

channels. Where an unsmoothed h.t. supply is used it is recommended that the decoupling capacitor should be increased in value to $250\mu\text{F}$.

Stability of output current setting. Some criticism has been voiced because there is no specific control over the output current value in the simplest form of this circuit, other than that due to the stability of the current gain of Tr_2 , whose performance determines this parameter. In order to meet this point (in anticipation) a circuit was described in the original article which allowed precise control over the operating 'quiescent' current without detriment to the performance of the amplifier.

However, measurements made on an amplifier without this addition have shown no significant change in operating current in somewhat over two years use, and there is also little measurable difference in current from a minute or so after switch-on to the end of a six-hour period of continuous use. In practice therefore, in temperate climates at least, the simplest form of the circuit is adequate in this respect. If any user cares to experiment with an alternative and somewhat more elegant form of quiescent-current control another regulation circuit is shown in Fig. 3. The transistor used as Tr_5 requires to be somewhat more massive than that used for Tr_3 , since the mean collector current is twice that of Tr_3 and the maximum voltage and current occur simultaneously. The 2N2905A is just about adequate with a good heatsink, but a larger power device such as the 2N4919 is preferable.

Alternative transistor types. The amplifier has been built successfully with a wide variety of transistors, including fully complementary versions to operate from an existing negative h.t. line, and in one case two identical amplifiers have been made for use with the inputs in paraphase, in order to double the available output voltage swing. One constructor has, indeed, made a stereo 30W system using two such pairs of amplifiers plus input phase splitter, as shown in Fig. 4.

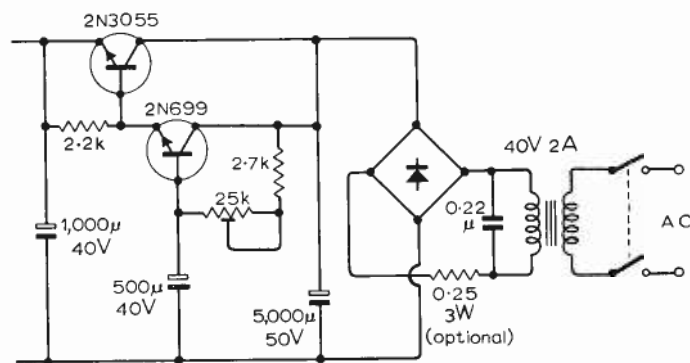


Fig. 2. Amended circuit of power supply for 15Ω systems.

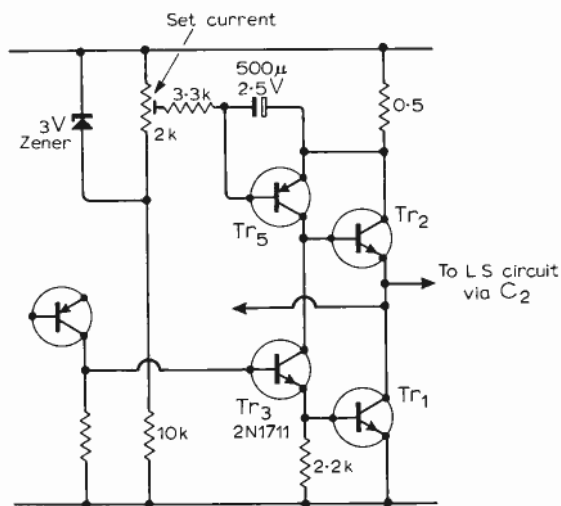


Fig. 3. Alternative method of quiescent-current control. R_1 , R_2 and C_1 in the original have been deleted. Tr_5 is 2N4919 on heatsink or alternative type.

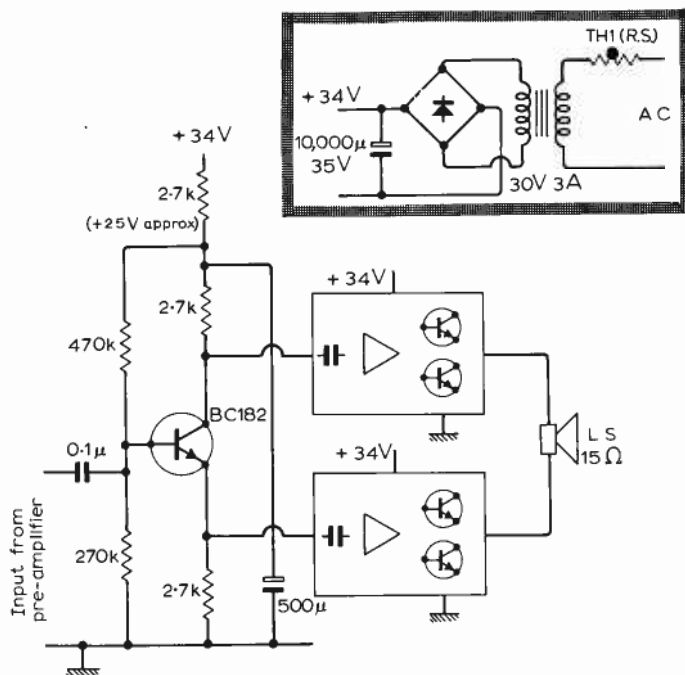


Fig. 4. Scheme for 30W class-A system. Two identical class-A amplifiers, each designed to give 15W into 8Ω (34V at 1.5A), are driven by a phase-splitter. The output capacitors have been removed.

However, one transistor change which is recommended is the use of a 2N1711 as Tr_3 . This has a high voltage capability equal to that of the 2N1613, and a current gain which is double that of either the 2N1613 or the 2N697. The use of the 2N1711 instead of the former types suggested for Tr_3 increases the feedback factor and approximately halves the typical distortion factor of the system (0.025% at 9W or 0.05% at full power) without detriment in other respects.

Also, a 2N1711 as Tr_3 allows the use of 2N3055 devices as Tr_1 and Tr_2 , with a final performance which is equal to that of the original specification below 100kHz. (The typical current gain of the 2N3055s is only half that normally found with the MJ480/1 output transistors, and their use was not originally recommended for this reason.)

Gain/frequency and power/frequency characteristics. These are, in fact, better than the curves published in April 1969. As mentioned in a letter to the editor published in October 1969, the h.f. fall-off shown was mainly due to an error in the measurement instrument. Although the performance at h.f. depends to some extent on the layout employed, the small signal voltage gain, with the component arrangement shown, is flat (within 1dB) to beyond 2MHz. This may be a snag in some cases because even a small feedback capacitance between output and input (as may happen, for example, if the output heatsinks are not earthed) may cause the amplifier to oscillate. A suitable circuit change to reduce the amplifier h.f. response to more normal levels was described in the letter above. This is not an essential modification—the author's own units are still exactly as described in April 1969.

The output power response of the unmodified amplifier is flat within 1dB to 200kHz.

The l.f. response shown in the original gain/frequency and power/frequency graphs was that determined for an earlier prototype of the amplifier. During the development of the circuit the values of some of the capacitors were increased to improve the l.f. performance, and by an oversight the graphs accompanying the article were not amended. In fact the gain and power graphs can be shown as 'flat' from 10Hz–200kHz. In this respect, and that of transient response, the class-A design is probably better than any circuit so far published. The i.m. distortion, at 10W output, (70Hz and 7kHz, 4:1) is less than 0.1%.

Miscellaneous. Surprise—and even alarm—has been caused to some constructors by the fact that the output transistors get hot.

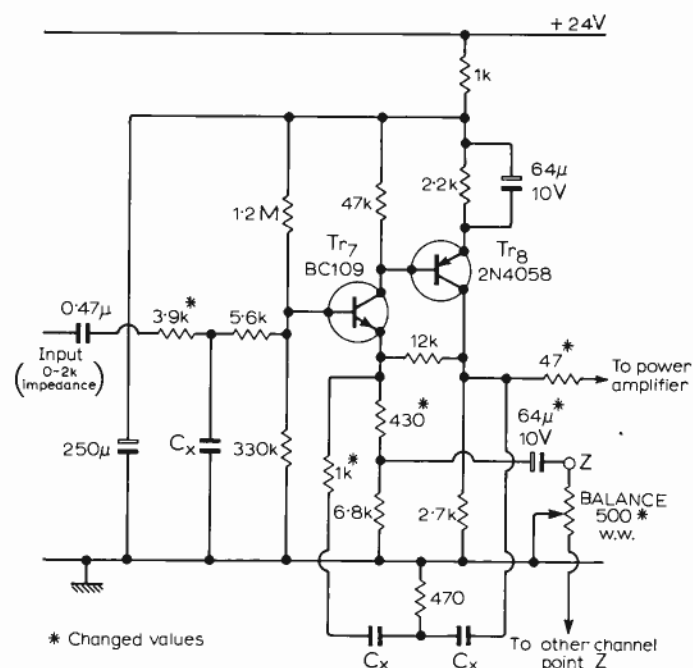


Fig. 5. Low-pass filter and driver stage (a) with gain of 20, (b) with a gain of 5. The cut-off frequencies, dependent on C_x , are the same as in the original article.

However, with adequate heatsinks, which should be black painted, the dissipations in the transistors are only a small fraction of the maker's permitted level, and provided that some care is taken in the layout to make sure that sensitive components, such as electrolytic capacitors, remain cool, no reduction in the working life of such a system, in comparison with an equivalent class-B unit for example, is to be expected.

Some difficulty has apparently been encountered by some constructors because the power supply regulation system is inoperative when the supply is operated without a load. If an equivalent resistive dummy load is connected for bench-testing, all should be found to be well.

Finally, it is prudent to wire a small resistor of about $2k\Omega$ across the loudspeaker terminals to make sure that the output capacitor charges even with the speaker disconnected. Charging of the capacitor by an accidental short-circuit could cause damage. This addition is shown in Fig. 1. No damage is caused by operating the amplifier on an o/c output.

Modular pre-amplifier

The intention in the original article was not to offer a complete pre-amplifier design, but rather to describe a series of versatile 'building blocks' from which the potential user could assemble a 'custom built' pre-amplifier to suit his own needs or preferences. To increase the scope of this some additional circuit modules are described below.

Steep cut low-pass filter. It is certainly prudent to include a low-pass filter somewhere fairly close to the input of the main amplifier whenever a wide-bandwidth main amplifier is to be used with a good-quality loudspeaker system. Doing so will prevent unwanted high-frequency components, arising from component noise, record surface noise, and similar causes, from impairing the long-term listening comfort of the user, and from producing avoidable intermodulation effects due to non-linearities in the loudspeakers.

The combination of such a steep-cut low-pass filter with a low-distortion, low-output impedance driver stage, with a gain of 50 and an output capability of some 2V r.m.s. at 0.02% t.h.d., appeared to provide the most versatile system for use with a wide variety of power amplifiers.

However, many power amplifiers require an input voltage of only 0.25–0.8V r.m.s., and there are snags in respect of hum and component noise if the stages following the volume control are operated at levels below some 50mV. The preferred level to

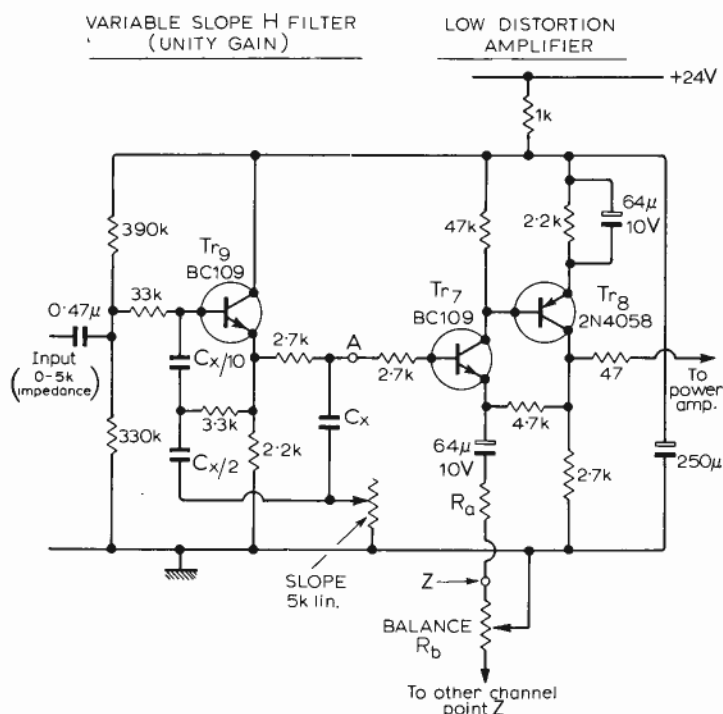
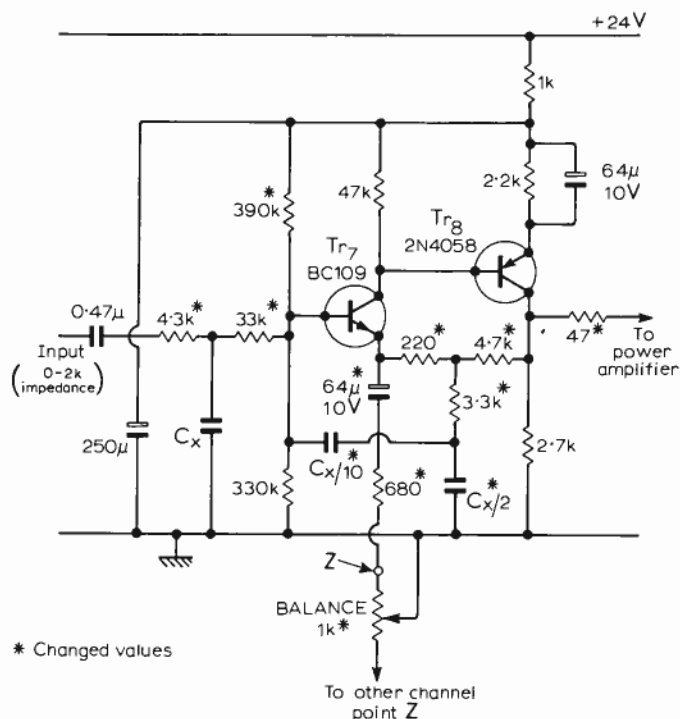


Fig. 6. Pre-amplifier driver stage incorporating a switched low-pass filter with slope variable from 6-18dB/octave. The gain of the filter is variable through the range 5-100 by choice of R_a and R_b .

$$\text{Gain} = 1 + \frac{4.7}{R_b + R_b/2}$$

Output can be taken from point A if only unity gain is required.

achieve an optimum balance of noise and distortion components is probably in the 100–200 mV region. In these circumstances a driver-stage gain of 50 is excessive, and much of the available gain must be removed by an input attenuator, and if a potentiometer is used for this it can introduce noise.

To meet this need more conveniently, two further versions of the driver amplifier, incorporating steep-cut low-pass filter characteristics which are identical to that of the original circuit, and having gains of 20 and 5, are shown in Figs. 5(a) and 5(b). An

alternative, three-transistor arrangement whose cut-off slope is variable over the range -6 to -18 dB octave, at any chosen (switchable) frequency, is shown in Fig. 6. This consists of a single transistor version of the 'H' filter used in the two previous pre-amplifier designs (the nomenclature derives from the shape of the component layout in the 'op-amp' form), followed by a very low-distortion two-transistor amplifier whose gain can be chosen, as required, over the range 5 to 100, by adjustment of R_a and R_b . If a unity-gain stage is all that is required (actually the gain is about 0.9) the output can be taken from the point marked 'A' on the diagram, and Tr_7 and Tr_8 omitted.

The response curve of the filter circuit, at any chosen turnover frequency is shown in Fig. 7. The slope is smoothly variable by adjustment to the $5k\Omega$ pot. If the slope pot. is open circuit the response is flat to $20kHz$ and beyond, but in this case the load impedance should not be less than $50k\Omega$.

For completeness, an equivalent single-transistor high-pass filter, having a cut-off slope approaching 18 dB/octave, and suitable for use as a 'rumble' filter or a pre-amplifier woofer/tweeter cross-over filter, is shown in Fig. 8. The frequency response characteristics of this filter are shown in Fig. 9. Both of these filter circuits should be driven from a source having a fairly low impedance—not higher than $6k\Omega$.

If single transistor 'H' filters are to be used at output signal levels exceeding $100mV$ a Darlington transistor, e.g. Motorola MPSA14, is to be preferred.

The apparent noise level, referred to the input, of the two-transistor driver amplifiers, using reasonably low noise transistors and an input impedance of the order provided in the normal circuit, is about $4.6\mu V$. The output noise voltage in the original circuit was 0.2 – $0.3mV$, which should be inoffensive. With a lower gain driver stage this noise will be reduced even further.

The use of a variable negative feedback type of balance control in these circuits is deliberate, in that it permits a low output impedance to be obtained from the driver stage. Measurements made with a wide range of published transistor-operated power amplifiers have shown that substantially lower distortion levels are often given by using a low-impedance drive circuit, and that there is frequently an advantage also in terms of hum, noise, and transient response.

Tone-control circuit. This stage has a worst case (bass and treble controls set to maximum 'lift') distortion level which is typically less than 0.1% at $1V$ r.m.s. output. It is perfectly capable of

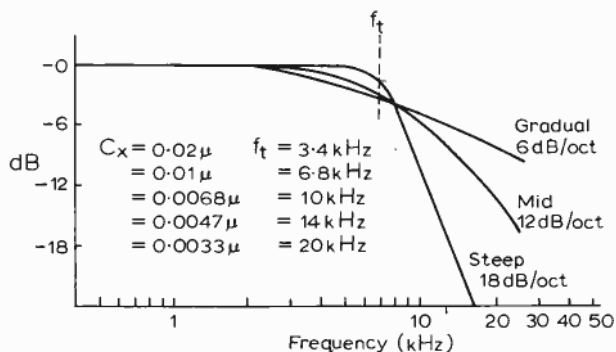


Fig. 7. Slope characteristics and turn-over frequencies of variable-slope 'H' filter.

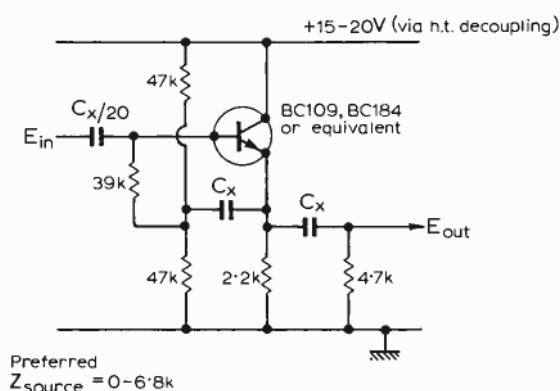


Fig. 8. Single transistor high-pass 'H' filter.

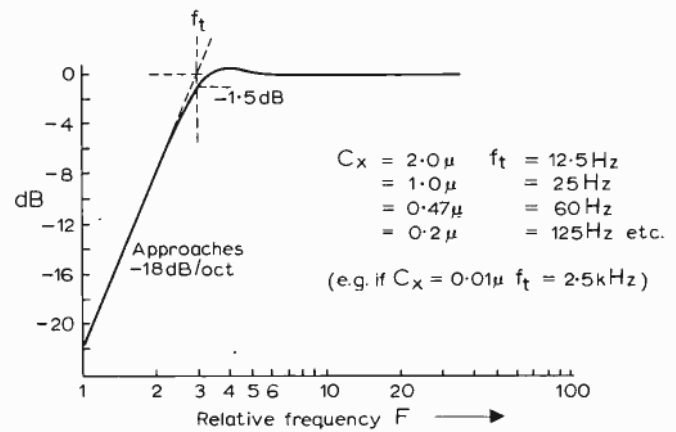


Fig. 9. Transmission characteristic of high-pass 'H' filter.

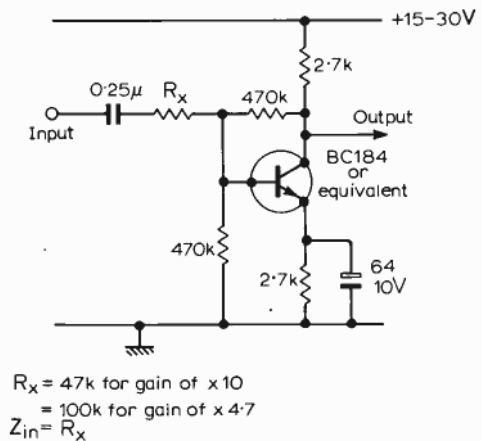


Fig. 10. Simple low-distortion single transistor amplifier.

driving a normal high-quality power amplifier without the interposition of other pre-amplifier stages. The required signal amplification could then be provided prior to the volume control. This is tending to be the normal practice in commercial 'hi-fi' amplifiers, in that it gives the highly-sought-after zero noise-level at minimum volume control settings, and makes for economies in the use of components.

Noise in the tone-control stage due to the f.e.t. has caused occasional troubles. This should not occur with the f.e.t. now recommended for this part of the circuit (the Amelco 2N4302), which appears to have a consistently low noise level. The necessary bias-adjustments were described in a letter to the editor published in April 1970.

The input impedance level suggested for the tone-control stage was $50k\Omega$, because it was thought that most of the other systems likely to be used with this unit would be transistor operated; and this would be of a suitable level for this purpose, while avoiding some of the hum pick-up problems likely to be encountered at higher impedance levels. However, if this impedance is too low, and if a high gain (beta greater than 400) transistor is selected for Tr_4 —in fact most BC109s will do—the base bias resistors can be increased to $1M\Omega$ and $560k\Omega$ (instead of $200k\Omega$ and $100k\Omega$) enabling the volume control and auxiliary control potentiometers to be increased to $25k\Omega$.

If an even higher input impedance is required, the f.e.t. impedance conversion shown in Fig. 4 in the original pre-amp. article can be substituted in its entirety for Tr_4 . To preserve the function of the rumble filter in this circuit, with the $0.47\mu F$ capacitor desired to feed the tone control network, a $4.7k\Omega$ resistor should be connected from the output side of this capacitor to the earth line. A low-noise f.e.t. is of course preferable.

If additional amplification is required on any signal source prior to the tone-control stage (if this is working at the $100mV$ level) a simple single-transistor feedback amplifier such as that shown in Fig. 10, can be used with confidence, in that its performance is stable, its noise level is low, it is almost impossible to damage by

an input overload, and its distortion is well below 0.1% at output voltages up to 0.25V r.m.s., and with gains up to 10.

Magnetic pickup equalization circuit. Some requests have been received for component values for the use of this circuit for tape-replay characteristic equalization. The author remains of the opinion that this type of provision is best left to the manufacturers of the tape recorder, in that the actual head characteristics can influence the replay frequency/voltage characteristics.

However, a fairly close approximation to the replay curve theoretically required for 7.5 i.p.s. is given if C_2 and R_2 in the original equalization network A are altered to 100pF and 27k Ω .

The noise level of this circuit is almost entirely determined by the performance of Tr_1 . The BC184C and 2N5089 transistor types may be of interest in this position.

The maximum output which can be obtained from this circuit at 0.02% t.h.d., is 2V r.m.s. If the normal input to the tone control circuit, or other following stages, is 100mV, this gives a 26dB overload capability. The gain of the equalization circuit can be increased by a factor of 3, (i.e. to 30 at 1kHz) without upsetting the rumble filter characteristics if R_3 is reduced to 68 Ω and C_4 increased to 100 Ω F.

Miscellaneous. An omission from the original article was the suggestion that high value resistors (2-5M Ω) should be

connected across the switch contacts, from slider to each C_x . This removes 'plops' on switching ranges.

A number of correspondents have queried the need for a separate h.t. power supply for the pre-amp. (The reservoir capacitors for the unit shown should have read 35V working, not 25V). It is always possible to run the pre-amp via a suitable voltage-dropper circuit from the main amplifier power supply, and if a zener diode is included in this line, this scheme may be satisfactory. However, measurements on channel separation, and harmonic and i.m. distortion, with identical amplifier systems invariably show some advantage, particularly at the low-frequency end of the audible spectrum, in the use of a separate power supply for the pre-amp (even when the electrolytic bypass capacitors are still new) and this arrangement is still recommended by the author as well worth the small additional cost.

One point which has not been published, to the best of the author's knowledge, concerns the particular advantage conferred by the feedback pair amplifier using complementary transistors, such as that used in the low-pass filter circuit, in comparison with the more usual n-p-n/n-p-n pair, where the bias for the first transistor is derived from the h.t. line. In the case of the n-p-n/p-n-p pair, any h.t. line feedback, due to inadequate h.t. line bypass, will be negative rather than positive, and this can assist in obtaining good t.h.d. figures down to low signal frequencies.

Letters to the Editor

The Editor does not necessarily endorse opinions expressed by his correspondents

Class AB amplifiers again

Mr. Linsley Hood's reply in the October issue to my letter (August) does indeed clear up the difficulties I experienced in following his article and his reply to Mr. Gibbs (August issue), but I feel bound to justify my objections more fully. I understand the mutual conductance of a transistor or a pair of transistors to be $\frac{dI_c}{dV_{be}}$.

measured between the input base and output emitter, under precisely those near zero source impedance conditions to which he refers. With values of less than an ohm the shape of the basic mutual characteristic of the MJ481 is preserved. The curve obtained with 100- Ω source resistance looks much more like the current gain characteristic, except at low collector currents. If the effect of the 10- Ω resistor is removed from Mr. Linsley Hood's curve A, the slope does become steeper than that

of curve C. Consider an MJ481 with and without a 0.2- Ω emitter resistor and with and without a 40361 driver in the Darlington pair configuration, with zero source impedance (Fig. 1), with modifications where appropriate. It is easier to work in terms of mutual resistances than conductances, and representative values of these are shown in Table 1 (R is infinite here).

The mutual resistance of combinations of these three, including the MJ481, is the sum of these resistances seen at the output

emitter. The MJ481 is assumed to have a current gain of 100; this does not prejudice the argument as the characteristic of the 40361 is nearly exponential, so that the slope is approximately inversely proportional to I_c . The results for the four cases are shown in Table 2. The optimum quiescent current for a voltage driven stage is normally the collector current at which the resistance slope is twice its high current value.

It can be seen that the addition of an

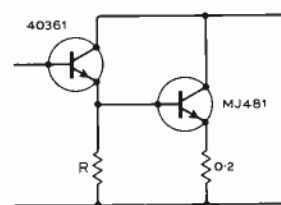


Fig. 1.

Table 1

$\frac{dV_{be}}{dI_c}$ of MJ481	I_c of MJ481	I_c of 40361	$\frac{dV_{be}}{dI_c}$ of 40361
			100
Ω	mA	mA	Ω
0.16	1,000	10	0.04
0.25	200	2	0.15
0.32	100	1	0.30
0.50		0.5	0.50

Table 2

Combination of components	Slope at 1 A output current	Optimum quiescent current	Slope at this current
	Ω	mA	Ω
MJ481	0.16	100	0.32
MJ481 + 0.2 Ω	0.36	50	0.70
MJ481 + 40361	0.20	200	0.40
MJ481 + 40361 + 0.2 Ω	0.40	100	0.82