A Monolithic IC for 100MHz RMS-DC Conversion

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Previous monolithic circuits that converted waveforms to their DC-RMS equivalents utilized logarithmic techniques. This method limits bandwidth to below 1MHz and crest factor performance to about 10:1. Practically speaking, a waveform's RMS value is defined as its heating value in the load. Specialized instruments employ thermally based assemblies that compute the RMS value of the input. The thermal method provides substantially improved bandwidth and crest factor capability compared to logarithmically based converters.

Applications such as wideband RMS voltmeters, RF leveling loops, wideband AGC, high crest factor measurements, SCR power monitoring and high frequency noise measurements require the advantages of thermally based conversion.

Thermal RMS-DC converters are direct acting, thermoelectronic analog computers. The thermal technique is explicit, relying on "first principles." The simple operation permits wideband performance unattainable with implicit, indirect methods based on logarithmic computing.

Previously, thermally based converters were large and expensive to produce. A new IC, the LT1088, brings the advantages of thermal conversion to the circuit board in a 14 pin DIP, and at reasonable cost. Before discussing the LT1088, it is worthwhile reviewing thermal RMS-DC conversion.

Figure 1 shows a conceptual thermal RMS-DC converter. The input waveform warms the heater, resulting in increased output from the temperature sensor. The heating is related to the RMS value of the input waveform. The temperature sensor's DC output represents this heating.

Although simple, this method has some problems. The temperature sensor cannot distinguish between signal and ambient induced temperature changes. This issue could be addressed by summing in ambient temperature information, but a more significant problem remains. Even if the electrical portions of the design are perfectly linear, overall response is not. The power produced by the heater is non-linearly related to the input voltage ($P = I^2R$); hence temperature rise is similarly non-linearly proportioned.

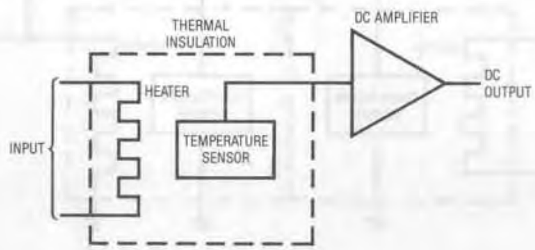

Figure 1. Conceptual RMS-DC Converter (with some problems)

Additional non-linear signal conditioning is necessary for an output which linearly corresponds to the input voltage.

Figure 2 shows a classic scheme which corrects both of Figure 1's deficiencies. Here, the 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 non-linear, the input-output 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, ambient temperature terms will affect the pairs equally in phase and amplitude. The DC amplifier will reject 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 2's output is linear because the matched thermal pair's non-linear voltage-temperature relationships cancel each other.

The advantages of this approach have made its use popular in thermally based RMS-DC measurements. Typically, the assembly is composed of matched heater resistors, sensors and thermal insulation. These assemblies are relatively large and expensive to produce. In theory, monolithic IC techniques can be used to replace such assemblies, but the thermal insulation requirements present problems.

A simplified monolithically based circuit which accomplishes this function appears in Figure 3. It is quite similar to Figure 2's generalized approach. Here, the input drives R1, producing heating which lowers the value of D1's voltage. A1 responds by driving R2 to heat D2, closing a loop around the amplifier. Because the transistors and resistors are matched, A1's DC output equals the RMS value of the input, regardless of input frequency or waveshape. The aforementioned thermal terms limit the circuit's practical performance. In particular, thermal cross-coupling between the R1-D1 and R2-D2 pairs degenerates gain, degrading available signal. Also, differences in the dissipation constants and thermal capacity of the R-D pairs

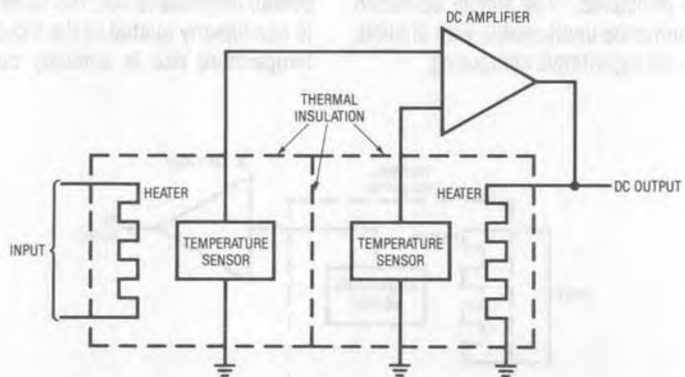

Figure 2. (A Better) Conceptual Thermal RMS-DC Converter

result in overall gain errors. Additionally, thermal resistance to ambient must be high to maximize D1-D2 signal output for reasonable input drive. Finally, the thermal path between the mated resistor-transistor pairs must be designed for efficient, low loss heat transfer.

Although the converter's basic principle is a straightforward extension of Figure 2, the electro-thermal design must be carefully addressed to produce a practical monolithic circuit. These thermal considerations dominate the design and form of the circuit.

Figure 4 shows a simple electro-analog of the thermal terms in the converter. The overall lumped matching of these terms heavily influences achievable performance. In particular, the die attach thermal resistance dominates the thermal impedance path. If this resistance is made very high, the effects of mismatch in the other terms are minimized.

Thermal cross-coupling is almost entirely eliminated by using separate, identical die for the diode-heater pairs. This eliminates cross heating more effectively than any possible single die approach. A gain error, which is corrected by introducing a corresponding gain trim, is caused by residual mismatch in thermal terms. These include die size, dissipation constant, and thermal capacity differences. The most significant term is differing amounts and distribution of the die attach material. The gain correcting trim is introduced by altering the gain of the output stage in Figure 3.

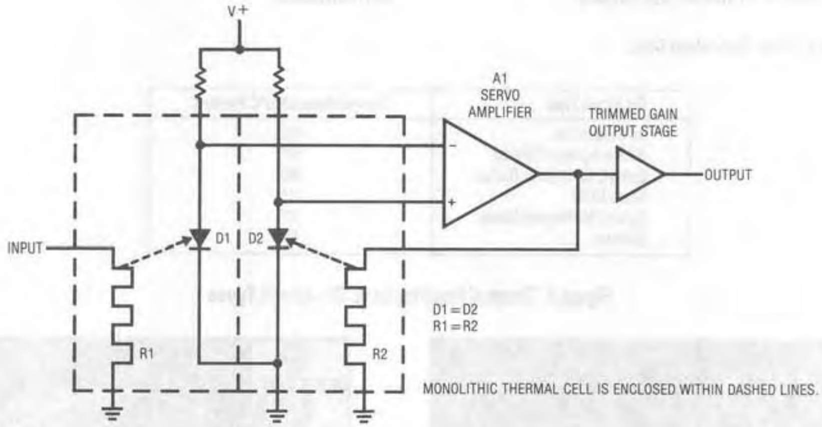

Figure 3. Simplified Monolithic Thermal RMS-DC Converter

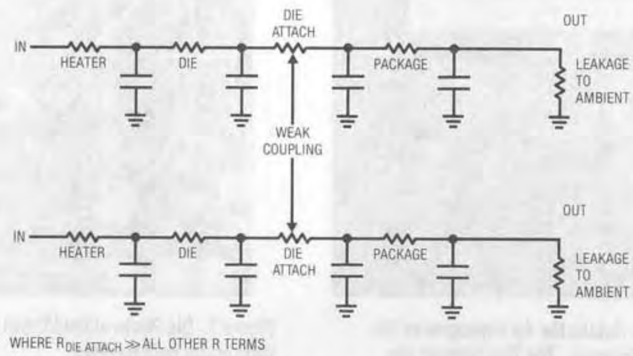

Figure 4. Simplified Electro-Analog of the LT1088

Because the die attach resistance is so important (see Box Section, "Measuring Thermal Resistance") it must be carefully considered. Figure 5 shows results for various die attach methods. As might be suspected, die suspended in free air offer the highest thermal resistance, although the approach is impractical. Conversely, standard eutectic bonding gives low thermal resistance but is easy to produce. Another die attach method, air impregnated polymer, is nearly as good as air suspension, and is practical. The OTT™ process (Oracular Thermal Transfer) was developed to allow use of air impregnated polymer die attach. Figure 6 is a side-on die photo showing the results of OTT processing beneath the die. Large areas beneath the die are filled with air, resulting in the high thermal resistance noted in Figure 5. Sufficient amounts of polymer attach material ensure a reliable die attach.

Figure 7 is a die shot of one heater resistor-diode pair of the LT1088. The circular, concentric heater resistors promote evenly distributed, isothermal characteristics. The placement and aspect ratio of the heater rings is optimized for an even thermal flow across the die. The sensing diode is actually a paralleled quad located symmetrically about the die center. This quad arrangement provides improved temperature sensing characteristics over a single device. The separate heater rings allow the user to select either a 50Ω or 250Ω input. The test structure in the die center is not used. It is designed to offset effects described by Counts Theorem (see References). Note that the IC contains only the basic thermal components to maintain isothermal conditions. Inclusion of support circuitry would add thermally based error terms, degrading performance.

Die Attach Type	Thermal Resistance°C Per Watt
Air Suspended	460
Air Impregnated Polymer	300
Epoxy/2 Mil Polymer Barrier	250
Glass-10 Mil	115
Epoxy/1 Mil Polymer Barrier	107
Eutectic	54

Figure 5 Thermal Resistance vs Die Attach Types

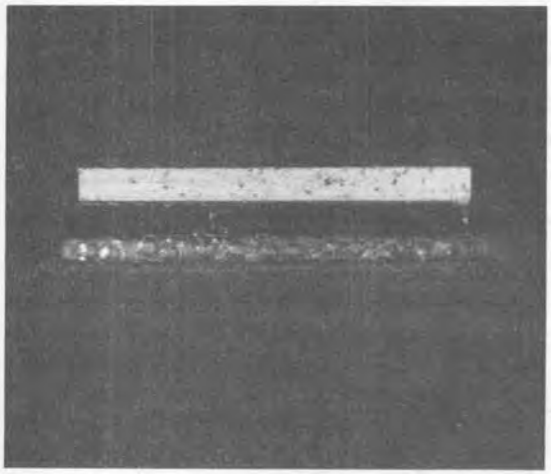

Figure 6. Side-On Chip Photo Details the Air Impregnated Die Attach Produced by the OTT Process. The Two Distinct Die Regions are Caused by Scribe and Break Operations.

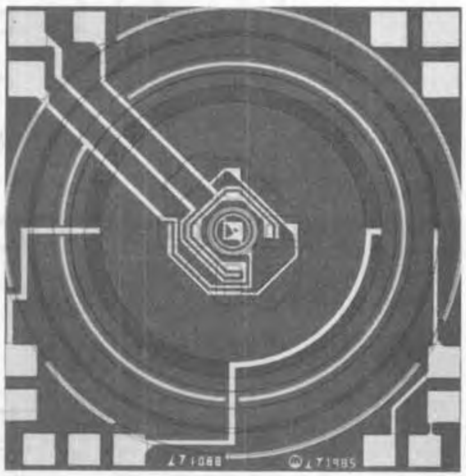

Figure 7. Die Photo of the LT1088 Showing 50Ω (inner ring) and 250Ω (outer ring) Heaters

Figure 8 shows a detailed circuit using the LT1088. Typical performance specifications are given in Figure 9. The LT1088's temperature sensing diodes are biased from the supply. A1, set up as a differential servo amplifier with a gain of 9000, extracts the diode's difference signal and biases Q1. Q1 drives one of the LT1088's heaters, completing a loop. The 330pF capacitor gives a stable roll-off. The 1.5M-0.022μF combination improves settling by reducing

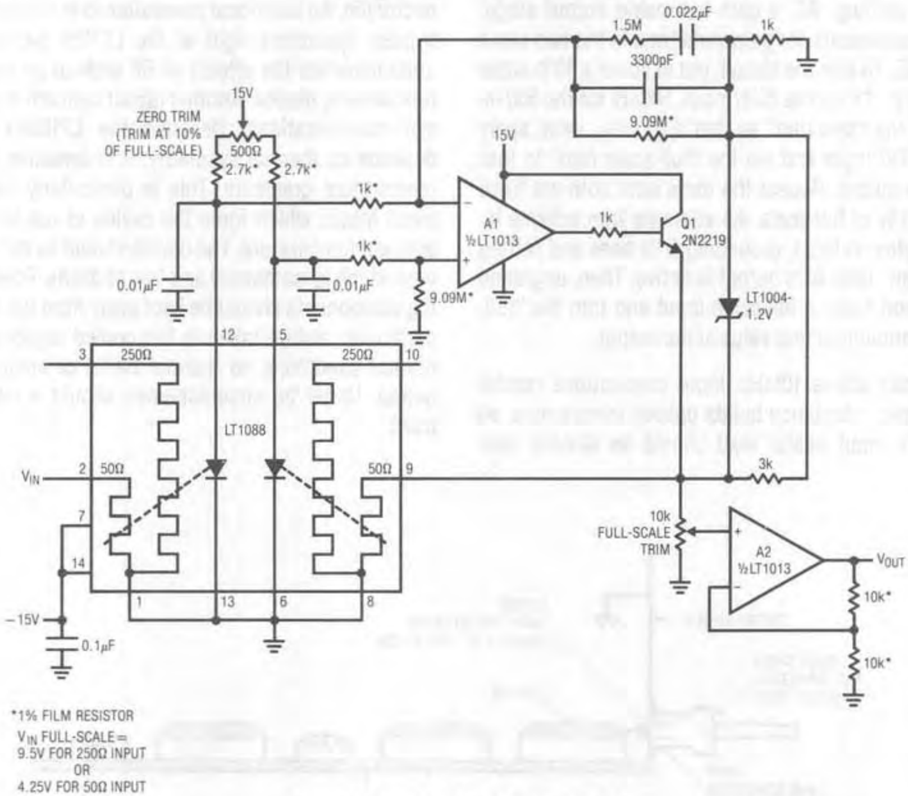

Figure 8. Practical Circuit for the LT1088

Accuracy: 50Ω Input DC to 50MHz .... 1% FS DC to 100MHz .... 2% FS 250Ω Input DC to 20MHz .... 1% FS Temperature Effect on Accuracy .... 100ppm/°C Dynamic Range 20:1	Crest Factor: 50Ω Input .... 50:1 250Ω Input .... 40:1 3dB Bandwidth .... 300MHz Full-Scale Settling Time (1%) .... 500ms Input Voltage Range 50Ω Input .... 4.25V 250Ω Input 9.5V

Figure 9. Typical Specifications for Figure 8's Circuit

gain during output slew. The square-law thermal gain of the LT1088 means overall loop gain is lower for small inputs. Normally, this would result in slow settling for values below about 10-20% of scale. The LT1004 1k-3k network is a simple breakpoint, boosting amplifier gain in this region to improve settling. A2, a gain trimmable output stage, serves to compensate for gain variations in the two sides of the LT1088. To trim the circuit, put in about a 10% scale DC signal (e.g., 1V for the 250Ω input, 500mV for the 50Ω input). Adjust the "zero trim" so that $V_{OUT} = V_{IN}$. Next, apply a full-scale DC input and set the "full-scale trim" to that value at the output. Repeat the trims until both are fixed well within 1% of full-scale. An alternate trim scheme involves applying no input, grounding Q1's base and setting the "zero trim" until A1's output is active. Then, unground Q1's base and apply a full-scale input and trim the "full-scale" adjustment for that value at the output.

At frequencies above 10MHz, input connections require care. Parasitic inductance builds quickly in wire runs, so the LT1088's input heater lead should be *directly* connected to the source to be measured. It is also wise to shield the input line from the rest of the circuit. Figure 10 shows one way to do this. A simple copper RF shield isolates the circuitry from the input. The LT1088 is mounted so the input pin is as close as possible to the input connector pin. An additional precaution is to mount the 0.01μF bypass capacitors right at the LT1088 package. These units minimize the effects of RF pick-up by the temperature sensing diodes. Another layout concern involves thermal considerations. Because the LT1088's operation depends on thermal symmetry, it is sensitive to external temperature gradients. This is particularly the case for small inputs which force the device to run very close to ambient temperature. The device should be mounted in an area which is isothermal and free of drafts. Power generating components should be kept away from the LT1088 and particular caution taken in fan cooled equipment. Under normal conditions no thermal baffle or enclosure is required. Under no circumstances should a heat sink be used.

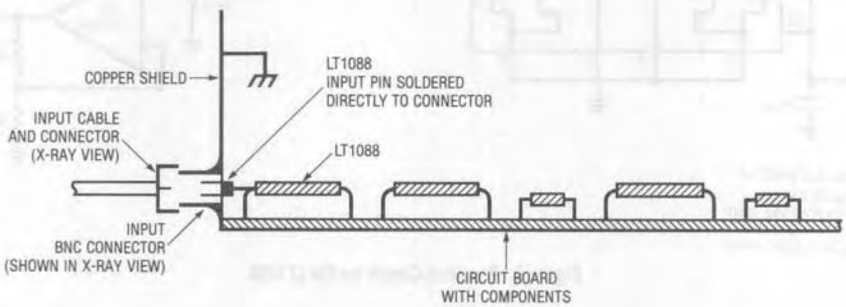

Figure 10. Recommended Layout

Figure 11 is a plot of output error vs input frequency using the 50Ω input at full-scale. There is no degradation of bandwidth for smaller inputs. Thermal transfer standards (Fluke Model 540B with A-55 converters) certified to 50MHz were used as references. The data above 50MHz was also taken with these references, although the individual units used had not been certified at these frequencies. The accuracy of units of this type which have been certified is normally inside the tolerances listed, so there is good probability the data is valid. Figure 12 is a similar plot but over extended ranges of frequency and flatness. Unfortunately, equipment and test set-up limitations imposed uncertainties at the highest frequencies, but the data probably approximates actual performance. The peaking followed by the steep roll-off is most likely due to the LT1088's high frequency limitations. In particular, bond wire inductance becomes increasingly significant as frequency increases. Also, capacitance between the heater and sensing diode permits RF pumping of the diode, almost certainly causing deleterious results. The 1% error point using the 250Ω range is lower. The parasitics described combine with the higher input voltage swing to limit 1% bandwidth to about 20MHz.

The low end of the frequency spectrum is limited by loop time constants. For Figure 8's values, the circuit begins to follow the input below about 50Hz. Lower frequency operation requires longer loop time constants (e.g., increasing the 3300pF value), increasing settling time.

Crest factor performance is set by IC breakdown limits and the usable low input power range. Breakdown limits are a function of processing. The usable low input power range is a basic signal-to-noise conflict. Low input power produces small amounts of signal. This makes accurate, stable discrimination between desired inputs and ambient thermal phenomena uncertain and noisy.

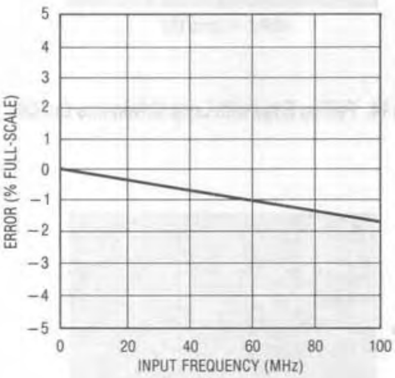

Figure 11. Error vs Frequency

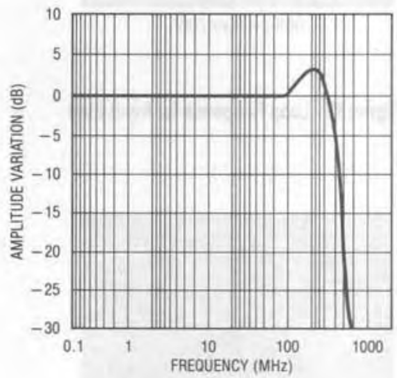

Figure 12. Extended Plot of LT1088 Frequency Response

Step response is determined by the servo electronics roll-off. Figure 13 shows response for a full-scale step into the 250Ω heater. Loop response is nicely damped. The small glitching at the beginning of the step is due to the gain breakpoint in A1's feedback loop. Figure 14 shows the negative going step. Although the response appears clean (again, the gain glitch is due to A1's breakpoint network), Figure 15 reveals the loop coming to a new value well after it appears to have settled. This photo is a slower version of Figure 14 (initial negative going step is just visible at the extreme left). Almost 5 seconds after settling apparently occurs the loop abruptly assumes a new value. This effect is due to the loop being driven into saturation. The allowable low range operating area (defined by the dynamic range specification) has been exceeded, forcing the servo into saturation. This causes thermal imbalances in the LT1088, resulting in the extended length of time before the loop becomes active again.* Figure 16 shows response when the input is kept within the specified operating range. Settling on both edges is clean, and the loop quickly assumes and maintains its final value.

*As an interesting exercise, consider what would happen if the servo amplifier could somehow extract, as well as supply, heat to the system.

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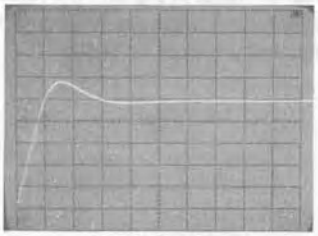
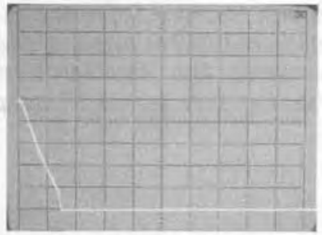


HORIZ = 50ms/DIV

Figure 13. Loop Response for Input Step

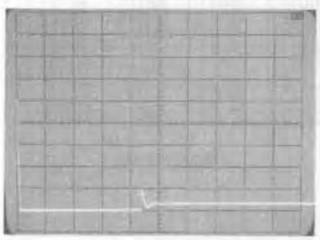
A = 2V/DIV



HORIZ = 50ms/DIV

Figure 14. Falling Edge with Loop Driven into Cut-Off

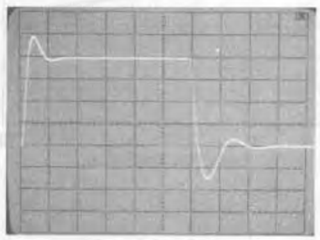
A = 2V/DIV



HORIZ = 1sec/DIV

Figure 15. Loop Recovery from Cut-Off

A = 2V/DIV



HORIZ = 200ms/DIV

Figure 16. Loop Response with Input Kept in the Linear Region

Some applications may drive the LT1088 outside its operating range, forcing thermal imbalance. If fast settling is required, Figure 17's circuit is useful. This scheme speeds settling by applying an open loop heating correction when thermal imbalance occurs. When the input (Trace A, Figure 18) steps negatively, A1's output (Trace B) slews. Loop delays cause A1 to overshoot, turning off heater drive and thermally imbalancing the LT1088. Diode steering at the LT1010 buffer's output sinks current from the 250Ω input heater (Trace C). This produces heating, tending to compensate the thermal imbalance, thereby decreasing settling time. Trace D, A2's output, settles fairly quickly after it recovers from the open loop correction.

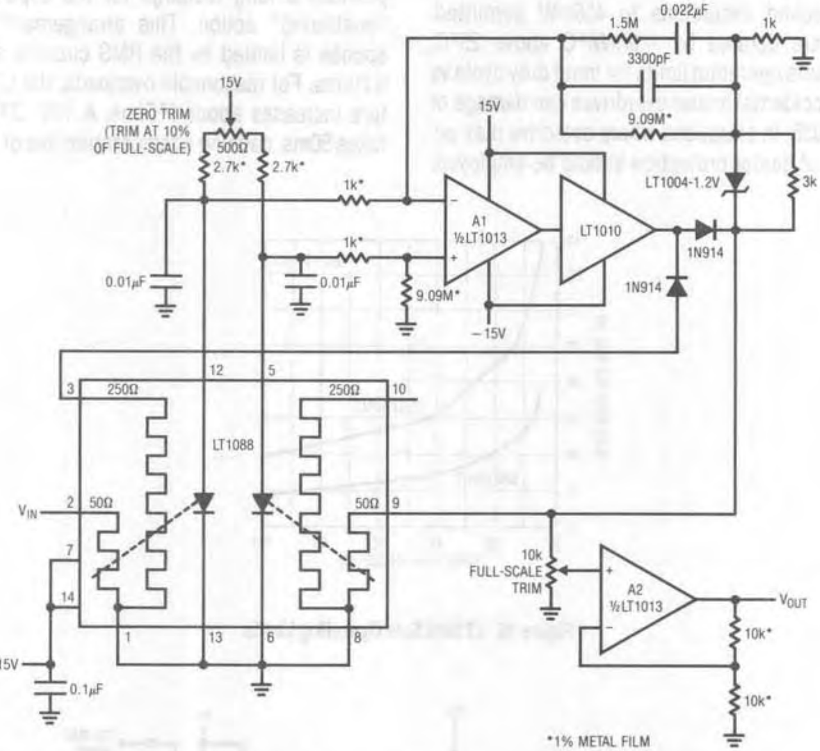


Figure 17. Fast Recovery Circuit (diode steering provides compensation for thermal imbalance)

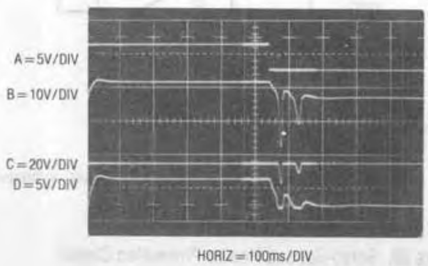


Figure 18. Waveforms for Fast Recovery Circuit

Any study of LT1088 dynamics must consider that although the output voltage has settled, internal die temperature may still be moving towards final value. It is important to distinguish between voltage and thermal dynamics when observing LT1088 operation.

Most LT1088 failures will be caused by excessive heater drive. Input power (25°C) is specified at 375mW continuous with 30 second excursions to 435mW permitted. These figures are derated by −3mW/°C above 25°C. Figure 19 plots safe operating limits for input duty cycle vs input voltage. Accidental heater overdrives can damage or destroy the LT1088. In situations where overdrive may occur, some form of heater protection should be employed.

Figure 20 shows a heater protection circuit which responds quickly enough to prevent damage from most overloads. C1's input is connected to the output of the LT1088 servo circuit. If the LT1088 circuit's output exceeds the threshold at C1's other input, C1 trips, discharging the 2μF capacitor. This causes C2's output to go low, energizing the relay and breaking the heater circuit. The 560k resistor provides a long recharge for the capacitor, preventing "chattering" action. This arrangement's speed of response is limited by the RMS circuit's slew rate, about 0.2V/ms. For reasonable overloads, the LT1088's temperature increases about 1°C/ms. A 10V LT1088 output step takes 50ms, causing a temperature rise of about 50°C.

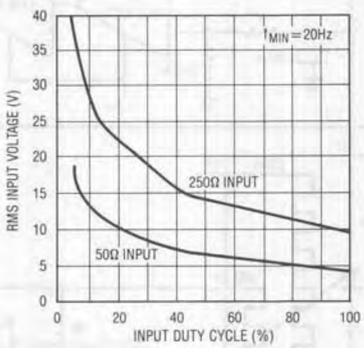


Figure 19. LT1088 Safe Operating Limits

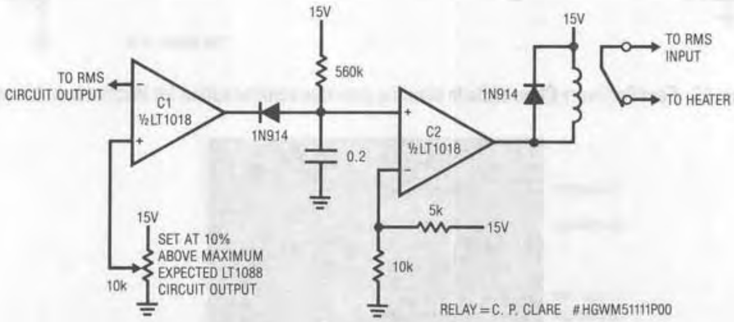


Figure 20. Servo-Sensed Heater Protection Circuit

Severe overloads will cause faster heating, and this circuit may not act quickly enough to prevent damage. Figure 21's circuit is faster, but requires a trim. It works by directly sensing the temperature of the LT1088, instead of the servo amplifier output. Circuit action is similar to Figure 20, except that the input is taken from the LT1088's input temperature sensing diode. Figure 22 shows waveforms. Excessive drive to the LT1088 (Trace A, Figure 22) forces the servo amplifier (Trace B) into slew. The sensing diode, responding more quickly than the servo, causes the circuit to switch the relay (Trace C), removing heater drive in 15ms. Because loop response lags temperature, the servo amplifier's output peaks at only 6V, about one-third of the input step. This circuit has the disadvantage of requiring a trim, due to initial diode tolerances. To trim, measure diode output at 25°C and set C1's negative input to the desired temperature cut-off point. Assume a diode slope of 1.8mV/°C.

Some applications may require buffering the LT1088's relatively low input impedance. This is not easy if the device's wide bandwidth and accuracy must be preserved. With an LT1010 buffer, bandwidths in the low megahertz region are achievable. Figure 23's circuit, a FET input, complementary emitter follower output design, extends bandwidth out to 25MHz. If gain is desired, Figure 24 furnishes low megahertz performance at a gain of ten. Figure 25's design, although complex, has 32MHz bandwidth at a gain of ten. Detailed discussion of Figures 24 and 25 appears in Application Note 21, "Composite Amplifiers." Figure 26's table summarizes performance of the buffer amplifiers.

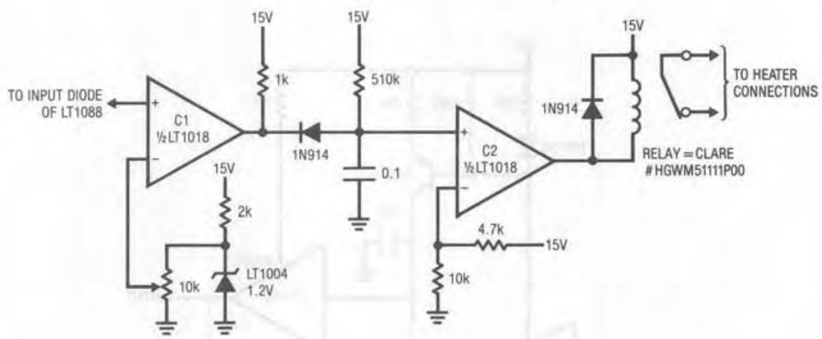


Figure 21. Diode Sensed Heater Protection Circuit

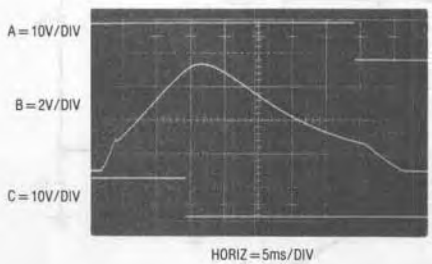


Figure 22. Response of Diode Sensed Heater Protection Circuit

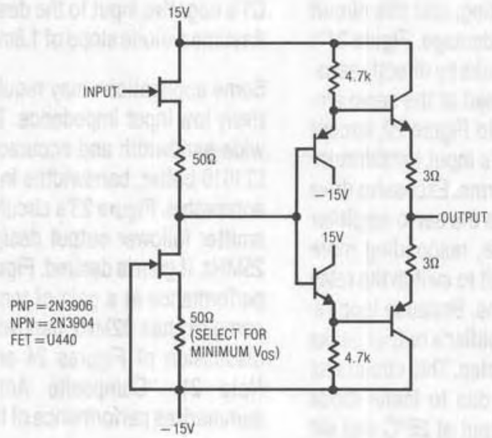


Figure 23. Input Buffer for the LT1088

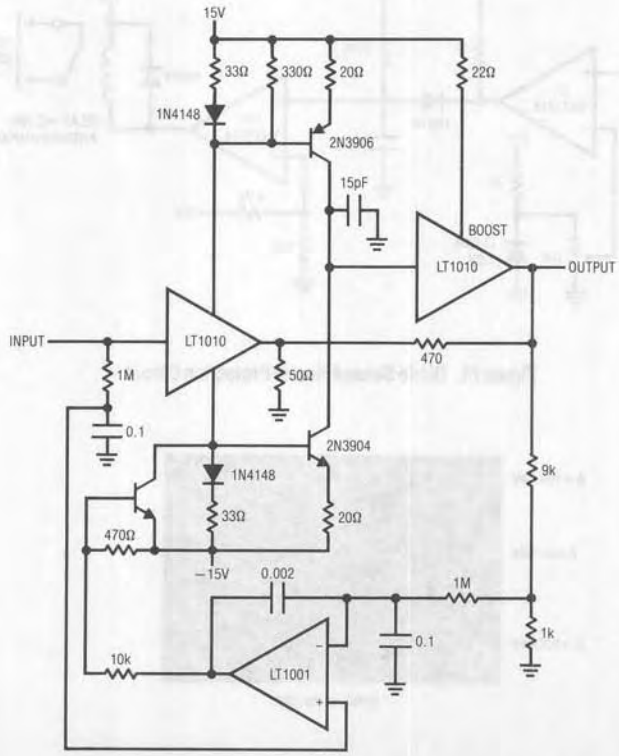


Figure 24. Fast, "Current Mode" Amplifier-Buffer

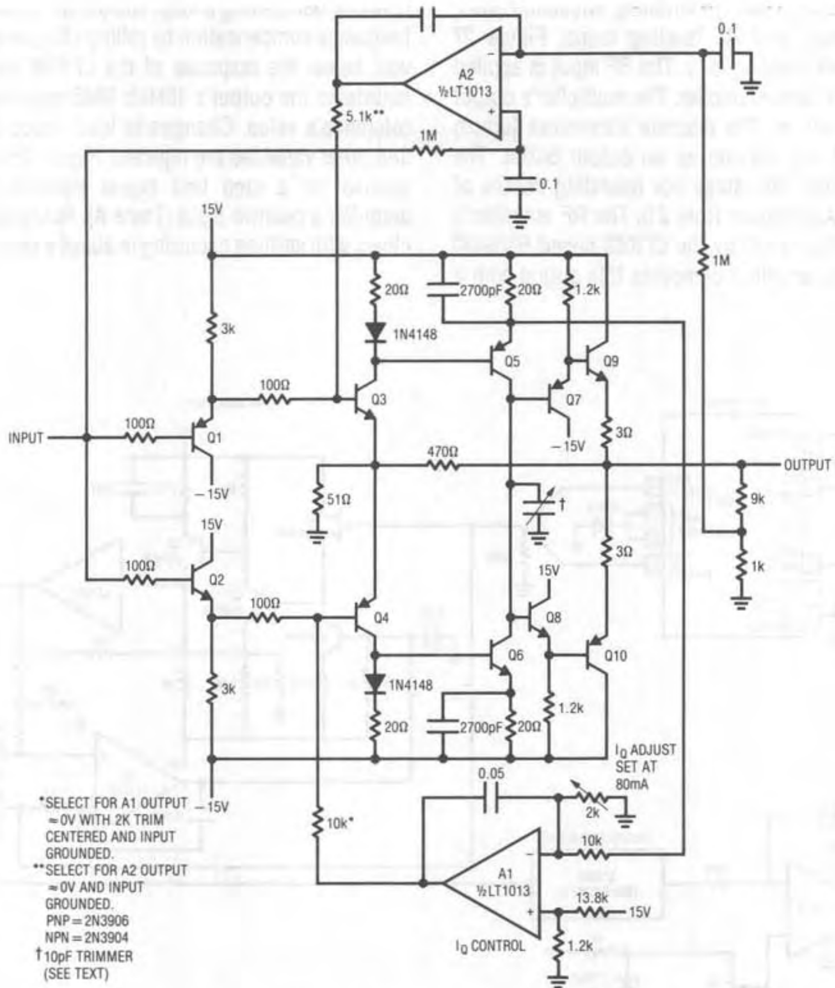


Figure 25. Ultra-Wideband "Current Mode Feedback" Buffer Amplifier

Type of Buffer	Slew Rate	1% Error Bandwidth 250Ω Load (±10V_{OUT})	1% Error Bandwidth 50Ω Load (±5V_{OUT})
Discrete—A = 10	3000V/μs	25MHz	32MHz
LT1010 Based—A = 10	100V/μs	0.75MHz	2MHz
Discrete—A = 1	2000V/μs	15MHz	25MHz
LT1010 Based—A = 1	100V/μs	0.75MHz	2MHz

Figure 26. Summary of Buffer Characteristics

Applications for the LT1088's wideband capability exist in AC voltmeters, SCR power monitoring, wideband AGC, noise measurement and RF leveling loops. Figure 27 shows a 10MHz RF leveling loop. The RF input is applied to the AD539 wideband multiplier. The multiplier's output drives an RF amplifier. The discrete transistors furnish gain with the LT1010 serving as an output buffer. The LT1012 DC stabilizes the stage (for operating details of this circuit, see Application Note 21). The RF amplifier's output is converted to DC by the LT1088 based RMS-DC converter. A servo amplifier compares this output with a settable DC reference and biases the multiplier's control channel, completing a loop. The 0.33μF capacitor provides frequency compensation by rolling off gain at a frequency well below the response of the LT1088 servo. The loop maintains the output's 10MHz RMS amplitude at the DC reference's value. Changes in load, input, power supply and other variables are rejected. Figure 28 shows loop response for a step test signal injected at the servo amplifier's positive input (Trace A). Response (Trace B) is clean, with settling occurring in about a second.

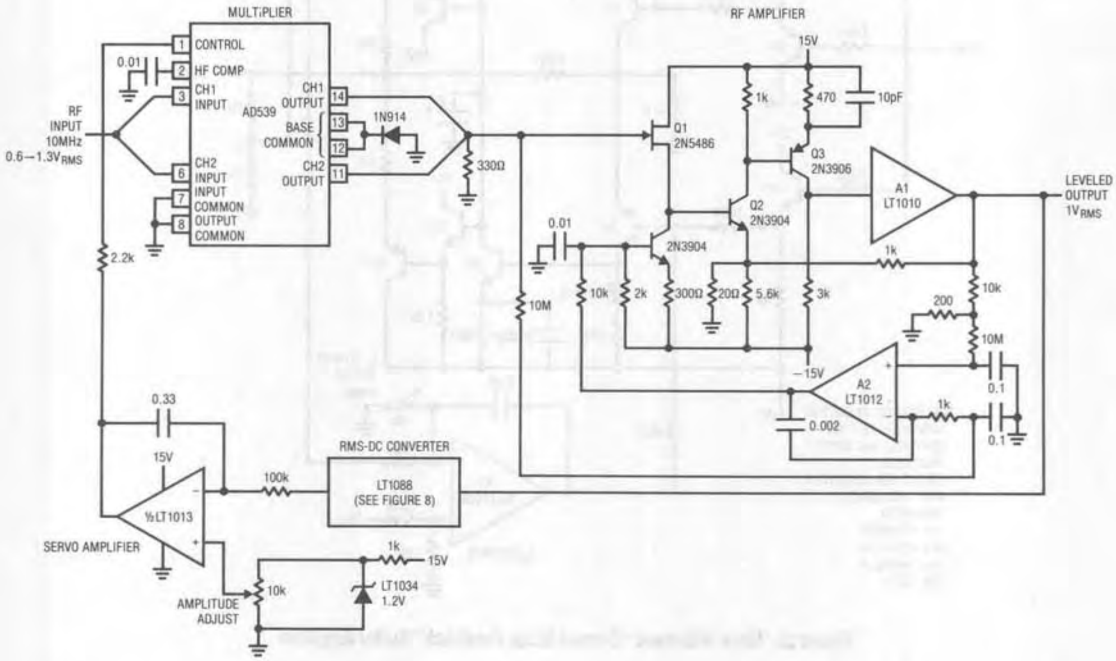


Figure 27. RF Leveling Loop

Figure 28. RF Leveling Loop Response

BOX SECTION

Measuring Thermal Resistance

Establishing, maintaining and verifying proper thermal resistance between the LT1088 and ambient is extremely important to its operation. Because thermal resistance is a gain term, it must be constant with time, cycling and power level. Additionally, it must fall within limits. Low thermal resistance results in poor sensitivity to downscale inputs, limiting dynamic range. Too high a thermal resistance causes excessive die heating, leading to failure.

Thermal crosstalk, the heat conduction between the two die, must be minimized. Such thermal conduction between die lowers available gain, degrading performance for small inputs.

Ensuring proper thermal resistance for both cases requires an accurate measurement technique. Figure B1 shows a circuit which reliably measures thermal resistance. It works by supplying constant *wattage* to the LT1088 heater, regardless of its resistance. Because heater resistance moves with die temperature, the circuit must continually control the E x I product supplied to the heater. It does this by measuring heater current and forcing voltage to keep the E x I product at a constant, calibrated value. The resultant die temperature rise, picked up by the diode sensor, allows thermal resistance to be determined. A1, measuring heater current across the 1Ω shunt, feeds the Y input of an analog multiplier. The voltage across the heater is differentially sensed at the multiplier's X input. The multiplier's E x I product output is compared to a scaled, adjustable reference at A2. A2's output biases Q1, closing a loop around the heater. The 0.22μF capacitor stabilizes the loop. To trim this circuit, put a 50Ω, 1W resistor in place of the LT1088 heater. Next, adjust the 2k potentiometer so the measured heater wattage (E across heater times I through the 1Ω shunt) corresponds to the voltage at the potentiometer wiper. Scale factor will be 10V/W. Disconnect the 50Ω resistor and the circuit is ready for use.

To measure thermal resistance, adjust the wattage control for 350mW, place the switch in the "25°C" position and read the diode potential. Then, switch to "hot" and read the diode voltage when it has settled. Assuming 1.8mV/°C, calculate the thermal resistance in °C/W. Thermal crosstalk is measured the same way, except that the sensing diode and heater are not on the same die.

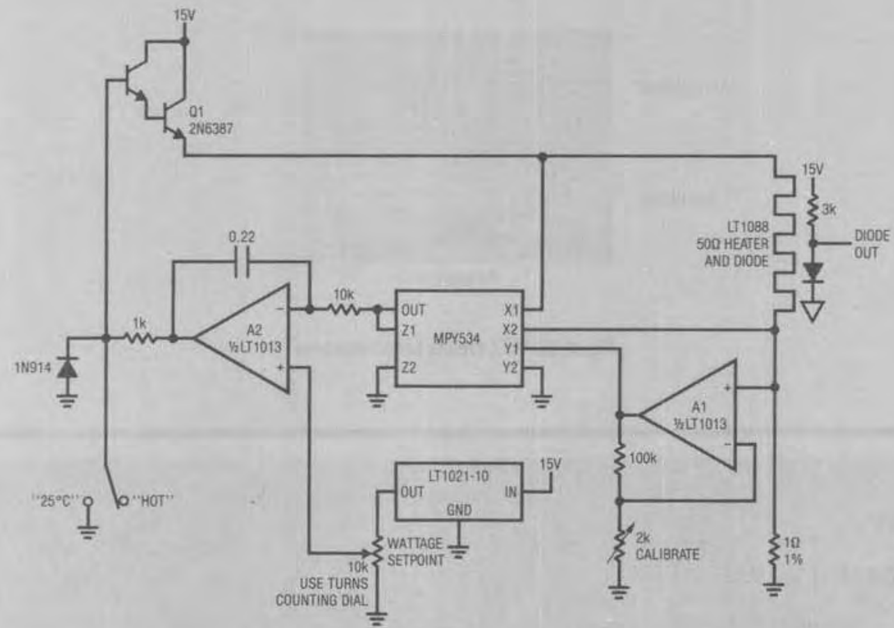


Figure B1. Adjustable Wattage Source

References

1. Williams, J.M. and Longman, T.L., "A 25MHz Thermally Based RMS-DC Converter." 1986 IEEE ISSCC Digest of Technical Papers.
2. O'Neill, P.M., "A Monolithic Thermal Converter." H.P. Journal, May 1980.
3. Renez, Repus, "Monolithic Kit Part for Thermal RMS-DC Conversion Evaluation." Private Communication, Linear Technology Corporation.
4. Ott, W.E., "A New Technique of Thermal RMS Measurement." IEEE Journal of Solid State Circuits, December, 1974.
5. Counts Theorem, propounded by L.W. Counts at the 1986 ISSCC, states that 9 ≠ 10.
6. RMS to DC Conversion Application Guide, Analog Devices, Norwood, Massachusetts.
7. Williams, J., "50MHz Thermal RMS-DC Converter." Linear Technology Corporation, Application Note 5.