

Simple Class A Amplifier and Modular Pre-amp

A postscript to designs published last year

by J. L. Linsley Hood

The author has had the benefit of an extensive and frequently helpful correspondence with readers following the publication of the two circuit designs.* Attention has been drawn to some obscurities in the original articles and to certain possible improvements in the design. Details are given below.

Class-A amplifier

Power supply. Although much interest was aroused among constructors by the good technical (and audible) performance given by the amplifier, it is clear that the principal feature in the eyes of many users was the relative simplicity of the circuit. This being so it must have seemed a pity that the power-supply unit was somewhat complex. However, the purpose of this power-supply design was to avoid possible degradation of the amplifier performance by h.t. ripple. The sawtooth ripple voltage across the reservoir capacitor in a class-A amplifier drawing some 2·3 A will be many times greater than in a class-B system, particularly at the audibly important low-signal levels.

However, a number of measurements have been made since the publication of the original article on an amplifier of this type, operated from a simple power supply unit of the type shown in Fig. 1. There is little difference in the performance above 100 Hz either in total harmonic distortion or in intermodulation distortion, although the shape of the output power/distortion curve at the onset of overload is modified, as would be expected, by the ripple on the h.t. line. Below 100 Hz the distortion curve rises more steeply to about 0.2% at 20 Hz.

A thermistor is necessary, in this case, to slow down the rate of rise of the h.t. voltage. This will get hot in use.

It now appears that the mains transformers used in the development of the prototype of this amplifier were not as efficient in respect of apparent secondary circuit resistance or secondary leakage reactance as some of those which have been supplied for this purpose since the publication of the article. The reservoir voltage found with the 15Ω system may be above that given by the author. At switch-on this can cause a transient overloading of the transistors specified for the series regulator circuit in the original article. In view of this, it is suggested that these should be an MJ481 or 2N3055, used in conjunction with a 2N1613 or, better still, a 2N699. These amendments are shown in Fig. 2.

Adjustment of amplifier output current and centre-line potential.

The author had supposed, somewhat naively, that most constructors of the circuit would have somewhere in their workshops a collection of odd-value resistors needed for trimming circuit parameters, and it was mentioned in the original article that the desired quiescent levels could be set by adjustment to R_2 and R_5 or R_6 . This sort of comment is unhelpful if one is writing away for a kit of parts. In view of this it is suggested that R_2 should be replaced by a resistor in series with a potentiometer, as shown in Fig. 2. The necessary value of resistors R_5 and R_6 to give an entirely adequate accuracy in mid-point voltage setting can be predicted, and the suggested amended values are shown.

Some obscurity arose, inadvertently, in the original diagram concerning the reason for the different values of input decoupling resistor quoted for mono and stereo use. This was because it was intended that the one decoupling circuit should serve both

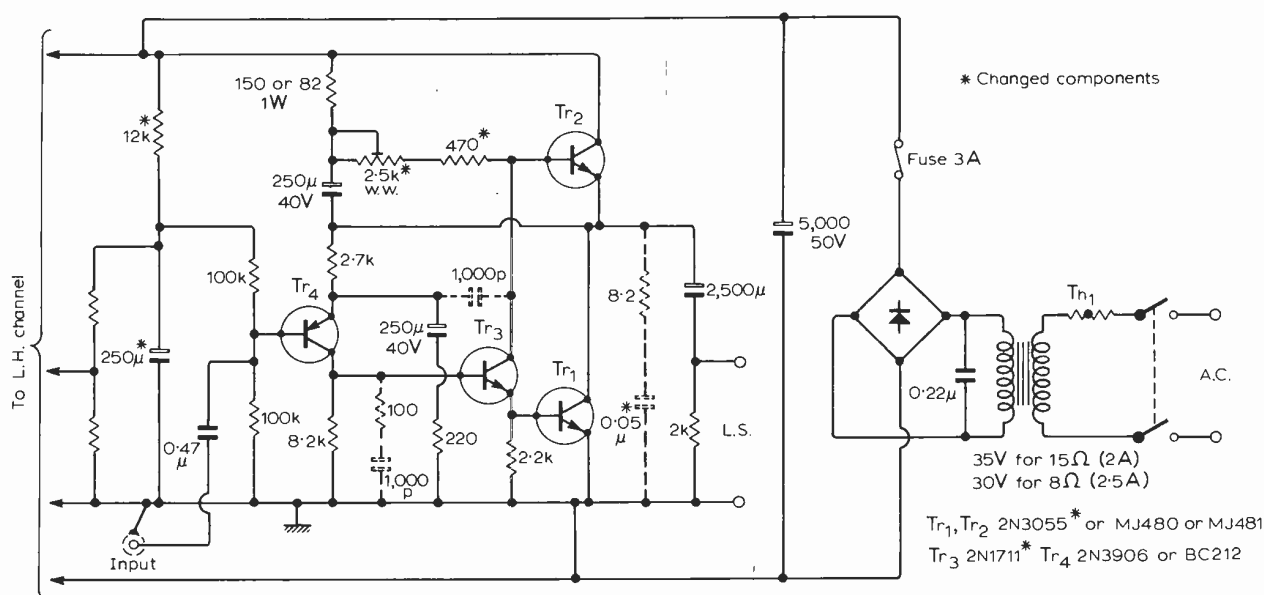


Fig. 1. Suggested amended circuit for 8 or 15Ω, use employing a simplified power supply. The dotted components reduce the h.f. response and should be used with capacitive loads.

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.

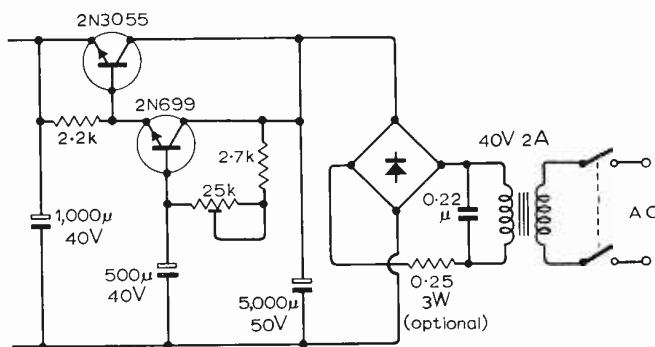


Fig. 2. Amended circuit of power supply for 15Ω systems.

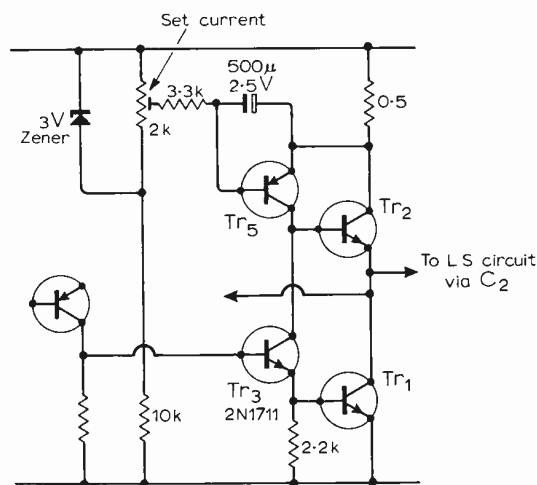


Fig. 3. Alternative method of quiescent-current control. R_1 , R_2 and C_1 in the original have been deleted. Tr_5 is 2N4919 on heatsink or alternative type.

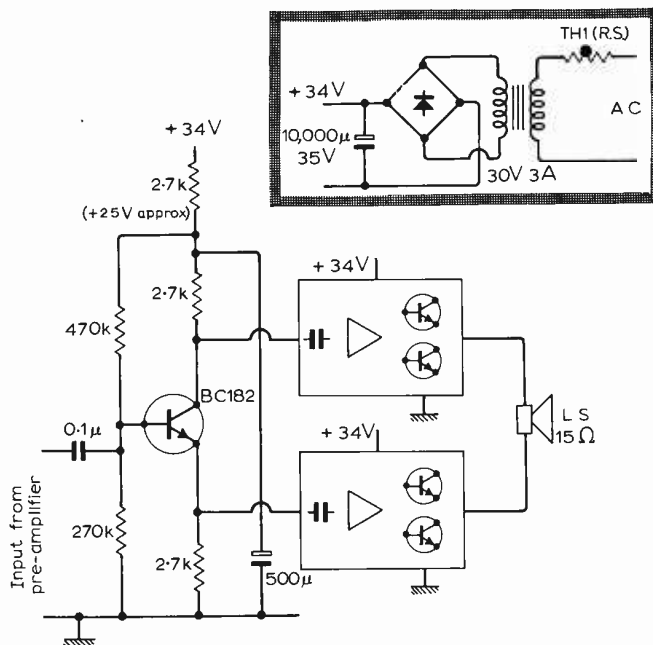


Fig. 4. Scheme for 30W class-A system. Two identical class-A amplifiers, each designed to give 15W into 8Ω (34V at 1.5A), are driven by a phase-splitter. The output capacitors have been removed.

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.

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

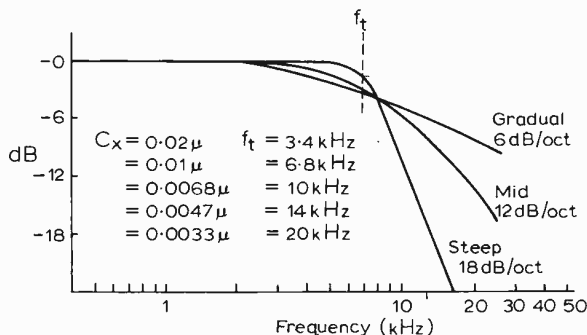


Fig. 7. Slope characteristics and turn-over frequencies of variable-slope 'H' filter.

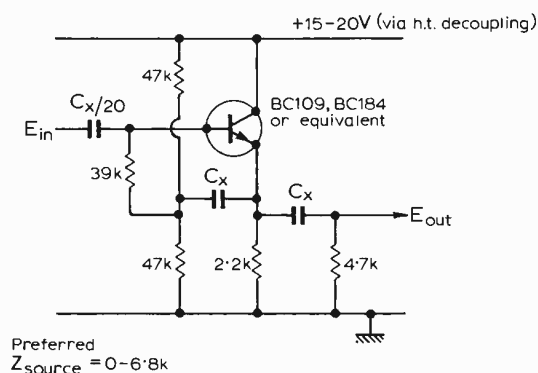


Fig. 8. Single transistor high-pass 'H' filter.

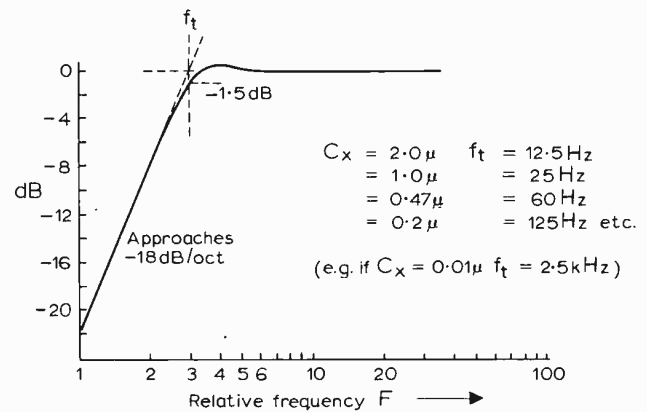


Fig. 9. Transmission characteristic of high-pass 'H' filter.

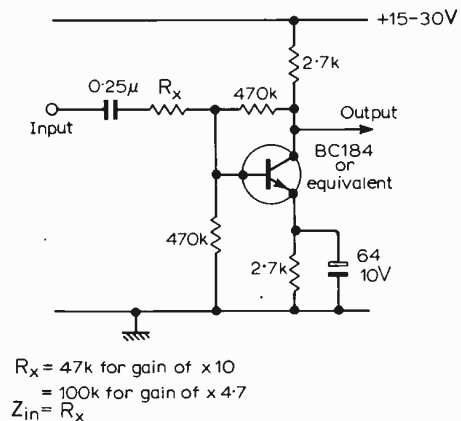


Fig. 10. Simple low-distortion single transistor amplifier.

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_5 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 $\frac{dI_c}{dV_{be}}$. V_{be} is

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

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

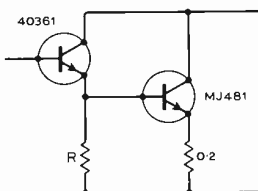


Fig. 1.

emitter resistor reduces the optimum quiescent current and of a driver increases it, although either addition reduces the overall mutual conductance at all currents. The effect of finite values of R is to reduce the change introduced by the driver.

The p-n-p/n-p-n configuration is more complicated (c.f. Mr. Baxandall's letter in the September 1969 issue), but in general it has a higher mutual conductance (Fig. 2, $r=0$) than the simple output transistor. With common values of r the combination is linear down to much lower

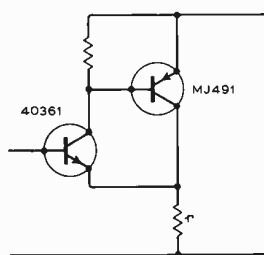


Fig. 2

collector currents in the output transistor, giving a lower half-slope current. With a high source impedance the optimum quiescent current for a complementary or quasi-complementary output stage is not so readily defined. It may well be Mr. Linsley Hood's experiences in these circumstances which leads him to the conclusion (August issue) that the optimum quiescent current varies inversely with the absolute magnitude of the current gain in half of the output stage.

The "circumstances which would exist in a practical design" are precisely those put there by the designer; source impedances of under 1Ω are perfectly feasible. It begs the question to insert resistors in the base lead before even measuring the basic properties of the transistors. The mutual characteristic so obtained is only relevant to a complete amplifier which has these impedances in series with each half of the output stage—resistors R_1 & R_2 in Figs. 3 & 4—excepting pure class B using transistors which cut off perfectly and do so with zero base-emitter voltage. If R_1 & R_2 are

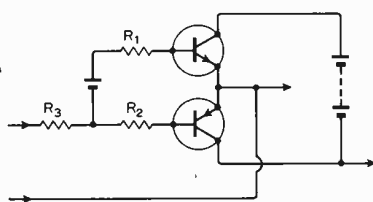


Fig. 3

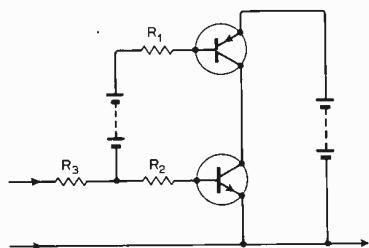


Fig. 4

zero, and R_3 is finite, the overall transfer characteristic of the complete output stage is best not looked at in terms of the mutual conductance measured when one transistor is omitted.

I apologize for making objections in terms of the article, since it does not convey the sense that the author intended, but I based my arguments on the design itself. The source impedance to the output stage is genuinely low. The minimum current gain of an MJE521 at 50 mA collector current is about 80, giving a drive impedance of 70Ω at the most (derived from the $6.8\text{ k}\Omega$ resistor). The input impedance of the output stage varies between 50 and 100 ohms in the 15Ω version with output transistors of current gain 100. It is the inappropriate ratio between these two quantities which is responsible for the effects to which I referred.

It would be convenient if the bootstrap capacitor could supply the extra current required to drive low gain MJ491s which need a base current in excess of the standing current in the driver stage. This could only occur if the bootstrap capacitor temporarily sustained a greater voltage than it does under static conditions. This situation arises during a short negative transient (MJ491 on) a short time after a long positive excursion (MJ481 on). Short and long are referred to the time constant of the bootstrap capacitor and R_4 in Fig. 3 of the article. Quite how common these conditions are in music (with whatever d.c. components there might have been removed well before bootstrap capacitor has its say) I can't imagine.

The other points I should like to make are best left to a future date—we both appear to be drawing on material which should see the light of day in articles rather than in letters.

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In defence of the amateur

I was glad to see the present situation regarding some of the v.h.f. amateur bands put so lucidly by Dr. Saxton in "Letters", October issue. It is obvious that for short-term gains, many commercial interests as well as national ones around the world would like to insert the thin end of the wedge into amateur frequency allocations. Clearly there is an increasing ignorance in commercial circles of the unique long-term value of amateur radio. When industry itself profits so much by the (admittedly immeasurable) boost to its own manpower by the early influences of amateur radio, then the movement can still ask for and I think expect support from this quarter. The

influence of such a movement on young people not only develops the interest of future technical employees but also broadens the experience of non-technical staff into the hardware field, to say nothing of the enrichment of social life for every one. I feel that industry and commerce that have an interface with amateur radio should offer technical assistance, market small quantities of materials even if inconvenient, and give general encouragement for the maintenance and extension where possible of the wavebands allocated to the amateur service.

The matter of early influences particularly interest me and when assisting in the work of setting up the R.S.G.B. Education Committee, I voiced the opinion that an Hon. Youth Officer, or some such worker, should be established. It would be excellent if, during his presidential year, Dr. Saxton could bring a little influence to bear for inaugurating a youth programme, as I know he is very interested in the educational aspects of amateur radio.

It is to be deplored that in the same issue we read in paragraph two of the report on the R.S.G.B. Exhibition that *Wireless World* itself is suggesting that the amateur frequency bands be re-allocated to a more deserving cause. This is a terrible generalization from insufficient data. (one stand in one exhibition!). Come now Mr. Editor, ask your reporter to call in at any club constructional evening, R.A.E. lecture, or R.A.E.N. practice session etc. etc. KEN SMITH, G3JIX, Canterbury, Kent.

"Toothpaste bass"

Your readers all over the world think of *Wireless World* as being above the controversy*, as being a legitimate source of unbiased information on the world of electronics, and music reproduction in particular. Therefore, it pains me to hear (in your report on Sonex 70) our IMF loudspeakers characterized as having "toothpaste bass"—a highly emotional characterization of the only loudspeakers in England which have any bass, to my knowledge!

In the cause of scientific truth, and also be it known, of British exports, I do believe your reporter should be called to order a bit. The role of criticism is not to destroy, or to create personal vendettas, but to inform the public and to instruct the public and the industry on current techniques for furthering the art—and not some individual's pique!

IRVING M. FRIED,
IMF Products,
Philadelphia, Pa.

* The publication of this letter shows that we are not; we do allow both sides to have their say. ED.