

# New Approach to Class B Amplifier Design

by Peter Blomley\*

The class B amplifier has established itself as the most versatile and lowest cost amplifier known. This is mainly due to the excellent work in the field of semiconductor circuit design by H. C. Lin<sup>1</sup>, R. C. Bowes<sup>2</sup>, R. Tobey and J. Dinsdale<sup>3</sup>, A. R. Bailey<sup>4</sup> and P. J. Baxandall<sup>5</sup>. In this article it is hoped to complement the work of these designers by putting forward a new approach† which may solve some of the problems inherent in present designs.

## Conventional approach

A definition of a class B amplifier could be 'one in which the operating point of each output device is set at the lower extreme of its transfer characteristic. Hence in a push-pull design, for any symmetrical input signal, each output device conducts only one half of the output waveform'. This method of operation gives the amplifier zero (or nearly zero) quiescent power consumption, high efficiency and excellent peak current drive capability. It is unfortunate that the sacrifice paid for these virtues is the problem of ensuring a linear transfer of signal drive from one output device to the other.

So that the class B system can be analysed it is useful to approach the output circuit as two separate amplifiers (labelled X and Y in Fig. 1 for convenience), the outputs of which combine to give the complete signal. This is shown in diagrammatic form in Fig. 1, where it is assumed that the blocks representing the amplifiers form the equivalent of a complementary output stage.

The transfer characteristic for one of these 'sub-amplifiers' is shown in Fig. 2, where above the bias point A the characteristic is extremely linear and below it becomes a combination of linear and exponential relationships. The designer's task is to define this last region so accurately that when it is combined with that of the other sub-amplifier, the overall gain will remain constant (i.e. as the gain of one sub-amplifier decreases, the other increases equally to compensate).

The workings of a class B output circuit can be clarified by the use of  $g_m$  diagrams, these being a plot of gain—or in this case

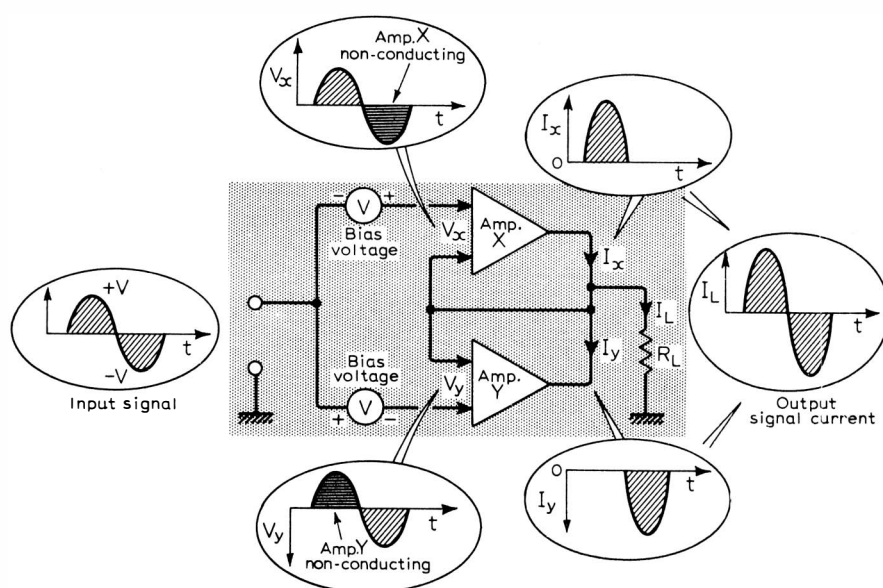


Fig. 1. Block diagram of conventional class B amplifier with the two halves of a complementary output stage represented by sub-amplifiers X and Y.

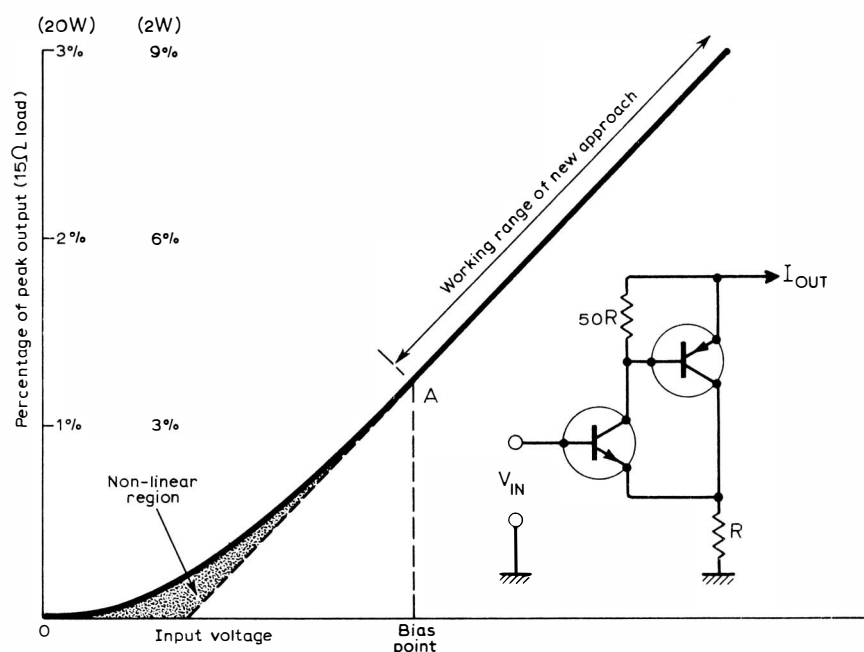


Fig. 2. Transfer characteristic of sub-amplifier X which is linear above bias point A and non-linear below. In conventional class B, non-linear regions of X and Y sub-amplifiers have to be accurately matched to give overall linearity.

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†The subject of patent application 53916/69.

mutual conductance—of the complete output circuit against drive voltage. The ideal would, of course, be a straight line parallel to the input voltage axis (indicating there is no change of gain with input swing), but regrettably this is not the case with designs popular at the moment. To provide a comparison of the different types of output circuits, I have prepared gain plots showing the effects of different bias levels, these being illustrated in Figs. 3 and 4. From these it is now easy to see the characteristic change in gain which can occur during the transfer from one sub-amplifier to the other. Referring to Fig. 3 about a 10% gain change occurs during transfer, whatever the bias level is set at.

The output circuit in Fig. 4 is a quasi-complementary type giving most interesting results. The main conclusion is that it is impossible to bias this circuit for symmetrical gain change and in practice it proves very difficult to establish which biasing point would give the best results concerning the *rate* of gain change.

This method of describing a class B amplifier can give an insight into the problems involved with a conventional design. First, each sub-amplifier has to have two regions in its transfer characteristic:

- the constant gain region (above bias point A in Fig. 2)
- the non-linear region (below this point).

Second, the non-linear region of each sub-amplifier has to be complementary to its partner, otherwise the situation shown in the  $g_m$  diagrams (Figs. 3 & 4) will occur. An interesting point is that the only reason why the non-linear region of the transfer curve is important is because the input signal normally traverses this region as well as the linear portion. If this was not the case most

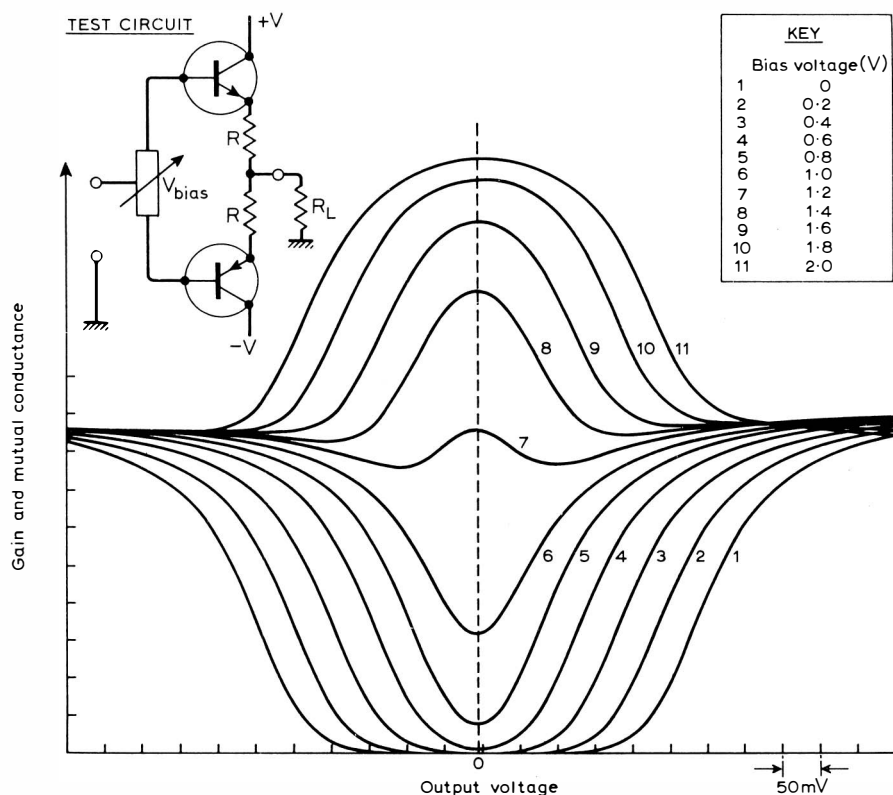


Fig.3. Gain—or mutual conductance—of simple symmetrical output circuit showing change in gain which can occur during transfer from one sub-amplifier to the other. Effect of different bias levels is shown.

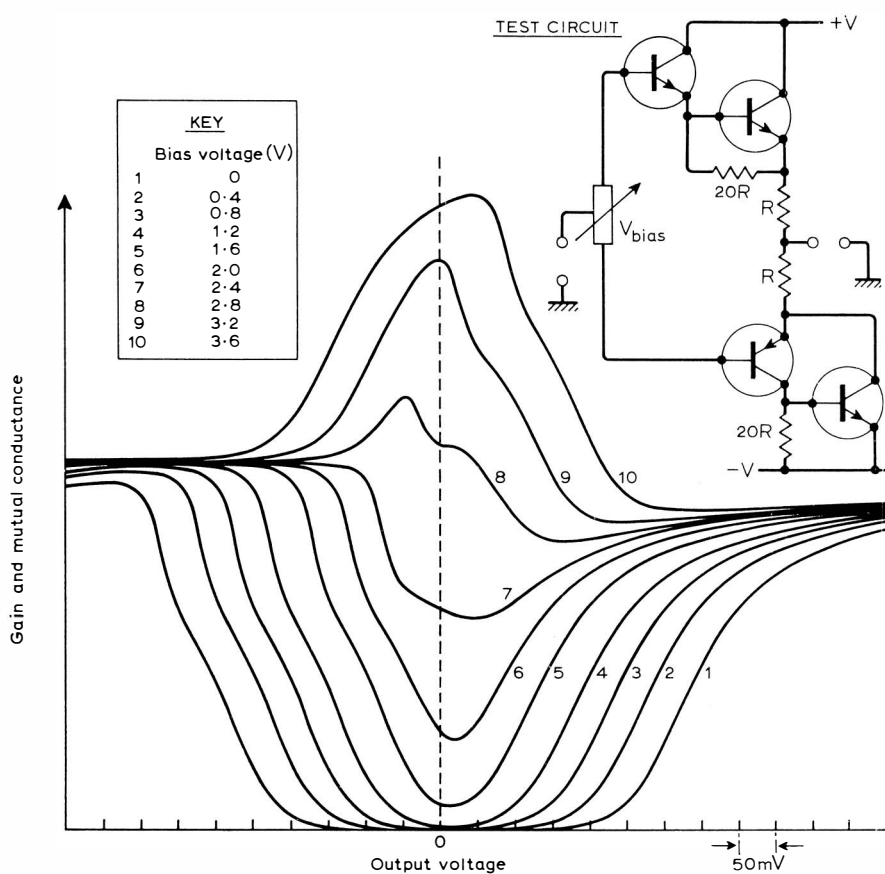


Fig.4. Curves for quasi-complementary output circuit show impossibility of biasing circuit for symmetrical gain change.

of the design problems in class B amplifiers would be solved.

It is difficult to realize at first that a class B amplifier has to have this non-linear region in the sub-amplifier characteristic so that the two halves of the waveform can

be separated. With conventional designs this is a built-in feature, but it need not be so. Assuming we define class A operation to include any amplifier where the input signal never traverses the non-linear region, the sub-amplifiers of a class B amplifier can operate in class A as long as the input signals are uni-directional. To accomplish this the required non-linear element is placed before the sub-amplifier inputs.

### New approach

Now the key to the problem is in the proposition that each sub-amplifier should be considered as a separate class A design, hence distortion generated by each of these units can be held to an extremely low level as long as the input signal can be prevented from driving the amplifier into the cut-off region. In the new approach, the output sub-amplifiers are biased *above* the non-linear region and *uni-directional* signals are fed into the input. This arrangement is illustrated in Fig. 5, where the necessary circuit changes are shown by comparison with Fig. 1. The obvious difference is the addition of the two diodes at the input which produce the uni-directional signal to drive the output sub-amplifiers. The linear transfer of signal between the two amplifiers is now dominated by this signal splitter.

**Signal splitter.** As the name implies the task of the signal splitter in a class B amplifier is to segregate the top and bottom halves of the signal waveform. Normally this is achieved by using the non-linear characteristics of each half of the output stage, but as this particular approach leads to

problems, the new approach separates the two functions of amplifying and signal splitting completely.

To explain the problems involved with the design of a signal splitter it is usual to establish the ideal and see how this can be approached practically. As it happens there are two ideal 'half' characteristics which will give a linear cross-over when they are combined. The first, and obvious one, has a conduction path only in one direction and absolutely zero in the other. The other is more complicated and has three regions—linear region (large positive inputs), a non-linear region (transfer coefficient is proportional to signal) and a reverse region (transfer coefficient is zero).

The difficult region is the non-linear one. This will only give a linear crossover when it is combined with another conjugate characteristic. Not only this, but the relationship between the linear and non-linear portion has to be accurately defined. Normally this is achieved by altering the quiescent current in the signal splitter for minimum crossover distortion. Thus using this approach in the signal splitter means that the non-linear region has to be complementary to its partner and also that the linear and non-linear regions have to be accurately related. If additional constraints are imposed—due to device spreads and temperature changes—the situation can become very difficult unless a simple approach is used.

Returning to the first type of signal splitter, the immediate comparison which can be drawn is the simplicity of the characteristic. There are no interactions between each element and only one region has to be accurately defined. Ideally, therefore, this type of characteristic should be easy to control once a suitable device configuration is found.

**Ideal element.** The simple p-n diode fabricated in silicon can have a forward-to-reverse current ratio of  $10^{10}$ ; thus it approaches the ideal almost within the boundaries of measurement. This is however only considering the forward characteristic under conditions of *current drive*. If a voltage source were used the forward transfer would revert to the familiar exponential relationship between input voltage and output current (Fig. 6a). If a signal splitter is now made of two of these diodes and a *current* of changing direction fed into the common point, then from Kirchhoff's second law the current must flow either in diode  $D_1$  or diode  $D_2$  depending on the direction of signal current flow. The transfer coefficient for the diode must be unity, as it is only a two-terminal device, hence this type of signal splitter is extremely linear under the conditions of current drive (Fig. 6b).

**Transistor signal splitter.** The use of a transistor as a signal splitter (Fig. 7) logically follows that of a p-n diode simply because the emitter-base junction has almost identical characteristics to that of a diode. Exactly where the transistor is superior to that of the diode depends on

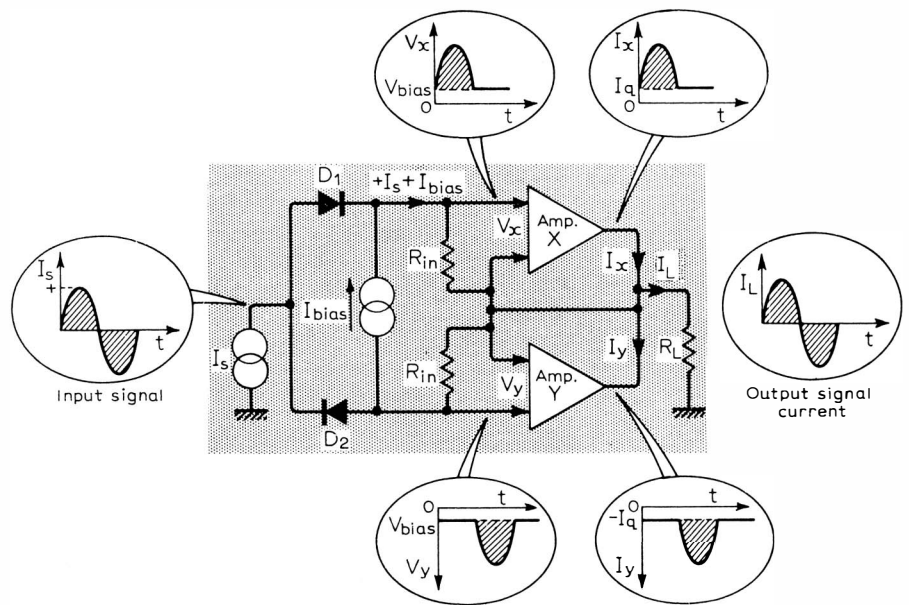


Fig. 5. New approach to class B amplifier in which sub-amplifiers are biased above non-linear region and fed with uni-directional signals produced by the diodes. This effectively transfers signal splitting from the sub-amplifiers to a separate part of the circuit.

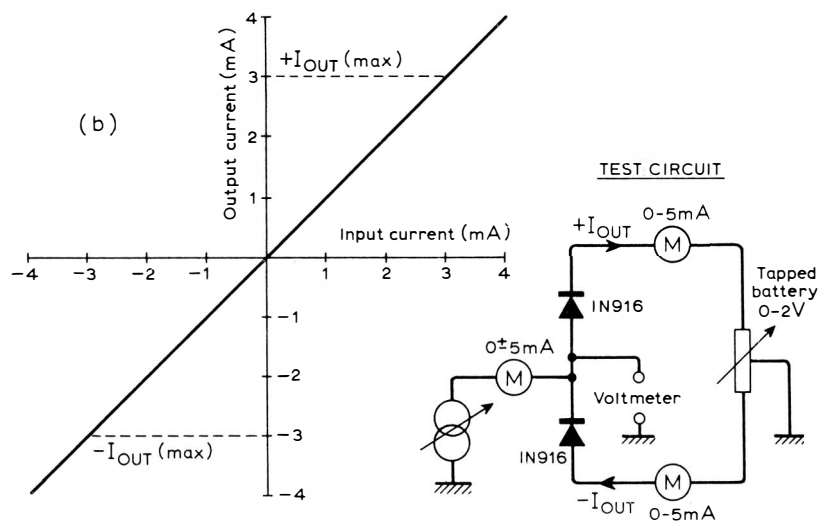
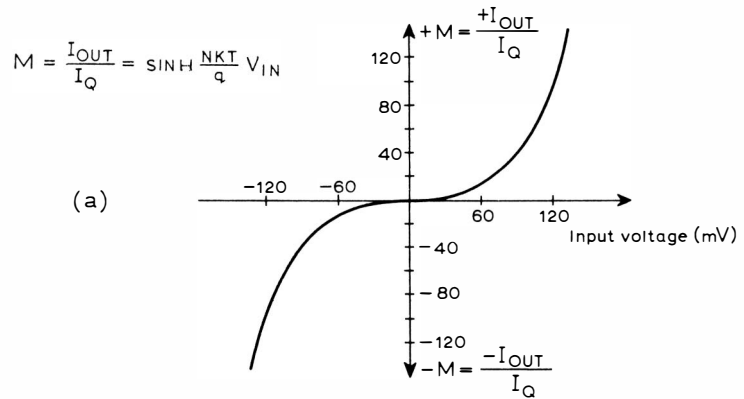


Fig. 6(a). Transfer characteristic of voltage driven diode signal splitter.

Fig. 6(b). Linear transfer characteristic of current driven diodes.

the design approach but in most cases the level-shifting property of a bipolar device is the main reason. This is very useful in a practical design but care has to be taken in the selection of the type of device. There is a problem with the use of transistors as signal splitters due to the emitter-base depletion capacitance. Under conditions of

low injection this can add an additional phase lag during the crossover period. The problem can be overcome by using silicon planar devices with very high transition frequency ( $f_T$ ) or by selecting devices in which the  $f_T$  is dominated by the diffusion capacitance as  $f_T$  remains constant down to very low emitter currents.

**Synchronous signal splitter.** There is a limit to the speed at which the diode or transistor signal splitter will transfer the signal path between the sub-amplifiers. If a synchronous signal splitter is used the time taken can be reduced to a few nanoseconds. This makes true class B operation possible at frequencies far higher than the audio spectrum. The system diagram is shown in Fig. 8(a). Instead of using the characteristics of the devices, as in the signal splitter which separates the two halves of the waveform, switches  $Tr_1$  and  $Tr_2$  are turned on and off at the required time by another amplifier labelled

ST. This is a high-gain amplifier with a small amount of hysteresis, and as soon as the input exceeds a predetermined level the output from the trigger (amplifier ST) will change its polarity and turn on  $Tr_1$  or  $Tr_2$ , depending on the signal direction, Fig. 8(b). This therefore gives almost the ideal signal splitting characteristics but the added complication might spoil its commercial possibilities.

**Performance of the new design**

The transistor signal splitter and the output stage circuit have a combined charac-

teristic shown in Fig. 9 which demonstrates the excellent gain linearity. It is only when the bias of the sub-amplifiers is decreased below its optimum, allowing the output signal excursions to trace the non-linear region of the characteristic, that distortion begins to rise sharply. Further studies of these curves reveal that increasing the quiescent current through the output devices *does not* degrade, or for that matter improve, the crossover performance of the output circuit. Keeping this in mind it is therefore possible to design a class B amplifier *without any bias adjustment*. This assumes that the designer can guarantee that spreads in active devices and resistance values do not permit the quiescent current to fall below the level where the mutual conductance of the amplifier begins to decrease.

In this discussion about the performance of the design as a whole it would be fitting if the sub-amplifier design is mentioned. With conventional designs this two- or three-transistor element is fraught with compromises, one of the most serious being the decision on the inclusion of a base-emitter 'turn-off' resistance for the power transistor. Such a combination generates what can be called 'dead zone' distortion, mainly due to the change in slope of the transfer characteristic at zero crossing. One example of this is shown in Fig. 10 where, as predicted, the lower the value of resistance the more pronounced is the effect. It is very tempting to exclude this resistance altogether, especially if the current drive approach has been adopted, but the penalty would be a poor high-frequency performance coupled with overload recovery problems. This dilemma is aggravated if the designer decides to use homotaxial base powder devices (chosen for the robust nature of their construction and freedom from secondary breakdown) because the input diffusion and depletion capacitance is very high, hence the gain-bandwidth product of the device is relatively low (e.g. silicon planar  $f_T \approx 90\text{MHz}$ , homotaxial base  $f_T \approx 1\text{MHz}$ ). In the latter case it is essential that the resistor is included. However, if the approach suggested in this article is adopted the sub-amplifier will never enter this non-linear region, thus the base-emitter turn off resistance can be included in the circuit to improve the performance without undue complications.

Once the decision has been taken to use the new approach the best circuit configuration has to be found and here again nature's swings and roundabouts create a difficult situation where compromise seems necessary. One of the criteria I used was that of thermal performance, following an initial consideration of the electrical properties of each configuration. The power transistor chip can change its temperature by tens of degrees centigrade during a power cycle, this being reflected by a corresponding change in the base-emitter voltage ( $V_{BE}$ ) of the device. If the voltage bias to the sub-amplifier is applied directly to the power device (Fig. 11a), any change in the  $V_{BE}$  will cause a considerable change in quiescent current and in turn an

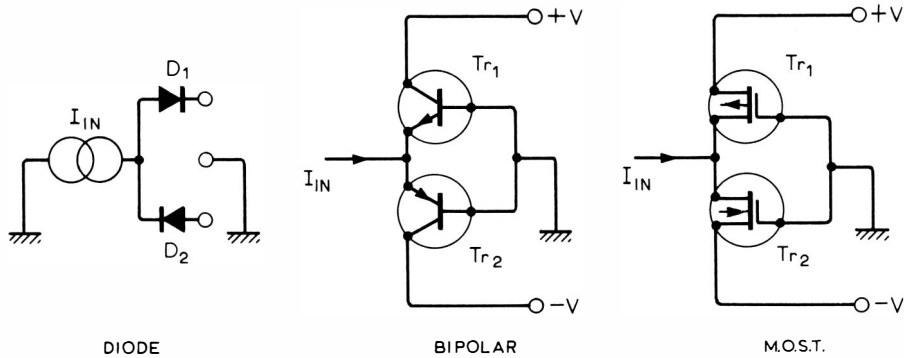


Fig. 7. Types of signal splitter. Transistor type has the advantage of level shifting.

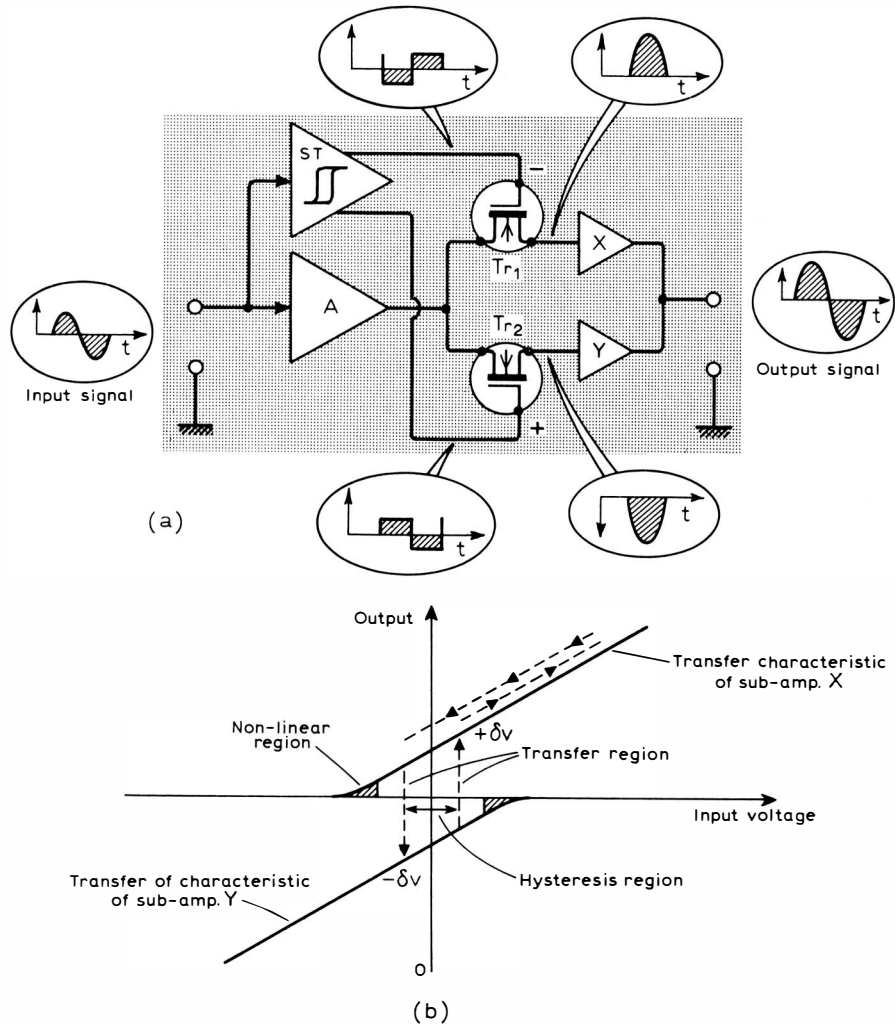


Fig. 8(a). Synchronous signal splitter, with fast switching time, allows new approach to be used at frequencies well above audible range.  
Fig. 8(b). Operation of synchronous splitter of Fig. 8(a). When input level exceeds a pre-determined level, output changes polarity and turns on  $Tr_1$  or  $Tr_2$ .

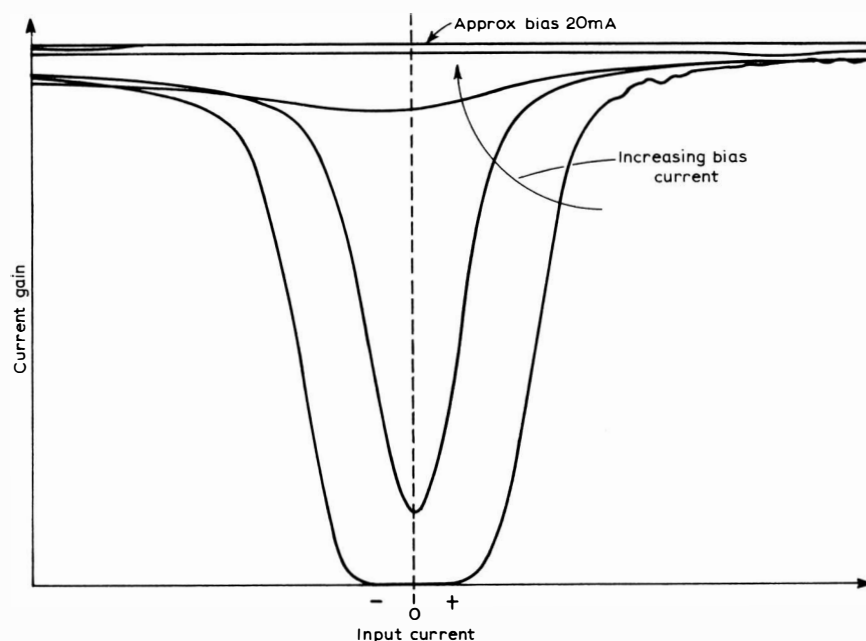


Fig.9. Combined characteristic of transistor splitter and output circuit shows excellent gain linearity.

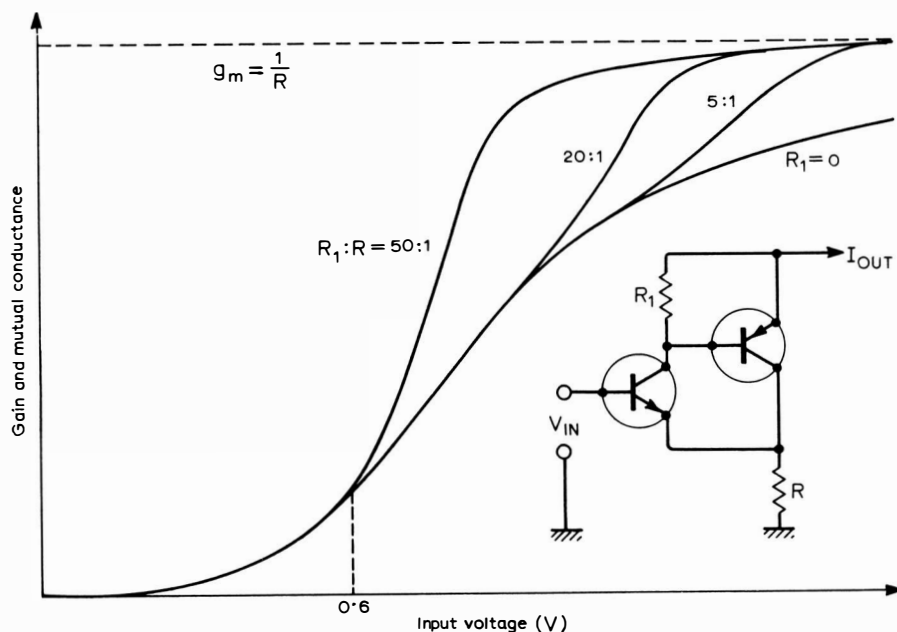


Fig.10. Transfer characteristic for conventional two-transistor sub-amplifier showing worsening effect of reducing the base-emitter 'turn-off' resistance of the power transistor. This normally generates 'dead zone' distortion due to the change in slope at zero crossing but is avoided in the new approach.

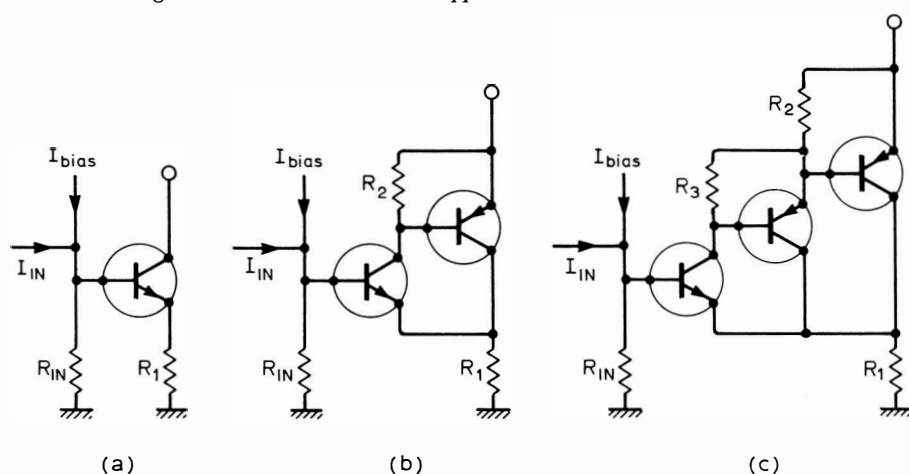


Fig.11. (a) Change in  $V_{BE}$  with temperature causes considerable change in bias current which could adversely affect intermodulation distortion. Circuits in (b) and (c) avoid this.

increase in distortion at low frequencies which could adversely affect the intermodulation performance of the amplifier as a whole. An improved design is shown in Fig. 11(b) and a more elegant version similar to that used in the Quad amplifier<sup>6</sup>, in Fig. 11(c). It is on this latter example that I have concentrated most design effort, mainly because the performance advantages tend to outweigh the increased cost of pre-driver devices.

Returning now to an examination of the performance of the whole amplifier—the total distortion through the audio range can readily be made less than 0.1% before applying feedback, this performance being repeatable at almost any level of quiescent current.

### Future designs

The use of class B amplifiers is not, of course, confined to the field of audio and in fact the ideas set out in this article lend themselves to applications in the high-frequency (> 1MHz) spectrum. The poor cross-modulation performance of present designs is usually due to the presence of non-linearities in the crossover region, hence substantial improvements can be expected in this direction.

Other applications where an ultra-low distortion amplifier of low stand-by power and high output capability is needed can be seen, examples of such devices being portable standard oscillators and meter calibration amplifiers. In the next article a practical design for a 30-watt audio amplifier is discussed in detail and future proposals developed in diagrammatic form.

(To be concluded)

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## Our 60th Birthday

The first issue of this journal, which for two years was entitled *The Marconigraph*, appeared in April 1911 and we therefore celebrate our 60th birthday this year. To mark the occasion we plan to have an enlarged April issue including several contributions reviewing the past 60 years in various fields. Further details will be given in our next issue.

# New Approach to Class B Amplifier Design

by Peter Blomley\*

(Concluded from February issue)

This article describes a 30-watt amplifier design which embodies the author's approach to class B design, outlined last issue. Although further work on this approach is still needed, the design illustrates the kind of problems involved. The author also discusses the application of integrated components in future designs.

The general design of a complete amplifier using the new approach is relatively conventional except for the inclusion of the signal splitter (described last month). In principle, the design of each half of the output stage is made simpler as there is no cut-off, hence

removing the necessity for predicting the performance in the cross-over region.

Examination of the circuit (Fig. 1) shows that the amplifier consists of three sections, the input amplifier, signal splitter and output amplifier.

**Input amplifier.** This converts the input voltage into a proportional output current

which drives the signal splitter. To enhance the performance of the amplifier as a whole, this section should have a reasonable mutual conductance (1A/V) and good linearity (1%). The latter does not represent a serious problem as the input amplifier is a low-level class A amplifier, but care is needed to control the maximum value of  $g_m$  otherwise frequency compensation problems arise.

**Signal splitter.** As many fundamental details of the signal splitter were described last month, further details are confined to the biasing system. If perfect bipolar devices were available and ideal current sources existed, voltage bias across the emitter-base junction would not be needed, but such situations do not exist and distortions due to conditions falling short of the ideal can be rendered negligible by employing simple bias diodes (Fig. 2). This reduces the voltage excursion at the input to the signal splitter from 1.2V to 300mV pk-pk. The waveform with a sinusoidal output current is shown in Fig. 3.

**Output stages.** This now is one of the easiest to design. As long as the gain remains constant throughout the output cycle all is well. In the initial version, used to evaluate system performance, a compromise was reached between complexity, performance and cost. Thus individual adjustment potentiometers were used instead of the matched devices.

The output sub-amplifiers are similar to the Quad triples, these giving excellent linearity down to very low output currents, coupled with outstanding thermal stability. To compensate for the effect of ambient temperature changes on the quiescent current of the amplifier, diodes  $D_1$  and  $D_2$  cancel  $V_{BE}$  changes in transistors  $Tr_7$  and  $Tr_8$ . It may have occurred to the reader that diodes in the forward path of the amplifier loop could generate appreciable distortion. However, in practice the maximum change in current is about 4:1 and thus almost corresponds to the change in collector current of transistors  $Tr_7$  and  $Tr_8$ . In this way the change in voltage drop across the transistors compensates for the change in the diodes. Even if this did not occur, the resultant gain change for the output sub-amplifier is less than 4% for  $I_{out}$  values between 0 and 2A. The problem can be alleviated by increasing the current into

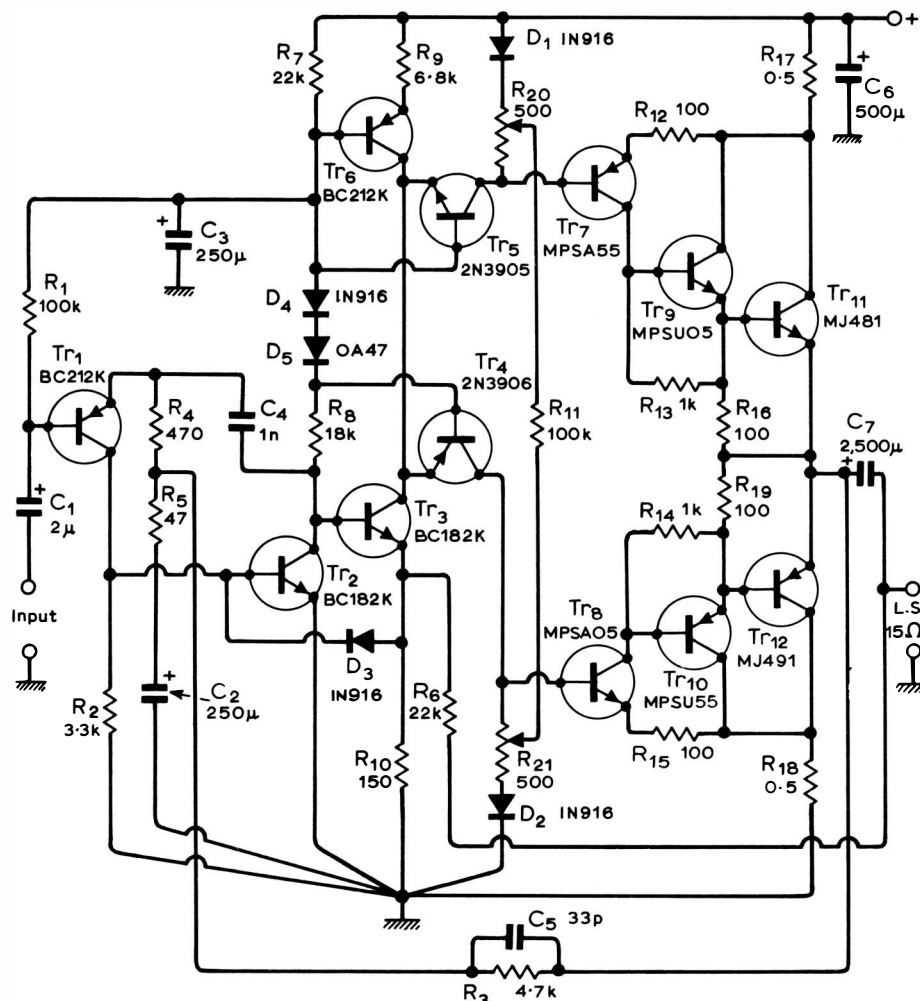


Fig. 1. Complete power amplifier circuit using new approach. Design gives harmonic distortion of 0.01% at all power levels and intermodulation distortion of 0.003%.

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diodes  $D_1$  and  $D_2$  and adding one resistor, but the advantages gained from this are negligible.

Circuit description

The function of  $Tr_1$ ,  $Tr_2$ , and  $Tr_3$  is to convert an error voltage—the difference between the input and feedback voltage—into a proportional output current. Now to produce the required mutual conductance of this stage (1A/V) without sacrificing either noise performance or linearity, the design in Fig. 1 was used. Starting at the input transistor  $Tr_1$ , this p-n-p type is used mainly as a level shifter. If we assume that the

current gain of  $Tr_2$  was extremely large ( $> 500$ ), then this input stage would have a maximum voltage gain of about five—not very much! If voltage gain was increased to the theoretical maximum of 30 (by decreasing the value of  $R_2$  and  $R_4$ ) problems would arise with the voltage offset at the speaker output due to increased emitter current flowing through  $R_1$  and base current flowing through  $R_1$ .

Assuming for the moment that this first stage gain is a reasonable compromise, it now becomes obvious that the noise and distortion performance is dictated by the next stage. This stage ( $Tr_2, R_8$ ) is a straightforward class A amplifier with very high

gain (typically 400) and low distortion due to the limited modulation index of the collector current (0.04 max). The peak 2nd harmonic voltage generated is about  $10\mu V$  and, assuming this is referred to the input of the first stage, it represents less than 0.001% 2nd harmonic distortion with feedback. Thus this second stage is the work horse of the input section, the third device  $Tr_3$  being used both as a buffer to reduce the loading of  $R_{10}$  on  $R_8$ , and to convert the voltage changes across  $R_8$  into an output current to drive the emitters of the signal splitter.

Resistor  $R_{10}$  performs two functions in this last stage of the input section. It defines the conversion constant  $Engmen$  for the stage, and it governs the maximum current which can be driven out of the collector of  $Tr_3$ . (This maximum current is defined by using the conducting voltages of  $D_3$  and  $Tr_2$  and the value of  $R_{10}$ .) Therefore this input section seems to have excellent performance during normal operation, but what can happen during an overload?

If the input transient was negative all would be well due to  $Tr_2$  entering saturation. But if the transient was positive  $Tr_1$  would turn off completely, the potential across  $R_{10}$  rising toward that at the end of  $R_8$ . ( $Tr_2$  would also be completely cut off.) This would cause excess currents to flow in  $Tr_3$ , upsetting the bias chain  $R_7, D_4, D_5, R_8$ . After the excessive input signal is removed some time would elapse before recovery would take place, hence diode  $D_3$  clamps the voltage and maintains  $Tr_2$  in full conduction to reduce recovery time and improve amplifier stability.

While discussing the problem of recovery from overload, the charge across the compensation capacitor  $C_4$  has also to be taken into account. The time for the accumulated charge to decay is a function of the amount of charge and the rate of decay. If the rate of decay is constant, the only way to reduce the recovery time is to limit the accumulated charge (in terms of voltage). Diode  $D_3$  performs this function as well as clamping the voltage across  $R_{10}$  at 1V thus limiting drive current into the signal splitter.

The second section is the signal splitter, unique to this approach, and consists of transistors  $Tr_4$  and  $Tr_5$  plus a current source transistor  $Tr_6$ . The signal current into the emitter of  $Tr_4$  or  $Tr_5$  is derived by subtraction of two current levels, one constant and set by the voltage across  $R_9$ , and the other the output current of the input section. This signal current either appears at the collector of  $Tr_5$ —causing a voltage change across  $R_{20}$ —or at the collector of  $Tr_4$ —causing a voltage change across  $R_{21}$ . These voltage changes are converted into positive and negative output currents in the output section, which are then added together to give the final waveform. The current gain of the output sections which are conventional triples are governed by the ratio of  $R_{20}$  to  $R_{17}$  and  $R_{21}$  to  $R_{18}$ , and in this case the gain of 1000 seemed reasonable.

To keep the output triples above the minimum conduction level a bias current is provided by  $R_{11}$ . The procedure adopted for setting the standing current is to first set  $R_{20}$  and  $R_{21}$  to minimum (diode end).

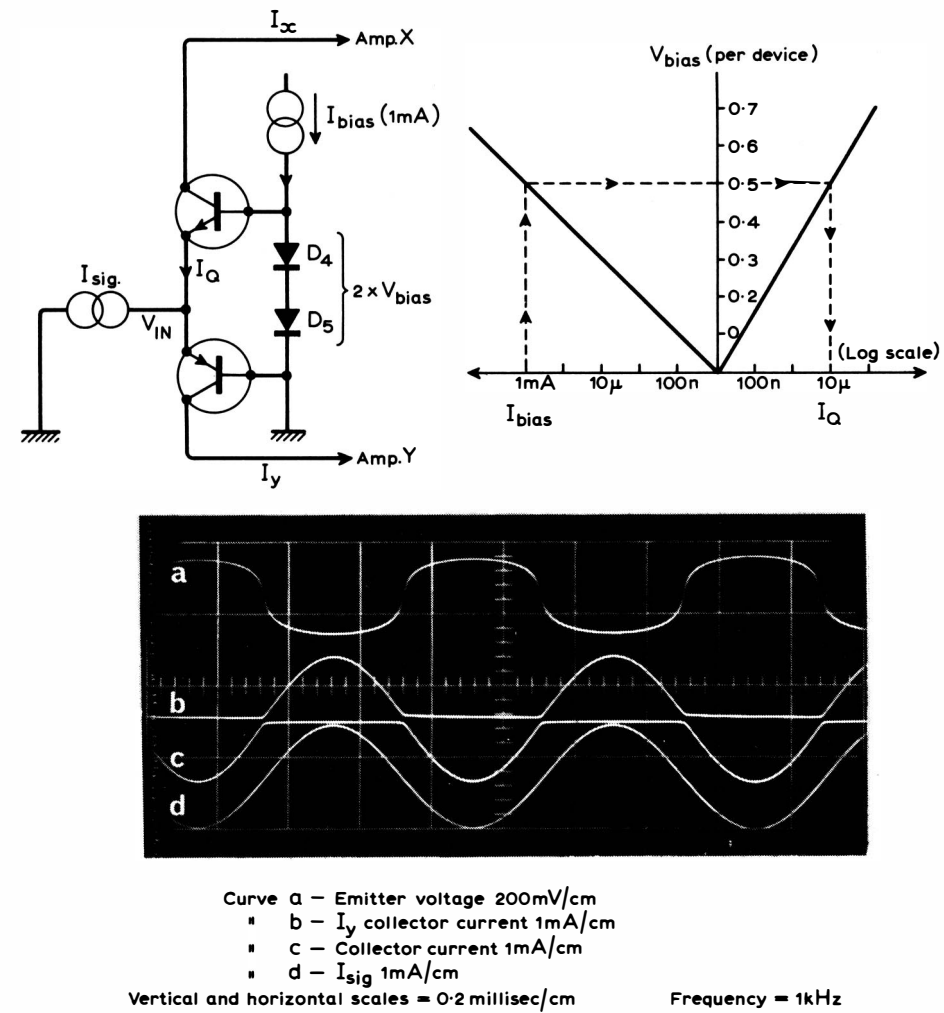


Fig. 2. Input amplifier converts signal voltage to a proportional current to feed transistor signal splitter. Bias diodes reduce voltage excursion from 1.2V to 300mV pk-pk. Bottom trace is current signal input to splitter.

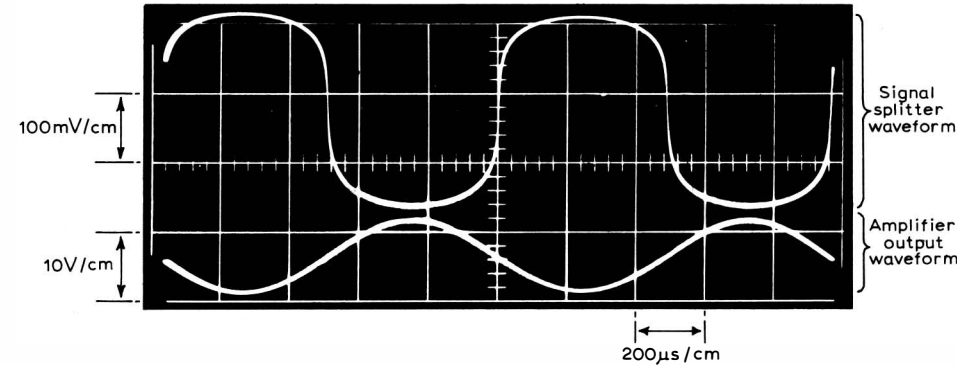


Fig. 3. Voltage excursion at signal splitter input with corresponding sinusoidal amplifier output current ( $R_L = 15\Omega$ ).

Set quiescent current with  $R_{20}$  and increase  $R_{21}$  until there is a small increase in current.

The only part still to be described is the biasing chain  $R_7$ - $D_4$ - $D_5$ - $R_8$ - $C_3$ . This provides the half supply voltage for the base of  $Tr_1$  (decoupled by  $C_3$ ), a load for the class A stage  $Tr_2$ , and sets devices  $Tr_4$  and  $Tr_5$  at the minimum conduction level required for good phase response during cross-over—by using the voltage across  $D_1$  and  $D_2$ . By increasing the value of  $C_3$  it is possible to reduce the rate of charging of the speaker coupling capacitor, eliminating 'thump', but capacitor size becomes very large.

Returning for a moment to the input section,  $Tr_2$  is in a similar position to that used in many amplifiers, but instead of driving another stage ( $Tr_3$ ) which only requires a limited voltage swing, it is the prime mover for the output section. To have sufficient drive capability the quiescent current in this stage may well need to be 10mA—instead of the 1mA in mine—and the voltage swing on the collector will be the full supply voltage (50 volts).

It now seems clear why the distortion of many amplifiers rises at low frequencies. The dissipation change of this device during a voltage cycle could be 500mW pk-pk in the case I have quoted giving an emitter-base voltage change at low frequencies of about 100mV. This change, even if we assumed it is basically a linear function of voltage, will cause a non-linear change in the input device and hence a considerable rise in distortion at low frequencies. In my amplifier the maximum dissipation change in  $Tr_2$  will be less than 1mW, thus eliminating this form of distortion and improving intermodulation performance.

## Performance

The measurement of distortion created some difficulties especially when con-

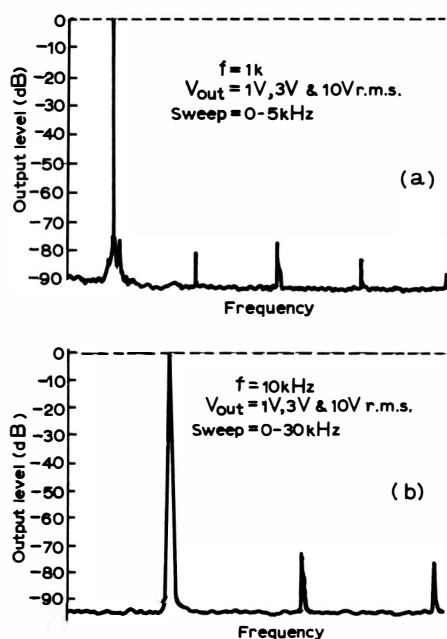


Fig. 4. Spectra made with a wave analyser showed no difference between spectra of outputs from oscillator and amplifier. Plots were made with (a) 1kHz and (b) 10kHz signals and were identical at all three power levels.

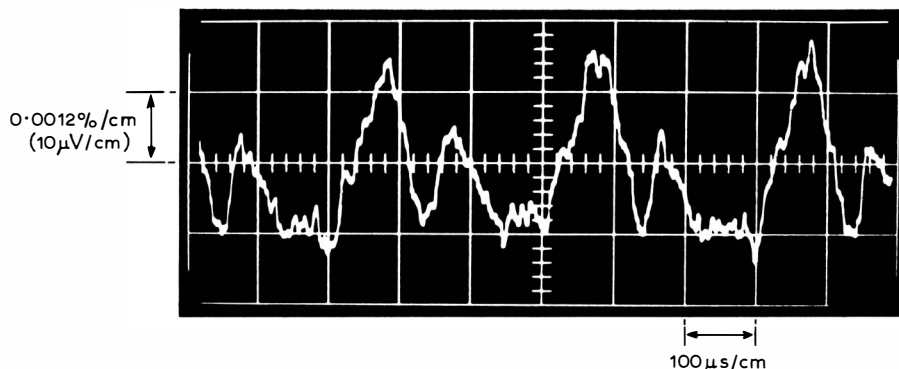


Fig. 5. Null method of assessing amplifier distortion shows distortion products to be well down in the noise. Deflection of 4cm represents 0.003% peak distortion at 10 watts (3kHz, 15Ω load).

sidering the range of frequencies over which this amplifier operates. The methods employed can be separated into two distinct techniques—spectrum analysis and nulling methods. To realize the first technique, an oscillator with a pure, single-line spectrum was needed, but the only one available at the time, approaching a reasonable performance, was the Si 451 produced by J. Sugden & Co, having a range up to 30kHz. This was found (excellent as it is) to be inadequate to permit the measurement of amplifier distortion.

So difficult in fact was the problem that it is impossible to publish distortion curves with any degree of confidence in their truth, but it can be said that using the Hewlett Packard 3590 wave analyser there was *no discernible difference* between a plot of the distortion of the oscillator and that taken after the oscillator output had been passed through the amplifier. Plots were taken over the frequency range 100Hz to 20kHz and powers of 100mW to 25W. As a matter of interest the spectrum plots of the amplifier are shown in Fig. 4 for 1kHz and 10kHz and at several power levels. The second method attempted was rather more successful but unfortunately does not present information in a usable form because it involves a comparison of output and input signals. It is also not a sequential test as in the previous method and as a result problems were encountered in successfully nulling the output against the input of the amplifier, due to the phasing of the signals and the earth loops generated by the measurement method. After considerable adjustment of the phase compensation and spurious pick-up difficulties the photograph Fig. 5 was obtained. Here the distortion generated is right in the noise (−120dB down from 20V r.m.s.) and the total deflection of 4cm represents 0.003% peak distortion at 10 watts and a frequency of 3kHz, chosen for easiest phase cancellation. The spikes usually evident in the difference waveform with this type of amplifier are completely absent, even with reactive loads, indicating that stability in the cross-over region must be excellent.

## Intermodulation performance

The use of these two techniques is limited in one way or another to the evaluation of

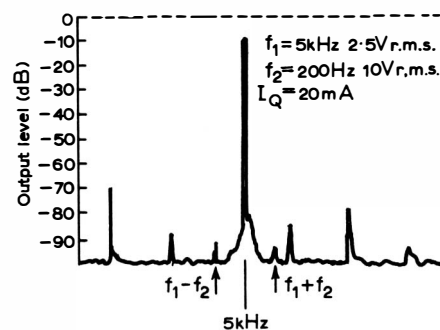


Fig. 6. Result of feeding 5kHz and 200Hz signals in a 16:1 power ratio into amplifier. Intermodulation products  $f_1 + f_2$  and  $f_1 - f_2$  are 90dB below 200 Hz signal. Other spectral lines are due to generator distortion.

amplifier linearity. The main advantage is, of course, that a direct numerical value of distortion is obtained which can be used in comparison with other amplifiers.

The intermodulation test does not rely on low-distortion oscillators or signal cancelling techniques—in fact the only component which limits the measurement accuracy is the wave analyser itself. The real drawback is seen when an interpretation of the results is necessary! The method adopted is to “sweep” the transfer characteristic of the amplifier with a low-frequency signal of large amplitude, and to “measure” the slope of the characteristic with a low-level high-frequency signal. The two frequencies selected were 200Hz and 5kHz in a power ratio of 16:1.

The results not only ease the assessment of the amplifier performance in an absolute sense but also give some form of subjective measurement for comparison with other elements in the system. The results obtained in Fig. 6 indicate an excellent performance, the intermodulation products  $f_1 + f_2$  and  $f_1 - f_2$  are −90dB below the sweeping signal (200Hz) all other spectral lines being due to generator distortion.

## Amplitude-frequency response

The type of frequency compensation used for this amplifier is unusual, mainly as a result of the system design. The open-loop gain begins to fall off at about 4kHz and continues on a 6dB/octave roll-off to about

500kHz where the second pole of the output section starts to contribute excess phase shift. The choice of the position of the dominant compensation was a difficult one. If it was placed in the output section, as is normally the case, the gain of the input amplifier would have to be restricted at low frequencies, affecting the distortion performance of the amplifier.

Another choice was using the dominant lag to encompass the output section as well as part of the input amplifier. This would lead to instability internal to the loop enclosed by the dominant lag and thus an internal pole would have to be introduced to remedy this condition. The final choice (shown in Fig. 7) gives the single-pole compensation needed for unconditional stability coupled with minimal high-frequency distortion. The inherent pole in the output section is subdued by the feedback resistance  $R_3$  (so far as the main loop is concerned) but gives the required unconditional stability of the output section.

The performance with reactive loads will be spoilt if the output impedance of

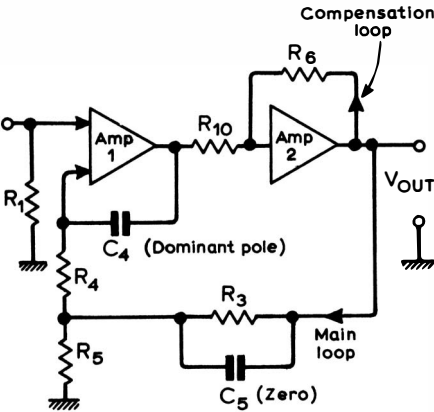


Fig. 7. Single-pole frequency compensation method used gives unconditional stability coupled with minimal h.f. distortion.

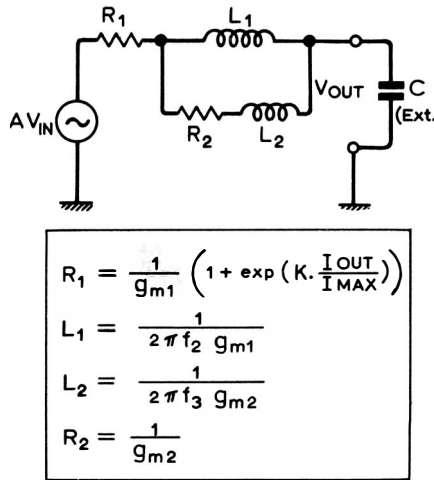


Fig. 8. Power amplifier equivalent circuit. Simple analysis shows output impedance is controlled by main feedback loop, but in practice  $R_6$  generates another loop effectively placing a damping resistance across the apparent output inductance.

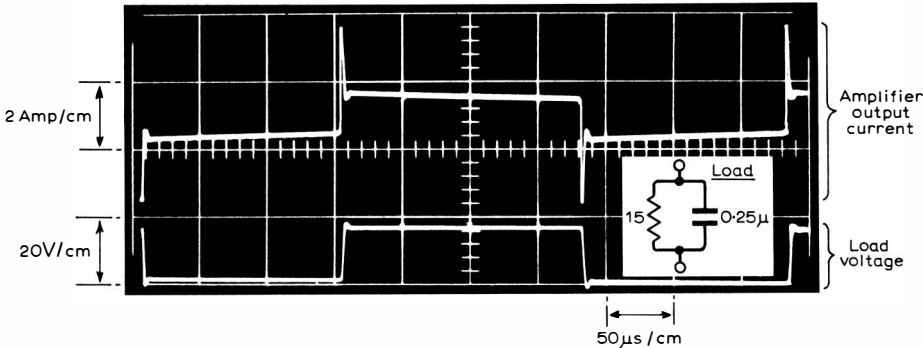


Fig. 9. Performance with a capacitive load. Capacitor in feedback loop effectively reduces maximum rate of change of voltage across load. Overshoot is much less when fed from a pre-amplifier.

Performance—with 60V regulated supply	
output power	20 watts into 15 ohms 30 watts into 8 ohms
power response	30Hz to 100kHz (–3dB)
output impedance	0.1 ohm at 1kHz
total harmonic distortion	< 0.01% throughout audio band and all power levels
intermodulate distortion	< 0.003%
voltage gain	100
noise level	–120dB below full power
maximum peak output current	±3 amps, approx.

the amplifier is controlled by the overall feedback loop, i.e

$$Z_{out} = \left(1 + \frac{f_1}{f_2}\right) / g_m$$

where  $f_1$  is the signal frequency and  $f_2$  the open-loop –3dB frequency. This expression has a simple analogy with a series inductance and resistance, where  $R = 1/g_m$  and  $L = 1/2 \pi g_m f_2$ .

A little more work† shows that if a capacitive load is used the amplifier would have a response given by

$$G = \frac{1}{p^2 T^2 + a p T + 1}$$

This is the equation of a second-order system, where  $a = (1/g_m) \sqrt{(C/L)}$ , and the natural frequency of oscillation is  $\omega_o = 1/T = 1/\sqrt{(LC)}$ . If the amplifier has an overshoot it must be due to the overall amplifier having an  $a$ -value approaching zero. If we now assume typical values and examine the worst case condition,  $g_m = 10A/V$ ,  $f_2 = 4kHz$  and  $a = 0.1$  (20dB peak), then  $C = 4\mu F$  and  $\omega_o = 250kHz$ .

If this was a perfect model for the amplifier the overshoot would be excessive, but in practice the output impedance is not only a function of frequency but also of output current. Thus  $a$  gets larger (less overshoot) as the output current increases. The basic assumption of this simple analysis is that the output impedance is controlled by the main feedback loop, but in this amplifier resistor  $R_6$  generates another loop which effectively places a damping resistance across the apparent output inductance (Fig. 8).

The only remaining improvement to the transient performance of the amplifier is by pole-zero cancellation using the feed-

back element. If this term seems somewhat academic, an alternative is to study the overshoot with a second-order system with various inputs. If the input is an ideal step the amplifier will give theoretical overshoots, but if the rate of rise of the input waveform is decreased the overshoot will reduce and eventually disappear. The capacitor (a zero) in the feedback loop is really reducing the maximum rate of change of the voltage across the load and hence the degree of excitation given to this inherently oscillatory system. By using this type of compensation excellent performance with reactive loads has been finally achieved (Fig. 9). The overshoot with capacitive loads, such as  $4\mu F$ , is about 50% with an ideal step input and far less when fed via a preamplifier, thus no difficulties should be experienced with any normal load.

**Electrostatic loads.** The distortion characteristic with this type of load was still insignificant below 10kHz and gave a gradual rise up to 20kHz where it was still less than 0.05% at maximum output ±. Square-wave performance is shown in Fig. 10 at maximum ± output. The ringing is due to the finite output impedance converting the ringing current in the inductance and capacitance of the load into ripples in the output, plus the overshoot of the amplifier itself.

### Future developments

The amplifier design is hopefully only a source of ideas which may encourage further research into the whole approach to design. So that the trend may be continued, future proposals are outlined in Fig. 11. Here, the main difference is that

† See for instance "Active filters" F. E. J. Girling and E. F. Good, *Wireless World*, vol. 75, Sept. 1969, pp. 403-8. [www.keith-snook.info](http://www.keith-snook.info)

± Maximum output is dictated by peak current output capability.

the output subamplifiers are oriented toward the use of integrated components. It has become obvious that past problems with class B amplifiers originated with the stabilization of the quiescent current to give zero cross-over distortion. Attempts were made to use diodes to compensate for device  $V_{BE}$  changes with fluctuations in the ambient temperature—the independent variations due to device dissipation could not be eliminated. Most of the time the diode did its job and the voltage defined by the combination of transistor and diode remained constant. This constant voltage was used in conjunction with low-value resistors to set the quiescent current in the output circuits.

If now an integrated component is used both the diode and the transistor are on the same chip and, apart from minor fluctuations, the combination is isothermal. As a result the quiescent current is a function *only* of the setting voltage and not ambient temperature or differential device temperatures. The accuracy with which the current can be set is largely governed by the offset voltage of the transistor pair. Typical values of  $\pm 4\text{mV}$  which would represent a  $\pm 8\text{mA}$  inaccuracy in the quiescent current using 0.5-ohm feedback resistors are readily obtained. With such an arrangement a reasonable quiescent current for the sub-amplifiers would be 30mA, the worst case figures would be 24mA and 38mA. Both of these values are well above the low conductance current level (5mA) which is required for good linearity of the sub-amplifiers.

The advantage of the new approach is fairly evident when it is realized that as long as the amplifiers are above the non-linear region, the spreads introduced in the sub-amplifier quiescent current will not cause the class AB situation of over-biasing (shown last month) characteristic of present designs. It is now possible to design an output stage without the normal trim potentiometers, thus giving a degree of freedom in production not possible with current amplifiers. The performance of the amplifier, once checked at the end of a

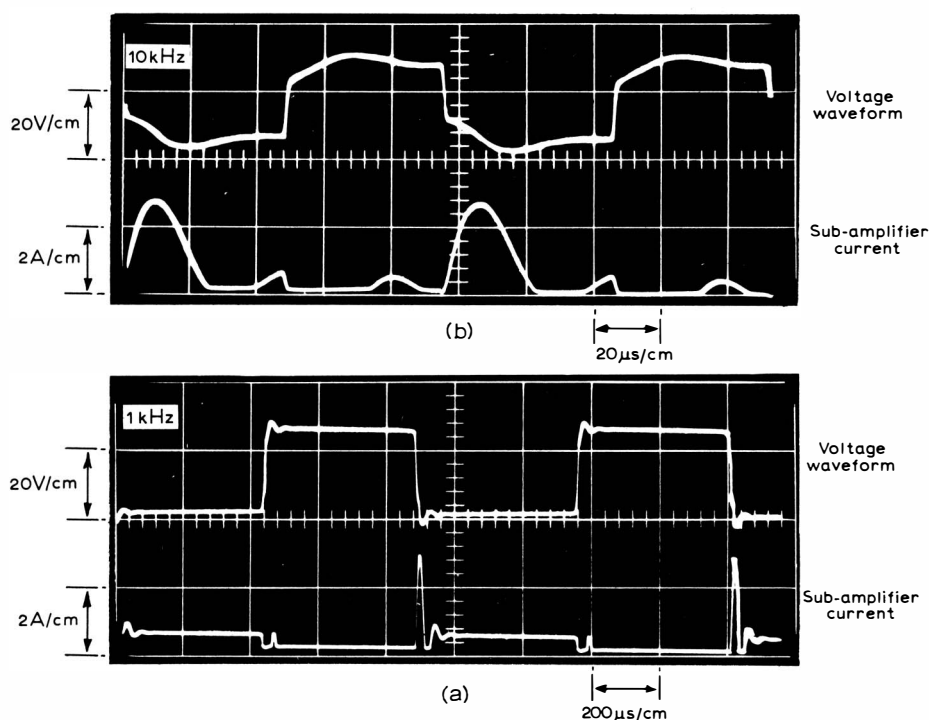


Fig. 10. Square-wave performance when driving electrostatic load at 1kHz (a) and 10kHz (b). Top traces are voltage and lower traces current out of sub-amplifier. Ringing is due to output impedance converting ringing current in  $L_2$  and  $C_2$  into ripples in the output.

production line can be guaranteed for operation in any climate and for any period of time.

### Possible applications

The performance of an amplifier of this calibre is, in my opinion, wasted in a conventional audio set-up. In most cases, the transducers will be the weakest link.

The approach used in the design of the output sub-amplifiers does not rely on complementary matched devices—in fact, in most cases n-p-n devices are preferred for their superior secondary breakdown characteristics. This represents considerable reduction in amplifier costs especially in the 100-watt region as presently available devices boast a  $V_{CEO}$  of 120V with

100 watts dissipation at a cost of less than 75p.

The ultimate use for this amplifier would appear to lie with the high-power professional market where the performance of cascaded amplifiers in a system would have to be excellent. Use in other fields would be mainly governed by the expected gain in performance or reduction in cost. A possible application would be as a portable standard oscillator, perhaps meter calibration amplifiers, or even high-frequency low-distortion class B transmitter amplifiers. However, these are only inspired guesses which may interest those working in these relevant fields.

Thanks are due to Peter J. Baxandall for his advice and encouragement and to Hewlett Packard and the Plessey Co. for use of their facilities.

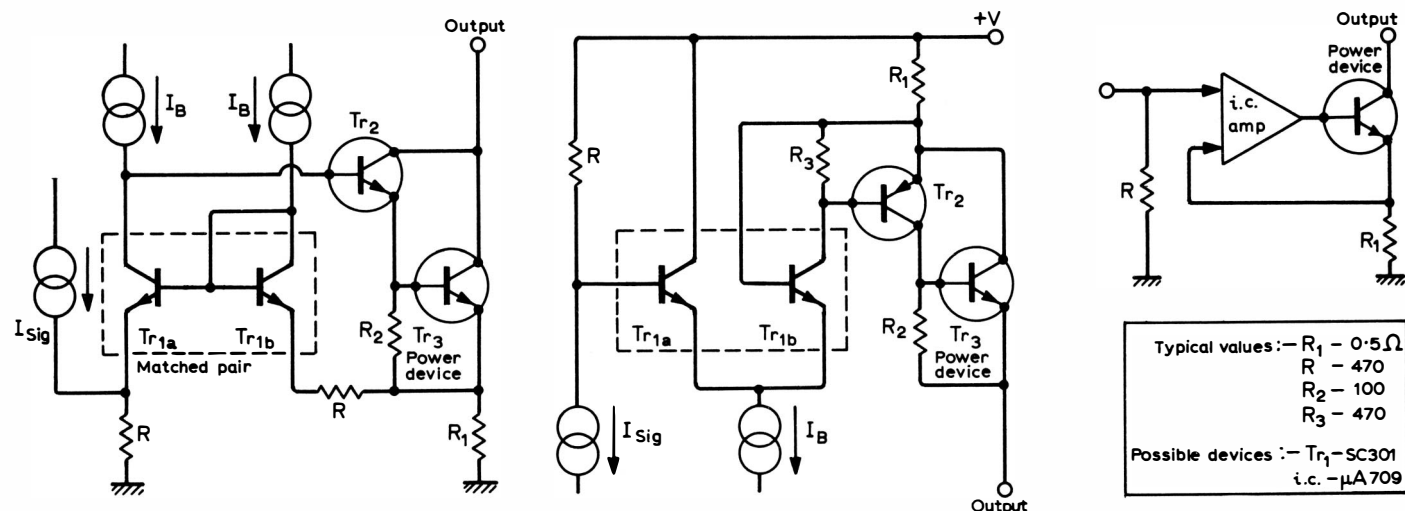


Fig. 11. Proposals for integrated components in output sub-amplifier.