ETI-480

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1. Design Derivation and Development History

1.1 Original Publication

The forerunner of the ETI-480 power amplifier design first appeared as project ETI-422[1] in ETI May 1974 p.73, as a 50W amplifier. It was intended to supersede the 100W ETI-413 Guitar Amplifier of ETI December 1972 (see also ETI Top Projects Vol. 2, p.6), and was designed for 8Ω loads only (not 4Ω as well).

1.2 ETI-422 Re-issue

The ETI-422 design was reprinted in ETI October 1975 p.56 with some changes. 0.1µF capacitors had been added across the existing local feedback resistors in the output stage and the 100pF in parallel with the feedback resistor had been increased to 330pF. No explanation for the changes was given, but clearly some stability problems may have either been suspected or found to have occurred.

1.3 ETI-480 Publication

The ETI-480 first appeared in ETI December 1976 p.64[2,3]. The reworked ETI-422 design was intended to improve ease of construction with a simplified layout and output transistors mounting directly onto the PCB. It was designed in two versions, for 50W into 8Ω[4], and 100W into 4Ω by addition of a second pair of 2N3055/ MJ2955 output transistors[5]. The explanation included with the 1976 article as regards how the circuitry actually worked remained essentially the same and no more than that which was published previously in May 1974.

1.3.1 ETI-480 Changes Compared to ETI-422

From the preceding ETI-422 design the ETI-480 had the following changes.

- Muting (“de-thump”) FET Q1(ETI-422) 2N5485 removed.
- Separation of grounds by addition of 10Ω resistor R3[6].
- Polarity of C3 (ETI-422) reversed (as C2 in the ETI-480).
- R43, R45 (ETI-422), 0.5Ω 2W[7] replaced in each case by two 1Ω 1W resistors in parallel.
- V_{be} multiplier transistor Q13 (ETI-422) PN3643 replaced as Q6 (ETI-480) with the more common BC549.

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[1] For the schematic, refer Appendix 2.2, Figure 2.

[2] For the schematic, refer Appendix 2.2, Figure 1.

[3] For performance specification table and distortion vs. power output curve, see Appendix 2.1, Fig's 2 and 3.

[4] Ref. Appendix 2.1, Fig. 1.

[5] Ref. Appendix 2.1, Fig's 4, 5 and 6.

[6] “The 10Ω resistor provides a high frequency connection between the quiet ground track and the ground ends of the bypass capacitors; 10Ω is less than the reactance of a 100nF bypass at frequencies up to 1MHz, but is large enough to reduce hum current in the quiet ground track to the order of a milli-ampere” E.M. Cherry, JAES Vol. 29, No. 5, May 1981, pp. 327~328.

[7] “If difficult to obtain, these resistors may be fabricated from a short length of electric jug element – about 90mm is sufficient for each. Wind securely around a 1watt resistor (100Ω or higher) and solder into place” ETI October 1975 p.58.

1.3.2 20 Years On

From first publication in 1976 over twenty years have elapsed. Literally thousands of these amplifiers have been built in Australia and elsewhere, mostly as kits sourced from the major Australian electronics retail suppliers,[8] but also from blank PCB's sourced separately[9] and so on. ETI-480 kits have remained on retailer shelves in almost uninterupted supply throughout that twenty year period when many many other designs and kits have come and gone. This must surely be an incredible testiment to a basic but sound, reliable, tolerant and application adaptable circuit design.

1.4 Other Articles

1.4.1 ETI Circuits No. 4, 1983

On page 15, ETI Circuits No. 4, 1983 appeared details on operating paired ETI-480's in bridged configuration.[10] The article also gave a table of output powers for various "single ended" and bridged configurations, as variations from the original design.[11]

1.5 Superceding Projects

1.5.1 ETI-470

"Designed by Phil Wait from an original circuit by [West Australian] Trevor Marshall", the 60W[12] ETI-470 appeared in ETI May 1979, p5. It was intended to "replace the ETI-480 and features simpler mechanical construction, low distortion (particularly TID[13]) and generally better performance."

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[8] **Jaycar's** 1995 catalogue includes ETI-480 modules KE-4042 (100W) at $35.95 and KE-4050 (50W) at $28.95 on p.9. Recommended heatsinks were HH-8592 (p. 136, 2RU, 1.2deg. C/W thermal resistance, $14.95) and HH-8590 (p.136 also, 1.4deg. C/W, $9.95). The ETI480 Power Supply was listed at p.9 as KE-4080, $29.50 and suggested transformers were MF-1095 (p.116, but listed as MM-2095 at p.9, 28-0-28V, 2A, 112VA, referenced as being equivalent to type (Fergusson) PF3577 and type JT144, $44.95), or toroidal MT-2115 (30-0-30V, 2.66A, 160VA, $49.95 reduced from $66.50). The ETI-480 made Jaycar's 1996 catalogue but not their 1997 one.

Altronics 1994 and 1997/98 catalogues do not list the ETI-480.

Rod Irving Electronics, 1995~1996 catalogue, p.175 lists K10040 (50W) @ $27.95 and K10045 (100W) @ 34.95. Advertising in Electronics Australia, September 1997, p.47 lists both and at the same prices, and also includes K10050 ETI-480 Power Supply @ $28.95.

Dick Smith's 1997/98 catalogue, p.189 lists K-3440 (50W) @ $28-50 and K-3442 (100W) @ 35-50. Their 1996/97 catalogue, p.173, lists these same prices but in the 1995/96 issue, p.160, they were higher, at $29.95 and $36.95 respectively. Dick Smith lists the ETI power supply PCB kit K-3438 at $29.50 in the 1997/98 catalogue p.189, and references as suitable for two 50W or one 100W module, the transformer M-0144, p.205, @44.90, 28-0-28V, 120VA.

[9] For example, **R.C.S. Radio** Pty. Ltd., 651 Forest Rd., Bexley, NSW 2207, ph: (02) 9587 3491.

[10] Ref. 3.1.2.1 and ETI Circuits No. 4, 1983, p.15.

[11] Ref. 3.1.1 and ETI Circuits No. 4, 1983, p.15.

[12] 60W into 8Ω only; "If the amplifier is to be used with a 4Ω speaker system the supply voltage must be limited to about 30 volts maximum, otherwise the output devices will attempt to deliver over 100 watts followed by rapid self destruction!"

[13] *T*ransient *I*ntermodulation *D*istortion.

The amplifier featured Darlington device outputs[14] (Philips BDV65B and BDV64B, or alternately Texas Instruments TIP142 and TIP147[15]), and the design focused heavily on minimisation of TID.

A 0.47μF tantalum connected the bases of the input differential amplifier pair. "Input lag compensation is provided by C3, limiting the slew rate of the amplifier to reduce high frequency intermodulation". This gave the amplifier a very specific requirement to be driven from a very low impedance source to avoid distortion at higher audio frequencies. ETI needed themselves to publish[16] an "interface" circuit (again by Phil Wait) to allow ETI-470's to be appropriately driven in their Series 4000 stereo amplifier project[17]. The follow-up article also included a "Hints and Tips" section for the ETI-470, emphasising particularly the need for good constructional practice in regard to thermal considerations and the output Darlingtons.

There seems little to recommend this amplifier project to any but the most experienced amateur constructor, the presumably superb performance with respect to TIM distortion notwithstanding.

1.5.2 ETI-466

Project ETI-466, by Barry Wilkinson, appeared in ETI February 1980, as successor to both the ETI-480 and later ETI-470. Much more sophisticated, complex and expensive to construct, it is rated at 200W into 8Ω, and a little over 300W into 4Ω. It thus represents around the limit of conventionally connected, reasonably common BJT outputs[18]. The ETI-466 is a solid, reliable amplifier, and as good a foundation for uprating, variation or plain sustained dependable operation, as the ETI-480.

Despite at least one letter to ETI following the original article[19] complaining of stability problems, these are easily avoided by good construction practice, and problems encountered early on with particular kits

[14] Douglas Self (Audio Amplifier Design Handbook, Newnes, 1996, p.96) describes such devices thus; "an integrated device that includes driver, output, and assorted emitter resistors in one ill-conceived package" and certainly does not endorse their use in output stages.

[15] ETI July 1980, p.23.

[16] ETI February 1980, p.34~ 35.

[17] ETI-471 "High Performance Stereo Preamp Control Unit", Phil Wait, ETI June 1979 p.12 and SERIES 4000 "The "Series 4000" stereo amplifier", Trevor Marshall, Phil Wait, Bill Crump, ETI July 1980.

[18] Higher power variations are however indeed possible. Five MJ15024/MJ15025 pairs, twin 500VA or 650VA toroidal power supply transformers at 70-0-70Vac producing +/-100Vdc rails with various other circuit changes to uprate the ETI-466, are reported to produce 400W into 8Ω, 650W into 4Ω. However around 200W into 8Ω, 300W into 4Ω remains about the limit for conventional circuits using readily available BJT's due to secondary breakdown difficulties. Beyond these levels series outputs, Class G or H operation and so on, needs to be implemented.

[19] In ETI December 1981 p.69, a letter from A. Stewart, Gumsdale, Queensland, identified a long serious history of difficulty with an ETI-466 suffering from high frequency instability and detonating the output stage repeatedly. Roger Harrison, ETI's then editor responded with various points, also noting publication of information responding to similar troubles with the ETI-477 Mosfet amplifier, in ETI August 1981, p.11.

- Zobel network capacitor C15 needs to be of a type that exhibits little or no inductance at high frequencies; apparently some parts supplied with kits at the time were found to be misbehaving in this regard.

- Output transistor emitter resistors must, as always for any amplifier, be low inductance types; "Noble" and "IRH" brand 5W wirewound resistors work reliably, but some "noname" brands do not.

* "Noname" brand MJ15003's and MJ15004's being supplied at around that time apparently caused quite severe problems that original Motorola manufactured devices did not.

related to particular parts supplied in those kits. While minor modifications will improve reliability, there are no vital major changes to the original ETI-466 circuit necessary for operation as originally published.

2. How it Works

2.1 Input Differential Amplifier

2.1.1 General Operation

Q2 and Q3 form a differential pair, a topology that today appears at the input of all conventional amplifiers. It confers a number of advantages (over a single ended input), including an inherant distortion cancellation mechanism independent of transistor parameters such as h_{fe}. It is critical however that the design makes the collector currents of the pair equal, and this isn't quite the case in the ETI-480. Maintaining or achieving as much symmetry between the two legs of a differential amplifier is important for establishing high CMRR (Common Mode Rejection Ratio); ability to ignore input signals (noise etc.) appearing equally on both inputs.

The differential amplifier is referred to as a "*transconductance*" amplifier, meaning it takes a *voltage* (difference) input signal and produces a *current* drive output (ref. also 2.2.1). It is operated at low gain, ensuring bandwidth much wider than the Voltage Amplifier Stage ("VAS" for short) following, thereby minimising the amount of lag compensation required for stability (ref. section 2.2.2). Also critical to the design of this section is operation at "sufficient bias current in order to drive the high-frequency capacitive input impedance of the second stage [the VAS]. If this bias current is insufficient, the slewing rate of the amplifier will be degraded."[20]

The gain of the input differential amplifier is given by $R4/2r_e$[21] (since there are no emitter degeneration resistors R_E for Q2 and Q3). r_e is approximately $26/I_e$ where I_e is in milli-amps, here 1mA (ref. section 2.3.1), i.e. r_e is 26Ω. The differential amplifier gain is therefore approximately 5K6/(2 x 26) = 108.

The input impedance of the following VAS (Q5) is approximately $Z_{in} = h_{fe}(r_e + R_E) = 25 \times [(26/I_c(Q5)mA) + 470\Omega] = 25 \times [26/10.6 + 470]$[22] , or approximately 11K8 as a minimum. This impedance lowers Q2's collector load impedance to 5K6//11K8 = 3K8, and the differential amplifier stage gain to 3K8/(2 x 26) = 73.

2.1.2 DC Output Offset

The input differential amplifier provides, (for a DC coupled amplifier with advantages inherent in avoiding AC coupling), a minimal DC offset. To achieve lowest practical output DC offset the following are steps that can be taken.

[20] Prof. W. M. Leach "Build a Double Barreled Amplifier", Audio, April 1980, p.40.

[21] Gdiff = $R_c/2(r_e + R_E)$ p.99, "The Art of Electronics", 2nd. Ed., Cambridge University Press, 1989, P. Horowitz and W. Hill.

[22] For Q5 = BD139 gain ref. Appendix 3, BD139 device data, section 3.1.2. Ic(Q5) = 10.6mA as shown in section 2.3.2, and R_E = R13 = 470Ω.

- Q2 and Q3 should be gain[23] matched using a transistor h_{fe} tester, as closely as practicable, to within say 10%. The gain of each should be high, and BC549's are therefore possibly a better choice than the BC547's originally specified. Mismatched gains of say 150 and 250 give an offset of about 25mV. A gain difference between the two transistors of say 200 and 210 gives a DC offset at the output of around 2mV.

- R8 and R2 should have the same value. Where overall gain changes are needed, for example as specified to alter the 50W design for 100W operation, change R7 (and C2 to preserve the same low frequency roll off) rather than R8. R2 = 10kΩ and R8 = 4.7kΩ gives a DC offset of around 20mV.

- I_c(Q2) and I_c(Q3) should have the same nominal value. Following section 2.3.1 shows calculation of these currents, and their theoretical values. Bringing R5 closer to 2.5kΩ than the original 2.7kΩ would reduce the offset by some 5 to 10mV. Adding 33K across the existing 2K7 achieves a suitable result.

DC offset at the output occurs when there is an inbalance between the base current flowing in Q3 and then R8, and that flowing in Q2 and then R2 to ground. These currents are;

$I_b(Q3) = I_c(Q3)/h_{fe}(Q3)$ and

$I_b(Q2) = I_c(Q2)/h_{fe}(Q2)$.

These currents give rise to the following voltage drops across the respective base resistors R8 and R2;

$V(R8) = R8 \times I_c(Q3)/h_{fe}(Q3)$ and

$V(R2) = R2 \times I_c(Q2)/h_{fe}(Q2)$.

Partial or complete cancellation effectively occurs between these two, as both base currents flow via their respective resistors to ground. The DC voltage appearing at the bases is V(R8) for Q3 and V(R2) for Q2, and OV_{dc} appears (in theory) at the grounded end of R2 and at the output end of Q3.

From the above equations however it can readily be seen that imbalances will offset this relationship. If R8 is not the same as R2, if I_c(Q2) is not the same as I_c(Q3) and if h_{fe}(Q2) is not the same as h_{fe}(Q3) then output DC offset occurs to some degree. As a general rule this should be no more than +/-50mV maximum, and this figure should be considered relatively high as amplifier design and construction goes. Anything above this should be considered a "fault" and corrected. Unmodified ETI-480's will in practice show DC output offsets right up to +/-50mV. Reducing this to within +/-20mV is normally quite practicable and should be attempted. It is definitely not unreasonable to try for closer limits still.

2.2 Voltage Amplifier Stage

[23] Vbe matching of the *input* pair is not warranted. At most it gives rise to about 5mV of output DC offset (Douglas Self, Audio Power Amplifier Design Handbook, p.74). For the input pair it is *gain* that is important for matching. The reverse is the case for matching output transistors for good output stage current sharing.

2.2.1 General Operation

The Voltage Amplifier Stage (VAS), consisting of Q5 and associated components, provides almost all the voltage gain of the amplifier. Its design is critical, as it must be able to provide almost the full output voltage swing of the complete amplifier. It is also (quite conventionally) the primary site for controlling and ensuring overall stability of the amplifier.

It is important that the voltage gain of the Q5 circuit is high and as linear as possible. For these reasons an active load, constant current source[24] Q4 and associated components, is used rather than a simple resistor collector load. (The alternate and almost as well thought of design method is bootstrapping the collector resistor). The active load also enables the VAS to drive the output stage higher up to the positive supply rail in the positive direction, than a simple resistive collector load could.

The output from the differential pair is current (not voltage) drive to VAS Q5 (ref. 2.1.1). In fact the voltage appearing at the base of Q5, as observed with a CRO, can be quite misleading. It is a very low amplitude (a few millivolts and substantially less than the voltage amplitude of the signal being input to the whole amplifier in the first place), distorted triangle shape (for a sinewave test signal input).

The VAS is a "*transresistance*" amplifier, converting *current* drive at the base of Q5 to *voltage* drive at Q5 collector. The *transresistance* amplifier thus has the reverse current-voltage relationship to the input differential *transconductance* amplifier, considered earlier.

2.2.2 Dominant Pole Compensation

C7 provides what is termed "Miller dominant pole compensation" and sometimes is referred to as "C_{dom}" for short. "Compensating a three stage amplifier is relatively simple; since the pole at the VAS is already dominant, it can be easily increased to lower the HF negative-feedback factor to a safe level."[25] In other words, the VAS already has the poorest high frequency response, compared to the diff. amp. and output stage coming before and after it respectively. It's a relatively simpler matter therefore to push its high frequency roll-off point further downward, as a means of controlling overall amplifier stability.

As frequency increases the gain of an amplifier falls and phase shift imparted on the amplified signal increases. Each amplifying stage adds phase shift due to junction capacitance (this applies to both voltage amplifiers and current amplifiers in the output stage). The -3dB point for each of these amplifying stage frequency roll-offs is called a "pole". At a pole frequency "f" phase shift has already increased to 45^0. At f/10 phase shift begins and at 10f phase shift has reached the maximum for a single pole, 90^0. The rate of frequency roll-off for a single pole is "6dB/octave" (meaning halving of amplitude per doubling of frequency), or "20dB per decade" (i.e. a tenfold decrease in amplitude per tenfold increase in frequency). Each additional stage adds another pole. In a typical or conventional topolgy, such as the ETI-480, there are three consecutive stages; diff. amp., VAS, and output stage. Each contributes phase shift.

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[24] Bootstrapping is an alternative, quite acceptable, method of achieving this in amplifiers, generally speaking. Its use depends on output stage unity gain however and therefore it would not have been a suitable design option for the ETI-480.

[25] Douglas Self, Audio Power Amplifier Design Handbook, p.31.

When this phase shift at high frequencies accumulates to 180^0 then the amplifier feedback, which is 180^0 *negative* feedback anyway, becomes $180^0 + 180^0 = 360^0$ which is a complete cycle so the feedback is now *positive*. If the gain of the amplifier is anything greater than unity at the frequency at which such positive feedback occurs, then the signal will grow each time it goes round the loop; the whole circuit will oscillate. The gain of the amplifier must therefore be controlled at high frequency; deliberately rolled off early, to ensure such oscillation can't occur.

This isn't a problem provided that, by a healthy margin, gain has already fallen below unity. The margin refered to is "phase margin"; the amount of phase angle in degrees between the closed and open loop unity gain points. 45^0 is a conventional design phase margin[26]. This means that the frequency compensation added to the VAS should result in the overall amplifier gain falling to unity by the time that phase shift has reached no more than $180^0 - 45^0 = 135^0$.

If the gain at some high frequency at which amplifier phase shift has reached 180^0 is still above unity, then the amplifier becomes an oscillator, self sustaining even the slightest circuit disturbance (e.g. the effect of switching it on) leading promptly to self destruction in all likelihood. Close to but before this point the amplifier stability is marginal - high frequency peaking and overshoot starts occuring. Neither are desirable effects.

Bode showed it is also necessary, to ensure stability, that the amplifier response is being rolled off at no sharper rate than about -10dB/octave, by the time the unity gain frequency is reached. This is less than the -12dB/octave rate of a second order filter; the roll-off rate at unity gain in practice must therefore be first order; -6dB/octave.

The conventional method for ensuring overall stability is the addition of C_{dom} which acts as this "first order" filter, rolling off the amplifier response earlier than it otherwise would, at –6dB per octave. This is referred to as inserting a "*pole*" in the amplifier's frequency response, and since its value is deliberately structured to precede all naturally occurring poles, it is referred to as the "*dominant* pole". In parallel with the negative feedback critical "*Miller*" base-collector capacitance of voltage amplifier Q5, the term "Miller" is used to indicate how or where C_{dom} is connected.

As frequency rises, C_{dom} does its job causing gain to decrease. This also means available feedback around the whole amplifier decreases. However C_{dom} then continues to provide feedback around Q5 in a purely local fashion. Q5 as a single stage can have any amount of feedback applied without inherent instability problems. The position chosen to add C_{dom}, i.e. around the Voltage Amplifier Q5, thus confers additional benefit

"Feedback through C_{dom} lowers VAS input and output impedances, minimising the effect of input-stage capacitance, and the output stage capacitance. This is often known as pole splitting; the pole of the VAS is moved downwards in frequency to become the dominant pole, while the input stage pole is pushed up in frequency."[27] Adding C_{dom} decreases the load impedance at high frequencies that the input stage diff. amp. is working into. The effect of input stage (Q2 collector-base) junction capacitance on high frequency response is thereby decreased, and input stage frequency

[26] There are many texts giving explanation of this and related subjects. That in Philips Semiconductors, Application Note AN166, Basic Feedback Theory, Dec. 1988,is notably concise and readable.

[27] Douglas Self, Audio Power Amplifier Design Handbook, p.50.

response inproved. In summary, negative feedback to the VAS not only decreases the frequency of the pole associated with the VAS, but also has the effect of increasing the pole associated with the preceding stage. This pushing of the one down and the other up is called "pole splitting". The further these poles are apart the better; increasing the assurance that the C_{dom} has total control over the amplifier's complete response.

2.2.3 Other Stability Measures

"Additional HF phase correction components will almost certainly need to be used … to ensure adequate overall loop stability, but their position and value will need to be determined specifically for each new design."[28] There are four main positions in a conventional three stage amplifier for the addition of lead-lag compensation components. Unusually, all four positions are used in the ETI-480; C4, C3, C5 and C7, C5 provides phase advance or "lead" compensation by being in the feedback path, while C3, C4 and C7 are all phase "lag" compensation methods, in the forward gain path.

Conventional stabilisation technique is "always to put in one dominant lag to attenuate the loop gain at 20dB/decade (6dB/octave), starting from a corner frequency which is sufficiently low to ensure that the loop gain is reduced to unity before the other lags inevitably present at high frequencies have produced too much further phase lag."[29] A capacitor across the first stage collector load is suggested as the most straightforward method of achieving this; C4 in the ETI-480. Connecting such a capacitor, as C7 across the base-collector junction of the voltage amplifier transistor is an alternative, with a number of secondary advantages (but possible slew rate limiting difficulties, that the C3 position avoids).

Again, conventional stabilisation technique suggests that the dominant compensation pole should have a time constant as removed as possible from all others present. It appears doubly unusual therefore to see such a variety of compensation sites filled in the ETI-480 design. Without a doubt however, the ETI-480 is a notably stable circuit; the phase compensation, its complexity notwithstanding, is certainly effective!

2.2.3.1 R9 and C3

The series network R9, 1K2, and C3, 100pF, is connected between the collector of the voltage amplifier Q5 and the base of differential amplifier transistor Q3, and thereby reduces the gain of the Voltage Amplifier Stage as frequency increases, to assist overall loop stability. With R9 limiting the charge current of C3, this method of stabilisation is intended to impose less current demand on the voltage amplifier than C_{dom}, and hence less slew rate limiting.[30]

2.2.3.2 C4

C4, 3n3, is connected across differential amplifier load resistor R4, reducing the gain of the stage as frequency increases, to assist overall loop stability.

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[28] John Linsley Hood, "The Art of Linear Electronics", p.163.

[29] Audio Power Amplifier Design - 4, P.J. Baxandall, Wireless World, July 1978, p.76.

[30] "Regrettably, [*the Cdom topology*] is normally used in commercial audio amplifier designs because it gives slightly better harmonic distortion figures at high frequencies." John Linsley Hood, "The Art of Linear Electronics", p.163.

2.2.3.4 C5

C5, 330pF, is connected in parallel with feedback resistor R8, increasing feedback at high frequencies and thereby reducing gain. "…it is often advantageous to a capacitor of quite small value across the feedback resistor … sufficient to cause a little phase advance around the unity-loop-gain frequency and a reduction in the rate of attenuation of loop-gain at frequencies above this".[31]

2.3 Constant Current Sources

The input differential pair Q2, Q3 and the Voltage Amplifier Q5 are operated with constant current sources. These increase the effective load impedance at which Q2, Q3 and Q5 are operated, without requiring a correspondingly large $V_{(B+)}$ to provide the same collector currents.

R6 sets a reverse current through Zener ZD1 between about 7.5mA and 10mA[32]. ZD1 in turn fixes the voltage across the series combinations of R5 and Q1 base–emitter junction, and Q4 base-emitter junction.

2.3.1 Q1, for the Differential Pair

Q1 is a constant current source supplying around 2mA, which is shared by Q2 and Q3. Use of a constant current source to control the current for the input differential pair results in improved rejection of power supply ripple on the $V_{(B+)}$ supply[33], and better rejection of common mode input signals[34] than a simple tail resistor.

The 5.6V set by Zener ZD1 across R5 and Q1 base-emitter, results in a fixed or constant current through R6, and hence Q1 emitter-collector, and hence the differential pair Q2 and Q3, of;

$$I(Q1) = (V_z - V_{be\ Q1}) / R5$$

$$= (5.6 - 0.6)V / 2{,}700\Omega$$

$$= 1.85mA$$

Constant current I(Q1) is shared between Q2 and Q3. Ignoring transistor base currents, R4 determines the quiescent current through Q2, and 1.85mA minus this Q2 current is that which flows through Q3. Observing from 2.3.2 that the I(Q4) constant current set up by Q4 flows through R13, the quiescent R4 current is determined by;

$$I(Q2) = [V(R13) + V_{be}(Q5)] / R4$$

$$= [I(Q4) \times R13 + V_{be}(Q5)] / R4$$

$$= [10.6mA \times 470\Omega + 0.6] / 5600$$

$$= 1mA$$

[31] Audio Power Amplifier Design - 4, P.J. Baxandall, Wireless World, July 1978, p.77.

[32] For V(B+)max = 40Vdc and V(B+)min = 30Vdc, I(R6) = (40-5.6)V/3K3Ω = 10.4mA, and (30-5.6)V/3K3Ω = 7.4mA.

[33] The Art of Linear Electronics, John Lindsley Hood, Butterworth Heinemann, 1993, p.90.

[34] Douglas Self, Audio Power Amplifier Design Handbook, p.60.

Since $I_c(Q2)$ = 1mA, then I(Q3) = 1.85mA – 1mA = 0.85mA.

There are benefits in terms of substantially less (second harmonic) distortion to be gained from maintaining strict balance between the Q2 and Q3 currents in the differential amplifier[35]. Changing R5 from 2K7Ω to 2K5Ω (perhaps as a series combination of 1K5Ω + 1KΩ) increases I1 to 2mA and hence, since the change has no effect on I(Q2), increases I(Q3) also to 1mA.

2.3.2 Q4, for the Voltage Amplifier

The higher dynamic collector load impedance, provided for voltage amplifier Q5 by constant current source Q4, provides greatly increased gain and linearity. The 5.6V set by Zener ZD1 across R5 and Q4 base-emitter results in a constant current through R10, and hence Q5 emitter-collector and R13, of;

$$\begin{aligned} I(Q4) &= (V_z - V_{be\ Q4}) / R10 \\ &= (5.6 - 0.6)V / 470\Omega \\ &= 10.6mA \end{aligned}$$

2.4 V_{be} Multiplier

"The resistors R11 and R12 together with potentiometer RV1 control the voltage across Q6 and maintain it at about 1.9V. But as Q6 is mounted on the heatsink, this voltage will vary with heatsink temperature. Assuming that the voltage on the bases of Q7 and Q8 is equally spaced about zero volts (i.e. 0.95V) the current will be set at about 12mA through Q7 and Q8. The voltage drop across the 47Ω resistors (R14, R18) will be enough to bias the output transistors Q9 and Q10 on slightly to give about 10mA quiescent current adjustable by means of potentiometer RV1."

"Temperature stability is attained by mounting Q6 on the heatsink and this transistor automatically adjusts the bias voltage."[36] Its collector-emitter current increases with temperature rise caused by heat dissipation in the output transistors. Increased Q6 collector-emitter current causes the voltage dropped between Q6 collector and emitter to therefore decrease. Reducing the voltage difference between these two points means reducing the forward DC or quiescent bias on Q7 and Q8, and hence the output devices. Thus the increased quiescent current produced in the output transistors by the temperature increase that caused the change in the first place, is countered[37].

[35] "Exact DC balance of the input differential pair is essential in a power amplifier. It still seems almost unknown that minor deviations from equal Ic in the pair seriously upset the second-harmonic cancellation….." Douglas Self , Audio Amplifier Design Handbook, Newnes, 1996, p.65.
[36] ETI December 1976.

[37] Critical however is the tracking of this thermal feedback path. Inevitably time delay occurs between a change in audio drive signal such as to cause a change in temperature in the output transistors and hence a change in their bias point, and thermal sensing by Q8 with consequent countering of the quiescent bias change.

2.5 Output Stage

2.5.1 General Operation

The ETI-480 uses one of the most ubiquitous complimentary output transistor pairs of all; the 2N3055 and MJ2955[38]. V_{cer} is 70V, I_c is 15A, P_{diss} is 115W and they are TO 3 package devices; one pair in the 50W version[39], and two pairs in parallel for the added requirements of 100W operation[40].

The circuit topology employed provides *voltage* gain with local negative feedback applied, as well as current gain, and is a variation on the conventional Complimentary Feedback Pair (CFP)[41] configuration[42].

The CFP confers significant output stage linearity advantages over a purely emitter follower topology, lower quiescent dissipation and better thermal stability.

2.5.2 Design Calculations

Design calculations for 50W and 100W output into 8Ω and 4Ω respectively, may be summarised as follows[43].

2.5.2.1 Peak Voltage Delivered to the Load

[38] For 2N3055 and MJ2955 component data see Appendix 3, Fig's. 1,2 and 3.

[39] See Appendix 2.1, Fig. 1.

[40] See Appendix 2.1, Fig's 4, 5 and 6.

[41] "There seems to be only one popular configuration [*of the CFP*], though versions with gain are possible" Douglas Self , Audio Amplifier Design Handbook, Newnes, 1996, p.99. Ref. also p.121 "The CFP topology has good quiescent stability and low LSN [*Large Signal Nonlinearity*]; its worst drawback is that reverse-biasing the output bases for fast switchoff is impossible without additional HT rails".

Prof. W. L. Leach in "Build a Double Barreled Amplifier", Audio, April 1980, p.40, provides an opposite view. "Because this [*CFP*] connection forms a negative feedback path from the collectors of the output transistors back into the emitters of the driver transistors, a large reduction of static distortions in the output stage can be realised. It is felt, however, that the [*emitter follower output topology*] results in a more stable and better sounding amplifier. This is because the output transistors in the [*CFP*] are operated in their slowest configuration. The driver transistors are forced to supply a higher and higher share of the load current as the frequency is increased, which in turn causes the high-frequency output impedance of the amplifier to increase, resulting in a reduced high-frequency damping factor."

[42] Also called a "Sziklai Pair"

[43] The design procedure that follows is drawn directly from an exceptionally interesting, practical and readable article by David Eather, "A Practical Approach to Amplifier Output Design", Silicon Chip, February 1991, pp14~18 and April 1991, pp.64~67.

$V_{max\ load}$[44] $= \sqrt{(2 \times P \times Z)}$

For 50W/8Ω, $V_{max\ load} = \sqrt{(2 \times 50 \times 8)}$

$= 28.3V_{pk}$

For 100W/4Ω, $V_{max\ load} = \sqrt{(2 \times 100 \times 4)}$

$= 28.3V_{pk}$ also.

2.5.2.2 Peak Current Delivered to the Load

$I_{max\ load} = \sqrt{(2 \times P/Z)}$

For 50W/8Ω, $I_{max\ load} = \sqrt{(2 \times 50/8)}$

$= 3.54A$

For 100W/4Ω, $I_{max\ load} = \sqrt{(2 \times 100/4)}$

$= 7.07A$

2.5.2.3 Selection of Emitter Resistors

"As a guide you would normally try for about 0.6 volts across the emitter resistors at Imax load … These resistors help provide thermal stability of the output stage bias current and, in designs with output transistors in parallel, they help to ensure equal current sharing. The higher the resistance the better the thermal stability and current sharing but the more power they waste. The final value is a compromise."[45]

This suggests R19//R25 and R24//R26 in the 50W version equalling 0.6V/3.54A = 0.17Ω ,say preferred and readily obtainable value 0.22Ω[46].

In the ETI-480 the designer clearly has needed to make allowance for possible wide variation in output transistor characteristics and no V_{be} or gain matching at all, poor heatsinking and so on, and has opted for a conservative 0.5Ω value. In addition, by paralleling two 1W carbon composition resistors to achieve this value, all possibility of inductive wirewound instability problems is removed at a stroke. Reducing the combined value of the parallel resistors in each case to anything above 0.18Ω 1W is therefore a suitable modification where effective heatsinking and parallel output transistor V_{be} or gain matching is undertaken.

2.5.2.4 Required supply voltage

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[44] From $P = V^2/R$, ⇒ $Prms = Vrms^2/Rload$, and where $Vrms = Vpk/\sqrt{2}$, then by inserting this and rearranging, $Vpk = \sqrt{(2 \times Prms \times Rload)}$.

[45] David Eather, "A Practical Approach to Amplifier Output Design", Silicon Chip, February 1991, pp.14~18 and April 1991, pp.64~67.

[46] With R27//R29, R28//R30 in the 100W version 0.6V/(7.07Ax0.5) = 0.17Ω again. The calculation is based on the assumption that basically equal current sharing occurs between the *two* parallel pairs of output devices used.

$V(B+/-) = V_{overhead} + V_{ripple} + V_{max\ load}$

$V_{overhead} = I_{max\ load}$ x R19//R25 $+V_{ce\ sat(Q9)}$ [about 0.6V at 4A I_c, about 0.9V at 8A I_c];

$V_{overhead}$ = (3.54A x 0.5Ω) + 0.6V

= 2.4V [for the 50W case], and

= ([7.07Ax0.5[47]] x 0.5Ω) + 0.9V

= 2.7V [for the 100W amp.]

V_{ripple} = 6,300 x $I_{max\ load}$/C, where V_{ripple} is $V_{pk\text{-}pk}$ ripple on the power supply, and C is the filter capacitor size in µF on *each* rail. For the original design the capacitance on each rail (ETI-480 Power Supply[48] C2//C4 and C3//C5) is 5,000µF.

V_{ripple} = 6,300 x 3.54A/5,000µF

= 4.5V for the 50W case, and

= 6,300 x 7.07A/5,000µF

= 8.9V for the 100W version, and thus

$V_{(B+/-)}$ = 2.4 + 4.5 + 28.3

= 35.2V for the 50W amplifier, and

= 2.7 + 8.9 +28.3

= 39.9V for the 100W amplifier.

Since the V_{cer} of the output transistors limits the supply to little more than $70V_{dc}/2$ = +/- $35V_{dc}$, the 50W case is achievable but the 100W case is not. In practice, with a fairly stiff power supply, around 60W is achievable into an 8Ω resistive load and about 90W into a 4Ω resistive load.

From this, but taking into account quite clearly the V_{cer} limitation[49] of the output pair, $V_{(B+/-)}$ has to be +/-$40V_{dc}$ to +/-$42V_{dc}$ (but definitely no more than this), falling to +/-$35V_{dc}$ on full load. Power supply voltage sag any further than this reduces available output power, and failure to droop in this way risks failure of the output devices.

In practice the ETI-480 is much more tolerant of variations than the above apparently strict supply voltage "window" suggests. The relatively low VA rating of transformers typically used to power these amplifiers usually ensures sufficiently poor voltage regulation (and hence voltage sag at high load currents) to avoid exceeding V_{cer} limits.

[47] Allowing for *two* devices sharing the output current this time.

[48] For the schematic of the original power supply design, see Appendix 2.2, Figure 3.

[49] Ref. 2.8.2.

2.5.2.5 Transistor Load Lines

"…we need to make an estimate of the maximum phase shift caused by the inductive portion of the speaker load. 45^0 seems to be the accepted standard…"[50]

Tables 1~3, Appendix 1, show the results of calculations needed to plot graphically the load lines for the output transistors for the 50W amplifier, with 45^0, 60^0 and 90^0 maximum phase shift. Fig. 1, Appendix 1, shows these results plotted graphically. Tables 4~6 show the corresponding data for the 100W version and Fig. 2, Appendix 1, shows these results plotted too.

The tables are constructed as follows.

- $\omega t-\theta$ is stepped from 0^0 to 180^0 in 15^0 increments.

- ωt is then $\omega t-\theta$ with θ, the maximum phase angle for which the calculation is being done, added back in.

- $I_c = I_{max\ load} \times \sin(\omega t-\theta)$
  where Imax load is 3.54A and 7.07A for the 50W and 100W versions respectively.

- $V_{ce} = V_{[B+]} - [I_{max\ load} \times R_{load} \times \sin(\omega t)] - [I_c \times R_e]$
  where R_e is the emitter resistance value, $1\Omega//1\Omega = 0.5\Omega$

- $P_{pk} = V_{ce} \times I_c$

The tables show the output device V_{cer} and I_c parameters being exceeded only marginally in the 45^0 case, but certainly in the 60^0 and 90^0 cases[51]. Practical usage shows this to be at an acceptable degree.

The same calculations are now done for the driver transistors[52]. The load impedance presented to the drivers is $R_{load} \times H_{fe\ min}$[53]. $H_{fe\ min}$ is the minimum gain of the output transistors; a figure of 20 for the 2N3055. I_{max} for the drivers is $I_{max\ load} / H_{fe}$[54].

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[50] David Eather, "A Practical Approach to Amplifier Output Design", Silicon Chip, February 1991, p.15. However designing for 90^0 is fundamental to Douglas Self; reactive load impedances can "…double the Vce seen by the output devices. It is therefore necessary to select a device that can withstand at least twice the the sum of the HT rail voltages, and allow for a further safety margin on top of this." Audio Power Amplifier Design Handbook, Newnes, 1996, p.275. In practice this limits power output quite severely; 2N3055/MJ3055 pairs would serve for 12 to 15W into 8Ω only, off 18 to 20V rails, rather than the ratings seen in the ETI-480 for example. The message is clear however; protection circuitry is mandatory if the full Vcer capability of the 2N3055/MJ2955 pair is to be utilised, or output transistor series connection topology needs to be implimented to avoid secondary breakdown limitations.

[51] 45^0, 60^0 and 90^0 phase angles with respect to load impedances of 8Ω and 4Ω correspond to complex impedances as follows;

Impedance	Phase Angle	Complex Impedance
8Ω	45^0	5.65 + j5.65Ω
8Ω	60^0	4 + j6.9Ω
8Ω	90^0	0 + j8Ω
4Ω	45^0	2.8 + j2.8Ω
4Ω	60^0	2 + j3.5Ω
4Ω	90^0	0 + j4Ω

[52] The device data for the BD139 and BD140 driver transistors is shown in Appendix 3, Fig's. 4,5 and 6.

[53] 8Ω x 20 = 160Ω, and 4Ω x 20 = 80Ω.

[54] 3.54A / 20 = 0.18A, 7.07A / 20 = 0.35A.

Tables 7~9 and 10~12 show the data for the 50W and 100W versions respectively, at 45^0, 60^0 and 90^0 maximum phase shift, and Fig. 3, Appendix 1, shows these results plotted graphically in turn.

2.5.3 Output Stage Voltage Gain

Local feedback is applied to the output stage by the network R20~R23, giving the output stage a voltage gain of about four[55]. The overall feedback resistor, R8, gives the required gain control." The local voltage gain of the output is demonstrated by applying op-amp gain calculation methods to Q7 and Q9[56] jointly acting as a non-inverting voltage amplifier. R20 in parallel with R21, (220//220 = 110Ω), makes the feedback resistor connected between output and inverting input, and the relationship of this to R15 sets the gain;

$$A_v = (R20//R21 + R15) / R15$$

$$= (110 + 33) / 33$$

$$= 4.33 \ (+12.7\text{dB})$$

2.5.4 Zobel Network

The ETI-480 includes a shunt Zobel network, C11 0.1uF and R16 10Ω 1W, for stability into inductive loads, but no series output inductor with parallel damping resistor for stability into capacitive loads and long speaker cables.

The Zobel network resistor approximates to the expected load and can be a wirewound type without reducing its effectiveness. Use of a 5W resistor rather than the 1W original would be more likely to prevent its burn-up in the event of high frequency instability, and therefore is considered a useful modification to the original design. It is important that the capacitor in the Zobel network, C11, is a non inductive type, to avoid instability problems.

Lack of an output inductor in the ETI-480 should definitely be corrected in practice. All amplifiers should include such an inductor, as load capacitance contributes additional lag phase shift to the overall negative feedback loop. Note particularly that the Zobel network comes

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[55] "The almost universal choice in semiconductor power amplifiers is for a unity-gain output stage, and specifically a voltage follower. Output stages with gain are not unknown [*example cited*] but they have significantly failed to win popularity. Most people feel that controlling distortion while handling large currents is quite hard enough without trying to generate gain at the same time." Audio Power Amplifier Design Handbook, Douglas Self, Newnes, 1996, p.91.

[56] Both Q7 and Q9 are inverting, so when considered as a pair the amplifier is non-inverting. Hence the base of Q7 is the non-inverting terminal (of the pair jointly considered as an op-amp), and the emitter of Q7 is the inverting terminal.

before the output inductor; i.e. between the amplifier and the output inductor, *not* between the inductor and the load[57].

In practice, a 7μH to 14μH air-cored[58] inductor is used, with a 10Ω damping resistor in parallel to avoid resonance problems between the inductor and load capacitance. A wirewound type is okay but unnecessary, as the power requirement for this resistor is very low. A 1/2W or 1W type is more than sufficient, as the power dissipated even under worst case conditions is of the order of mW.

2.6 Protection

2.6.1 Fuses

2.6.1.1 DC Supplies

"Protection of the amplifier (against shorted output leads) is provided by fuses in the positive and negative supply rails to both amplifiers." These do not protect the output devices but are intended to minimise collateral damage when the output devices have already failed[59], and so should be of the *slow* blow type.

David Eather[60] however recommends a *quick* blow fuse for this position, and suggests $I_{max\ load}/3.18$ as a suitable empirically determined start point for value selection. This would suggest 3.54A/3.18 = 1.11A, say 1A or 1.2A for the 50W/8Ω case, and 7.07/3.18 = 2.22A, say 2.5A for the 100W/4Ω case. "This size fuse should allow the amplifier to produce a continuous sinewave output and allows a bit of clipping during music. Gross levels of clipping should blow the fuse." The values specified in the original design are 1.5A and 3A and although not specified appear to be *quick* blow types.

2.6.1.2 Speaker Lines

Speaker fuses are not included in the original design, are not dependable as a means of DC offset protection in the event of amplifier failure, invite wrong value replacement, and should not be considered as a possible useful addition to the ETI-480 (or any other amplifier) design.

2.6.1.3 Mains

The original design of the ETI-480 power supply did not include a mains fuse. Adding one is a sensible modification.

The mains fuse used with the amplifier module or modules should be a *slow* blow type. Its value is calculated by

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57 Strangely, amplifier designs are regularly seen, certainly from the same era as the ETI-480, either with this order reversed or an additional RC network placed after the output inductor. Literature explaining or supporting this topology is notably absent.

58 The inductor is preferably air cored to avoid possible distortion due to saturation of a ferrite core. This however means much greater power loss due to the coil resistance or a physically very much larger component or both. "The resistance of an air-cored 7uH coil made from 20 turns of 1.5mm diameter wire (this is quite a substantial component 3cm in diameter and 6cm long) is enough to cause a measurable power loss into a 4Ω load…" Audio Power Amplifier Design Handbook, Douglas Self, Newnes, 1996, p.159.

59 Audio Power Amplifier Design Handbook, Douglas Self, Newnes, 1996, p.276.

60 David Eather, "A Practical Approach to Amplifier Output Design", Silicon Chip, April 1991, p.67.

approximately determining the maximum current to be drawn from the mains, and then making suitable allowance for power drawn by other sections of the complete amplifier.

Maximum power to be delivered to the amplifier (in turn to be dlivered as power to the speaker or lost as heat) is V_{B+} x $I_{max\ load}$. V_{B+} is V_{sec} of the mains transformer x √2. Thus,

$240V_{ac}$ x $I_{fuse\ ratin}$g = V_{B+}/√2 x $I_{max\ load}$

$I_{fuse\ rating}$ = V_{B+} x $I_{max\ load}$ x 0.71/ 240[61]

This suggests a 500mA fuse for one ETI-480 50W module, 1A for a stereo version, and 1A for a 100W version, 2A for stereo.

2.6.2 Addition of Catching Diodes

Reversed biased diodes, e.g. 1N5404's, between each supply rail and the output, limit the voltage appearing at the output due to energy discharge back into the amplifier and the supply caused by inductive load effects, to one diode drop above each rail DC voltage. They are an easy and worthwhile addition to the original circuit design.

2.7 Gain

2.7.1 Calculation Method

The voltage gain of the amplifier as a whole is given by applying op-amp gain calculation methods, and treating the amplifier as a non-inverting op-amp, with input applied to the non-inverting terminal (base of Q2), and feedback applied to the inverting terminal (base of Q3). R8 is the feedback resistor connected between output and inverting input, and the relationship of this to R7 sets the gain.

2.7.2 As 50W into 8Ω

In this case;

$$A_v = (R8 + R7) / R7$$

$$= (10K + 220) / 220$$

$$= 46.5 \ (33.3dB)$$

For an output of $20V_{rms}$, being the voltage required to produce 50W into 8Ω, this suggests a required amplifier input voltage for full output of;

$$V_{in\ rms} = 20/46.5$$

$$= 430mV$$

2.7.3 As 100W into 4Ω

For the 100W version, R8 is changed to 4K7Ω[62];

[61] Ibid.

A_v = (4K7 + 220) / 220

= 22.4 (27dB)

For an output of $20V_{rms}$, being the voltage required to produce 100W into 4Ω, this suggests a required amplifier input voltage for full output of;

$V_{in\ rms}$ = 20/22.4

= 900mV

2.7.4 Other Components

C5 in parallel with R8 is present to help provide overall stability (ref. also 2.2.4.3 and C7, C_{dom}, 2.2.2).

C2, 100µF, in series with R7, sets low frequency roll off. Its impedance increases as frequency decreases, progressively increasing the effective value of R7 in the gain formula shown above. At DC "R7" therefore is an open circuit, there is 100% negative feedback via R8, and the amplifier therefore has unity gain.

The -3dB point occurs when closed loop voltage gain has fallen to 0.707 (1/√2) times that calculated previously (at 2.7.2 and 2.7.3). At 100µF C7 sets a -3dB frequency at abot 7Hz in both the 50W and 100W version cases.

2.8 Power Supply

2.8.1 General Operation

"The power supply uses a full wave rectifier and a centre tap to derive +/-$40V_{dc}$. " "The recommended power supply… gives about $40V_{dc}$ on no load, dropping to about $32V_{dc}$ on full output. This allows reproduction of transients beyond 50W (or 100W) whilst providing a degree of protection for the output transistors. If a regulated supply is used it should not be higher than +/- $35V_{dc}$."

A 28-0-$28V_{ac}$, (or $56V_{ac}$ centre tap), transformer is most suitable in this application, giving an off load DC voltage after rectification of slightly less than $28V_{ac}$ x √2 = +/-$39.6V_{dc}$. One such regularly supplied transformer has a rating of 120VA and is suggested by the supplier as being suitable for two 50W modules or one 100W module. Such loading certainly ensures sufficient supply voltage sag is going to occur at full load and hence safe operation in respect of V_{cer} for the output transistors, as well as maximum possible power output. The more commonly found 25-0-$25V_{ac}$ secondary voltage gives only $25V_{ac}$ x √2 = $35.3V_{dc}$. Such a transformer needs to have a substantially higher VA rating and hence correspondingly less voltage sag on full load, to fully exploit the power output capabilities of the amplifier(s).

As a rule of thumb, allow one and a half to two times the desired power output from the amplifier or amplifiers connected to the power

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[62] For the 100W version R8 is decreased from 10KΩ down to 4K7Ω, so R2 should also decrease from 10KΩ to 4K7Ω to minimise output DC offset. This halves Zin. Instead, increasing R7 from 220Ω to 470Ω gives a similar gain reduction (to 22.3, 27dB), leaving R8 (and R2) at 10KΩ and Zin as previously. The change to R7 however requires C2 reduction from 100µF to 47µF to give the same overall –6dB low frequency cutoff.

supply, as a suitable VA rating for the mains transformer. For professional or semi-professional use choose the higher of the two fugures. For example, a 75~100VA transformer would probably be suitable for a single 50W ETI-480 module where only 8Ω loads were envisaged. Two modules operating in bridge mode into 8Ω to produce 200W would certainly suggest a 300~400VA transformer. Opt for toroidal transformers at every opportunity as they surpass E-I types in all regards in this application, including availability and cost.

2.8.2 Output Transistor V_{cer}

A limiting factor for the power supply voltage, and hence the power output into 8Ω and 4Ω loads, (and therefore about which the rest of the amplifier has been designed), is the V_{cer}[63] of the 2N3055/MJ2955 pair, of 70V. When biased OFF, each output transistor Q9, Q10 must sustain the difference between the nominally fixed supply at its emitter, and full supply from the opposite rail when its opposite output transistor is fully ON.

Allowing for some sag or droop of the power supply voltage under these full drive conditions, from say +/- $40V_{dc}$ down to say +/- $35V_{dc}$, and making no allowance for V_{ce} drops or voltages dropped across emitter resistors, the supply is limited to $70V(V_{cer})$ /2 = +/- $35V_{dc}$. As an approximation, this suggests nominal power output capabilities of 50W into 8Ω and 80~85W into 4Ω. This suggests that 50W into 8Ω is achievable, but 100W into 4Ω requires a stiff supply (i.e. almost no droop below +/-$35V_{dc}$ on high load, but little more than this level off load).

2.8.3 Filter Capacitors

“The filter capacitors should be a minimum of 100µF-200µF per watt of output power for a class B amplifier with a full wave rectifier[64]". This suggests 5,000µF to 10,000µF per rail for the 50W version, and 10,000µF to 20,000µF per rail for the 100W version. This is considerably more than the two 2,500µF (i.e. 5,000µF) per supply rail provided in the original design, and an area of the amplifier suitable for modification accordingly. While home hi-fi probably only warrants the lower of the two totals suggested, the reverse is certainly true for professional or semi-professional applications.

3. Variations

3.1 Bridge Mode Operation

ETI-480 modules are certainly suitable for bridge mode operation and quite a number of examples exist of paired uprated modules with higher rated output transistors and higher DC supplies producing in excess of 200W into 8Ω. In bridge mode each amplifier effectively "sees" only half the load. Hence 8Ω is the minimum sensible load, meaning that each module "sees" 4Ω. Use of a 4Ω load in bridge mode means that each module "sees" 2Ω, which is courting failure unnecessarily.

It should further be clearly remembered that in bridge mode neither speaker output terminal is at ground potential; shorting either speaker connection to chassis or

[63] Vceo for both is 60V, and is the rating with Ib = 0, i.e. the base terminal open circuited. Vcer is a similar rating, but with Ib allowed to flow, by connection of Rb-e of 100Ω.

[64] David Eather, “A Practical Approach to Amplifier Output Design”, Silicon Chip, February 1991, p.15.

ground will produce a short circuit across the output of one module and associated failure.

3.1.1 Output Power

Output power for two amplifiers in bridge mode into 8Ω is theoretically four times the output power of one module into 8Ω. This assumes the power supply has sufficient VA rating for its output voltage to be unaffected by the substantial additional load, and no additional losses due to the significantly higher currents in each output stage. This does not occur in practice.

Realistically, the paired modules will produce into 8Ω twice what each will deliver into 4Ω, which makes simple sense as each module effectively "sees" half the load when connected in bridge mode. Unmodified ETI-480's indeed can deliver between 150 and 180W into 8Ω, depending on the DC supply voltage and how stiff is the supply.

ETI-480 TABLE OF OUTPUT POWER FOR VARIOUS CONFIGURATIONS[65]

Supply Voltage	**2N3055/MJ2955**	**MJ802/MJ4502**	**MJ15003/MJ15004**
+/-30V_{dc} single ended	35W into 8Ω 65W into 4Ω	35W into 8Ω 65W into 4Ω	35W into 8Ω 65W into 4Ω 100W into 2Ω
+/- 40V_{dc} single ended	50W into 8Ω 100W into 4Ω		
+/- 45V_{dc} single ended	65W into 8Ω	100W into 8Ω 150W into 4Ω (*)	100W into 8Ω 190W into 4Ω 300W into 2Ω (*)
+/- 30V_{dc} bridged	75W into 16Ω 150W into 8Ω(*)	75W into 16Ω 150W into 8Ω	150W into 8Ω 190W into 4Ω
+/- 45V_{dc}	100W into 32Ω	190W into 16Ω	195W into 16Ω 355W into 8Ω (*)

(*)Note; Not recommended, unless for home hi-fi.

At output powers above 100W_{rms} extreme heatsinking and/or forced air-cooling is recommended. Not all configurations are necessarily safe for continuous output power operation. When MJ802/ MJ4502 or MJ15003/ MJ15004 are used supply fuses F1, F2 must be changed to 5 amps.

3.1.2 Drive Methods

3.1.2.1 From the Output of One Module

"To bridge two modules, one must first get the modules working to specification, then add a 10KΩ, 1/2W resistor from the junction of R8 and R9 on module 2 to the output stage, junction of R22 and R21 on module 1. Audio input is then provided to module 1 and output is taken from module 1 and 2 output stages. The input to module 2 may optionally be shorted [*to input ground*] but in practice makes little difference."[66]

[65] ETI Circuits No. 4, 1983 p15, compiled by G.T. Dicker of Parkholme, S.A.

[66] ETI Circuits No. 4, 1983 p.15.

3.1.2.2 From a Separate Inverter Circuit

Various amplifier bridge operation[67] inverter/driver adaptor circuits have been published[68]. These provide a unity gain inverting amplifier stage that has its input the same signal as drives the first amplifier module, and its output driving the second power amplifier module.

3.2 Uprating

3.2.1 General

The ETI-480 tolerates uprating well, within sensible reason. Few if any stability problems, or problems effecting the viability of the circuit, are likely. Device changes are of course necessary as most of those used in the original circuit are working at or near their V_{ceo} limits. These are 60V (V_{ceo}) and 70V(V_{cer}) for the 2N3055/ MJ2955 output pair, and 80V(V_{ceo}) for the BD139/ BD140 drivers. This limits the rail voltage to about +/-40V_{dc} off load, and about 32~35V_{dc} at full load, and consequently the available power output into 8Ω and 4Ω loads.

How much it is worthwhile uprating the ETI-480 needs however to be given a little thought. The ETI-480 is a simple, reliable circuit. Attempting to turn it into something very much bigger than the original is inevitably going to require so much recalculation and redesign in all parts of the circuit, that effectively a completely new amplifier results; untested, unproven and probably unreliable too. Far better surely, to start with a design based originally around higher output power from the start.

Consequently there is a sensible limit to uprating ETI-480's before it becomes more practical to change to another (bigger and better suited to the purpose) design in the first place. Lifting the rail voltages to around 50~55V_{dc} to achieve power outputs at the 100W into 8Ω, 150W into 4Ω level, is about as much as it is really worthwhile attempting. The next significant increase in power is surely to 200W/8Ω, 300W/4Ω, and for this purpose very much better, purpose designed circuits are readily available.

3.2.2 Device Changes

All the transistors used in the ETI-480 need, unfortunately, to be changed to accommodate higher power supply rails due to the V_{ceo} limitations. This is in itself a further testiment to the original design, as it exploits the full capacity of all the devices used so well.

3.2.2.1 Outputs

There are quite a number of TO-3 packaged output pairs available to replace the 2N3055/ MJ2955 complimentary pair. Look for V_{ceo} to be greater than the sum of the power supply rails; for +/-50V_{dc} rails for example, V_{ceo} needs to be 100V or more. As a rule of thumb to guess at how many parallel output pairs are going to be necessary, allow 20% of the combined P_{diss} ratings of one pair, as useful output power. A 2N3055/ MJ2955 pair, at

[67] Refered to as amplifier "strapping" in the US in some literature, rather than bridging.

[68] As examples; "Build An Amp Strapping Circuit, W.L. Leach, Audio, February 1979, pp.40~43, "Build a Polarity Inverter", R.J. Kaufman, Audio, December 1987, pp.70~73, "Bridge Adaptor for Stereo Amplifiers", Leo Simpson, Electronics Australia, June 1985, pp.86~89, and "Project 479", G. Nicholls, D. Tilbrook, E.T.I. March 1982, pp.32~35.

combined P_{diss} of 115 + 150 = 265W, is good for about 20% x 265 = 53 say 55W; i.e. one pair is suitable for the 50W/8Ω version of the ETI-480 but two pairs are necessary for the 100W into 4Ω case. Similarly, one pair of MJ802/ MJ4502's is "worth" about 20% x (200+200) = 80W, and one pair of MJ15003/MJ15004's 20% x (250+250) = 100W.

If uprating an existing ETI-480 amplifier then the number of output devices is limited by the heatsink bracket present, to one or two output pairs as the case may be, if making another bracket (a fairly time consuming task requiring care and accuracy) is to be avoided. In practice there is little need to having three or more output pairs. Using other than TO-3 cased outputs seems unwarranted too; while flat pack plastic cased outputs are increasingly more readily available, the marriage of these to ETI-480 PCB's appears ineligant to say the least. Again there exists no shortage of amplifier circuits far more suited to such device packages in the first place.

The MJ802/ MJ4502 pair has V_{ceo} of 90V (but V_{cer} of 100V), I_c = 30A and P_{diss} of 200W. With rails of +/-50V_{dc} off load, +/-45V_{dc} at full load, power output of about 90W into 8Ω and 135W into 4Ω might be reasonably expected. One pair of output devices might suffice in the 8Ω case, but two would clearly be necessary for 4Ω operation with reasonable reliability.

The MJ15003/ MJ15004 pair costs little if anything more than the MJ802/ MJ4502 pair. Ratings are V_{ceo} at 140V, I_c at 20A and P_{diss} of 250W. Using 50~55V_{dc} supply rails, 100W into 8Ω using one pair and 150W into 4Ω using two pairs, is easily and reliabily achievable. While MJ15003's and MJ15004's remain as cheep, available and reliable as they currently are, they are certainly prime choice devices in this case.

3.2.2.2 Drivers

There are many good driver pairs to choose form. Try to use complimentary pairs rather than merely handy PNP's and handy NPN's that appear to have more than minimum required ratings; this is false ecomomy and, given the minimal cost of the devices, wholy unnecessary.

Some readily available examples include the eminantly suitable and minimal cost MJE340/ MJE350 pair, with V_{ceo} = 300V, I_c = 500mA, P_{diss} = 20W, h_{fe} = 30~240. For supply rails at no more than about +/- 50V_{dc}, BD237/ BD238 pairs, (V_{ceo} = 100V, I_c = 2A, P_{diss} = 25W, $h_{fe(min)}$ = 40), BD241C/ BD242C[69] pairs (V_{ceo} = 100V, I_c = 3A, P_{diss} = 40W, $h_{fe(min)}$ = 25) or TIP31C/ TIP32C[70] pairs, (V_{ceo} = 100V, I_c = 3A, P_{diss} = 40W, $h_{fe(min)}$ = 20) might well be suitable, provided some selection was carried out to pick resonably high gain; preferrably at least double that shown as minimum specification.

[69] Do not omit or ignore the "C" suffix in this case. The "C" version of these devices specifies Vceo rating at 100V; "A" and "B" are lower voltage ratings and hence not at all suitable for the purpose here.

[70] Again, the "C" suffix means 100V Vceo and alternatives are not suitable.

2SA958/ 2SC2168, 2SA1535/ 2SC3944 and 2SC2481/ 2SA1021 are some examples of JIS[71] designated devices that are suitable; there are many many more.

3.2.2.3 Small Signal

Q1, Q2 and Q3 have only the DC voltage of one rail less about 5V across collector-emitter, and there is minimal variation in this voltage with input signal level as the circuit is a transconductance amplifier (ref. 2.1.1). Therefore the BC547's originally used, with V_{ceo} of 45V are suitable, but also need substituting if higher supply rails are to be used and reliability is to be retained.

Complimentary small signal devices are, fortunately from a cost point of view, not required for uprating the ETI-480, as only PNP transistors are used for the small signal circuitry. It is desirable to have whatever device is to be used available in some quantity, in order to find a pair or pairs with closely matched, reasonably high gain, for Q2 and Q3.

There are many suitable examples, including 2SA1123 (V_{ceo} = 150V, I_c = 50mA, P_{diss} = 750mW, h_{fe} = 90~450), BC556 (V_{ceo} = 65V, I_c = 100mA, P_{diss} = 500mW, h_{fe} = 75~450), and MPSA56 (V_{ceo} = 80V, I_c = 500mA, P_{diss} = 625mW, $h_{fe(min)}$ = 100).

[71] JIS; Japanese Industry Standard.