relative lamp emission under controlled settings of frequency, waveshape and drive current in a "lossless" environment is invaluable for evaluating lamp performance. Evaluating display performance and correlating results with customers requires absolute light intensity measurements.

Calibrated light measurements call for a true photometer. The Tektronix J-17/J1803 photometer is such an instrument. It has been found particularly useful in evaluating display (as opposed to simply the lamp) luminosity under various drive conditions. The calibrated output permits reliable correlation with customer results.[2] The light tight measuring head allows evaluation of emittance evenness at various display locations.

Figure D6 shows the photometer in use evaluating a display. Figure D7 is a complete display evaluation setup. It includes lamp and DC input voltage and current instrumentation, the photometer described and a computer (lower right) for calculating optical and electrical efficiency.

Note 2: It is unlikely customers would be enthusiastic about correlating the "brights" units produced by the aforementioned glometer.
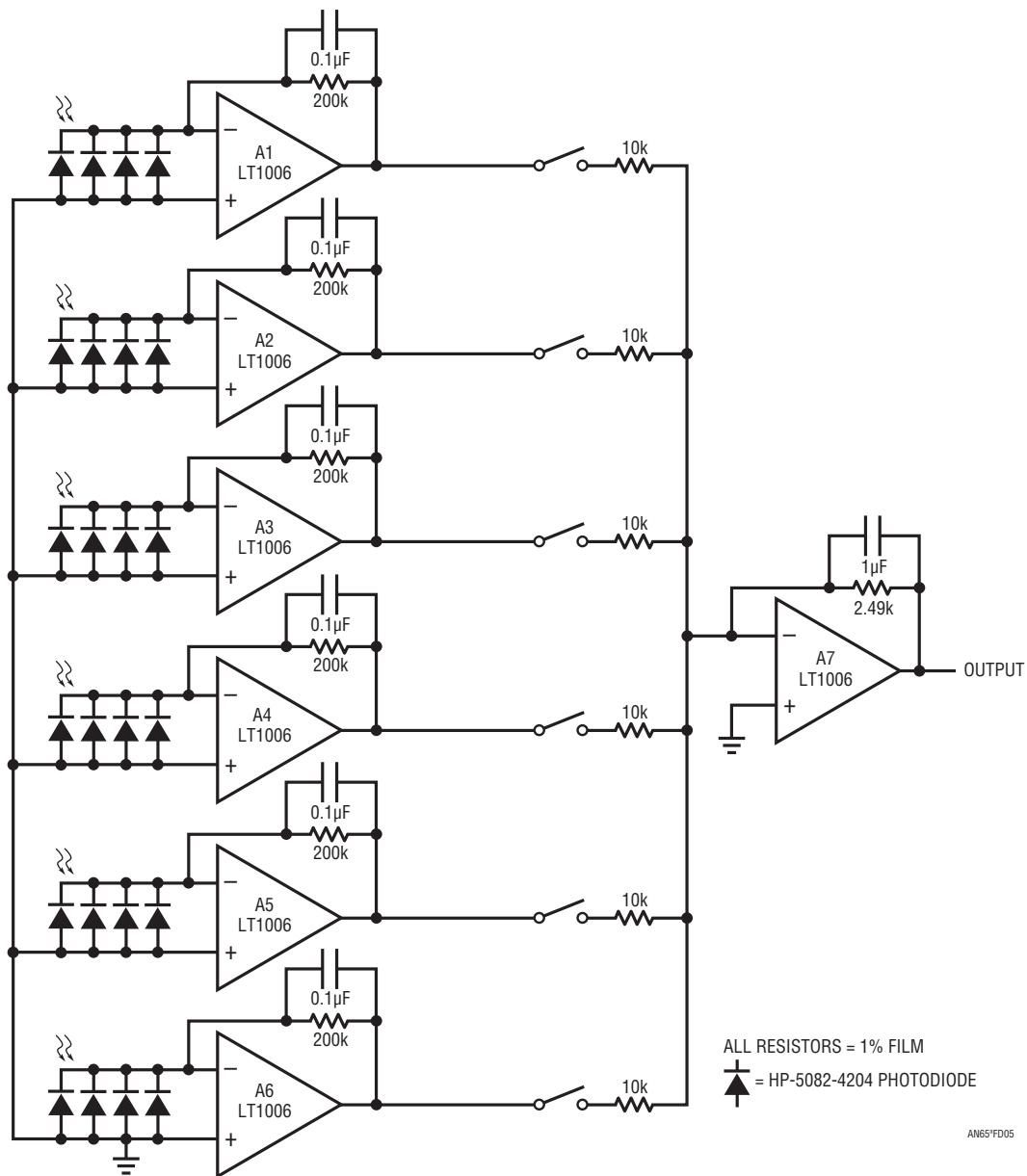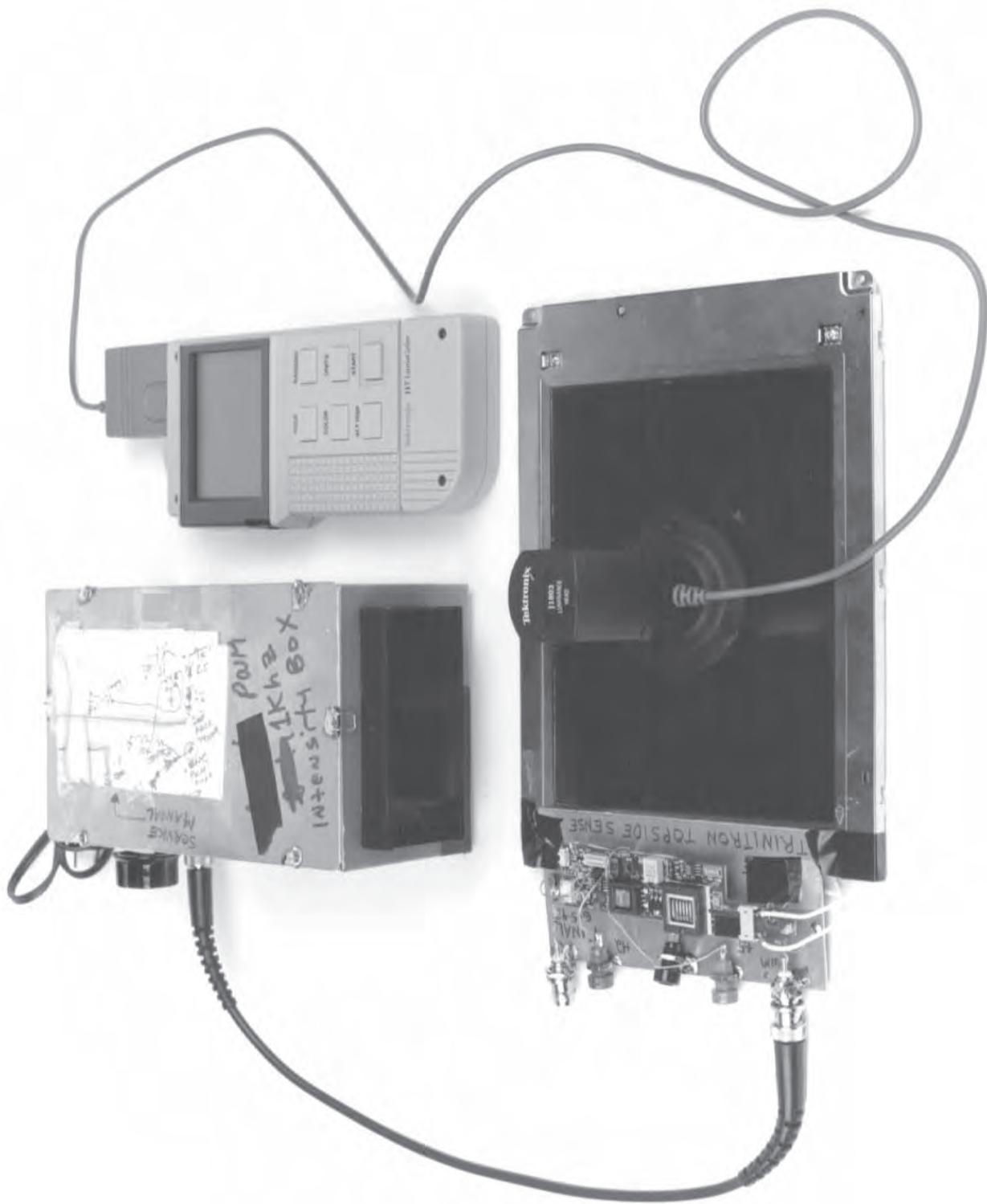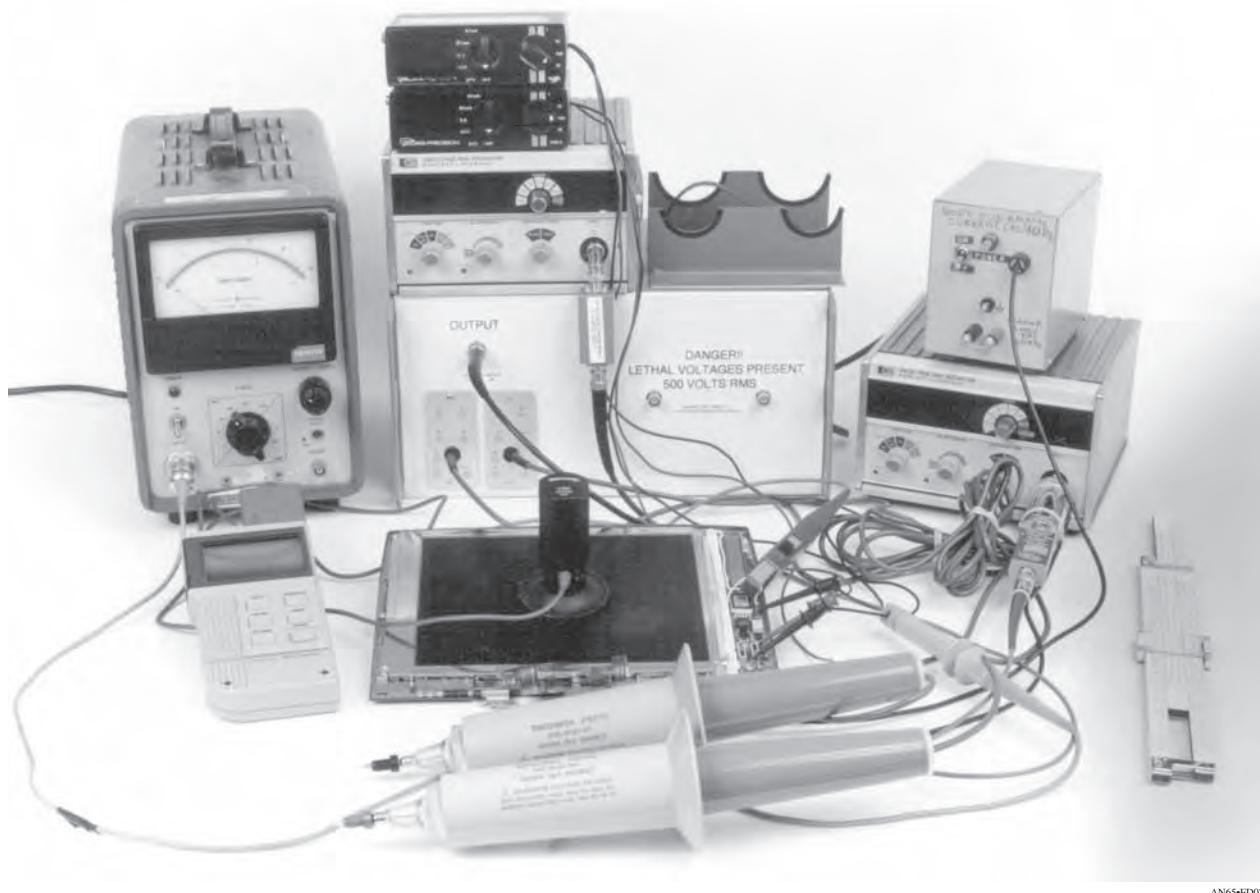
Figure D5. Glometer Photodiode/Amplifier Converts Lamp Light to Relative, Uncalibrated Electrical Output. Switches Permit Investigation of Individual Portions of Lamp Output

ALL RESISTORS = 1% FILM

= HP-5082-4204 PHOTODIODE

AN65•FD05

Figure D6. Apparatus for Calibrated Photometric Display Evaluation. Photometer (Upper Right) Indicates Display Luminosity Via Sensing Head (Center). CCFL Circuit (Left) Intensity Is Controlled by Figure F6's Calibrated Pulse Width Generator (Upper Left)

Figure D7. Complete CCFL Test Set Includes Photometer (Left and Center), Differential Voltage Probe/Amplifier (Upper and Lower Center), Current Probe Electronics (Right) and Input V and I DC Instrumentation (Upper Left). Computer (Lower Right) Permits Calculation of Electrical and Optical Efficiency

APPENDIX E

OPEN LAMP/OVERLOAD PROTECTION

The CCFL circuit's current source output means that "open" or broken lamps cause full voltage to appear at the transformer output. Safety or reliability considerations sometimes make protecting against this condition desirable. This protection is built into the LT118X series parts. Figure E1 shows a typical circuit. C5, R2 and R3 sense differentially across the Royer converter. Normally, the voltage across the Royer is controlled to relatively small values. An open lamp will cause full duty cycle modulation at the V_{SW} pin, resulting in large current drive through L2. This forces excessive Royer voltage at the C5/R2/R3

network, causing LT1184 shutdown via the "bulb" pin. C5 sets a delay, allowing Royer operation at high drive levels during lamp start-up. This prevents unwarranted shutdown during the lamp's transient high impedance start-up state.

The LT1172 and similar switching regulator parts need additional circuitry for open lamp protection. Figure E2 details the modifications. Q3 and associated components form a simple voltage mode feedback loop that operates if V_Z turns on. If T1 sees no load, there is no feedback and the Q1/Q2 pair receives full drive. Collector voltage rises

to abnormal levels, and V_Z biases via Q1's V_{BE} path. Q1's collector current drives the feedback node and the circuit finds a stable operating point. This action controls Royer drive and hence output voltage. Q3's sensing across the Royer provides power supply rejection. V_Z's value should be somewhat above the worst-case Q1/Q2 V_{CE} voltage under running conditions. It is desirable to select V_Z's value so clamping occurs at the lowest output voltage possible while still permitting lamp start-up. This is not as tricky as it sounds because the 10k/1µF RC delays the effects of Q3's turn-on. Usually, selecting V_Z several volts above the worst-case Q1/Q2 V_{CE} will suffice.

Additional protection for all CCFL circuits is possible by fusing the main supply line, typically at a value twice the largest expected DC current. Also, a thermally activated fuse is sometimes mated to Q1 and Q2. Excessive Royer current causes heating in the transistors, activating the fuse.

Overload Protection

In certain cases it is desirable to limit output current if either lamp wire shorts to ground. Figure E3 modifies a switching regulator-based circuit to do this. The current sensing network, normally series connected with the lamp, is moved to the transformer. Any overload current must originate from the transformer. Feedback sensing in this path provides the desired protection. This connection measures *total* delivered current, including parasitic terms, instead of lamp return current. Slight line regulation and current accuracy degradation occurs, but not to an objectionable extent.

Floating lamp circuits, because of their isolation, are inherently immune to ground referred shorts. Shorted lamp wires are also tolerated because of the primary side current sensing.
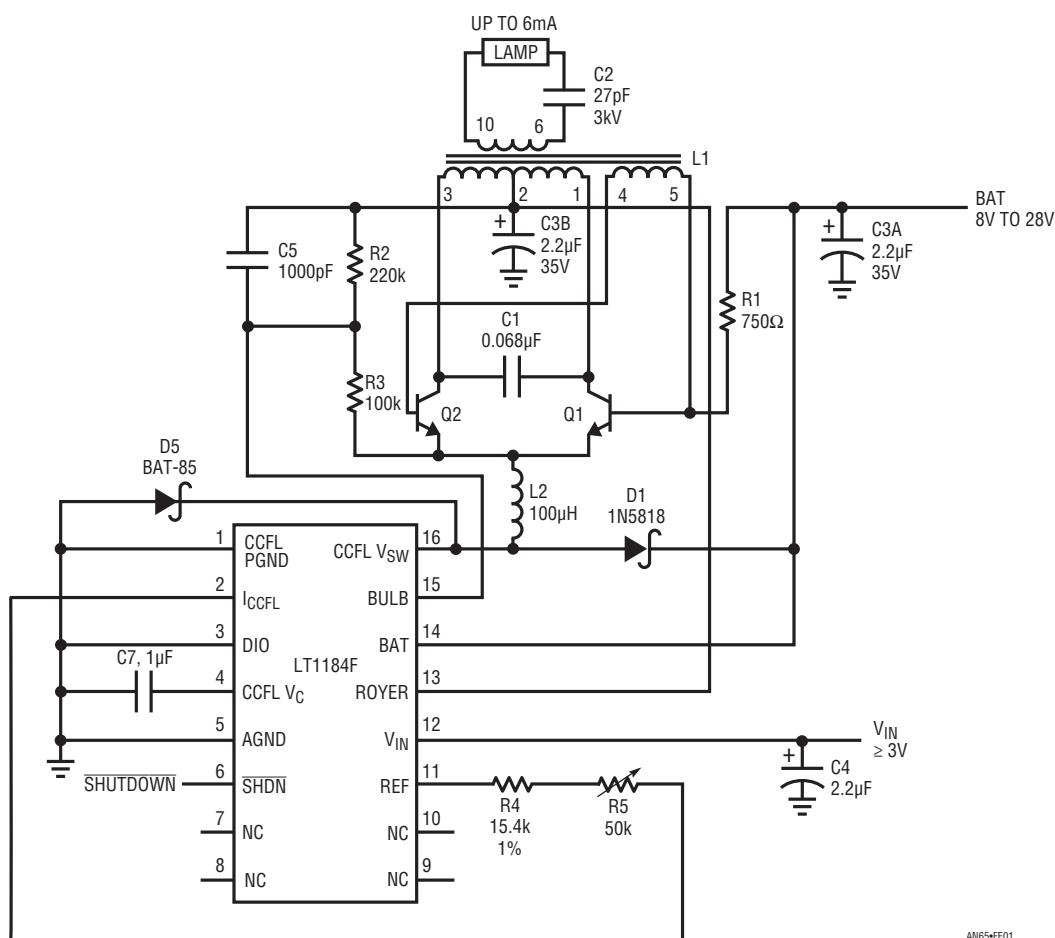

Figure E1. C5, R2 and R3 Provide Delayed Sensing Across Royer Converter, Protecting Against Open Lamp Conditions in LT118X Series ICs
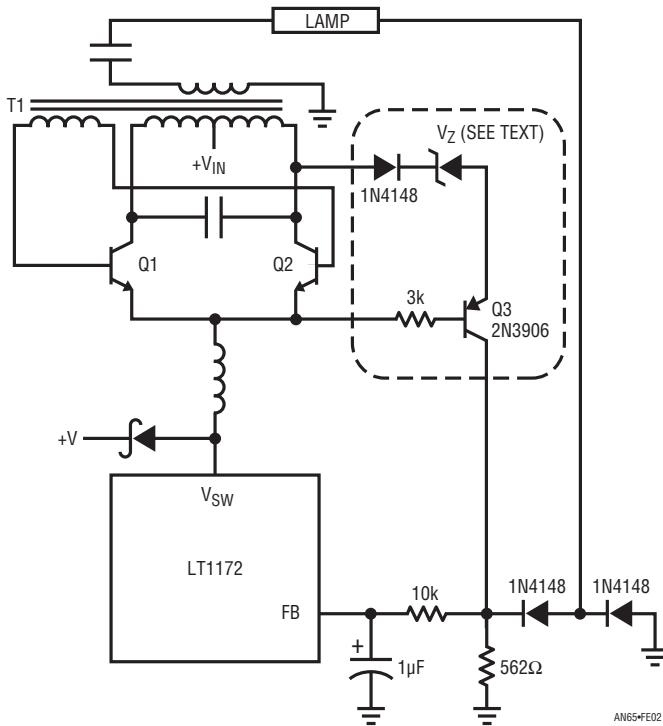
Figure E2. Q3 and Associated Components Form a Local Regulating Loop to Limit Output Voltage
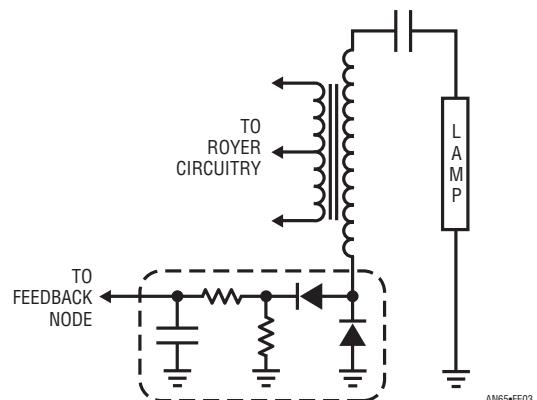

Figure E3. Relocating Feedback Network (Circled Components) to Transformer Secondary Maintains Current Control When Output Is Shorted. Trade-Off Is Slight Degradation in Line Regulation and Current Accuracy

APPENDIX F

INTENSITY CONTROL AND SHUTDOWN METHODS

The CCFL circuits usually require shutdown capability and some form of intensity (dimming) control. Figure F1 lists various options for the LT118X parts. Control sources include pulse width modulation (PWM), potentiometers and DACs or other voltage sources. The LT1186 (not shown) uses a digital serial-bit stream data input and is discussed in text associated with Figure 51.
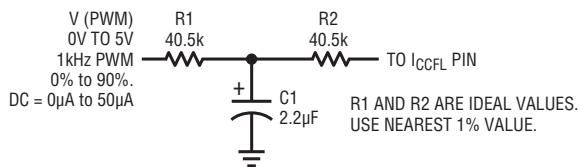
In all cases shown the average current into the I_{CCFL} pin sets lamp current. As such, the amplitude and duty cycle must be controlled in cases A and B. The remaining examples use the LT118X's reference to eliminate amplitude uncertainty-induced errors.

Figure F2 shows shutdown options for LT118X parts. The parts have a high impedance Shutdown pin, or power may simply be removed from V_{IN}. Switching V_{IN} power requires a higher current control source but shutdown current is somewhat lower.
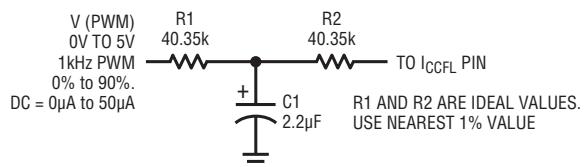
Figure F3 shows options for dimming control in LT1172 and similar regulator-based CCFL circuits. Three basic ways to control intensity appear in the figure. The most common intensity control method is to add a potentiometer in series with the feedback termination. When using this method ensure that the minimum value (in this case 562Ω) is a 1% unit. If a wide tolerance resistor is used the lamp current, at maximum intensity setting, will vary appropriately.

Pulse width modulation or variable DC is sometimes used for intensity control. Two interfaces work well. Directly driving the Feedback pin via a diode—22k resistor with DC or PWM produces intensity control. The other method shown is similar, but places the 1µF capacitor outside the feedback loop to get best turn-on transient response. This is the best method if output overshoot must be minimized. Note that in all cases the PWM source amplitude at 0%
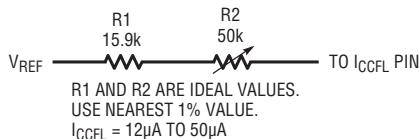
(F1a) LT1182/LT1183 I_CCFL PWM Programming

V (PWM)
0V TO 5V
1kHz PWM
0% to 90%.
DC = 0µA to 50µA

R1 40.5k   R2 40.5k

TO I_CCFL PIN

C1 2.2µF

R1 AND R2 ARE IDEAL VALUES.
USE NEAREST 1% VALUE.

(F1b) LT1184/LT1184F I_CCFL PWM Programming

V (PWM)
0V TO 5V
1kHz PWM
0% to 90%.
DC = 0µA to 50µA

R1 40.35k   R2 40.35k

TO I_CCFL PIN

C1 2.2µF

R1 AND R2 ARE IDEAL VALUES.
USE NEAREST 1% VALUE
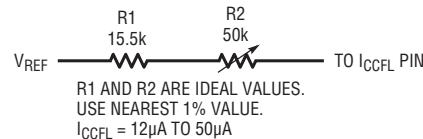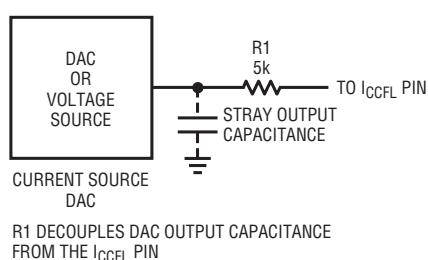
**(F1c) LT1183 I_CCFL Programming
with Potentiometer Control**

V_REF

R1 15.9k   R2 50k

TO I_CCFL PIN

R1 AND R2 ARE IDEAL VALUES.
USE NEAREST 1% VALUE.
I_CCFL = 12µA TO 50µA

**(F1d) LT1184/LT1184F I_CCFL Programming
with Potentiometer Control**

V_REF

R1 15.5k   R2 50k

TO I_CCFL PIN

R1 AND R2 ARE IDEAL VALUES.
USE NEAREST 1% VALUE.
I_CCFL = 12µA TO 50µA
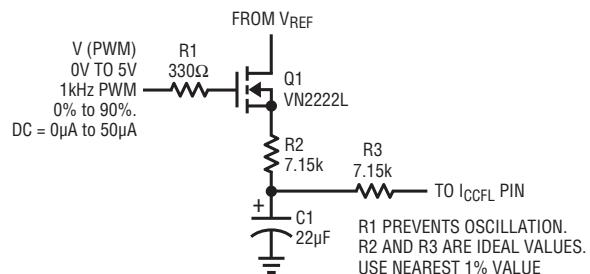
**(F1e) LT1182/LT1183/LT1184/LT1184F
I_CCFL Programming with DAC or Voltage Source Control**

DAC
OR
VOLTAGE
SOURCE

CURRENT SOURCE
DAC

R1 5k

TO I_CCFL PIN

STRAY OUTPUT
CAPACITANCE

R1 DECOUPLES DAC OUTPUT CAPACITANCE
FROM THE I_CCFL PIN

(F1f) LT1183 I_CCFL PWM Programming with V_REF

FROM V_REF

V (PWM)
0V TO 5V
1kHz PWM
0% to 90%.
DC = 0µA to 50µA

R1 330Ω

Q1 VN2222L

R2 7.15k   R3 7.15k

TO I_CCFL PIN

C1 22µF

R1 PREVENTS OSCILLATION.
R2 AND R3 ARE IDEAL VALUES.
USE NEAREST 1% VALUE
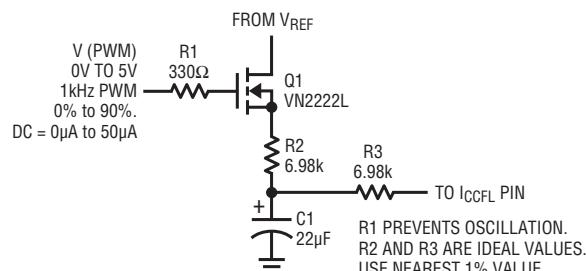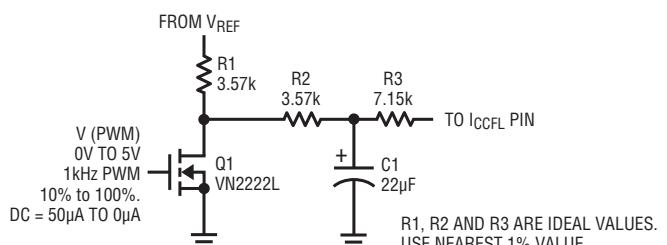
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(F1g) LT1184/LT1184F I_CCFL PWM Programming with V_REF

FROM V_REF

V (PWM)
0V TO 5V
1kHz PWM
0% to 90%.
DC = 0µA to 50µA

R1 330Ω

Q1 VN2222L

R2 6.98k   R3 6.98k

TO I_CCFL PIN

C1 22µF

R1 PREVENTS OSCILLATION.
R2 AND R3 ARE IDEAL VALUES.
USE NEAREST 1% VALUE

(F1h) LT1183 I_CCFL PWM Programming with V_REF

FROM V_REF

R1 3.57k

V (PWM)
0V TO 5V
1kHz PWM
10% to 100%.
DC = 50µA TO 0µA

Q1 VN2222L

R2 3.57k   R3 7.15k

TO I_CCFL PIN

C1 22µF

R1, R2 AND R3 ARE IDEAL VALUES.
USE NEAREST 1% VALUE

(F1i) LT1184/LT1184F I_CCFL PWM Programming with V_REF

FROM V_REF

R1 3.48k

R2 3.48k   R3 6.98k

TO I_CCFL PIN

V (PWM)
0V TO 5V
1kHz PWM
10% to 100%.
DC = 50µA TO 0µA

Q1 VN2222L

C1 22µF

R1, R2 AND R3 ARE IDEAL VALUES.
USE NEAREST 1% VALUE

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**Figure F1. Various Dimming Options for LT118X Series Parts. LT1186 (Not Shown) Has Serial-Bit
Stream Digital Dimming Input**

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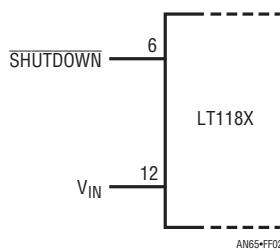
Figure F2. Shutdown Options for LT118X Series Parts Include Shutdown Pin or Simply Removing V$_{IN}$

duty cycle does not, by definition, effect full-scale lamp current certainty. See the main text section, "Feedback Loop Stability Issues" for pertinent discussion.

Figure F4 shows methods for shutting down switching regulator-based CCFL circuits. In LT1172 circuits pulling the V$_C$ pin to ground puts the circuit into micropower shutdown. In this mode about 50µA flows into the LT1172 V$_{IN}$ pin with essentially no current drawn from the main (Royer center tap) supply. Turning off V$_{IN}$ power eliminates the LT1172's 50µA drain. Other regulators, such as the LT1372, have a separate Shutdown pin.

About Potentiometers

Potentiometers, frequently used in CCFL dimming, require thought to avoid problems. In particular, resistance, ratio tolerances and other issues can upset an ill-prepared design. Keep in mind that ratio tolerances (see Figure F5)

are usually better than absolute resistance specifications. Because of this, it is sometimes advantageous to use the device as a voltage divider instead of a rheostat. The key issue in potentiometer-based dimming is usually ensuring that lamp overdrive cannot occur. This is why arranging dimming schemes for maximum intensity at "shorted" potentiometer positions is preferable. The "end resistance" tolerance, which should be checked, is often less significant and more repeatable than that for maximum resistance or even ratio setting. Other issues include wiper current capability, taper characteristic and circuit sensitivity to "opens." Always review circuit behavior for maximum wiper current demands. CCFL dimming schemes almost never require significant wiper current, but ensure that the particular scheme used doesn't have this problem. The
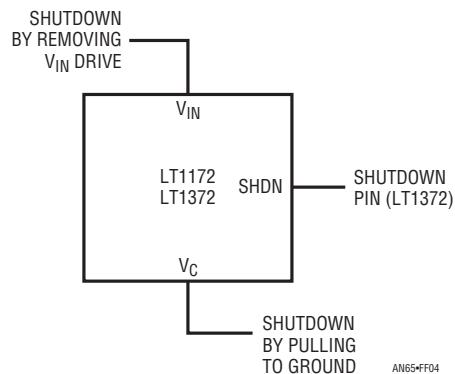

**Figure F4. Various Shutdown Options in LT1172/LT1372 Type CCFL Circuits**
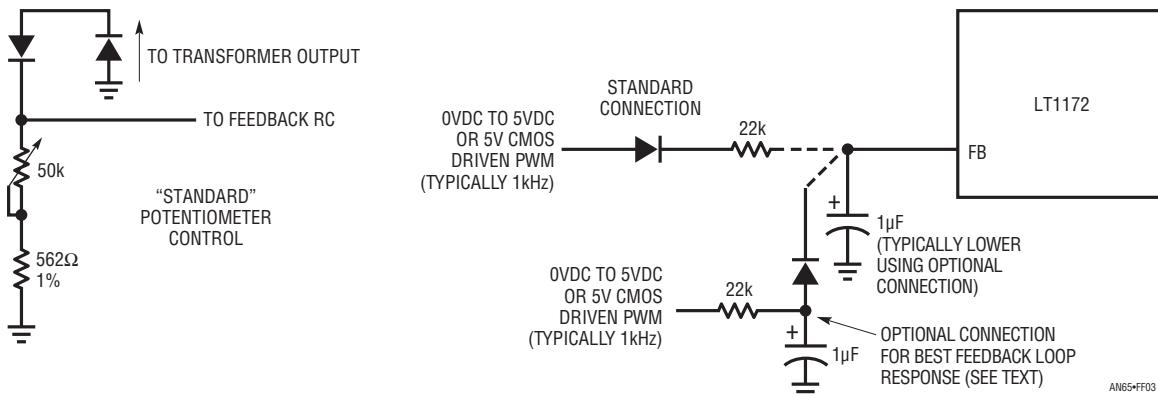
Figure F3. Various Options for Intensity Control in LT1172 and Similar Switching Regulator-Based CCFL Circuits

potentiometers taper, which may be linear or logarithmic, should be matched to the lamp's current-vs-light output characteristic to provide easy user settability. A poorly chosen unit can cause most of the useful dimming range to occur in a small section of potentiometer travel. Finally, always evaluate how the circuit will react if any terminal develops an open condition, which sometimes happens. It is imperative that the circuit have some relatively benign failure mode instead of forcing excessive lamp current or some other regrettable behavior.

Electronic equivalents of potentiometers are monolithic resistor chains tapped by MOS switches. Some devices feature nonvolatile onboard memory. These units have voltage rating restrictions which must be adhered to as with any integrated circuit. Additionally, they have all the limitations discussed in the section on mechanical potentiometers. Their most serious potential difficulty in backlight dimming applications is extremely high end resistance. In the "shorted" position the FET switch's on-resistance is typically 200Ω—much higher than a mechanical unit. Because of this, electronic potentiometers must almost always be set up as 3-terminal voltage dividers. This can usually be accommodated but may eliminate these devices in some applications.
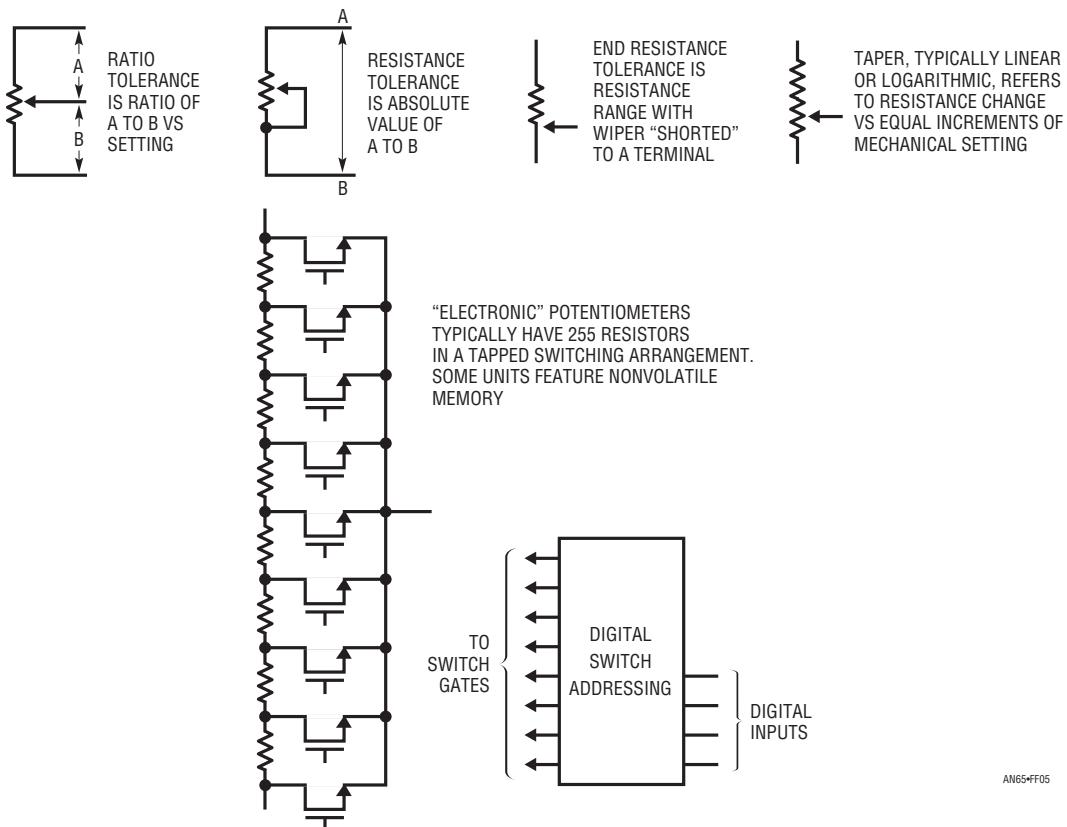

Figure F5. Relevant Characteristics of Mechanical and Electronic Potentiometers in CCFL Dimming Applications

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Precision PWM Generator

Figure F6 shows a simple circuit which generates precision variable pulse widths. This capability is useful when testing PWM-based intensity schemes. The circuit is basically a closed-loop pulse width modulator. The crystal controlled 1kHz input clocks the C1/Q1 ramp generator via the differentiator/CMOS inverter network and the LTC201 reset switch. C1's output drives a CMOS inverter, the output of which is resistively sampled, averaged and presented to A1's negative input. A1 compares this signal with a variable voltage from the potentiometer. A1's output biases the pulse width modulator, closing a loop around it. The CMOS inverter's purely ohmic output structure combines with A1's ratiometric operation (e.g., both of A1's input signals derive from the 5V supply) to hold pulse width constant. Variations in time, temperature and supply have essentially no effect. The potentiometer's setting is the sole determinant of output pulse width. Additional inverters provide buffering and furnish the output. The

Schottky diodes protect the output from latchup due to cable-induced ESD or accidental events[1] during testing.

The output width is calibrated by monitoring it with a counter while adjusting the 2kΩ trim pot.

As mentioned, the circuit is insensitive to power supply variation. However, the CCFL circuit averages the PWM output. It cannot distinguish between a duty cycle shift and supply variation. As such, the test box's 5V supply should be trimmed ±0.01V. This simulates a "design centered" logic supply under actual operating conditions. Similarly, paralleling additional logic inverters to get lower output impedance should be avoided. In actual use, the CCFL dimming port will be driven from a single CMOS output, and its impedance characteristics must be accurately mimicked.

Note 1: "Accidental events" is a nice way of referring to the stupid things we all do at the bench. Like shorting a CMOS logic output to a −15V supply (then I installed the diodes).
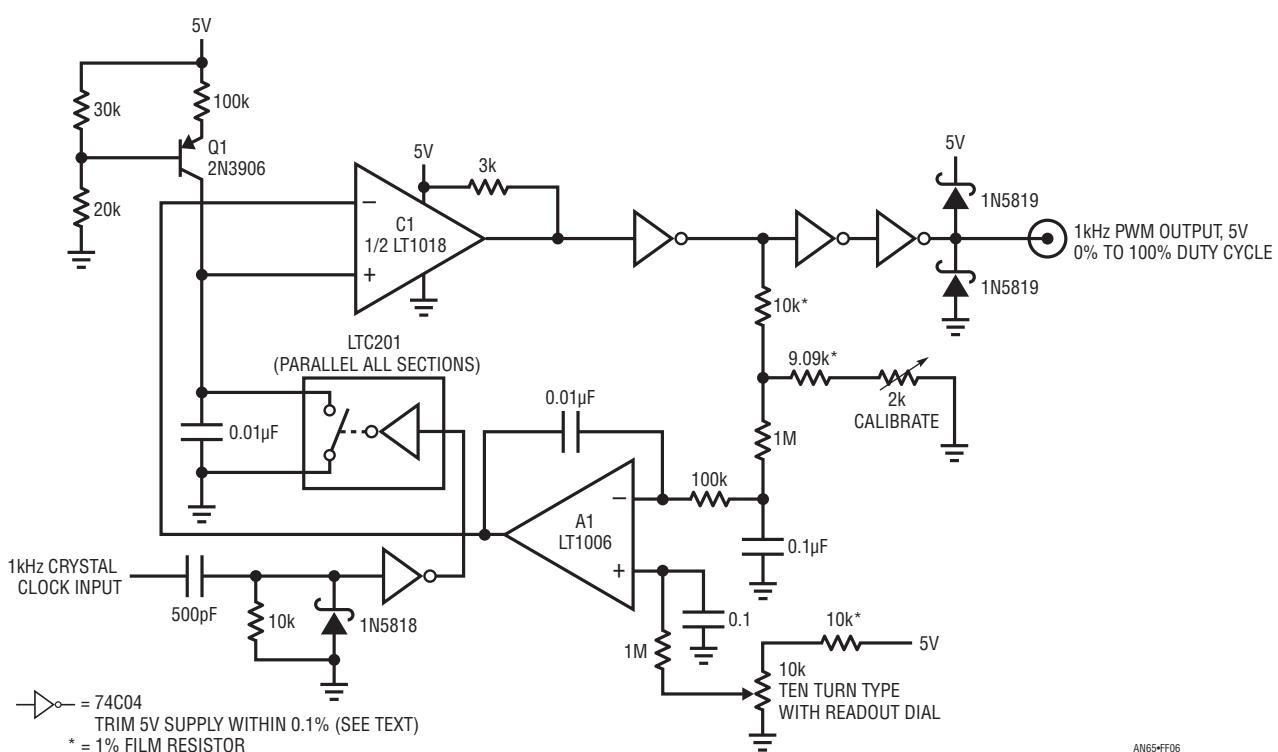
Figure F6. The Calibrated Pulse Width Test Box. A1 Controls C1-Based Pulse Width Modulator, Stabilizing Its Operating Point

APPENDIX G

LAYOUT, COMPONENT AND EMISSIONS CONSIDERATIONS

The CCFL circuits described in the text are remarkably tolerant of layout and impedance in supply lines. This is due to the Royer's relatively continuous current drain over time. Some review of current flow is, however, worthwhile. Figure G1 shows the more critical paths in thick lines for switching regulator-based CCFL circuits. In actual layout, these traces should be reasonably short and thick. The most critical consideration is that C1, T1's center tap and the diode should be connected *directly* together with minimum trace area between them. Similarly, C2 should be near the V_{IN} pin, although this placement is not nearly as critical as C1's.
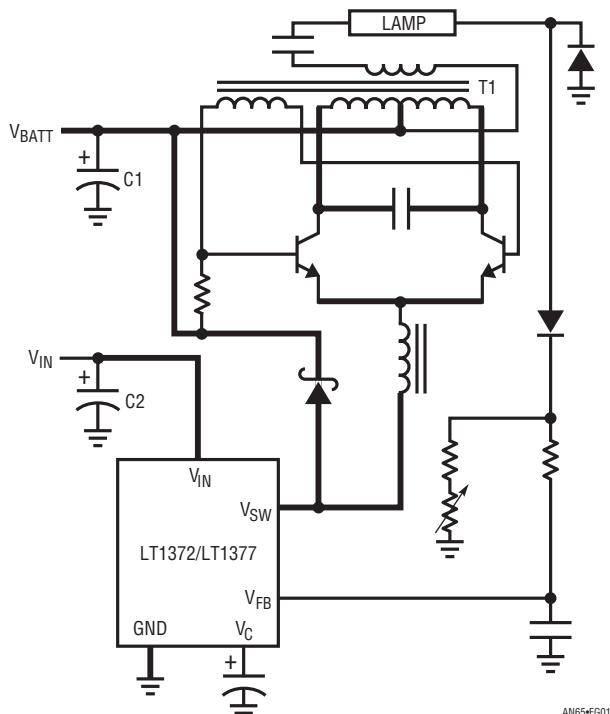
Figure G1. Thick Lines Denote PC Traces Requiring Low Impedance Layout in LT1172/LT1372 Type CCFL Circuits. Bypass Capacitors Associated with These Paths Should Be Mounted Near Load Point

Figure G2 indicates similar layout treatment for LT118X-based circuits. As before, the Royer and V_{IN} bypass capacitors should be near their respective load points, with the diode in close proximity to the Royer center tap.

Circuit Segmenting

In cases where space is extremely limited it may be desirable to physically segment the circuit. Some designs have placed a section of the circuit near the display with another portion remotely located. The best place for segmenting is the junction of the Royer transistor emitters and the inductor (Figure G3). Introducing a long, relatively lossy connection at this point imposes no penalty because signal flow into the inductor closely resembles a constant current source.

There are no wideband components due to the inductor's filtering effect. Figure G4 shows emitter voltage (Trace A) and current (Trace B) waveforms. There is no wideband component or other significant high speed energy movement. The inductor current waveform trace thickening due to Royer and switching regulator frequency mixing is not deleterious.

A very special case of segmentation involves replacing the transformer with two smaller units. Aside from space (particularly height) savings, electrical advantages are also realized. See Appendix I for details.

High Voltage Layout

Special attention is required for the board's high voltage sections. Board leakage, which can increase dramatically over life due to condensation cycling and particulate matter trapping, must be minimized. If precautions are not taken leakage will cause degraded operation, failures or destructive arcing. The only sure way to eliminate these possibilities is to *completely isolate the high voltage points from the circuit.* Ideally, no high voltage point should be within 0.25" of any conductor. Additionally, moisture trapping due to condensation cycling or improper board washing can be eliminated by routing the area under the
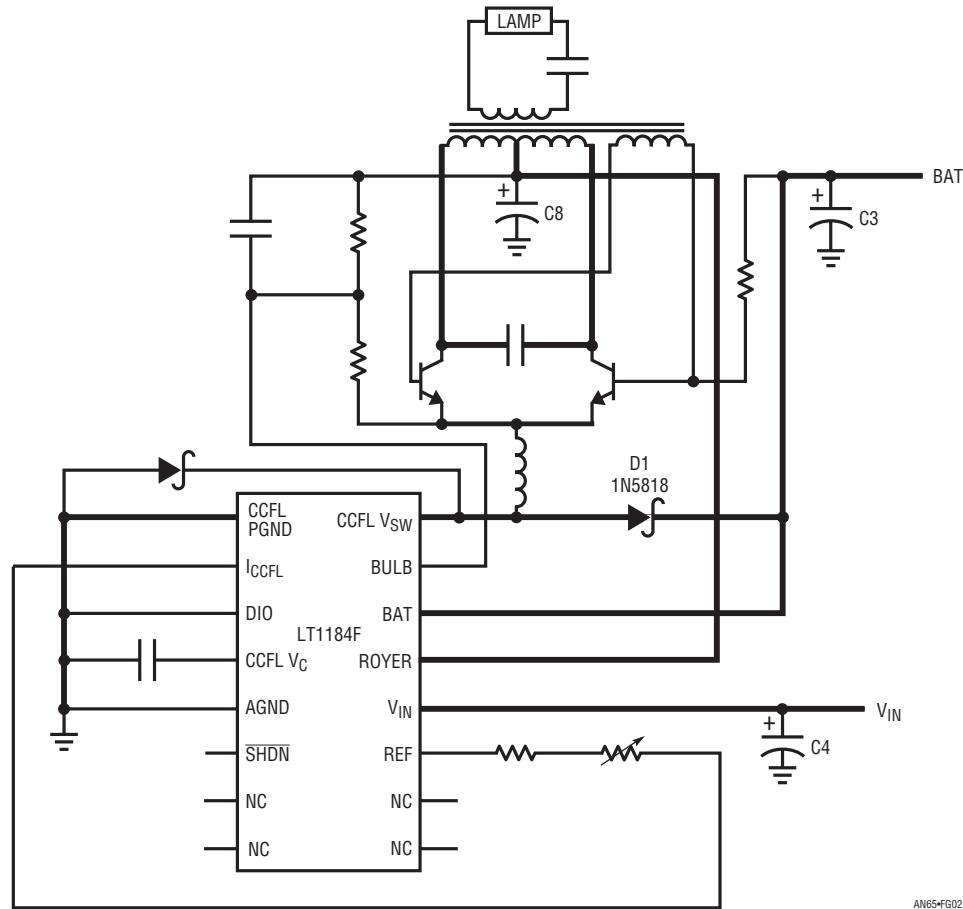
Figure G2. Critical Current Paths for LT118X Type Circuits. Thick Lines Denote PC Traces Requiring Low Impedance. Bypass Capacitors Associated with These Paths Should Be Near Load Points
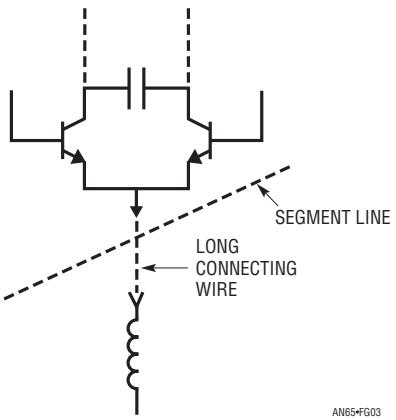

Figure G3. CCFL Circuit May Be Segmented in Limited Space Applications. Breaking at Emitter/Inductor Junction Imposes No Penalty
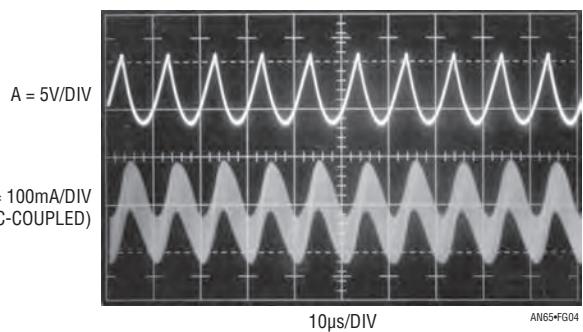

Figure G4. Royer Emitter/Inductor Junction Is Ideal Point for Segmenting CCFL Circuit. Voltage (Trace A) and Current (Trace B) Waveforms Contain Little High Frequency Content. Trace Thickening of Current Waveform Is Due to Frequency Mixing in Inductor

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transformer. This treatment, standard technique in high voltage layout, is strongly recommended. In general, carefully evaluate all high voltage areas for possible leakage or arcing problems due to layout, board manufacturing or environmental factors. Clear thinking is needed to avoid unpleasant surprises. The following commented photographs, visually summarizing the above discussion, are examples of high voltage layout.
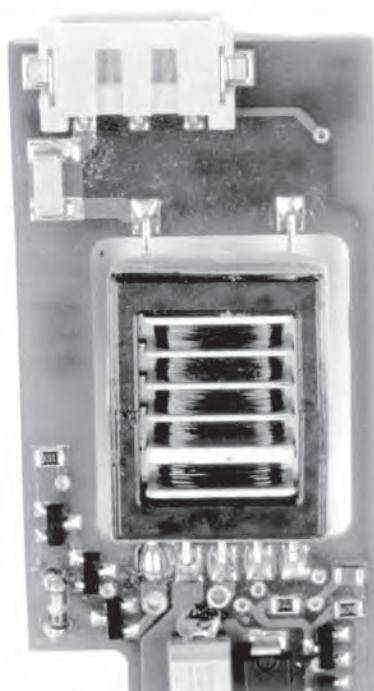

Figure G5. Transformer Output Terminals, Ballast Capacitor and Connector Are Isolated at End of Board. Slit Prevents Leakage


Figure G6. Reverse Side of G5. Note Routed Area Under Transformer, Eliminating Possibility of Moisture or Contaminant Trappings
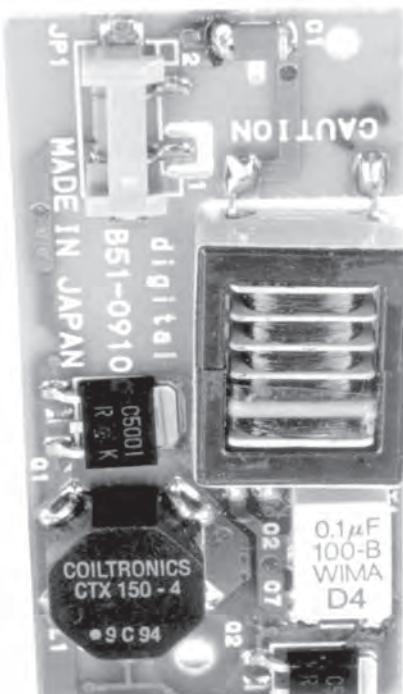
AN65•FG07

Figure G7. A Fully Routed Transformer, with High Voltage Capacitor Mounted Well Away from Transformer Ground Terminal (Right). Note Connector "Low Side" Trace Running Directly Away from High Voltage Points
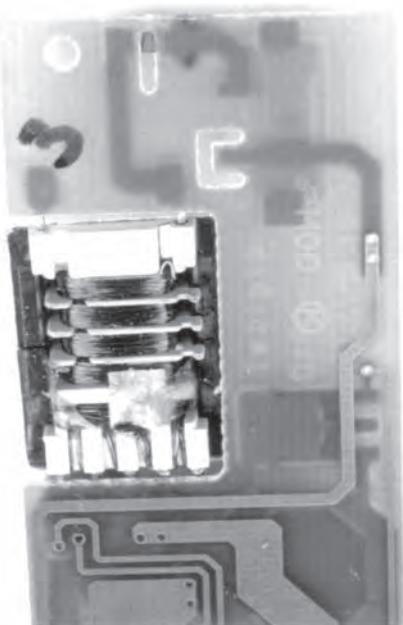

AN65•FG08

Figure G8. Routing Detail of Figure G7's Reverse Side. Transformer Header Sits Inside Routed Area, Saving Height Space. Board Markings Are Allowable Because HV Contacts Do Not Plate Through and Board Dielectric Strength Is Known

AN65•FG09

Figure G9. Very Thorough Routing Treatment Breaks Up Leakage. Routing Under Ballast Capacitor and Around Connector "Low Side" Pin Allows Tight Layout

AN65•FG10

Figure G10. Reverse Side of G9 Shows Offset Transformer Placement Necessitated by Packaging Restrictions. Transformer Header Sits in Routed Area, Minimizing Overall Board Height

AN65•FG11
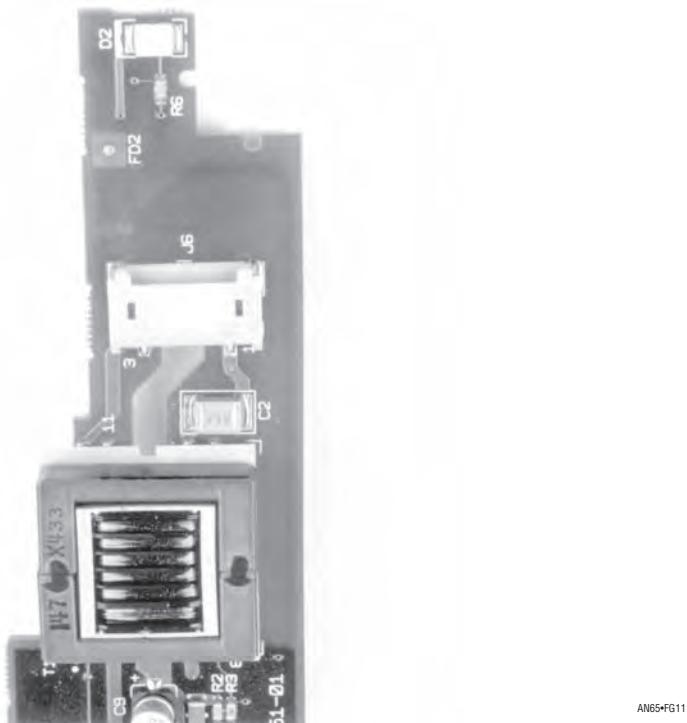
**Figure G11. Topside of Board Shows Isolation Slit Running to the HV Connector**
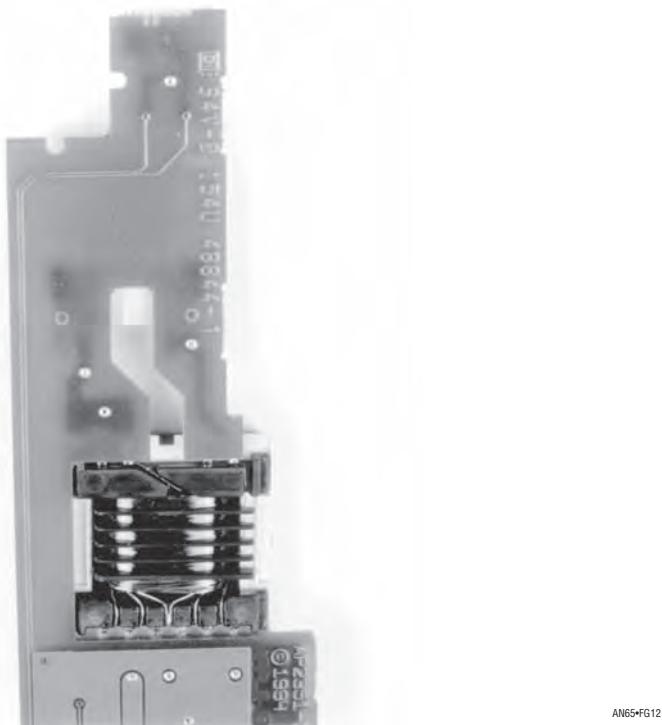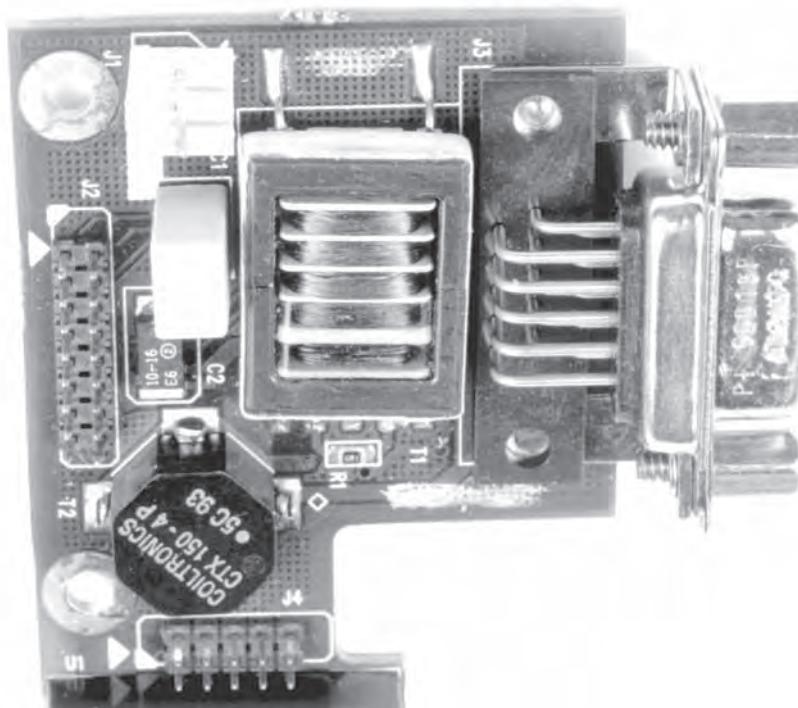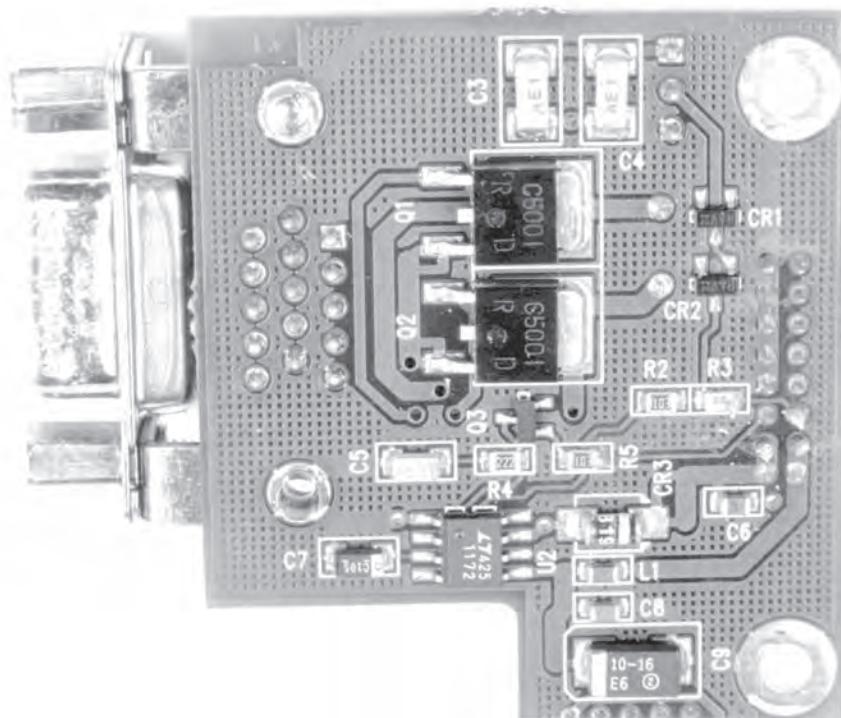
AN65•FG12

**Figure G12. Bottom of G11 Shows Routed Area. Traces at Board Extreme Left Are Undesirable
but Acceptable. Isolation Slit Between Traces and HV Points Would Be Preferable**

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AN65•FG13

Figure G13. A Disaster. Cross-Hatched Ground Plane Surrounds Output Connector and High Voltage Transformer Pins (Upper Center) in This Computer "Aided" Layout. Board Failed Spectacularly at Turn-On

AN65•FG14

Figure G14. Bottom Side of G13. Ground Plane in Region of Paralleled Ballast Capacitors (Upper Center) Caused Massive Arcing at Turn-On. Board Needs Complete Re-Layout. Computer Layout Software Package Needs E and M Course

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Discrete Component Selection

Discrete component selection is quite critical to CCFL circuit performance. A poorly chosen dielectric for the collector resonating capacitor can easily degrade efficiency by 5% to 8%. The WIMA and Panasonic types specified are quite good and very few other capacitors perform as well. The Panasonic unit is the only surface mounting type recommended although about 1% more lossy than the "through-hole" WIMA.

The transistors specified are quite special. They feature extraordinary current gain and V_{CE} saturation specifications. The ZDT1048, a dual unit designed specifically for backlight service, saves space and is the preferred device. Figure G15 summarizes relevant characteristics. Substitution of standard devices can degrade efficiency by 10% to 20% and in some cases cause catastrophic failures.[1]
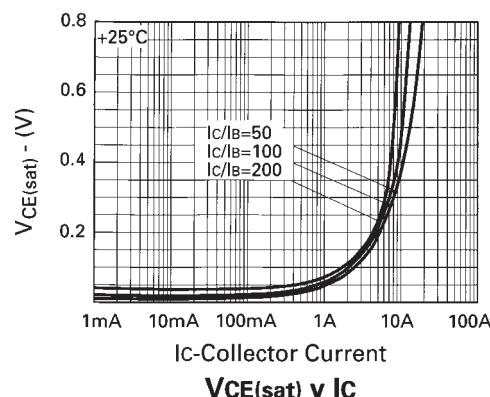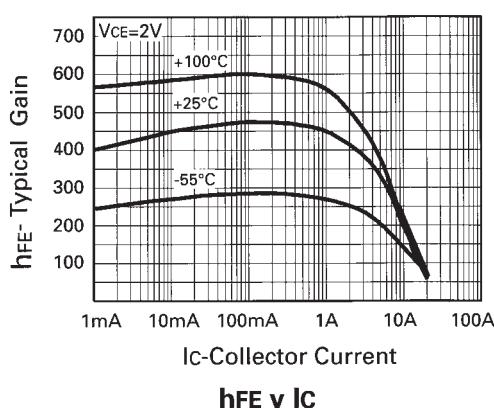
Note 1: Don't say we didn't warn you.

ELECTRICAL CHARACTERISTICS (at T_{amb} = 25°C unless otherwise stated).

PARAMETER	SYMBOL	MIN.	TYP.	MAX.	UNIT	CONDITIONS.
Collector-Emitter Breakdown Voltage	V_{CES}	50	85		V	$I_C=100\mu A$
Collector-Emitter Breakdown Voltage	V_{CEV}	50	85		V	$I_C=100\mu A$, $V_{EB}=1V$
Collector Cut-Off Current	I_{CBO}		0.3	10	nA	$V_{CB}=35V$
Collector-Emitter Saturation Voltage	$V_{CE(sat)}$		27 55 120 200 250	45 75 160 240 350	mV mV mV mV mV	$I_C=0.5A$, $I_B=10mA$* $I_C=1A$, $I_B=10mA$* $I_C=2A$, $I_B=10mA$* $I_C=5A$, $I_B=100mA$* $I_C=5A$, $I_B=20mA$*
Static Forward Current Transfer Ratio	h_{FE}	280 300 300 250 50	440 450 450 300 80	1200		$I_C=10mA$, $V_{CE}=2V$* $I_C=0.5A$, $V_{CE}=2V$* $I_C=1A$, $V_{CE}=2V$* $I_C=5A$, $V_{CE}=2V$* $I_C=20A$, $V_{CE}=2V$*
Transition Frequency	f_T		150		MHz	$I_C=50mA$, $V_{CE}=10V$ $f=50MHz$

ZETEX
U.K. FAX: 0161627-5467
U.S. FAX: 5168647630
HONG KONG FAX: 987 9595

*Measured under pulsed conditions. Pulse width=300μs. Duty cycle ≤ 2%

hFE v IC

VCE(sat) v IC

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Figure G15. Short Form Specifications for Zetex ZDT1048 Dual Transistor. Extraordinary Beta and Saturation Characteristics Are Ideal for Royer Converter Section of Backlight Circuits

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The following section, excerpted with permission from Zetex Application Note 14 (see Reference 28), reviews Royer circuit operation with emphasis on transistor operating conditions and requirements.

Excerpted from "Transistor Considerations for LCD Backlighting," Neil Chadderton, Zetex plc.

Basic Operation Of Converter

The drive requirements dictated by the CCFL tube's behaviour and preferred operating conditions can be achieved by the resonant push-pull converter shown in Figure G16. This is also referred to as the Royer Converter, after G.H. Royer who proposed the topology in 1954 as a power converter. (Note: Strictly speaking the backlighting converter uses a modified version of the Royer converter – the original used a saturating transformer to fix the operating frequency, and therefore produced a squarewave drive waveform). The circuit looks simple but this is very deceptive: many components interact, and while the circuit is capable of operation with widely varying component values, (useful during development) optimisation is required for each design to achieve the highest possible efficiencies.

Transistors Q1 and Q2 are alternatively saturated by the base drive provided by the feedback winding W4. The base current is defined by resistors R1 and R2. Supply inductor L1 and primary capacitance C1 force the circuit to run sinusoidally thereby minimising harmonic generation and RFI, and providing the preferred drive waveform to the load. Voltage step-up is achieved by the W1:(W2+W3) turns ratio. C2 is the secondary winding ballast capacitor, and effectively sets the tube current.

Prior to the tube striking, or when no tube is connected, the operating frequency is set by the resonant parallel circuit comprising the primary capacitance C1, and the transformer's primary winding W2+W3. Once the tube has struck, the ballast capacitor C2 plus distributed tube and parasitic capacitances are reflected back through the transformer, and the operating frequency is lowered.
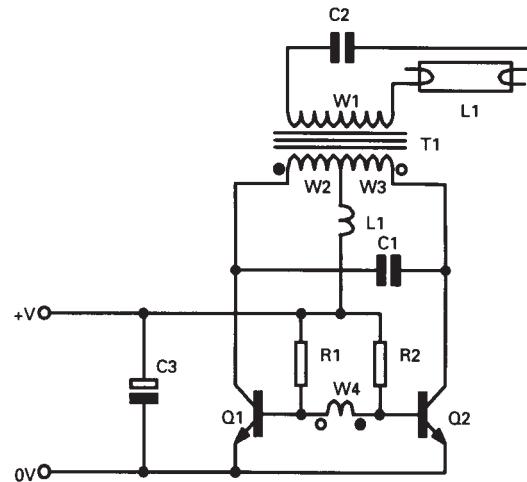

Figure G16. Generalised Royer Converter.

The secondary load can become dominant in circuits with a high transformer turns ratio, Eg. those designed to operate from very low DC input voltages.

Each transistor's collector is subject to a voltage= $2 \times \pi/2 \times V_S$, (or just $\pi \times V_S$) where V_S is the DC input voltage to the converter. (The $\pi/2$ factor being due to the relationship between average and peak values for a sinewave, and the x2 multiplier being due to the 2:1 autotransformer action of the transformer's centre-tapped primary). This primary voltage is stepped up by the transformer turns ratio Ns:Np, to a high enough level to reliably strike the tube under all conditions:- starting voltage is dependent on display housing, location of ground planes, tube age, and ambient temperature.

The basic converter shown in Figure G16 is a valid and useful circuit that has been utilised for many systems and indeed offered as a sub-system by several manufacturers.

Requisite Transistor Characteristics

The relatively low operating frequency as required by the backlighting Royer Converter (to minimise HV parasitic capacitance losses), and the ease of transformer drive, makes this circuit particularly suitable for bipolar transistor implementation. This isn't to exclude MOSFET based designs (some IC vendors have specified MOS as this suits their technology) but in terms of equivalent on-resistance and silicon efficiency, the low voltage bipolar device has no equal. For example, the ZETEX ZTX849 E-LINE (TO-92 compatible) transistor exhibits a $R_{CE(sat)}$ of $36m\Omega$. This can only be matched by a much larger (and expensive) MOSFET die, only available in TO-220, D-PAK, and similar larger packages.

The most important transistor characteristics are voltage rating, $V_{CE(sat)}$, and h_{FE}, and are considered in some detail below.

The voltage rating required deserves some thought with respect to the standard transistor breakdown parameters, as it is possible to over-specify a device on grounds of voltage rating, and thereby incur a reduction in efficiency due to unnecessary on-resistance losses. The primary breakdown voltage BV_{CBO}, of a planar bipolar transistor depends on the epitaxial layer - specifically it's thickness and resistivity. The breakdown voltage of most interest to the designer is usually that attained across the Collector-Emitter (C-E) terminals. This value can vary between the primary breakdown BV_{CBO} and a much lower voltage dependent on the state of the base terminal bias.
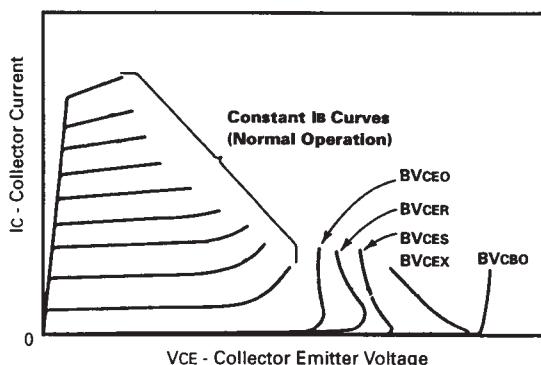
[The breakdown mechanism is caused by the avalanche multiplication effect, whereby free electrons can be imparted with sufficient energy by the reverse bias electric field such that any collisions can lead to ionisation of the lattice atoms. The free electrons thus generated are then accelerated by the field and produce further ionisation. This multiplication of free carriers increases the reverse current dramatically, and so the junction effectively clamps the applied voltage. The base terminal can obviously influence the junction current – thereby modulating the voltage required for a breakdown condition.]

Figure G17 shows how the breakdown characteristic is seen to vary for different circuit conditions. The BV_{CEO} rating (or when the base is open circuit) allows the Collector-Base (C-B) leakage current I_{CBO} to be effectively amplified by the transistor's β thus significantly increasing the leakage component to I_{CEO}. Shorting the base to the Emitter (BV_{CES}) provides a parallel path for the C-B leakage, and so the voltage required for breakdown is higher than the open base condition. BV_{CER} denotes the case between the open and shorted base options:- R indicating an external base-emitter resistance, the value of which is typically 100 to $10k\Omega$. BV_{CEV} or BV_{CEX} is a special case where the base-emitter is reverse biased; this can provide a better path for the C-B leakage, and so this rating yields a voltage close to, or coincident with the BV_{CBO} value. Figure G18 shows a curve tracer view of the relevant breakdown modes of the ZTX849 transistor, including a curve showing the device in the "on" state. Curves 1 and 2 are virtually coincident and show BV_{CBO} and BV_{CES} respectively. Curve 3 shows the BV_{CEV} case with an applied base bias (V_{EB}) of -1V. Curve 4 shows BV_{CEO} at approximately 36V. Curve 5 is a BV_{CE} curve, showing how the breakdown condition is affected by a positive base bias of 0.5V.

**Figure G17. Voltage Breakdown Modes of Bipolar**
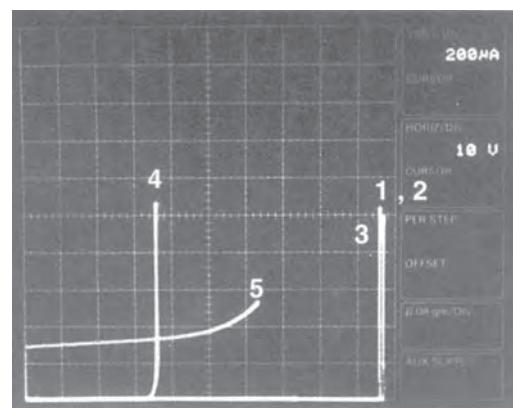
Figure G18. Breakdown Modes of ZTX849

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The BV$_{CEV}$ rating has particular relevance to the Royer Converter, as can be surmised from Figure 19. Examination of this will show that the transistor only experiences the high C-E voltage when the base voltage has been taken negative by the feedback winding, these events of course being in perfect synchronism. An expanded view of the C-E and B-E waveforms is shown in Figure G20.
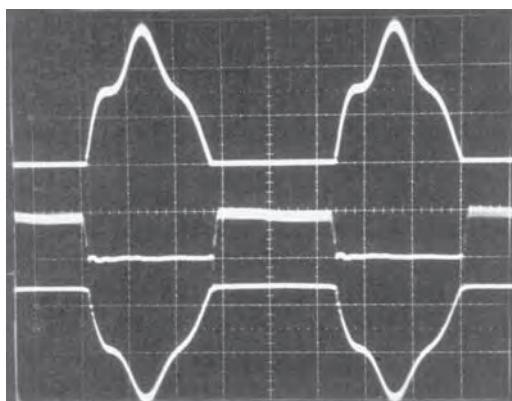

**Figure G19. Royer Converter Operating Waveforms:
V$_{CE}$ 10V/div; I$_E$ 0.5A/div; V$_{BE}$ 2V/div respectively,
2µs/div horizontal**
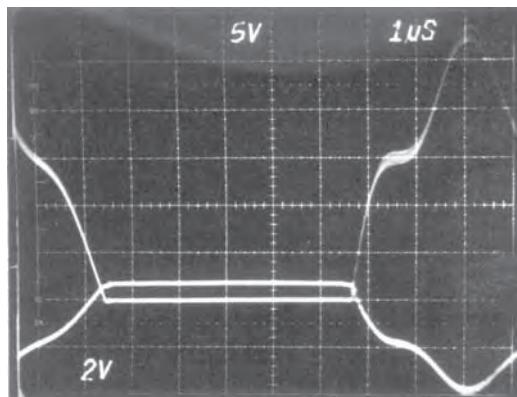
**Figure G20. Royer Converter: V$_{CE}$ and V$_{BE}$
Waveforms: 5V/div and 2V/div respectively**

[Note: The voltage applied by the feedback winding must not exceed the BV$_{EBO}$ of the transistor. This is specified at 5V usually, against an actual of 7.5 to 8.5V].

The V$_{CE(sat)}$ and h$_{FE}$ parameters have a direct bearing on the circuit's electrical conversion efficiency. This is especially true of low voltage battery powered systems, due to the high current levels involved. Selection of standard LF amplifier transistors provides far from ideal results; these parts are for general purpose linear and non-critical switching use only. The high V$_{CE(sat)}$ inherent to these parts, and low current gain could reduce circuit efficiency to less than 50%. For example, the stated V$_{CE(sat)}$ maximum measured at 500mA, for the FZT849 SOT223 transistor, and a LF device sometimes quoted as a suitable Royer Converter transistor are 50mV and 0.5V respectively. Eg.

	V$_{CE(sat)}$	@I$_C$	Ib
FZT849	50mV	0.5A	20mA
BCP56	0.5V	0.5A	50mA

To address the V$_{CE(sat)}$ issue, large power transistors are occasionally specified. Unfortunately their capacitance, and characteristic low base transport factor (a feature of Epitaxial Base devices) can lead to problems with cross-conduction losses due to long storage and switching times. The current gain is also important, as the losses in the base bias can be significant to the overall figure; judicious selection of the bias resistor to ensure a minimum V$_{CE(sat)}$ while preventing base overdrive needs to consider supply variation, maximum lamp current, and transistor h$_{FE}$ minimum value and range.

For the above reasons, transistors designed and optimised for high current switching applications offer the most cost-effective and efficient solutions. Figure G21 shows the V$_{CE(sat)}$ exhibited by the ZTX1048A for a range of forced gain values. This device is one of the ZTX1050 series of transistors that employ a scaled up variant of the highly efficient Matrix geometry, developed for the ZETEX "Super-SOT" series. This enables a V$_{CE(sat)}$ performance similar to the ZTX850 series at the low to moderate currents relevant to this application, though utilising a smaller die, and therefore providing a cost and possibly a space saving advantage.
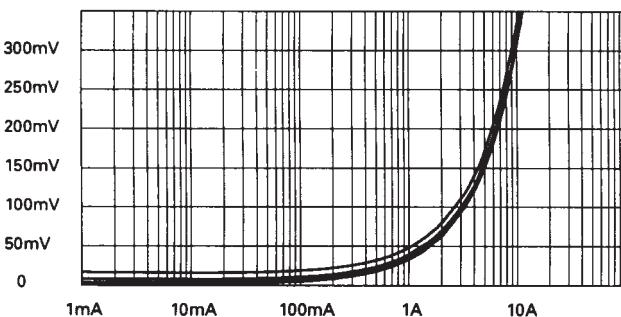

**Figure G21. V$_{CE(sat)}$ v I$_C$ for the ZTX1048A
Forced gains of 10,20,50,100**

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Additional Discrete Component Considerations

The magnetics specified have also been carefully selected and substitution can lead to problems ranging from poor efficiency to bad line regulation.

Bypass capacitors can be any type specified for switching regulator service, although tantalum types should be avoided for Royer bypassing if the supply is capable of delivering high current. As of this writing no tantalum supplier can guarantee reliability in the face of high current turn-on. If tantalums must be used, an X2 voltage derating factor *must* be enforced.[2]

The 2.2μF Royer bypass value used in LT118X-based circuits has been selected to ensure against any possible long-term damage to the IC's internal current shunt. Turn-on current surges can be large and this value limits them to safe excursions.

The high speed catch diode associated with the V_{SW} pin should be capable of handling the fast current spikes encountered. Schottky types offer lower loss than regular high speed units.[3]

Emissions

There are rarely emission problems with the CCFL circuits. The Royer circuit's resonant operation minimizes radiated energy at frequencies of interest. There is often more RF energy associated with the switching regulators V_{SW} node and minimizing exposed trace area eliminates problems. Incidental radiation from magnetics is reasonably low. Some (relatively rare) cases require consideration of magnetics placement to prevent interaction with other circuitry. If shielding is required its effects should be evaluated early in the design. Shielding in the vicinity of the Royer transformer can cause effects ranging from changing the inverter resonance to secondary arcing.

Note 2: See Footnote 1. Read it twice.
Note 3: Discussing utilization of 60Hz rectifier diodes (e.g., 1N4002) in this application qualifies as obscene literature. See also Footnote 1.

APPENDIX H

LT1172 OPERATION FROM HIGH VOLTAGE INPUTS

Some applications require higher input voltages. The 20V maximum input specified in the figures is set by the LT1172 going into its isolated flyback mode (see LT1172 data sheet), not breakdown limits. If the LT1172 V_{IN} pin is driven from a low voltage source (e.g., 5V) the 20V limit may be extended by using Figure H1's network. If the LT1172 is driven from the same supply as L1's center tap, the network is unnecessary, although efficiency will suffer. No other switching regulator discussed in the text is subject to this issue. Their operating voltage is set solely by voltage breakdown limits.


5V
39.2k
1%
1N4148
TO LT1172
V_{FB} PIN
16.9k
1%
AN65•FH01
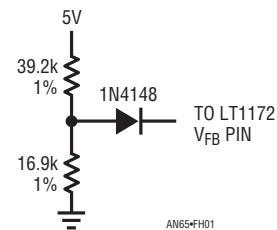
Figure H1. Network Allows LT1172 Operation Beyond 20V Inputs

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APPENDIX I

ADDITIONAL CIRCUITS

Desktop Computer CCFL Power Supply

Desktop computers, being line operated, can support higher power displays. High power operation permits high luminosity, enlarged display area or both. Typically, desktop displays absorb 4W to 6W and run from a relatively high voltage, regulated supply. Figure I1 shows such a display.

This "grounded lamp" LT1184-based configuration, similar to previously described versions, requires little comment. The transformer is a high power type, and scaling of the "I_{CCFL}" current programming resistors allows 9mA lamp current. In this case programming is via clamped PWM (see Appendix F), although all other methods described are feasible.
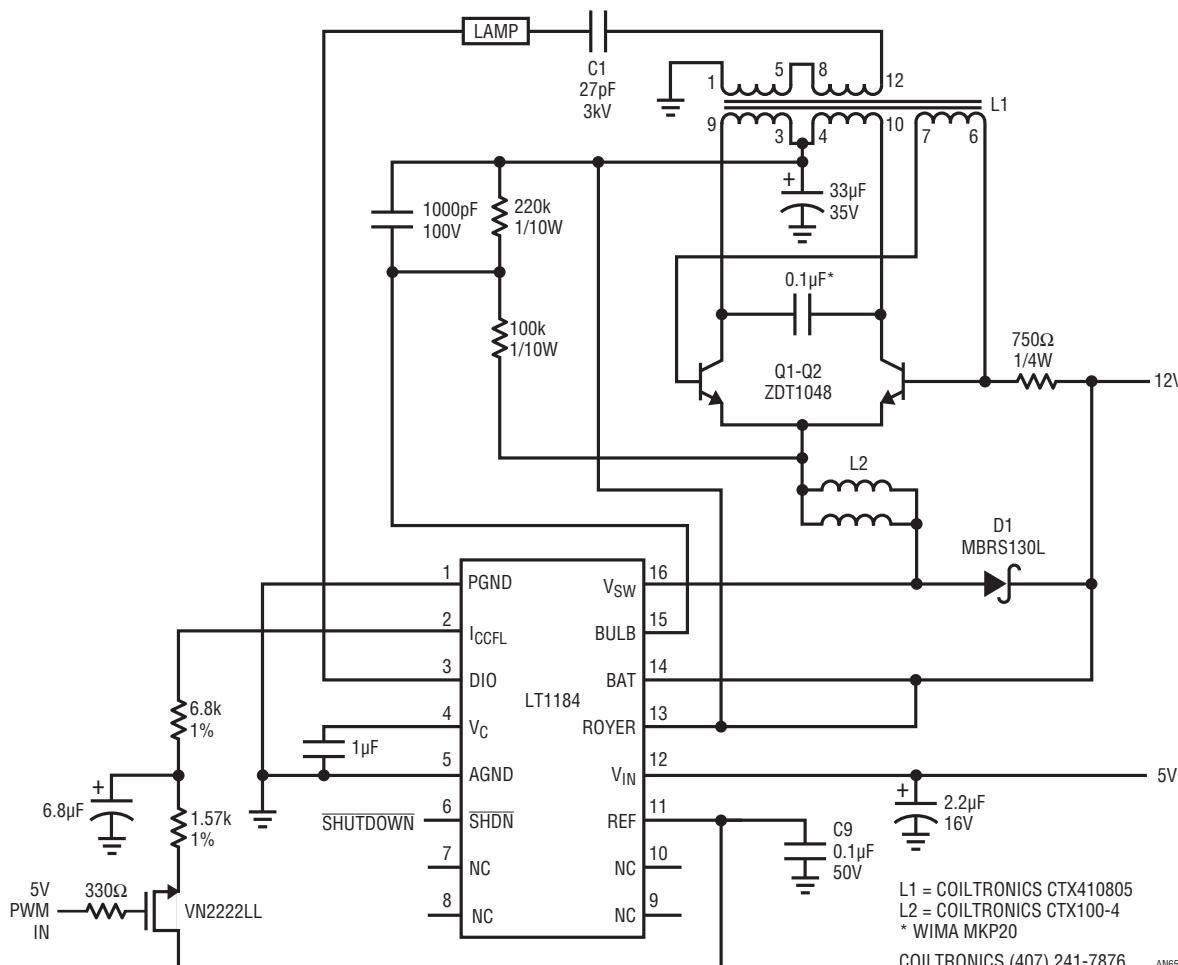
Figure I1. High Power Transformer and Scaled I$_{CCFL}$ Values Permit Desktop Computer LCD Operation

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Dual Transformer CCFL Power Supply

Space constraints may dictate utilization of two small transformers instead of a single, larger unit. Although this approach is somewhat more expensive, it can solve space problems and offers other attractive advantages. Figure I2's approach is essentially a "grounded lamp" LT1184-based circuit. The transistors drive two transformer primaries in parallel. The transformer secondaries, stacked in series, provide the output. The relatively small transformers, each supplying half the load power, may be located directly at the lamp terminals. Aside from the obvious space advantage (particularly height), this arrangement minimizes parasitic

wiring losses by eliminating high voltage lead length. Additionally, although the lamp receives differential drive, with its attendant low parasitic losses, the feedback signal is ground referred. Thus, the stacked secondaries afford floating lamp operation efficiency with grounded mode current certainty and line regulation.

L1 is directly driven, with winding 4-5 furnishing feedback in the normal fashion. L3, "slaved" to L1, produces phase-opposed output at its secondary. L1's and L3's interconnects must be laid out for low inductance to maintain waveform purity. The traces should be as wide as possible (e.g., 1/8") and overlaid to cancel inductive effects.
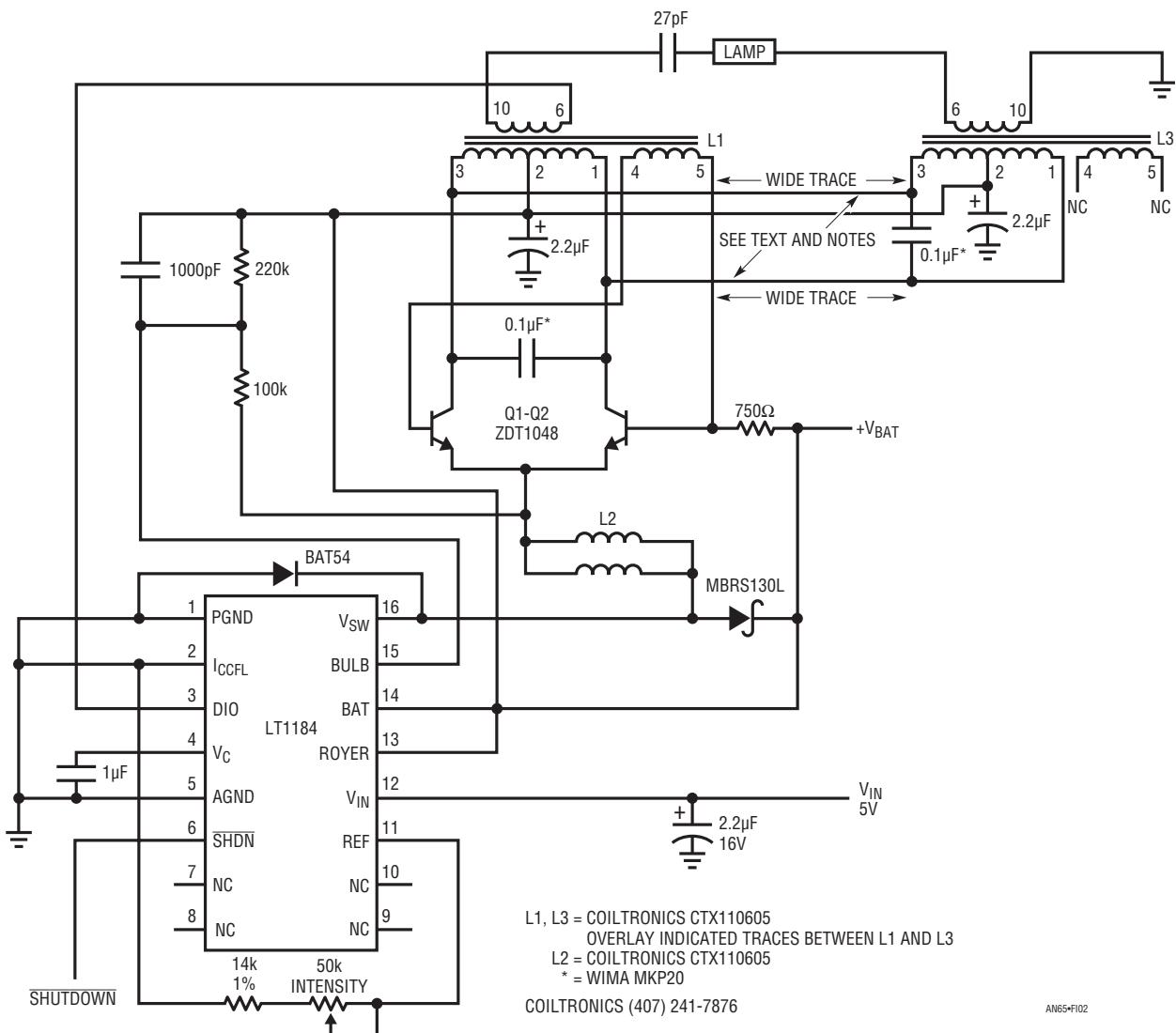

Figure I2. Dual Transformers Save Space and Minimize Parasitic Losses While Maintaining Current Accuracy and Line Regulation. Trade-Off Is Increased Cost

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HeNe Laser Power Supply

Helium-neon lasers, used for a variety of tasks, are difficult loads for a power supply. They typically need almost 10kV to start conduction, although they require only about 1500V to maintain conduction at their specified operating currents. Powering a laser usually involves some form of start-up circuitry to generate the initial breakdown voltage and a separate supply for sustaining conduction. Figure I3's circuit considerably simplifies driving the laser. The start-up and sustaining functions have been combined into a single closed-loop current source with over 10kV of compliance. The circuit is recognizable as a reworked CCFL power supply with a voltage tripled DC output.

When power is applied, the laser does not conduct and the voltage across the 190Ω resistor is zero. The LT1170 switching regulator FB pin sees no feedback voltage, and its Switch pin (V_{SW}) provides full duty cycle pulse width modulation to L2. Current flows from L1's center tap through Q1 and Q2 into L2 and the LT1170. This current flow causes Q1 and Q2 to switch, alternately driving L1. The 0.47µF capacitor resonates with L1, providing boosted sine wave drive. L1 provides substantial step-up, causing about 3500V to appear at its secondary. The capacitors and diodes associated with L1's secondary form a voltage tripler, producing over 10kV across the laser. The laser breaks down and current begins to flow through it. The 47k resistor


L1 = COILTRONICS CTX02-11128
L2 = COILTRONICS CTX150-3-52
Q1, Q2 = ZETEX ZTX849
0.47µF = WIMA 3X 0.15µF TYPE MKP-20
HV DIODES = SEMTECH-FM-50
LASER = HUGHES 3121H-P

COILTRONICS (407) 241-7876

AN65•FI03
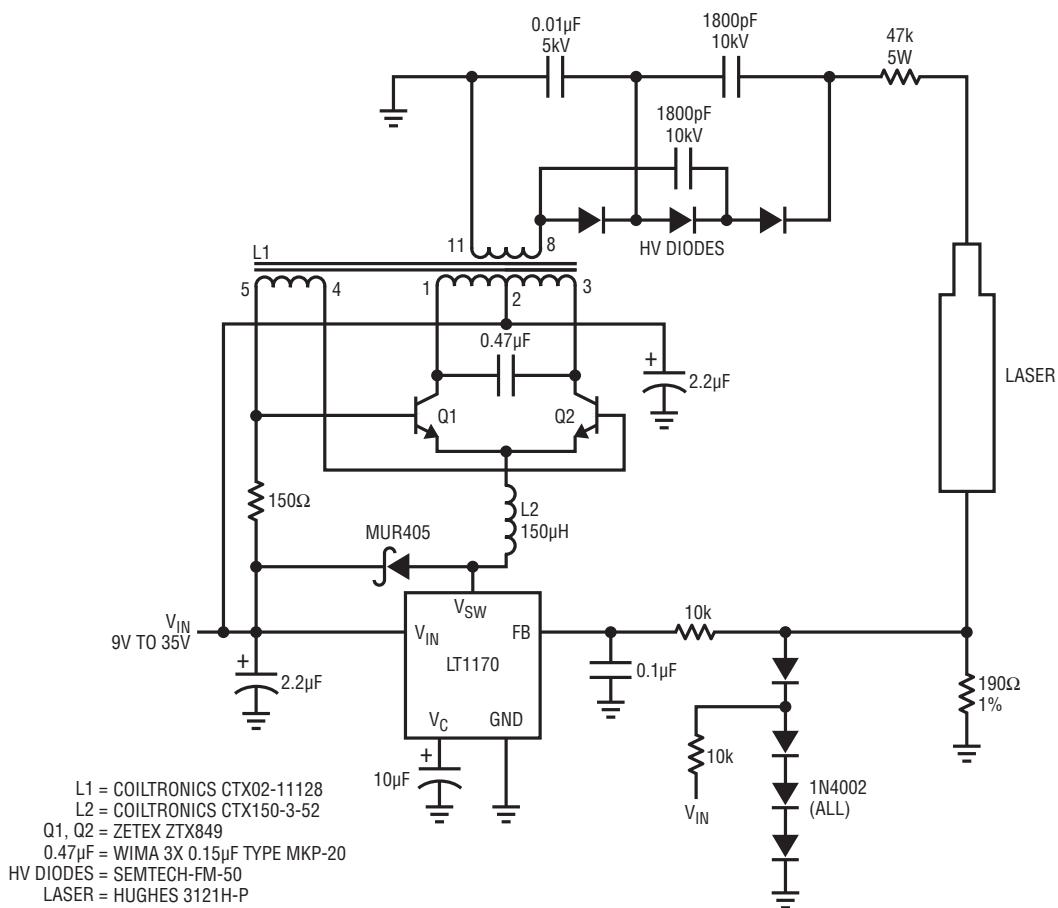
Figure I3. Laser Power Supply, Based on the CCFL Circuit, Is Essentially a 10,000V Compliance Current Source

limits current and isolates the laser's load characteristic. The current flow causes a voltage to appear across the 190Ω resistor. A filtered version of this voltage appears at the LT1170 FB pin, closing a control loop. The LT1170 adjusts pulse width drive to L2 to maintain the FB pin at 1.23V, regardless of changes in operating conditions. In this fashion, the laser sees constant current drive, in this case 6.5mA. Other currents are obtainable by varying the

190Ω value. The 1N4002 diode string clamps excessive voltages when laser conduction first begins, protecting the LT1170. The 10µF capacitor at the V_C pin frequency compensates the loop and the MUR405 maintains L1's current flow when the LT1170 V_{SW} pin is not conducting. The circuit will start and run the laser over a 9V to 35V input range with an electrical efficiency of about 80%.

APPENDIX J

LCD CONTRAST CIRCUITS

LCD panels require variable output contrast control circuits. Contrast power supplies of various capabilities are presented here.

Figure J1 is a contrast supply for LCD panels. It was designed by Steve Pietkiewicz of LTC. The circuit is noteworthy because it operates from a 1.8V to 6V input, significantly lower than most designs. In operation the LT1300/LT1301 switching regulator drives T1 in flyback fashion, causing negative biased step-up at T1's secondary. D1 provides

rectification, and C1 smooths the output to DC. The resistively divided output is compared to a command input, which may be DC or PWM, by the IC's I_{LIM} pin. The IC, forcing the loop to maintain 0V at the I_{LIM} pin, regulates circuit output in proportion to the command input.

Efficiency ranges from 77% to 83% as supply voltage varies from 1.8V to 3V. At the same supply limits, available output current increases from 12mA to 25mA.
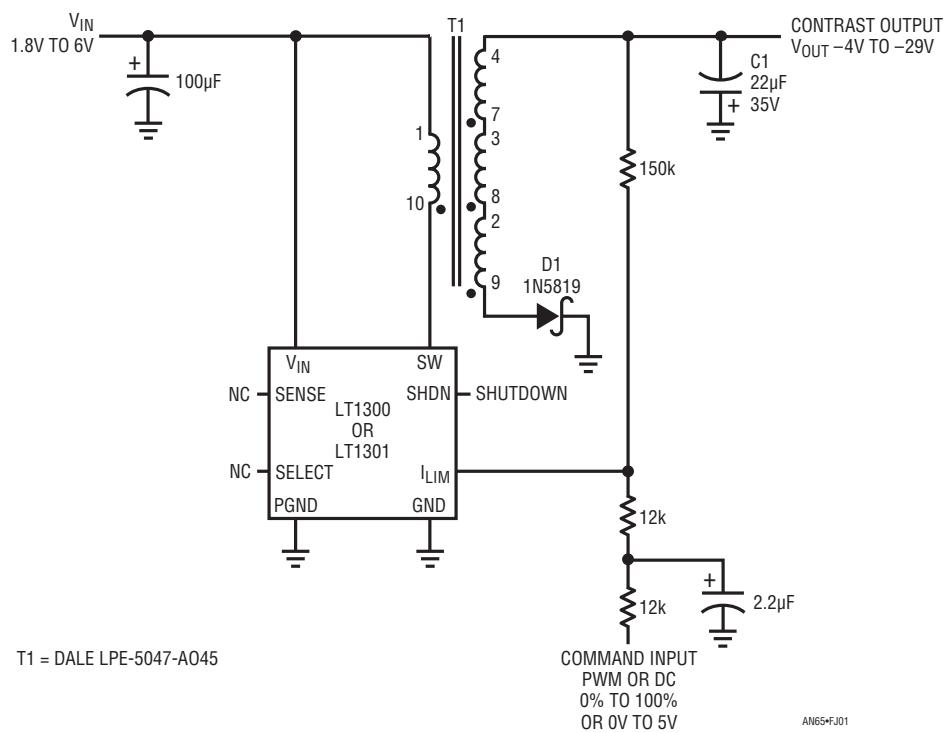

Figure J1. Liquid Crystal Display Contrast Supply Operates from 1.8V to 6V with −4V to −29V Output Range

Another LCD bias generator, also developed by Steve Pietkiewicz of LTC, is shown in Figure J2. In this circuit U1 is an LT1173 micropower DC/DC converter. The 3V input is converted to 24V by U1's switch, L2, D1 and C1. The Switch pin SW1 also drives a charge pump composed of C2, C3, D2 and D3 to generate –24V. Line regulation is less than 0.2% from 3.3V to 2V inputs. Load regulation, although suffering somewhat since the –24V output is not directly regulated, measures 2% from a 1mA to 7mA load. The circuit will deliver 7mA from a 2V input at 75% efficiency.

If greater output power is required, Figure J2's circuit can be driven from a 5V source. R1 should be changed to 47Ω and C3 to 47µF. With a 5V input, 40mA is available at 75% efficiency. Shutdown is accomplished by bringing D4's anode to a logic high, forcing the feedback pin of U1 to go above the internal 1.25V reference voltage. Shutdown current is 110µA from the input source and 36µA from the shutdown signal.

Dual Output LCD Bias Voltage Generator

The many different kinds of LCDs available make programming LCD bias voltage at the time of manufacture
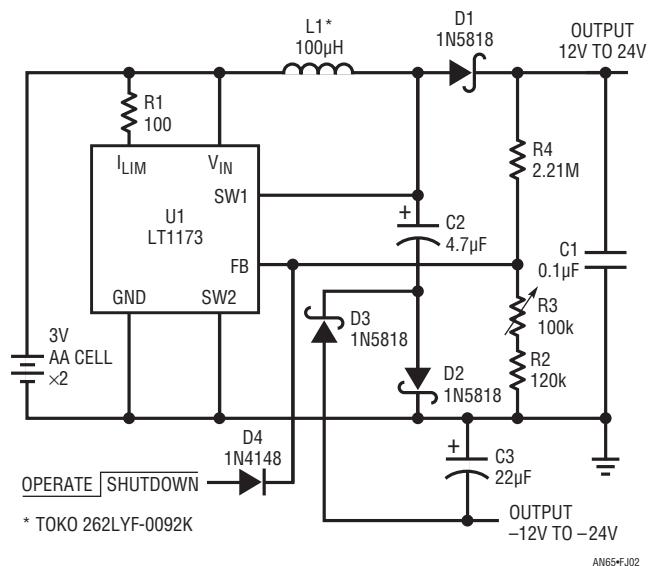

Figure J2. DC/DC Converter Generates LCD Bias from 3V Supply

attractive. Figure J3's circuit, developed by Jon Dutra of LTC, is an AC-coupled boost topology. The feedback signal is derived separately from the outputs, so loading does not affect loop compensation, although load regulation is somewhat compromised. With 28V out, from 10% to 100% load (4mA to 40mA), the output voltage sags about
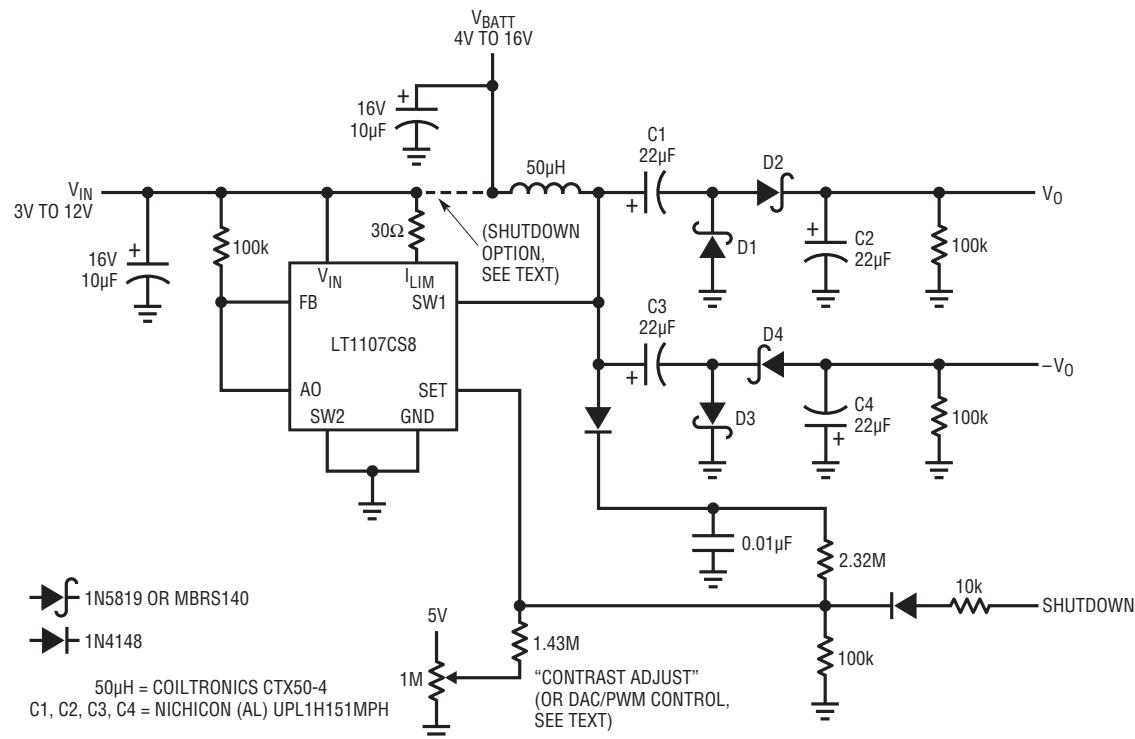

Figure J3. Dual Output LCD Bias Voltage Generator

0.65V. From 1mA to 40mA load the output voltage drops about 1.4V. This is acceptable for most displays.

Output noise is reduced by using the auxiliary gain block within the LT1107 (see LT1107 data sheet) in the feedback path. This added gain effectively reduces comparator hysteresis and tends to randomize output noise. Output noise is below 30mV over the output load range. Output power increases with V_{BATT}, from about 1.4W with $5V_{IN}$ to about 2W with 8V or more. Efficiency is 80% over a broad output power range. If only a positive or negative output voltage is required, the diodes and capacitors associated with the unused output can be eliminated. The 100k resistor is required on each output to load a parasitic voltage doubler created by D2/D4 shunt capacitance. Without this minimum load, the output voltage can rise to unacceptable levels.

The voltage at the Switch pin SW1 swings from 0V to V_{OUT} plus two diode drops. This voltage is AC-coupled to the positive output through C1 and D1, and to the negative output through C3 and D3. C1 and C3 have the full RMS output current flowing through them. Most tantalum capacitors are not rated for current flow. Use of a rated tantalum or electrolytic is recommended for reliability. At lower output currents monolithic ceramics are also an option.

The circuit may be shut down in several ways. The easiest is to pull the Set pin above 1.25V. This approach consumes 200µA in shutdown. A lower power method is to turn off V_{IN} to the LT1107 by a high side switch or simply disable the input supply (see option in schematic). This drops

quiescent current from the V_{BATT} input below 10µA. In both cases V_{OUT} drops to 0V. In the event $+V_{OUT}$ does not need to drop to zero, C1 and D1 can be eliminated. The output voltage can be adjusted from any voltage above V_{BATT} to 46V. Output voltage can be controlled by the user with DAC, PWM or potentiometer control. Summing currents into the feedback node allows downward adjustment of output voltage.

LT118X Series Contrast Supplies

Some LT118X series parts include a contrast supply based on a boost regulator. Figure J4 shows a basic positive output circuit. The V_{SW}-driven inductor provides voltage step-up with D5 and C11 rectifying and filtering the output to DC. The R12/R14 divider chain sets feedback ratio and hence output voltage. The connection to the LT1182 Feedback pin closes a control loop with R7 and C8 providing frequency compensation.

Figure J5 is similar, except that it uses charge pump techniques to reduce shutdown current. D4 and C12 are placed in L3's discharge path, AC coupling it to the output. In shutdown, no DC current can flow through L3, reducing battery drain over J4's DC-coupled approach.

Figure J6's transformer-fed output provides negative output voltages with the LT1183's "FBN" pin directly accepting the resultant negatively biased feedback signal. No level shift is required. In this case output voltage is set by a voltage control input, although potentiometer or PWM inputs could be accommodated (see Appendix F).
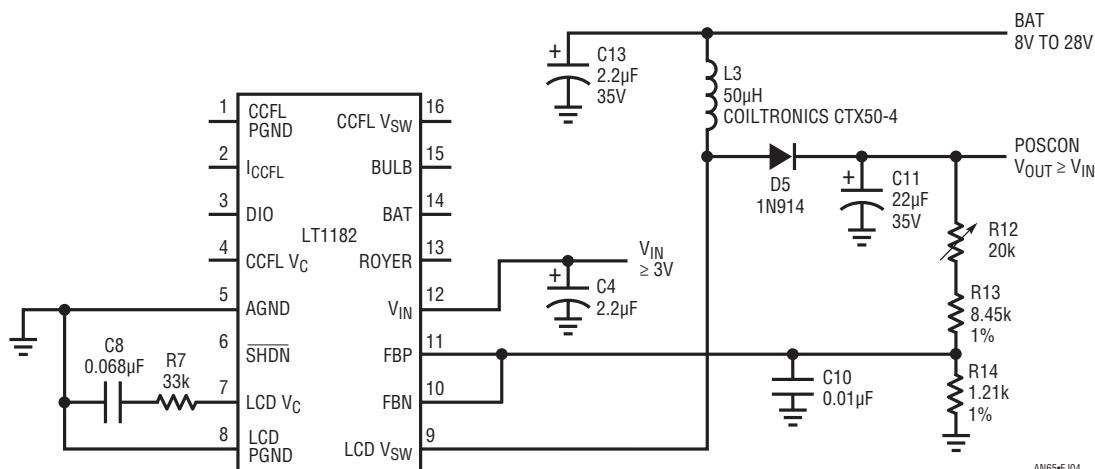

Figure J4. LT1182 LCD Contrast Positive Boost Converter. CCFL Circuitry Is Omitted for Clarity
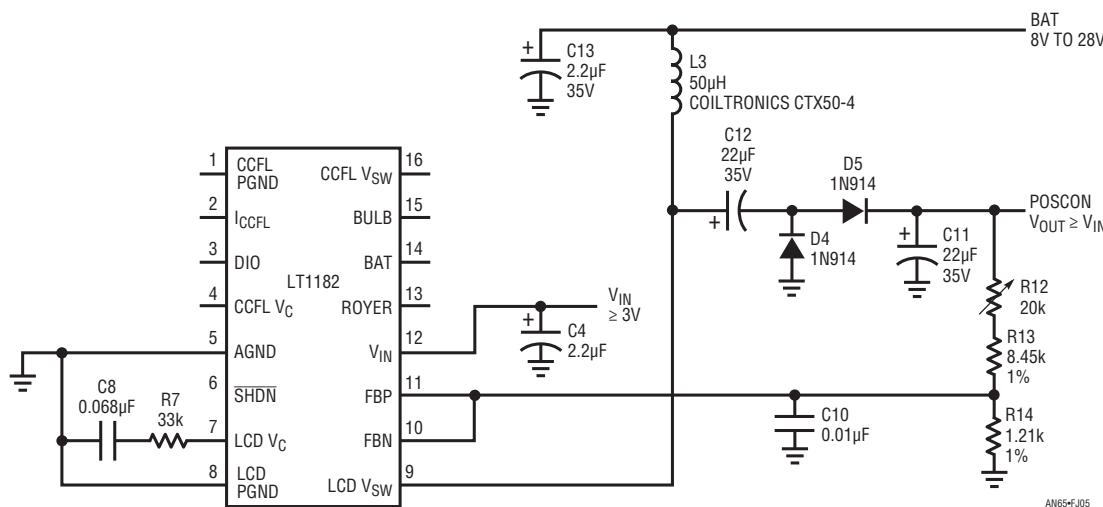
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**Figure J5. LT1182 LCD Contrast Positive Boost/Charge Pump Converter Reduces Battery Current in Shutdown**

C1 MUST BE A LOW LOSS CAPACITOR.
    C1 = WIMA MKP-20 OR PANASONIC ECH-U
    L1 = COILTRONICS CTX210605
    L2 = COILTRONICS CTX100-4
    L3 = COILTRONICS CTX02-12403
    Q1, Q2 = ZETEX ZTX849, ZDT1048 OR ROHM 2SC5001
*DO NOT SUBSTITUTE COMPONENTS
COILTRONICS (407) 241-7876
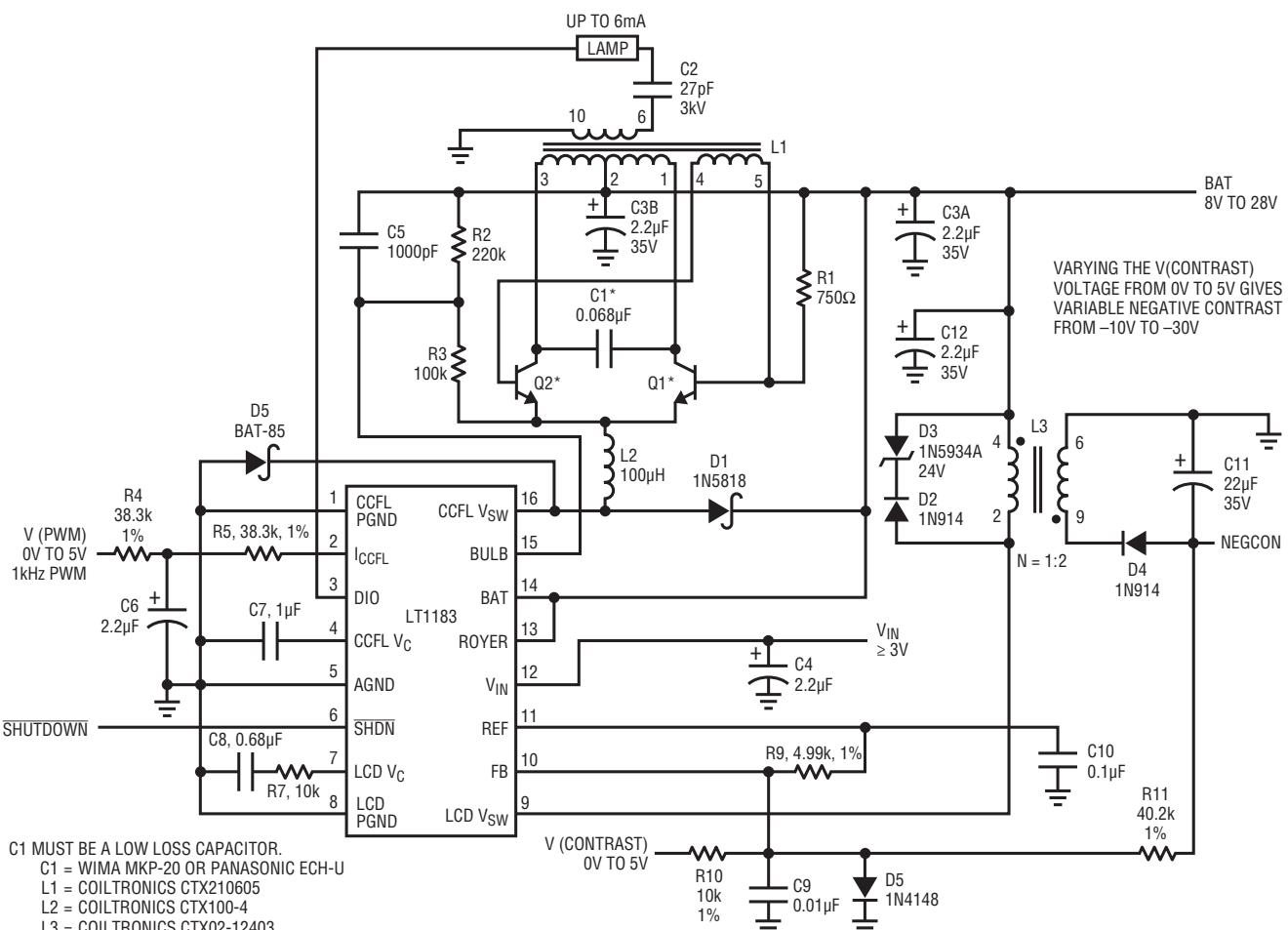
Figure J6. LT1183 Grounded Lamp CCFL Circuit with Negative Output LCD Contrast Supply
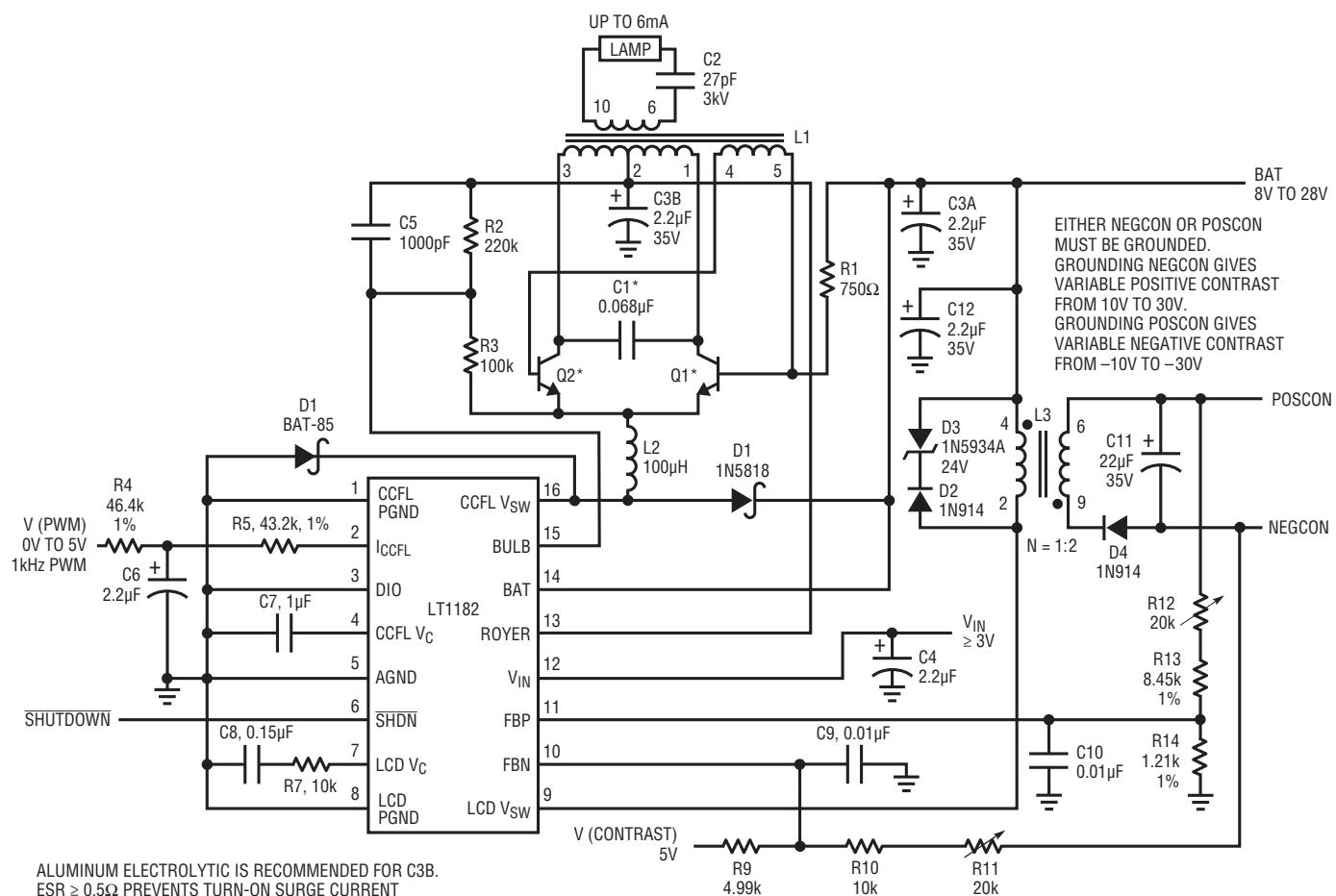
D3 and D2 damp L3 flyback amplitude to safe levels and the isolated secondary permits low shutdown current compared to a simple inductor-based circuit.

Figure J7 takes advantage of the LT1182's bipolar feedback inputs to provide selectable output polarity. This scheme permits the same circuit to be used with LCD's requiring either positive or negative bias. This can be a significant advantage in volume production involving different LCD panels. In operation the circuit is similar to Figure J6, except that L3's secondary winding feeds two separate feedback paths. Output polarity is selected by simply grounding the appropriate L3 secondary terminal.

ALUMINUM ELECTROLYTIC IS RECOMMENDED FOR C3B.
ESR ≥ 0.5Ω PREVENTS TURN-ON SURGE CURRENT
DAMAGE TO THE LT1182 HIGH SIDE SENSE RESISTOR.
C1 MUST BE A LOW LOSS CAPACITOR.
   C1 = WIMA MKP-20, PANASONIC ECH-U
   L1 = COILTRONICS CTX210605
   L2 = COILTRONICS CTX100-4
   L3 = COILTRONICS CTX02-12403
   Q1, Q2 = ZETEX ZTX849, ZDT1048 OR ROHM 2SC5001
*DO NOT SUBSTITUTE COMPONENTS
COILTRONICS (407) 241-7876

Figure J7. LT1182 Floating Lamp CCFL Circuit with Positive or Negative LCD Contrast Supply

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APPENDIX K

WHO WAS ROYER AND WHAT DID HE DESIGN?

In December 1954 the paper "Transistors as On-Off Switches in Saturable-Core Circuits" appeared in *Electrical Manufacturing.* George H. Royer, one of the authors, described a "d-c to a-c converter" as part of this paper. Using Westinghouse 2N74 transistors, Royer reported 90% efficiency for his circuit. The operation of Royer's circuit is well-described in this paper. The Royer converter was widely adopted and used in designs from watts to kilowatts. It is still the basis for a wide variety of power conversion.

Royer's circuit is not an LC resonant type. The transformer is the sole energy storage element and the output is a square wave. Figure K1 is a conceptual schematic of a typical converter. The input is applied to a self-oscillating configuration composed of transistors, a transformer and a biasing network. The transistors conduct out of phase, switching (Figure K2, Traces A and C are Q1's collector and base, while Traces B and D are Q2's collector and base) each time the transformer saturates. Transformer saturation causes a quickly rising, high current to flow (Trace E).

This current spike, picked up by the base drive winding, switches the transistors. This phase opposed switching causes the transistors to exchange states. Current abruptly drops in the formerly conducting transistor and then slowly rises in the newly conducting transistor until saturation again forces switching. This alternating operation sets transistor duty cycle at 50%.

Figure K3 is a time and amplitude expansion of K2's Traces B and E. It clearly shows the relationship between transformer current (Trace B, Figure K3) and transistor collector voltage (Trace A, Figure K3).[1]

**Note 1:** The bottom traces in both photographs are not germane and are not referenced in the discussion.
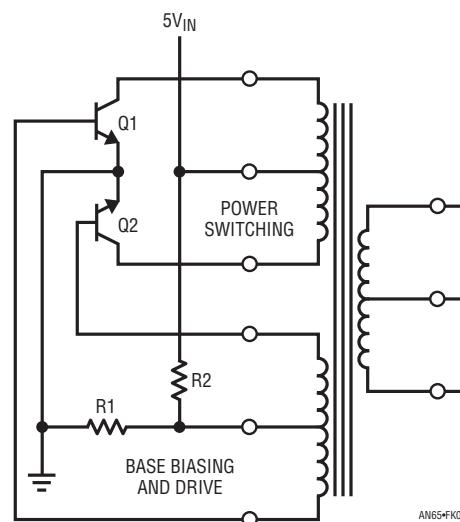
Figure K1. Conceptual Classic Royer Converter. Transformer Approaching Saturation Causes Switching


A = 20V/DIV
B = 20V/DIV
C = 2V/DIV
D = 2V/DIV
E = 5A/DIV

5µs/DIV
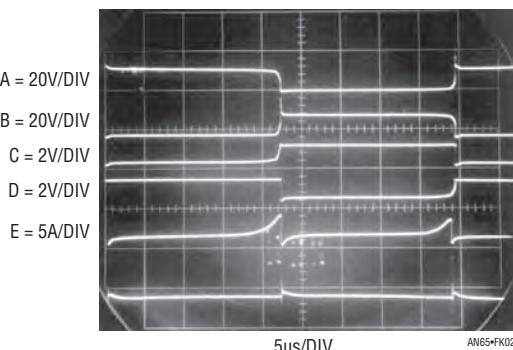
**Figure K2. Waveforms for the Classic Royer Circuit**

A = 10V/DIV

B = 2A/DIV

500ns/DIV
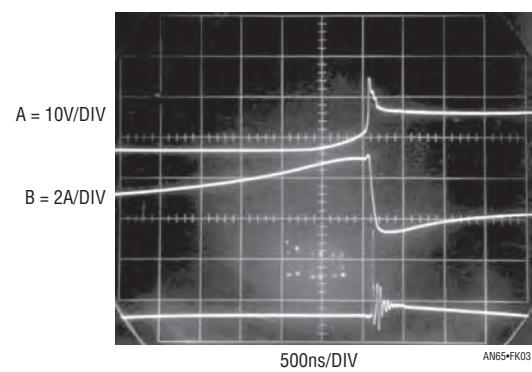
Figure K3. Detail of Transistor Switching. Turn-Off (Trace A) Occurs Just as Transformer Heads into Saturation (Trace B)

APPENDIX L

A LOT OF CUT OFF EARS AND NO VAN GOGHS
Some Not-So-Great Ideas

The hunt for a practical, broadly applicable and easily utilized CCFL power supply covered (and is still covering) a lot of territory. The wide range of conflicting requirements combined with ill-defined lamp characteristics produces plenty of unpleasant surprises. This section presents a selection of ideas that turned into disappointing breadboards. Backlight circuits are one of the deadliest places the author has ever encountered for theoretically interesting circuits.

Not-So-Great Backlight Circuits

Figure L1 seeks to boost efficiency by eliminating the LT1172's saturation loss. Comparator C1 controls a free running loop around the Royer by on-off modulation of transistor base drive. The circuit delivers bursts of high

voltage sine drive to the lamp to maintain the feedback node. The scheme worked, but had poor line rejection due to the varying waveform vs supply seen by the RC averaging pair. Also, the "burst" modulation forces the loop to constantly restart the lamp at the burst rate, wasting energy. Finally, lamp power is delivered by a high crest factor waveform, causing inefficient current-to-light conversion in the lamp and shortening its life.

Figure L2 attempts to deal with some of these issues. It converts the previous circuit to an amplifier-controlled current mode regulator. Also, the Royer base drive is controlled by a clocked, high frequency pulse width modulator. This arrangement provides a more regular waveform to the averaging RC, improving line rejection. Unfortunately, the improvement was not adequate. To avoid
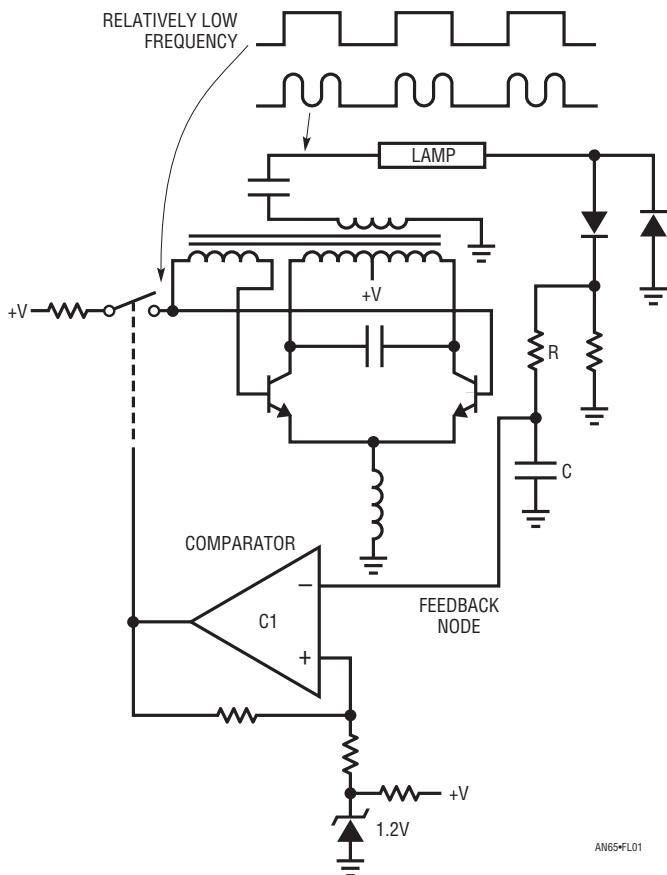

Figure L1. A First Attempt at Improving the Basic Circuit. Irregular Royer Drive Promotes Losses and Poor Regulation
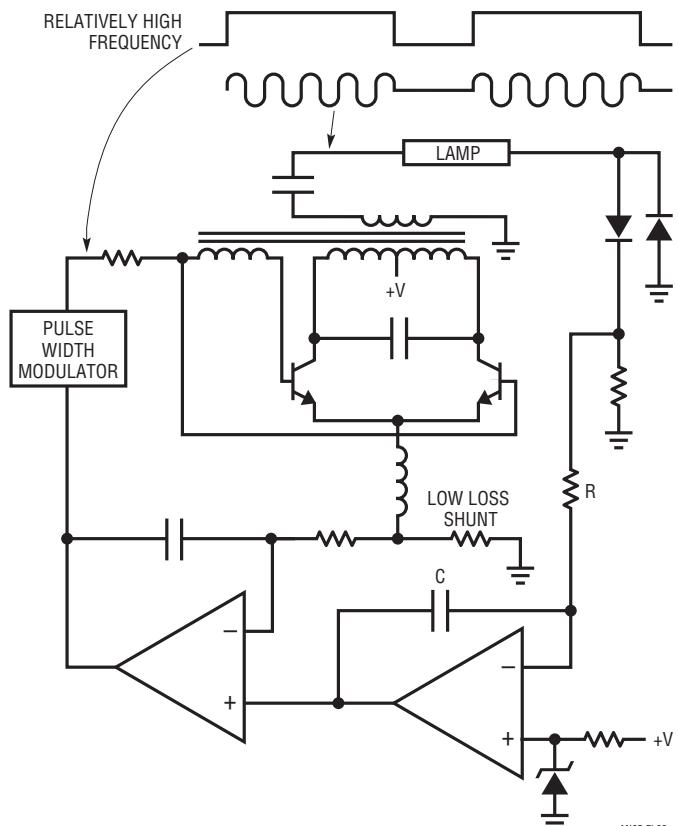

Figure L2. A More Sophisticated Failure Still Has Losses and Poor Line Regulation

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annoying flicker, 1% line rejection is required when the line moves abruptly, such as when a charger is activated. Another difficulty is that, although reduced by the higher frequency PWM, crest factor is still nonoptimal with respect to lamp emissivity and life. Finally, the lamp is still forced to restart at each PWM cycle, wasting power.

Figure L3 adds a "keep alive" function to prevent the Royer from turning off. This aspect worked well. When the PWM goes low the Royer is kept running, maintaining low level lamp conduction. This eliminates the continuous

lamp restarting, saving power. The "supply correction" block feeds a portion of the supply into the RC averager, improving line rejection to acceptable levels.

This circuit, after considerable fiddling, achieved almost 94% efficiency but produced less output light than a "less efficient" version of text Figure 35! The villain is lamp waveform crest factor. The keep alive circuit helps, but the lamp still cannot handle even moderate crest factors and lamp lifetime is still questionable.
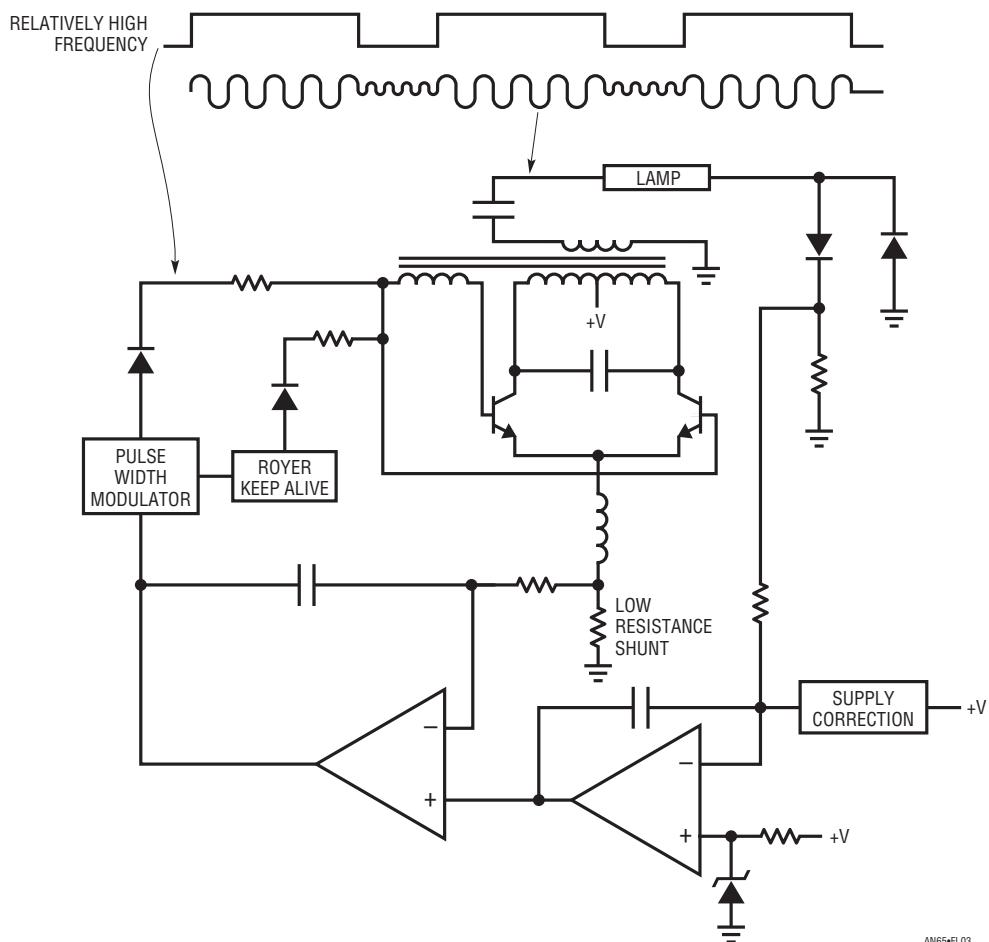

Figure L3. "Keep Alive" Circuit Eliminates Turn-On Losses and Has 94% Efficiency. Light Emission Is Lower Than "Less Efficient" Circuits

Figure L4 is a very different approach. This circuit is a driven square wave converter. The resonating capacitor is eliminated. The base drive generator shapes the edges, minimizing harmonics for low noise operation. This circuit works well, but relatively low operating frequencies are required to get good efficiency. This is so because the sloped drive must be a small percentage of the fundamental to maintain low losses. This mandates relatively large magnetics—a crucial disadvantage. Also, square waves have a different crest factor and rise time than sines, forcing inefficient lamp transduction.

RMS voltage across the shunt (e.g., the Royer current) is unaffected by this, but the simple RC averager produces different outputs for the various waveforms. This causes this approach to have very poor line rejection, rendering it impractical. L6 senses inductor flux, which should correlate with Royer current. This approach promises attractive simplicity. It gives better line regulation but still has some trouble giving reliable feedback as waveshape changes. Also, in keeping with most flux sampling schemes, it regulates poorly under low current conditions.


Figure L4. A Nonresonant Approach. Slew Retarded Edges Minimize Harmonics, but Transformer Size Goes Up. Output Waveform Is Also Nonoptimal, Causing Lamp Losses
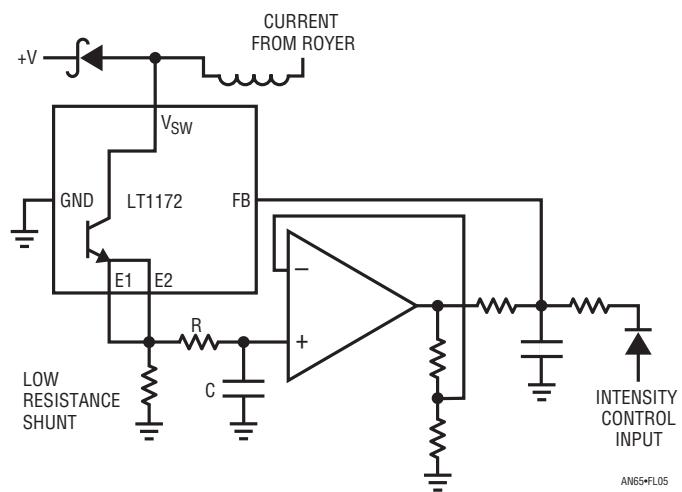
Figure L5. "Bottom Side" Current Sensing Has Poor Line Regulation Due to RC Averaging Characteristics

Not-So-Great Primary Side Sensing Ideas

Various text figures use primary side current sensing to control lamp intensity. This permits the lamp to fully float, extending its dynamic operating range. A number of primary side sensing approaches were tried before the "top side sense" won the contest.

L5's ground-referred current sensing is the most obvious way to detect Royer current. It offers the advantage of simple signal conditioning—there is no common mode voltage. The assumption that essentially all Royer current derives from the LT1172 emitter pin path is true. Also true, however, is that the waveshape of this path's current varies widely with input voltage and lamp operating current. The


Figure L6. Inductor Flux Sensing Has Irregular Outputs, Particularly at Low Currents

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Figure L7 senses flux in the transformer. This takes ad-vantage of the transformer's more regular waveform. Line regulation is reasonably good because of this, but low current regulation is still poor. Figure L8 samples Royer collector voltage capacitively, but the feedback signal does not accurately represent start-up, transient and low current conditions.

Figure L9 is a true, photometrically sensed feedback loop. In theory, it gets around all of the above difficulties

by directly sensing lamp emission and feeding back a representative electrical signal. In practice, it introduces severe drawbacks.

The loop servo controls current to whatever value is required to force lamp emission to the photodiode de-termined point. This eliminates the gradually increasing lamp output (see text Figure 4) at turn-on. Unfortunately, it also forces huge turn-on currents through the lamp for 10 to 20 seconds, greatly shortening lamp life. Typically, the display immediately settles to the final emission point, but turn-on current peaks at four to six times lamp rating. It is possible to clamp or limit this behavior, but a more insidious problem remains.

As the lamp ages its emissivity drops. Typically, a properly driven lamp will drop to 70% of its original emission level after 10,000 hours. In a photometrically sensed loop, the inverter will continually raise lamp current to counteract decreasing emissivity. Although lamp emission remains constant, life is greatly shortened by the continually in-creasing overdrive required to maintain output. This posi-tive feedback enforced degenerative spiral assures rapid, systematic lamp destruction. A five to eight times lamp lifetime reduction in this type of loop has been observed. As before, some form of limiting or 2-loop control scheme can mitigate the undesired characteristics, but advantages would be obviated. Finally, an economical photosensor with well-specified response is elusive.


Figure L7. Transformer Flux Sensing Gives More Regular Feedback, but Not at Low Currents
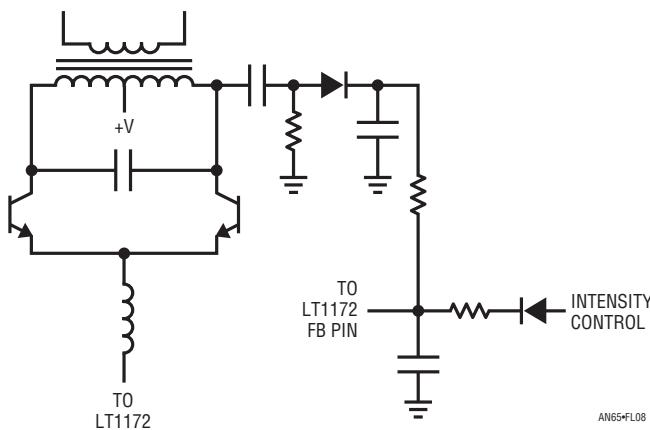

Figure L8. AC-Coupled Drive Waveform Feedback Is Not Reliable at Low Currents
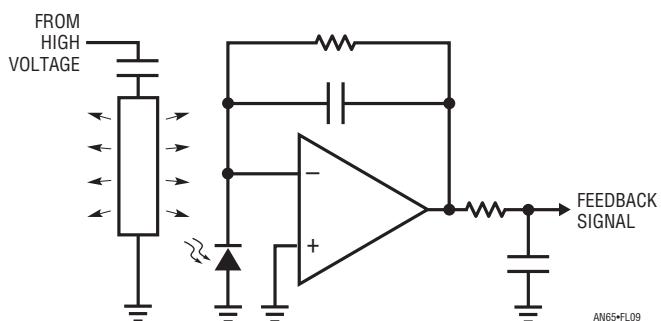

Figure L9. True Optical Sensing Eliminates Feedback Irregularities, but Introduces Systematic Lamp Degradation

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