(Part 2)

Transparent V-I protection in Audio Power Amplifiers

"The aversion cultivated by some designers to *V-I* limiting in audio power amplifiers is wholly illusory," argues Michael Kiwanuka in his second article on the topic of implementing no-compromise output protection for audio power amplifiers.

*Figs 1-25 were presented in last month's article, as were equations 1-9, Ed.

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Building on last month's discussion of single-slope, single-breakpoint non-linear foldback limiting, this next section covers a dual-slope alternative.

Introducing a resistor, R_d, in series with the diode in **Fig. 27** causes the voltage drop across the series combination to increase linearly above the diode's conduction threshold. In turn, this induces a net linear increase in potential across the voltage divider R_{2A}, and R_{2B}. This gives rise to segment B-D in the protection locus, **Fig. 26**, whose gradient can be varied linearly with R_d about point B. This allows greater flexibility with regard to optimal placement of the breakpoint. For brevity the essential diodes D_F and D_P described in part 1 of this paper are omitted in all subsequent figures.

As is the case with single-slope, linear foldback limiting, segment B-D must intersect the safe operating area's V_{ce} axis at a value greater than the sum of the moduli of the supply rails, if spurious limiter activation is to be prevented. Available current per output pair at $V_{ce}{}^{2}$4V, is further increased to 12A8 compared to 7A1 for the locus in Fig. 13*.

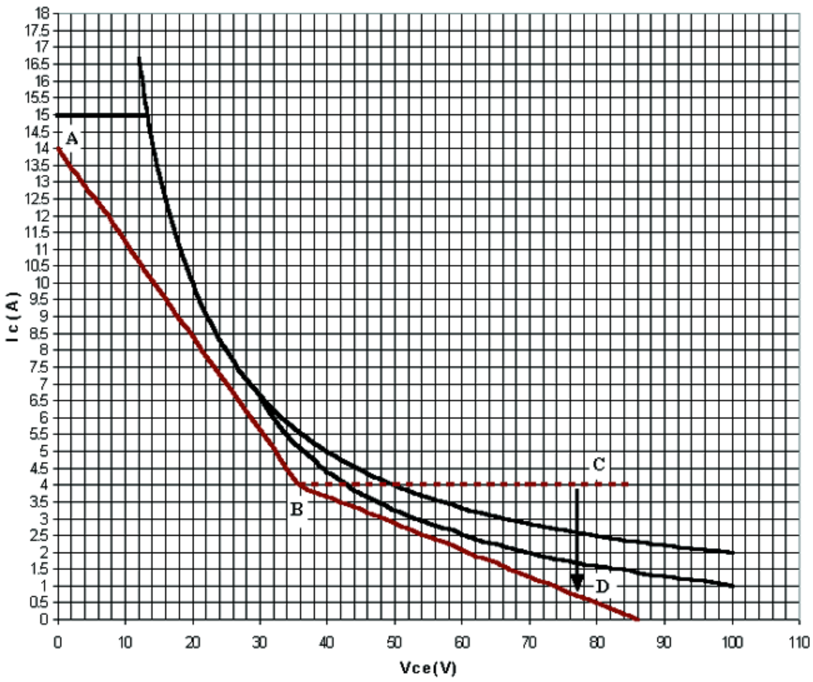


Figure 26: Dual slope, single breakpoint, non-linear foldback protection locus

Initially, resistor values without R_d are calculated for segment A-B-C, **Figs 28, 29,** and the value of R_d established *in situ*, **Fig. 30,** using any convenient set of points along B-D.

With reference to Fig. 28, and selecting R_1=8K2; I_d=1mA:

$$I_1 = I_d + I_2 + I_3 \tag{10}$$

And,

$$R_{2B} = \left(\frac{0.6}{0.88}\right) R_{2A} \tag{11}$$

From equation 10:

$$\frac{(40-4.6)}{8\text{k}2} = 1\text{mA} + \frac{0.88}{R_{2A}} + \frac{0.6}{R_3} \tag{12}$$

From Fig. 29, and invoking equation 11:

$$0.6 = \frac{3.08 R_{2A}(0.6/0.88)}{R_{2A}(0.6/0.88) + R_{2A} + 8K2 R_3/(8K2 + R_3)} \tag{13}$$

Solving (12), and (13), simultaneously:

$$R_3 \approx 704R7,$$

$$R_{2A} \approx 356R9.$$

And,

$$R_{2B} = (0.6/0.88)R_{2A} \approx 243R3.$$

With reference to Fig. 30:

$$I_2 = (0.6/R_{2B}) = (0.6/243R3) \approx 2.47\text{mA}$$

$$V_X = (I_2 R_{2A} - 39V4) \approx -38V52$$

$$V_{R3} = (V_X + 39V89) \approx 1V37$$

$$I_3 = (V_{R3}/R_3) \approx 1.94\text{mA}$$

But,

$$I_d = I_1 - (I_2 + I_3) \tag{14}$$

Where,

$$I_1 = (40 - V_X)/8K2 \approx 9.58\text{mA}$$

$$I_d = 9.58\text{mA} - (2.47\text{mA} + 1.94\text{mA}) \approx 5.17\text{mA}$$

$$R_d = (V_{Rd}/I_d) = (V_{R3} - 0.6)/I_d \approx 149R1.$$

It is worth considering that the forward voltage drop, V_f is around 0V6 when I_d is about a milliamp for most small signal diodes at 27°C. For a suitable device though, such as the 1N4148, V_f is around 0V65 when I_d is 5mA. This requires that R_d calculated above be revised downwards for enhanced precision. Thus,

$$R_d = (V_{Rd}/I_d) = (V_{R3} - 0.65)/I_d \approx 139R3$$

As previously recommended, the calculated resistor values should be made up from series, and/or parallel combinations of 1% components where necessary.

The dual-slope, single-breakpoint scheme in Fig. 31, sometimes erroneously[1] described as 'treble slope', (sic), is an amalgam of the circuits in Figs 5 and 18.

As in Fig. 18, the breakpoint occurs at V_{out}²0V, i.e. V_{ce}²V_{cc}, giving locus A-D-E-F, **Fig. 32.** However, segment D-E-F, being part of C-D-E-F, is established by R_1 and R_3. As a result, its efficacy is therefore dependant on the value of R_e as the network in Fig. 5.

Resistor R_2 merely pulls the base of the protection transistor low as required for 0V□V_{ce}<40V. This gives segment A-D, whose position in the safe operating area is

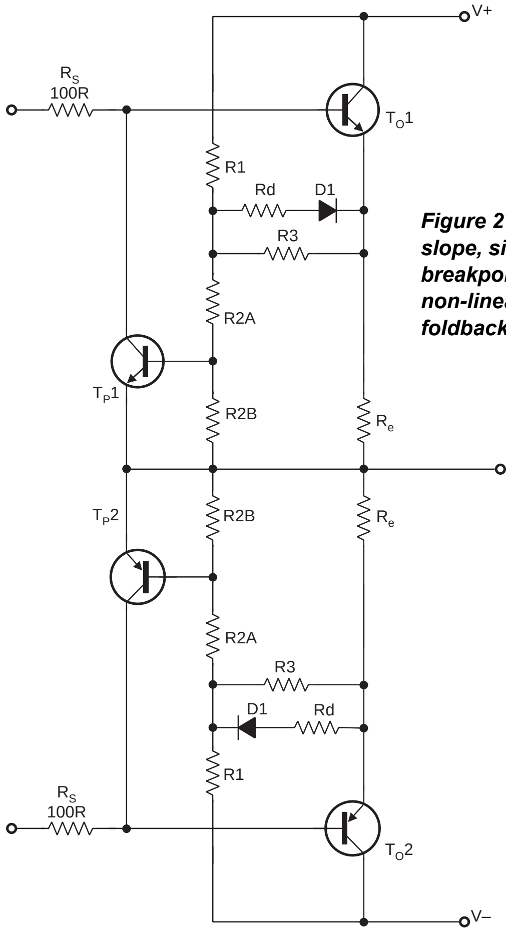


Figure 27: Dual slope, single breakpoint, non-linear foldback limiter

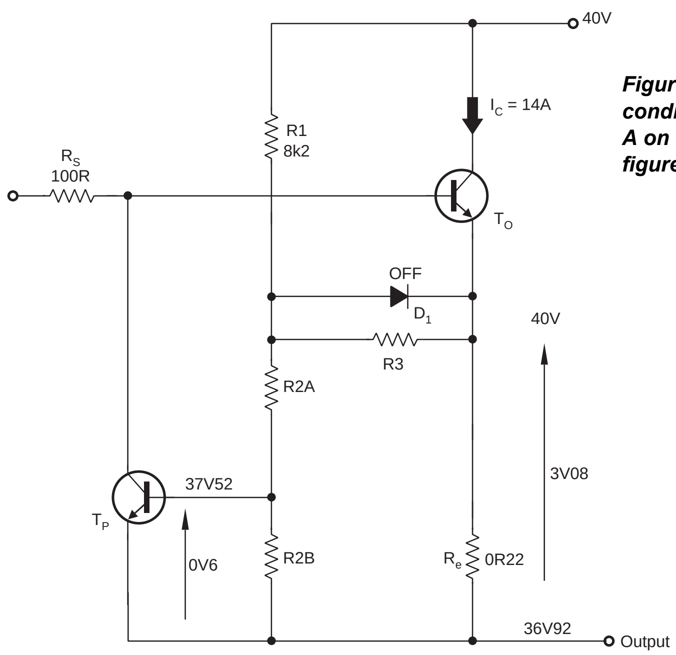


Figure 29: Output conditions at point A on the locus in figure 26

Figure 28: Output conditions at point B on the protection locus in figure 26

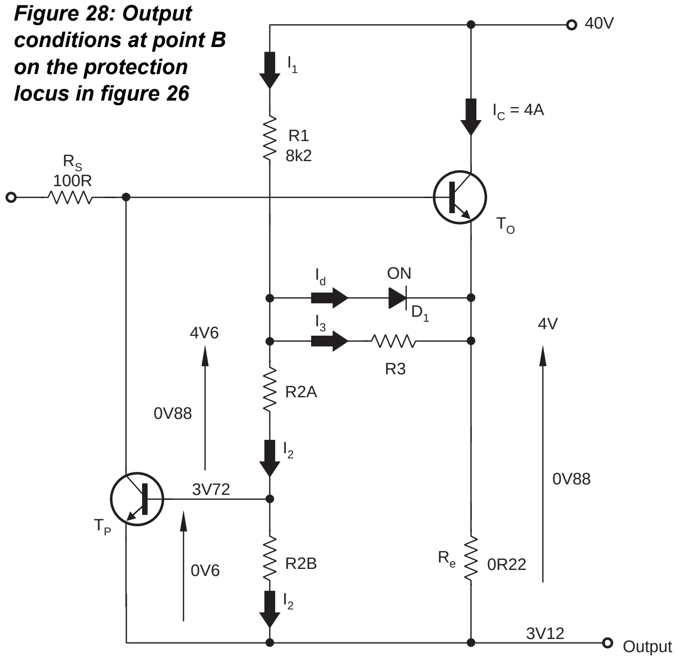
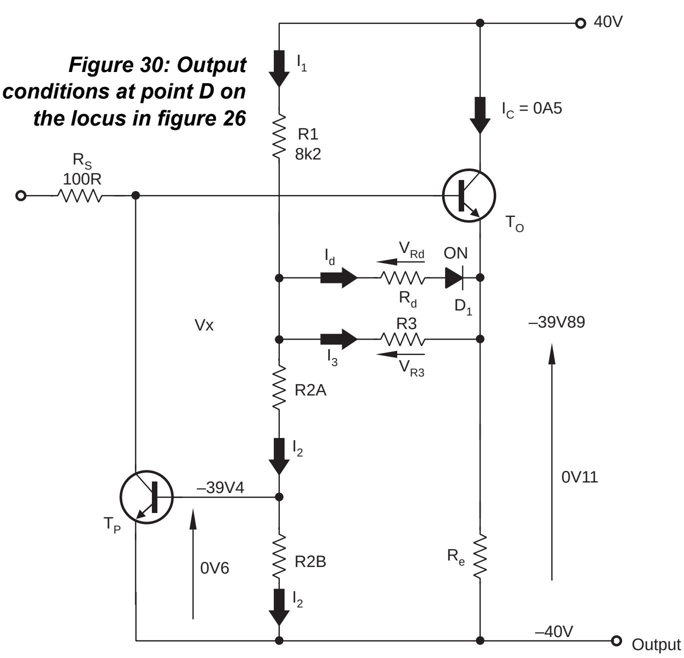


Figure 30: Output conditions at point D on the locus in figure 26



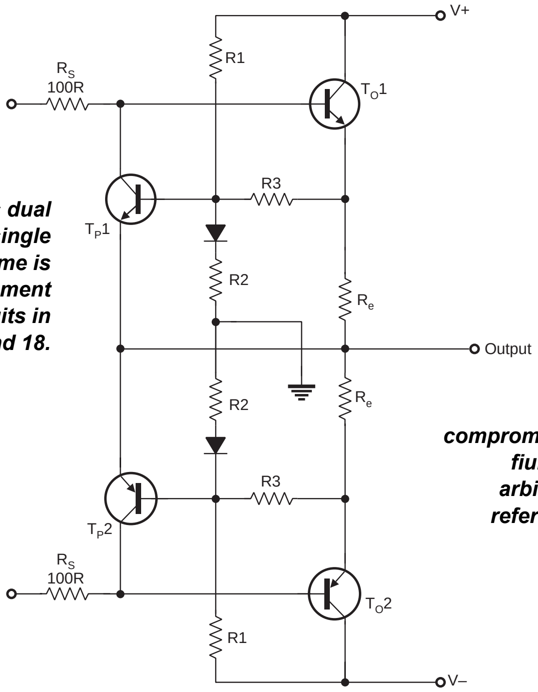


Figure 31: This dual slope single breakpoint scheme is a logical development of the circuits in figures 5, and 18.

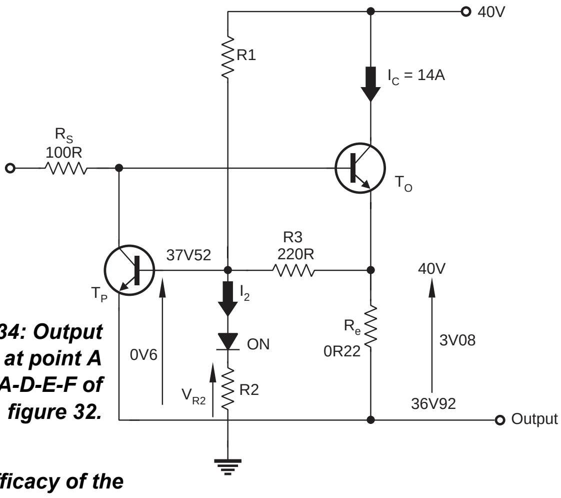
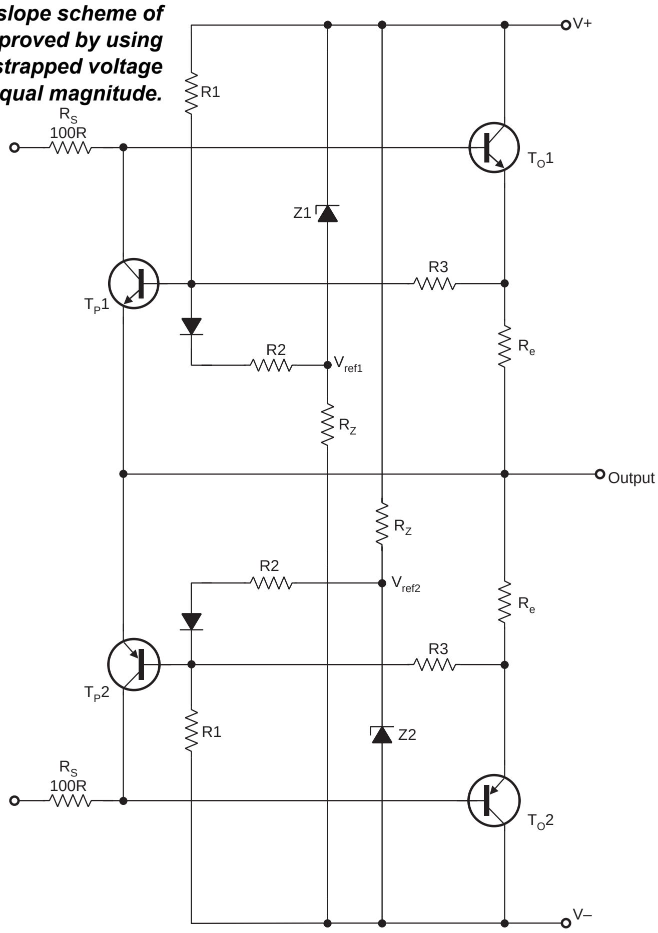


Figure 34: Output conditions at point A on locus A-D-E-F of figure 32.

Figure 35: Efficacy of the compromised dual slope scheme of fiure 31 is improved by using arbitary, bootstrapped voltage references of equal magnitude.



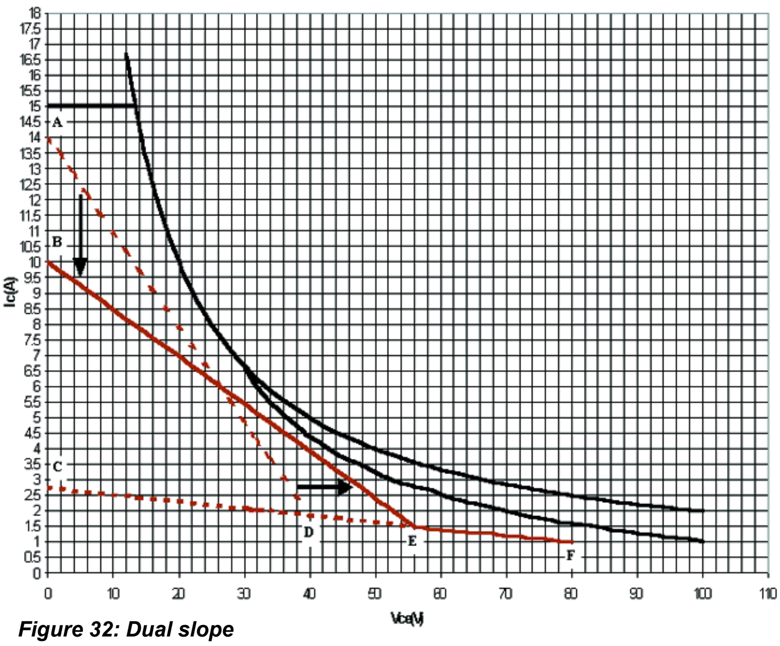


Figure 32: Dual slope single breakpoint loci described by the circuits in figure 31 and 35.

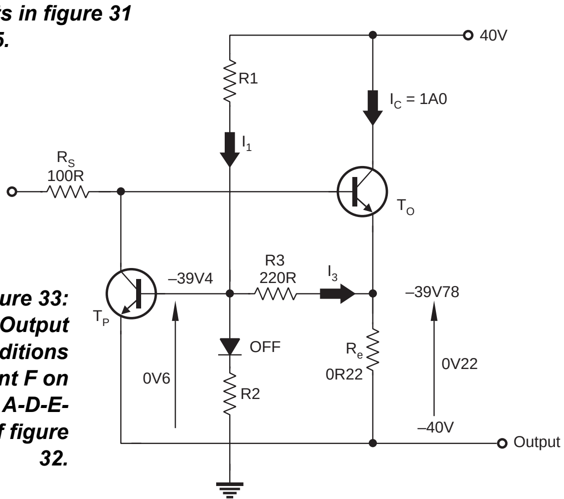


Figure 33: Output conditions at point F on locus A-D-E-F of figure 32.

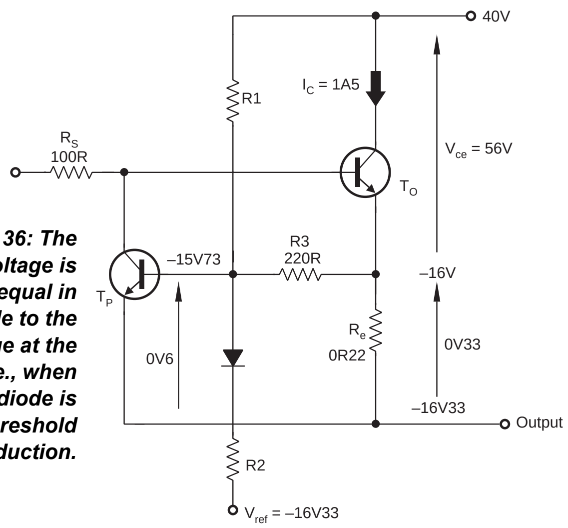


Figure 36: The reference voltage is made equal in magnitude to the output voltage at the breakpoint, (i.e., when Vce=56V); the diode is then at the threshold of conduction.

ill-defined for non-ideal supply rails, due to the use of an invariant voltage reference.

Since the breakpoint for this arrangement is fixed at $V_{ce}{}^{2}V_{cc}$, only points A and F on locus A-D-E-F are required to obtain a solution.

With reference to **Fig. 33,** let R_3=220R, and $V_{cc}{}^{2}40\text{V}$:

$$I_1 \approx I_3$$

Where,

$$I_3 = (-39.4 + 39.78)/220R \approx 1.73\text{mA}$$

and,

$$R_1 = (40 + 39.4)/1.73\text{mA} \approx 46K$$

With reference to **Fig. 34:**

$$I_2 \approx (40 - 37.52)/(R_1 // R_3) = 2.48/219R \approx 11.33\text{mA}$$

With $V_f{}^{2}0\text{V}7$ at 11mA,

$$R_2 = V_{R2}/I_2 = (37.52 - 0.7)/11.33\text{mA} \approx 3K3$$

This scheme is clearly inferior to the standard linear foldback arrangement of Fig. 1. It delivers only 1A5 at $V_{ce}{}^{2}45V97$, requiring a minimum of six output pairs for 4ohms ±60° load drive from ±40V supply rails.

As in Fig. 22, the network in Fig. 31 can be usefully improved, **Fig. 35,** by changing the diode reference from zero to an arbitrary voltage, V_{ref}, such that, 0V<$|V_{ref}|$<$|V_{cc}|$. This enhances the flexibility of the circuit, as the breakpoint can now be moved freely along segment C-F. This gives rise to a more efficient locus, B-E-F, Fig. 32, whose position in the safe operating area is unaffected by supply rail variation.

The reference voltage is established by determining the output conditions at the breakpoint, **Fig. 36**. Therefore for locus B-E-F in Fig. 32, $V_{\text{Ref}1}$=–16V33 and $V_{\text{Ref}2}$=+16V33. This calls for a nominal 56V33 zener diode. As previously recommended, multiple low-voltage devices should be used to minimise series impedance.

With reference to **Fig. 37:**

$$I_1 \approx I_3$$

Where,

$$I_3 = (-39.4 + 39.78)/220R \approx 1.73\text{mA}$$

$$R_1 = (40 + 39.4)/1.73\text{mA} \approx 46K$$

Referring to **Fig. 38:**

$$I_2 = (40 - 38.4)/(R_1 // R_3) = 1.6/219R0 \approx 7.3\text{mA}$$

With $V_f{}^{2}0\text{V}65$ at 7mA,

$$R_2 = V_{R2}/I_2 = (38.4 - 0.65 + 16.33)/7.3\text{mA} \approx 7K4$$

Note that there is no change in the value of R_1 and R_3 in the circuits of Figs 5, 31, and 35, with different values of R_2 required to merely pull the base of the protection transistor low as appropriate when the series diode is forward biased.

Although the efficacy of the protection locus is in part ameliorated by the means described above, the gradient of segment E-F, being part of C-D-E-F, is determined by resistors $R_{1,3}$, and limited by practical values of R_e – an affliction absent in the circuit of Fig. 27.

Complete independence from R_e of both segments of the dual slope protection locus described by the circuit in Fig. 35 can be accomplished by the introduction of a base-emitter resistor, R_2, **Fig. 39,** for each protection transistor. The result is in fact merely a union of the linear single slope scheme of Fig. 1, and the non-linear single slope circuit of Fig. 22.

The linear, single slope locus in Fig. 2 is reproduced in

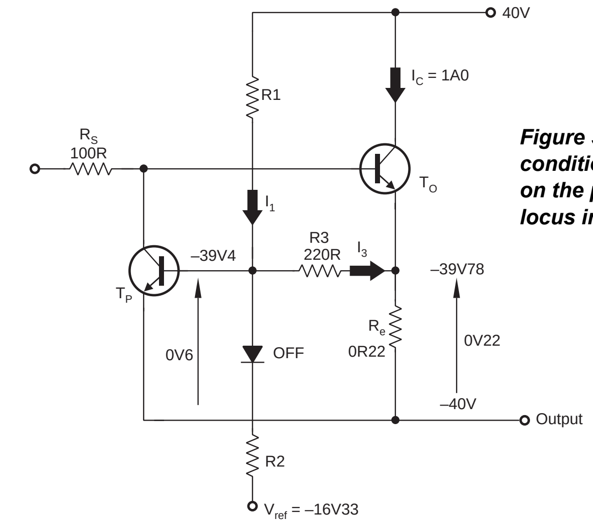


Figure 37: Output conditions at point F on the protection locus in figure 32.

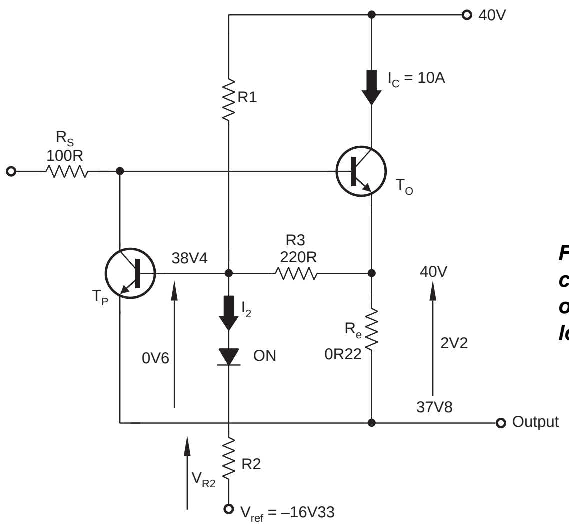


Figure 38: Output conditions at point B on the protection locus in figure 32.

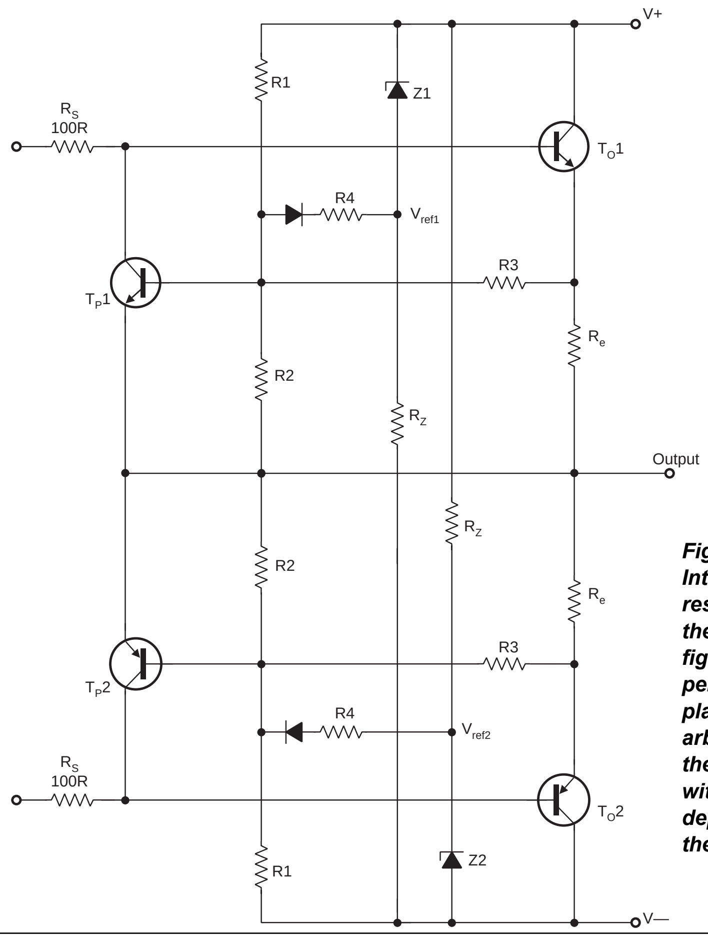


Figure 39: Introducing resistor R_2 into the circuit of figure 35 permits placement of an arbitary locus in the S.O.A., without undue dependence on the value of R_e.

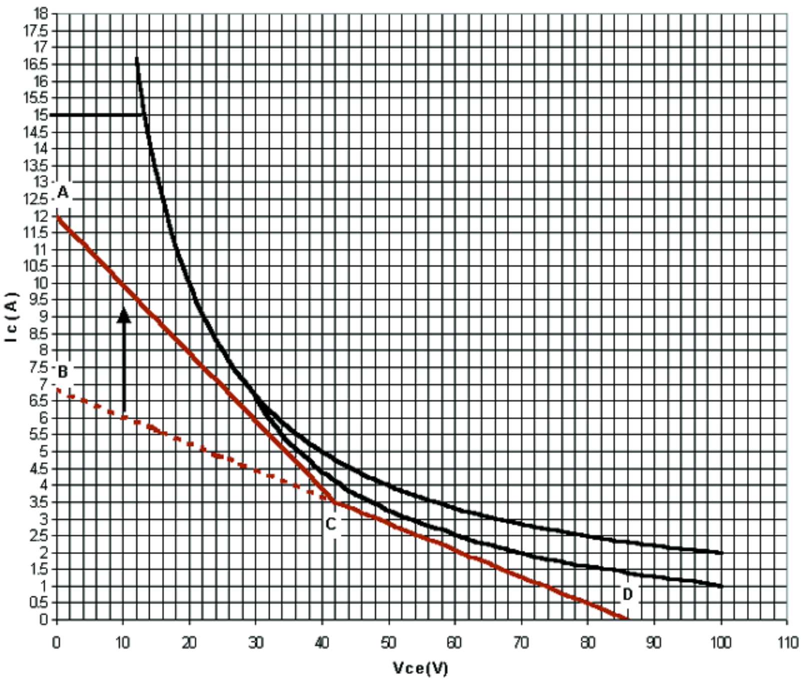
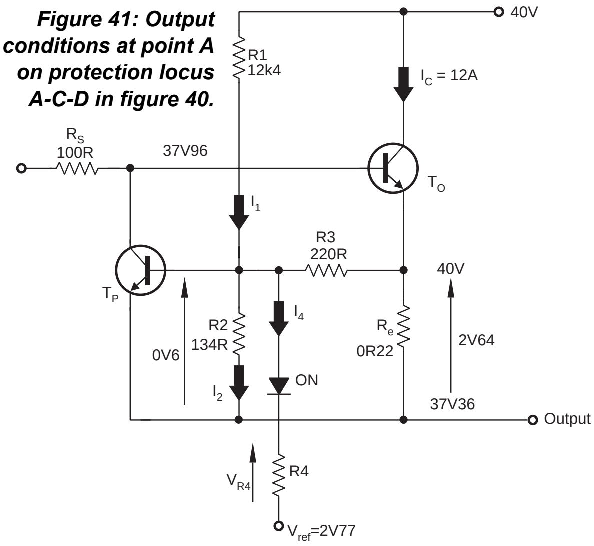


Figure 40: Dual slope, single breakpoint locus described by the circuit of figure 39. Resistor R4 modifies the linear single slope segment B-C-D, of figure 2 by effecting a vertical translation of segment B-C about point C.

Figure 41: Output conditions at point A on protection locus A-C-D in figure 40.



R_1=12K4kohms, and R_2=143R. Resistor R_4 pulls the base of the protection transistor low as required for $0V \leq V_{ce} \leq 42V$, giving segment A-C.

The reference voltage is equal to the output voltage when V_{ce}=42V, thus,

$$V_{\mathrm{Re}\,f1} = 40V - \{42V + (3A5 \times 0R22)\} = -2V77,$$

and $V_{\mathrm{Re}f2}$=+2V77.

Referring to Fig. 41:

$$(I_2 + I_4) \approx (I_1 + I_3)$$

$$I_4 \approx (I_1 + I_3 - I_2)$$

$$I_4 = (40 - 37.96)/12K4 + (40 - 37.96)/220R - 0.6/143R$$

$$I_4 \approx 5.24\text{mA}$$

With $V_f \approx 0V6$,

$$R_4 = V_{R4}/I_4 = (37.96 - 0.6 + 2.77)/5.24\text{mA}$$

$$R_4 \approx 7K7$$

The flexibility of the scheme in Fig. 39 is significantly improved relative to Fig. 35. However such flexibility is easily surpassed by the network in Fig. 27, whose accuracy is not compromised by dependence on discreet value zener references.

Treble-slope, (dual-breakpoint) non-linear foldback limiting

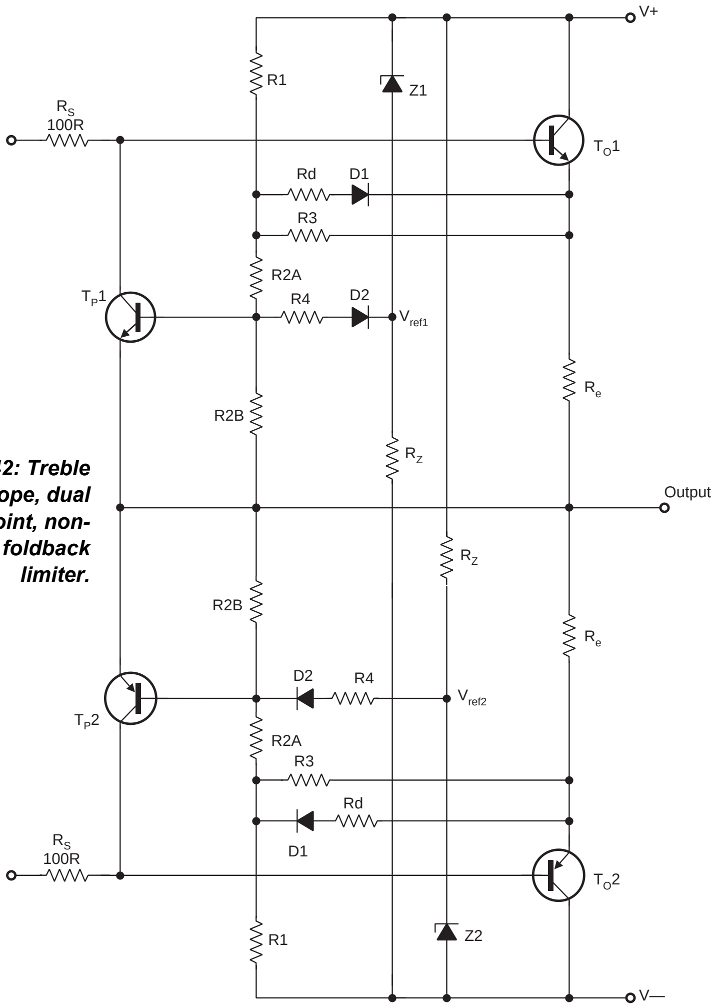
With modern power transistors and practical loudspeaker systems, an optimally located dual-slope protection locus realised by the limiter in Fig. 27 can hardly be improved upon with respect to efficiency in the critical $|V_{cc}| \leq V_{ce} < 2|V_{cc}|$ region.

However, for purely resistive laboratory loads with which a power amplifier's published specifications are obtained, the $0V \leq V_{ce} \leq |V_{cc}|$ region of the safe operating area is of primary interest, Fig. 10.

In a competitive market place therefore, even when the truth of the matter is known, an amplifier designed to maintain its rated voltage swing across resistive loads of decreasing magnitude – down to 1ohms – without limiter intrusion, may be commercially rewarding. A suitably robust power supply and conservative thermal management are assumed.

To this end the treble-slope design in **Fig. 42** is

Figure 42: Treble slope, dual breakpoint, non-linear foldback limiter.



presented. The circuit is a straightforward amalgam of the dual-slope scheme of Fig. 27, and the single slope, single breakpoint network of Fig. 22.

The circuit in Fig. 27 produces the dual slope characteristic B-D-F, **Fig. 43**, while resistor R_4 pulls the base of the protection transistor low as appropriate for $0\text{V} \leq V_{ce} \leq 42\text{V}$, giving segment A-C. Fifty-volt supply rails are assumed; a treble-slope locus with ±40V rails is vastly unnecessary.

The reference voltage is equal in magnitude to the output voltage V_{out} at breakpoint C, Fig. 43. Thus,

$$\left|V_{\mathrm{Re}\,f1}\right| = \left|V_{\mathrm{Re}\,f2}\right| = \left|V_{out}\right|_{V_{ce}=42V} = 7V23$$

with V_{Ref1} at $7V23$ and V_{Ref2} at $-7V23$.

As previously established for Fig. 27, component values without R_d are calculated for segment B-D-E, (**Figs 44** and **45**), and the value of R_d established *in situ*, **Fig. 46**, using any convenient set of points along D-F. Resistor R_4 is then calculated for a nominal V_{ce}=0V, at point A **Fig. 47**.

With reference to Fig. 44, let R_1=8K2, and $V_f \approx 0V6$ when I_d=1mA.

$$I_1 = I_d + I_2 + I_3 \tag{15}$$

And,

$$R_{2B} = \left(\frac{0.6}{0.44}\right) R_{2A} \tag{16}$$

From equation 15:

$$\frac{(50+11.4)}{8K2} = 1\text{mA} + \frac{0.44}{R_{2A}} + \frac{0.6}{R_3} \tag{17}$$

From Fig. 45, and invoking equation 16:

$$0.6 = \frac{1.474 R_{2A}(0.6/0.44)}{R_{2A}(0.6/0.44) + R_{2A} + 8K2R_3/(8K2+R_3)} \tag{18}$$

Solving (17) and (18) simultaneously:

$$R_3 \approx 160R7$$

$$R_{2A} \approx 159R8$$

$$R_{2B} = (0.6/0.44)R_{2A} \approx 217R9$$

With reference to Fig. 46:

$$I_2 = (0.6/R_{2B}) = (0.6/217R9) \approx 2.75\text{mA}$$

$$V_X = (I_2 R_{2A} - 49V4) \approx -48V96$$

$$V_{R3} = (V_X + 49V89) \approx 0V93$$

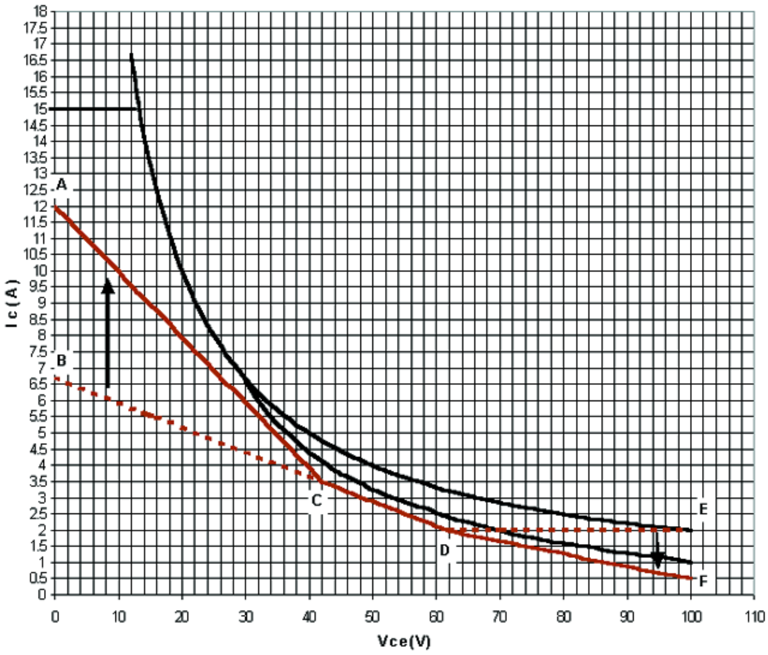


Figure 43: Treble slope, dual breakpoint protection locus described by the circuit of figure 42, Resistor R4 modifies the dual slope characteristic B-D-F by effecting a vertical translation of segment B-C about point C.

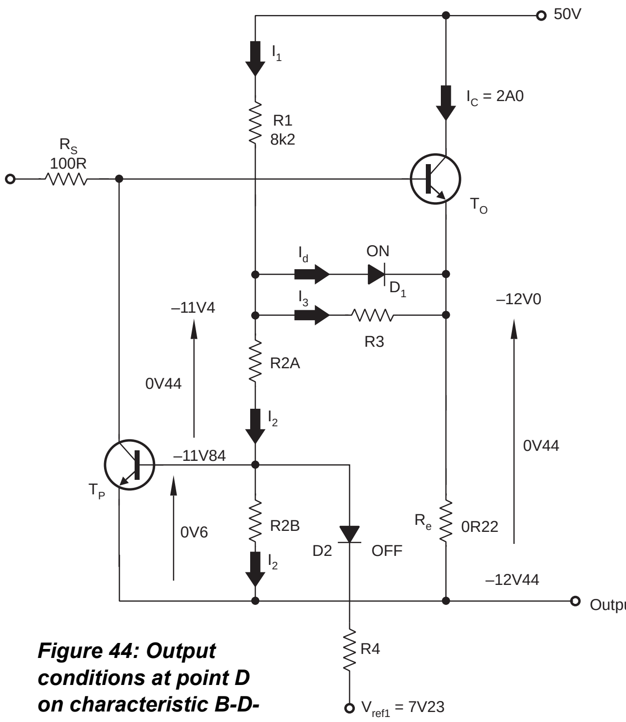


Figure 44: Output conditions at point D on characteristic B-D-E in figure 43.

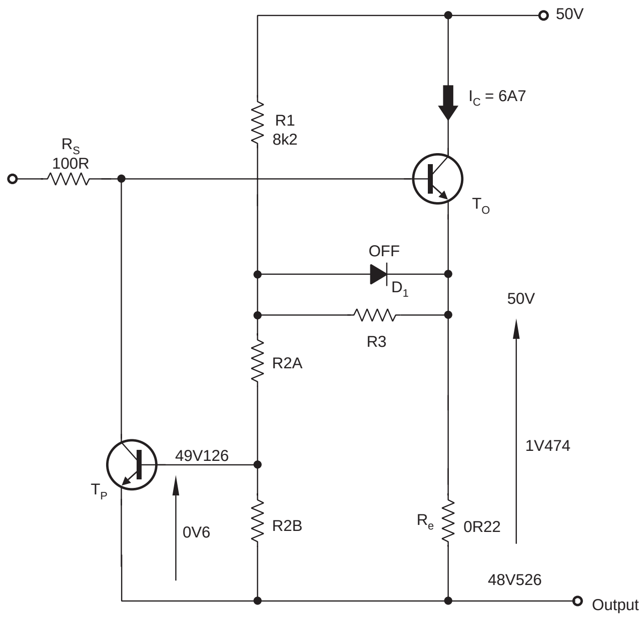


Figure 45: Output conditions at point B on characteristic B-D-E in figure 43.

But,

$$I_d = I_1 - (I_2 + I_3) \tag{19}$$

Where,

$$I_1 = (40 - V_X)/8K2 \approx 10.85\text{mA}.$$

$$I_d = 10.85\text{mA} - (2.75\text{mA} + 5.79\text{mA}) \approx 2.31\text{mA}$$

$$R_d = (V_{Rd}/I_d) = (V_{R3} - 0.6)/I_d \approx 143R0$$

From Fig. 47:

$$R_4 = (V_{R4}/I_4) \tag{20}$$

Where:

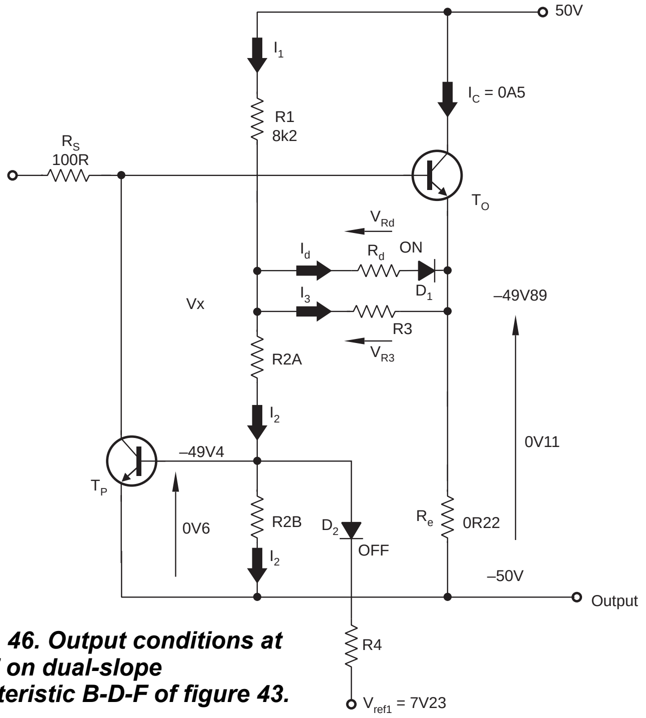


Figure: 46. Output conditions at point F on dual-slope characteristic B-D-F of figure 43.

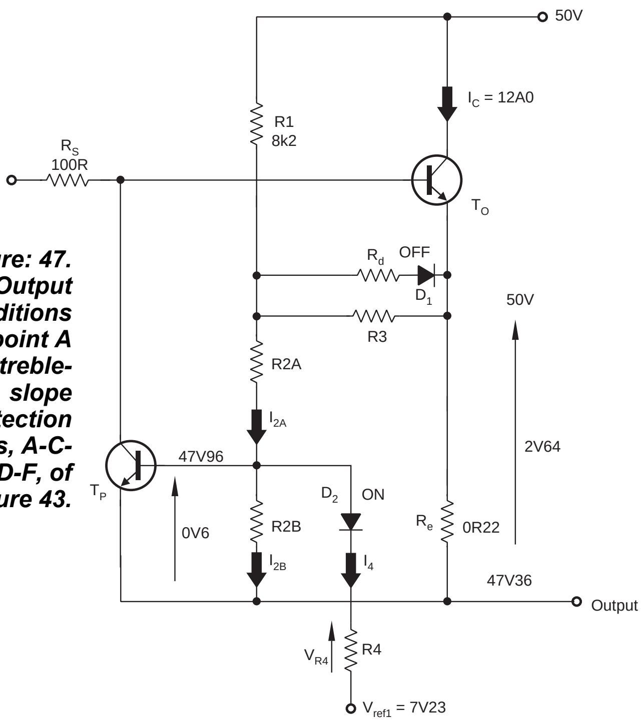


Figure: 47. Output conditions at point A on treble-slope protection locus, A-C-D-F, of figure 43.

$$I_4 = I_{2A} - I_{2B} \tag{21}$$

$$I_4 = \frac{(50 - 47.96)}{\{(R_1 // R_3) + R_{2A}\}} - \frac{0.6}{R_{2B}} \approx 3.67\text{mA}$$

$$R_4 = (47.96 - 7.23 - 0.6)/3.67\text{mA} \approx 10K9$$

A 4ohms ±60° load driven to ±50V rails requires $i_c{}^{2}9A5$ when $v_{ce}{}^{2}59\text{V}$, resulting in peak transistor dissipation, $P_{d(max)}{}^{2}561\text{W}$. The treble slope protection locus of Fig. 43 allows 2A at $v_{ce}{}^{2}59\text{V}$ for a single complementary transistor pair. Therefore, five complementary pairs are required to drive a notional 4ohms ±60° loudspeaker system from ±50V supply rails without intrusive limiter activation.

The required reference voltage calls for a nominal 42V77 voltage drop across Z_1 and Z_2. As previously recommended, the required voltage drop should be realised with multiple low-voltage devices, of 6V to 12V, as a series combination of these should collectively possess a significantly lower series impedance than a single high voltage device.

In practice, Z_1 and Z_2 may each consist of five ZPD6.8RL, in series with a single ZPD8.2RL, biased at a nominal quiescent current of 10mA by R_z.

A more elegant – if rather tedious – approach[2] compensates for variation in zener voltage drop with temperature. This calls for the introduction of typically two to four forward biased diodes in series with the zener diode.

The decreasing voltage of the forward biased p-n junctions with increasing temperature, (negative temperature coefficient), tends to counteract the increase in zener voltage with increasing temperature, (positive temperature coefficient), and conversely. Therefore Z_1 and Z_2 may each consist of a series combination of three IN961B 10V zeners, a single ZPD8.2RL 8.2V device, and seven 1N4148 forward biased diodes.

For brevity perhaps, in place of Z_1 and Z_2, the shunt-feedback circuit of **Fig. 48** may be used with a single, temperature compensated zener reference diode, such as the 6.2V 1N829A. This circuit permits the synthesis of a high voltage source without recourse to loose-tolerance, high voltage zener diodes, or indeed multiple small-value devices.

However, the variation in zener voltage drop due to

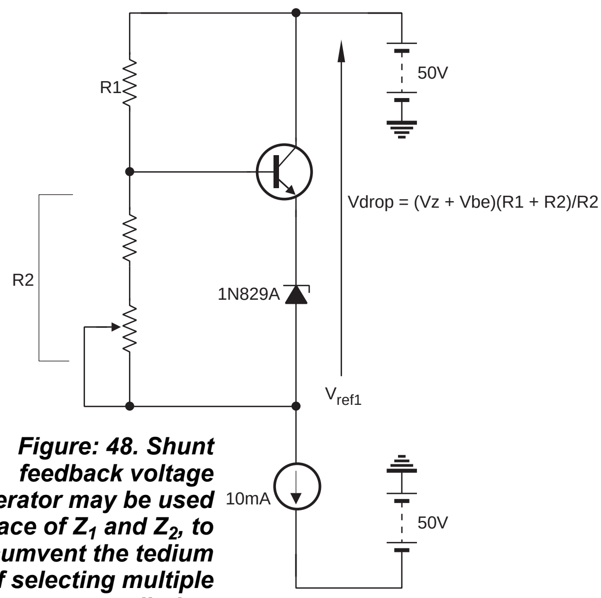


Figure: 48. Shunt feedback voltage generator may be used in place of Z_1 and Z_2, to circumvent the tedium of selecting multiple diodes.

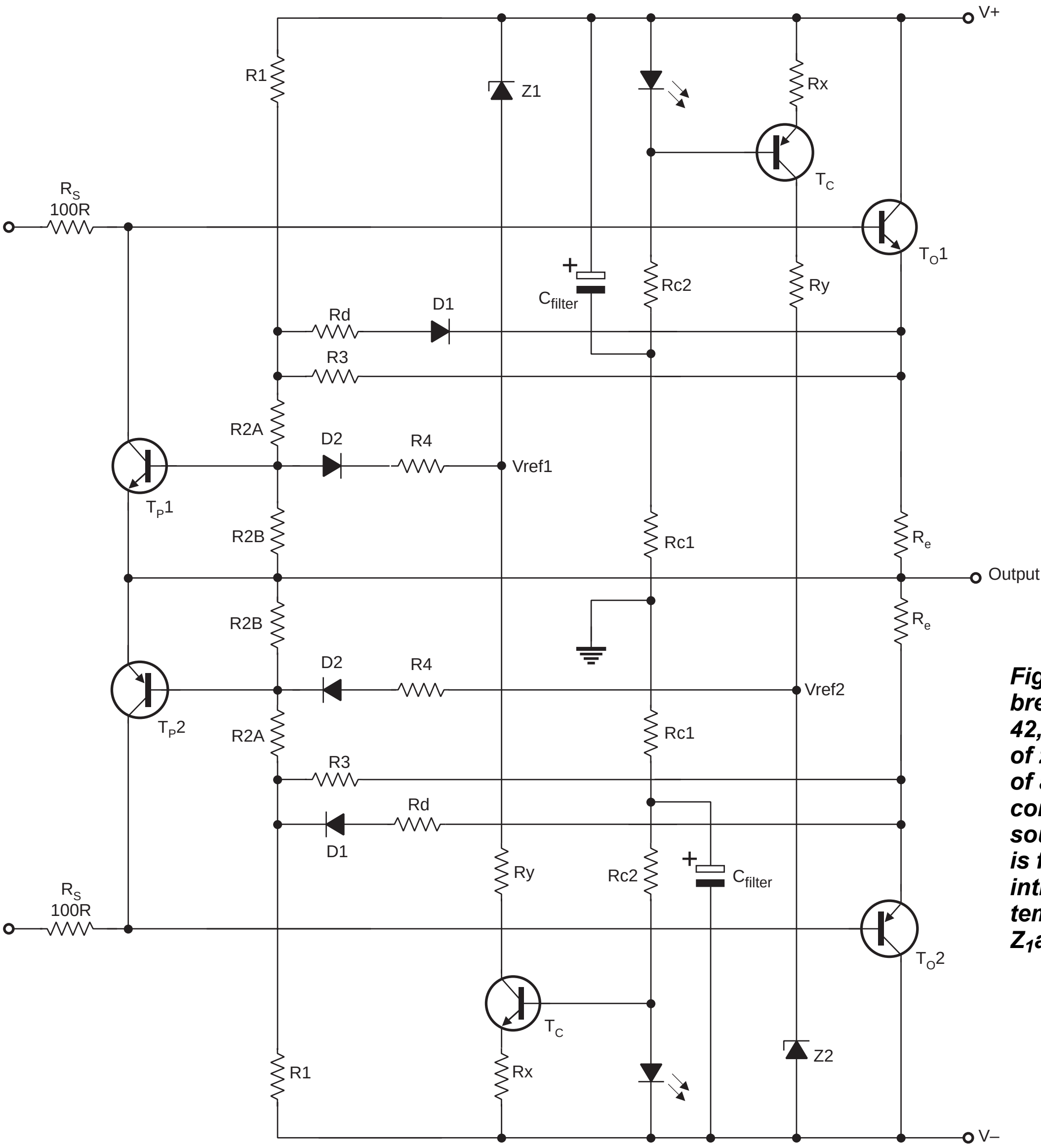


Figure: 49. Treble-slope, dual breakpoint limiter of figure 42, with improved regulation of zener voltage, V_z by means of a temperature compensated current source/sink. Such regulation is further enhanced by the introduction of a measure of temperature compensation to Z_1 and Z_2.

current fluctuation is invariably more significant than that due to change in temperature. Where cost is no object, R_z may be replaced with a temperature compensated[3,p.226] current source/sink, **Fig. 49.** This can be in the guise of a LED-biased transistor, T_c.

LED current limiting resistor R_c is split symmetrically into two components, R_{c1} and R_{c2}, whose intersection[4] is decoupled by capacitor C_{filter} to the supply rail. The single-pole filter comprising C_{filter} and R_{c1} across the LED's internal resistance, in series with R_{c2}, improves the regulation of the voltage drop across the LED by diminishing power supply ripple in the current established by R_{c1} and R_{c2}.

A time constant, $\tau_{filter}=C_{filter}R_{c1}$, of the order of two seconds is sufficient. Connecting C_{filter} directly across the LED is sub-optimal, as a commensurately larger component would then be required for the same time constant.

Resistor R_y minimises power dissipation in T_c; a collector-emitter voltage drop of the order of 20V for a collector current of 10mA should suffice with suitable small signal transistors, such as Motorola's 2N5551/2N5401.

Protecting paralleled complementary output transistors

Emitter resistor, R_e, performs current-voltage conversion for the *V-I* limiter. It also promotes thermal stability by maintaining equable current distribution in a paralleled pair output stage. For this reason some designers suggest[5,p.257] that it is only necessary to monitor transistor current in a single complementary pair in a multiple-pair output stage.

Alternatively, the calculated value of the current sensing resistor, R_3, for a single complementary transistor pair is multiplied by the number of paralleled output pairs, *N*, with each resistor of value NR_3, used to monitor the current in each transistor, as shown in **Fig. 50.**

Note that using the non-linear limiter of Fig. 27 in this fashion requires that each resistor of value NR_3 be shunted by the diode in series with resistor, R_d, whose value remains unchanged.

An obvious disadvantage inherent in both schemes is that the failure of a rogue transistor in one half of the output stage could result in the disastrous alteration of the protection locus for the remaining devices in that section. With modern power transistors though, this

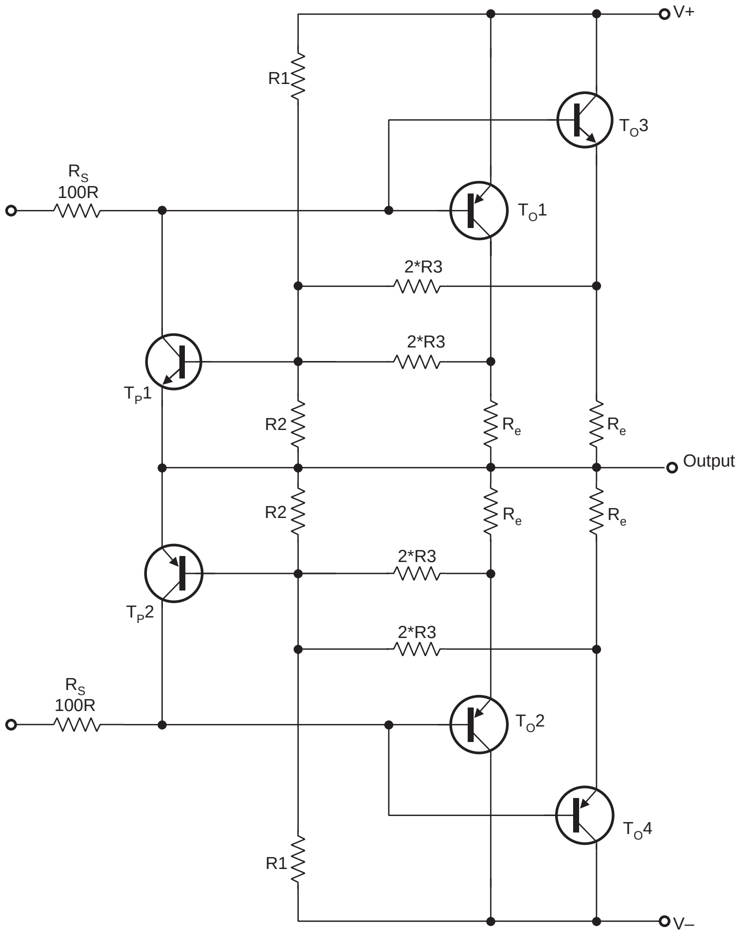


Figure: 50. In this single-slope, linear foldback scheme, voltage signals from multiple current sensing resistors are summed algebraically at the base of the protection transistor.

scenario is unlikely.

Using an independent *V-I* limiter for each complementary pair would eliminate this flaw, but such a solution would be financially indefensible for most commercial designs.

In summary

On grounds of safety and reliability, it is firmly recommend that all linear, complementary semi-conductor audio power amplifiers incorporate suitable *V-I* protection. The aversion cultivated by some designers to such is here shown to be wholly illusory.

A competently designed *V-I* limiter will remain demonstrably inert, and therefore completely unobtrusive with virtually all commercial loudspeaker systems, provided the output stage consists of sufficient complementary transistors to safely drive a 4ohms±60° load to the supply rails.

The dual-slope circuit of Fig. 27 represents a significant improvement in efficient safe operating area use relative to the single-slope topology of the previously published Fig. 1. Also, there's no significant penalty with regard to algebraic complexity.

As demonstrated in Fig. 26, the circuit's characteristic locus can be readily optimised to accommodate ±50V supply rails with MJL3281A/MJL1302A transistors. Higher supply rails are not recommended for worst-case reactive loads, as available collector current for these devices falls rapidly below 500mA with a V_{CE} of more than 100V.

Although e-MOSFETs are at least an order of magnitude less linear than bipolar transistors[4,p.273], they provide significantly greater scope for reliable design[4] at high device voltages, ($2|V_{supply}|>>100$V), with the promise of even greater efficiency in S.O.A. utilization, due to the absence of secondary breakdown. However, there is no need to endure the indignity of e-MOSFET non-linearity and on resistance voltage inefficiency in sub-200W into 8ohms designs.

More elaborate protection schemes are possible, with the use of as many diodes as the number of required breakpoints. However the increase in available current in the high voltage region, $|V_{cc}| <= V_{ce} < 2|V_{cc}|$, where it counts with respect to reactive load drive, is negligible in relation to the circuit complexity thus engendered. ■

References

1. Duncan, B. 'High performance audio power amplifiers'. Newnes, ISBN 0-7506-2629-1, p. 202, and p. 204, Fig. 5.23, respectively.
2. Motorola TVS/Zener device data book, DL150/D, REV1, Section 11. www.onsemi.com
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4. Crecraft, D.I., *et al*, 'Electronics', Chapman & Hall, ISBN 0-412-41320-5, p. 566.
5. Slone, R. S., 'High-power audio amplifier construction manual', McGraw-Hill, ISBN 0-07-134119-6, p. 244, and 260.

In Michael's previous article, "D_F" should have been written as "D_P" in three places – 6, 8 and 18 lines from the bottom of the right-hand column of page 46. Also, the words, "Diodes DP and DF are omitted in subsequent figures in the interest of clarity." should have appeared at the end of the last paragraph on page 46. In Fig. 4, the voltage at the base of T_p should have read 3V72, not 33V72, in Fig. 7, the voltage at the base of T_p should have read –39V4, in Fig. 15, the voltages at the base of T_p should have read –23V73 -> –23V4, and finally, in Figs 24 & 25, V_{ref} at the end of R_2 should have read –20V6. Apologies for these errors.