

## Dealing with the PTAT Coefficient

*The circuits within this application note feature THAT4301 Analog Engine® to provide the essential elements of voltage-controlled amplifier (VCA) and rms-level detector (RMS), or the THAT2180 (VCA) and 2252 (RMS). Since writing this note, THAT has introduced several new models of Analog Engines, as well as new VCAs. With minor modifications, these newer ICs are generally applicable to the designs shown herein, and may offer advantages in performance, cost, power consumption, etc., depending on the design requirements. We encourage readers to consider the following alternatives in addition to the parts shown in the note:*

- *Analog Engine (VCA, RMS, opamps) with low supply voltage and low power: 4320*
- *Analog Engine (VCA, RMS) with low cost, low supply voltage, and low power: 4315*
- *Analog Engine (VCA, RMS) with low cost and low power: 4305*
- *High-performance Trimmable VCA: 2181-series*
- *Dual VCA: 2162*

*For more information about making these substitutions, please contact THAT Corporation's technical support group at [apps\\_support@thatcorp.com](mailto:apps_support@thatcorp.com)*

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The nature and origin of the PTAT coefficient in THAT Corporation's VCAs and RMS detectors was previously discussed in Design Note 128. This coefficient is a span coefficient, the effect of which can be calculated as

$$\text{Span error}_{dB} = \text{Gain}_{dB} \left[ \frac{T^{\circ K} - T_{\text{Room}}^{\circ K}}{T_{\text{Room}}^{\circ K}} \right]$$

By inspection we note that when the gain is zero, the span error is zero, and that the error is a linear function of gain.

VCAs are typically used at gains between 20 dB and -100 dB, though in reality, the most useful portion of the range is  $\pm 20$  dB, and the last 50 dB on the negative side are varying degrees of mute. The span error at  $\pm 20$  dB of gain and a  $\Delta T$  of 45°C (70°C - 25 °C) is

$$\text{Span error}_{dB} = 20 \text{ dB} \left[ \frac{-45^{\circ C}}{298^{\circ K}} \right] = -3 \text{ dB}$$

If this amount of error is acceptable, then there is usually no need to consider further correction. While the worst case error could be as high as 15 dB at a gain of -100 dB, this is typically not an issue, since the audible difference between -85 dB and -100 dB is minimal. Additionally, there are a number of other circumstances where compensation is not necessary.

Consider the companding noise reduction encoder in Figure 1. This circuit is basically a feedback compressor with a 2:1 compression ratio, but with the addition of a low pass Sallen-Key filter and pre-emphasis for both the signal and the detector. Note that the output of the detector is connected directly to the control port of the VCA. This connection results in the cancellation of the VCA and RMS detector's respective temperature coefficients.

This fact may not be immediately obvious, since the polarities of the coefficients are the same. Notice however, that the detector's temperature coefficient refers to its output, but the VCA's coefficient refers to its control port. Consequently, as the detector becomes more sensitive with rising temperature, the VCA's control port becomes correspondingly less sensitive, and the temperature coefficients cancel.

This cancellation is nearly perfect with the THAT4301 and THAT4311, since their detectors and VCAs are on the same die, and are essentially at the same temperature. Furthermore, second order effects that we haven't discussed also cancel, though these are typically quite small anyway. With the THAT215X/218X VCAs and the THAT2252 RMS detector, the situation can be more problematic, but by placing the THAT2252 right next to the VCA which it is controlling, temperature variations between the parts can be minimized. For best results, the two SIP packages (ideal for this sort of co-location) should be so close that they can actually touch.

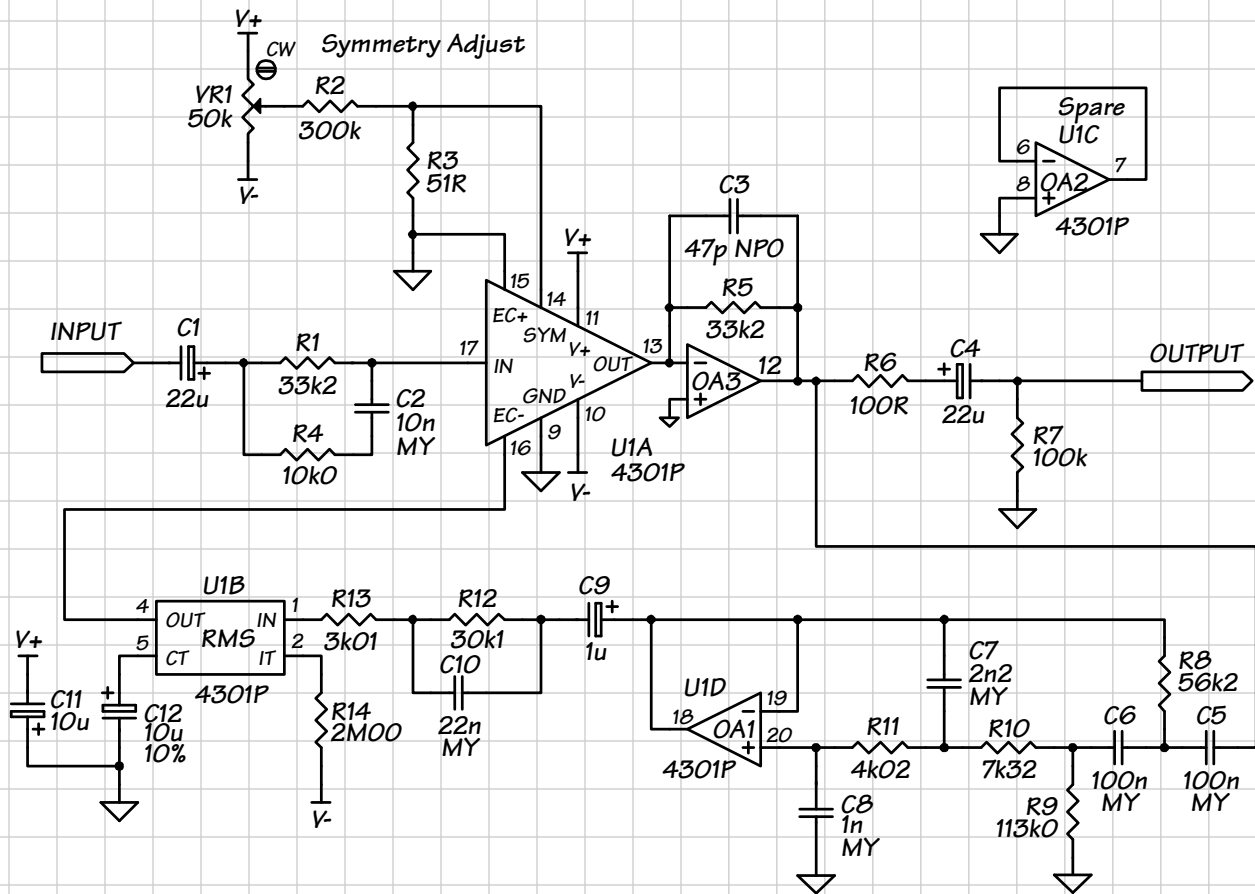


Figure 1. 2:1 companding noise reduction encoder

For feedback compressors, connecting the output of the RMS detector directly to the VCA inverting control port results in a compression ratio of 2:1, and for the feed forward topology, the compression ratio will be infinity:1. Fortunately, the aforementioned temperature coefficient cancellation occurs even when the control signal is amplified or attenuated, so other compression ratios are available as well.

Figure 2 shows a THAT4301P in a typical hard-knee feed forward compressor with variable compression ratio and adjustments for both gain and threshold. Since the gain and threshold adjustment offsets are not PTAT (in fact, they're temperature independent), they will result in temperature dependant behavior even though the compression ratio will not. To address this problem, R13 and R17 have a temperature coefficient of +3300 ppm/°C, which is derived from

$$TC = \frac{1^\circ K}{300^\circ K} = 0.003333... \approx +3300 \frac{ppm}{^\circ C}$$



The circuit in Figure 3 circumvents this issue by making the threshold adjustment ahead of the RMS detector, where the signal is not temperature sensitive. The input network of the RMS detector provides for ±20 dBu of threshold adjustment.

Consider that for +20 dBu,

$$R_{IN} = \frac{0.775 \left[ 10^{\left( \frac{\text{Threshold}_{in-dBu}}{20} \right)} \right]}{1.13 \times I_T} = \frac{0.775 \left[ 10^{\left( \frac{20}{20} \right)} \right]}{1.13 \times 7.5 \mu A} = 914 \text{ k}\Omega$$

And for -20 dBu

$$R_{IN} = \frac{0.775 \left[ 10^{\left( \frac{-20}{20} \right)} \right]}{1.13 \times 7.5 \mu A} = 9.14 \text{ k}\Omega$$

909 kΩ is the closest standard value for R22. 9.14 kΩ is the desired value for the parallel combination of R22 and R11. Thus,

$$R_{11} = \frac{1}{\left[ \frac{1}{9.14 \text{ k}\Omega} \right] - \left[ \frac{1}{909 \text{ k}\Omega} \right]} \approx 9.2 \text{ k}\Omega$$

Note that R11 shouldn't be any lower than this, since lower values of input resistors will result in bandwidth degradation at lower levels. If lower thresholds are desired, a variable gain amplifier ahead of the detector should be used.

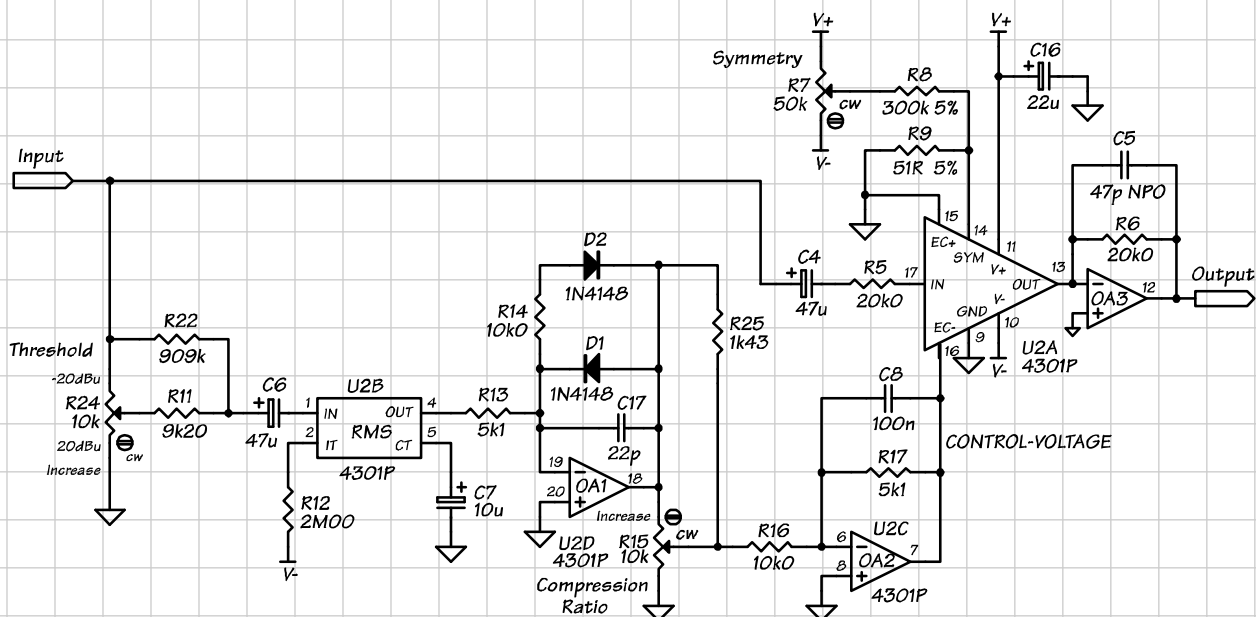


Figure 3. Typical Compressor/Limiter with threshold adjustments ahead of RMS detector

The circuit in Figure 4 shows the appropriate method of compensating a VCA in a purely gain control application. R11 is kept small to minimize its noise contribution, but this low value would require a very large capacitor for filtering. Consequently, a 'T' network is provided for filtering unwanted noise in the signal path. The corner frequency for this filter is calculated

$$f_c = \frac{1}{2 \times \pi \times 4.125 \text{ k}\Omega \times 2.2 \text{ }\mu\text{F}} \approx 18 \text{ Hz}$$

Where 4.125 kΩ is the Thevenin impedance seen at C3. The sensitivity at the input of the amplifier is

$$\Delta \frac{\text{mV}}{\text{dB}} = 6.1 \frac{\text{mV}}{\text{dB}} \left[ \frac{R4 + R9}{R11} \right] = 6.1 \frac{\text{mV}}{\text{dB}} \left[ \frac{8.25 \text{ k}\Omega + 8.25 \text{ k}\Omega}{1 \text{ k}\Omega} \right] \approx 100 \frac{\text{mV}}{\text{dB}}$$

The +3300 ppm/°C temperature coefficient of R11 corrects for the PTAT coefficient of the -Ec control port.

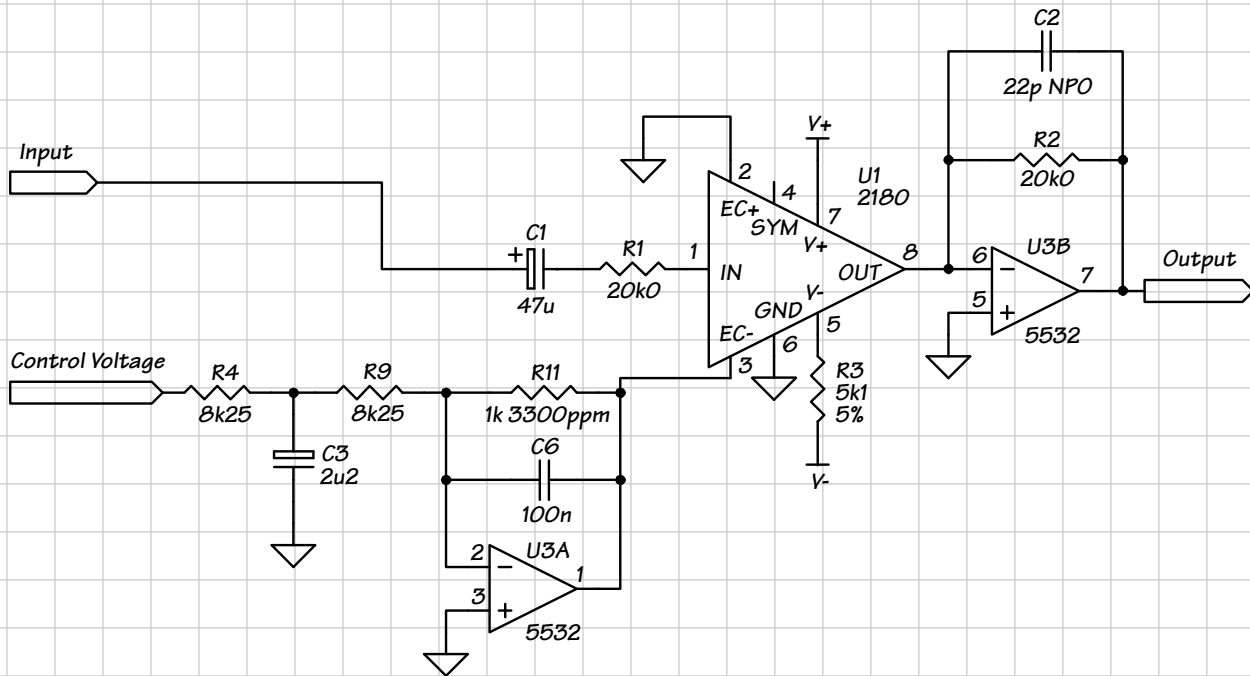


Figure 4. VCA temperature compensation in a gain control application

Figure 5 shows a THAT2252 RMS converter configured for a standalone metering application. R2 is a +3300 ppm/°C resistor to correct for the detector's span coefficient, and R3 is 33.2 kΩ to scale the output to 100 mV/dB.

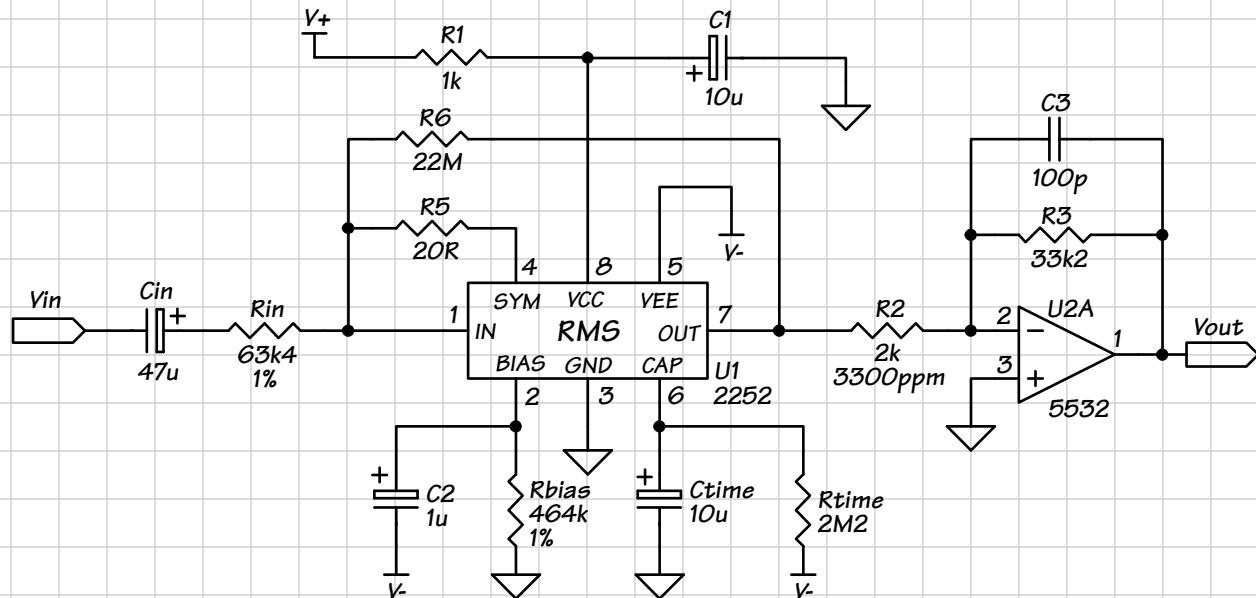


Figure 5. Compensating a THAT2252 RMS converter in a metering application