Wireless World Circard

Series 7: Power amplifiers

For the most part, these cards discuss circuits for audio-frequency amplifiers, including direct-coupled circuits, though r.f. amplifier circuits are included; video amplifiers are to be treated separately. The article gives a good introduction to power amplifiers and their configurations. Efficiencies and power output for different kinds of output stages are derived for class A and B circuits, and the brief survey concludes with a discussion of class C and D amplifiers for completeness.

Two class A and two class B circuits are described for output powers up to about 5 watts. The general-purpose wideband amplifier, with its 10-octave bandwidth up to 16MHz, is a useful circuit to have to hand. The other r.f. one—a class C circuit providing 1·4 watts at 7·2MHz —is more specialized and it was not possible to give design procedure on the card for other levels and frequencies (but refer to further reading).

The broad usefulness of the set is aided by cards describing a pulse amplifier, a pulse-duration modulated output stage, bridge amplifier techniques and a 100-volt amplifier circuit.

Power amplifiers

All amplifiers are power amplifiers in that the power delivered to the load is greater than that drawn from the source. Few are power amplifiers in the same sense that an operational amplifier with feedback may be said to be a voltage amplifier or a current amplifier. Thus in Fig. 1 the load voltage is defined for a given input signal and the load power is proportional to the conductance of R_L. For Fig. 2 feedback defines the load current while the corresponding power developed in the load is proportional to the resistance of R_L.

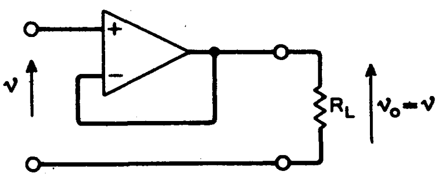
This suggests that as many power amplifiers use shunt-derived feedback to define their output characteristics, they could properly be regarded as voltage amplifiers which just happen to be capable of delivering large powers to a load of sufficiently low resistance. Operating these amplifiers from a constant supply voltage in the usual way fixes an upper limit on the load voltage. Practical imperfections in the transistors together with current limiting resistors or other protective circuitry reduce this upper limit but still leave the peak output voltage broadly proportional to the supply.

Output power depends equally on the maximum current that can be supplied to the load. The mean value of this power over a complete cycle for a sinusoidal output voltage and resistance load is $\hat{V}\hat{I}/2$, where $\hat{V}$ and $\hat{I}$ are the peak instantaneous values of voltage and current, and as $\hat{V} = \hat{I}R_L$ and $V_{rms} = \hat{V}/2$, alternative expressions are $V_{rms}.I_{rms} = \hat{V}^2/2R_L = I^2R_L/2 = V_{rms}^2/R_L = I_{rms}^2R_L$.

At this point, you might be asking: "what about r.m.s. power?". This, unfortunately, is on a level with the equivalent enquiry after the well-being of the workers. It can be answered in various ways none of which are useful. To interpret it properly it must be realized that, while the power developed in the load varies from instant to instant, it is the mean or average value of that power that determines, for example, the loudness of the sound produced by a given loudspeaker. The r.m.s. value of the power can be defined mathematically in the same manner as the r.m.s. value of the voltage, but it has no comparable physical significance, i.e. it would require the instantaneous power to be squared, integrated to obtain the mean value of the "power squared", and then the square root taken of the result. The confusion arose because mean power happens to equal the product of r.m.s. voltage and r.m.s. current for certain specified conditions which commonly occur. Hence the "r.m.s." term has become firmly attached to the power measurement itself particularly in the advertising for audio equipment. It should be detached.

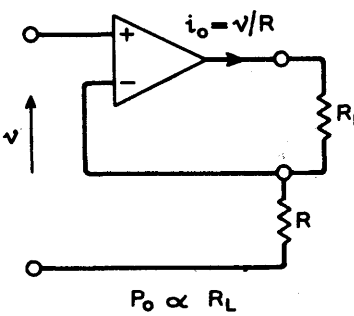
While the voltage term in the output power is fixed by the supply voltage, the current term is a property of the amplifier. Consider first the amplifiers based on a single transistor as shown in Fig. 3. In each case the transistor is assumed to be dealing with an a.c. signal which has equal positive and negative magnitudes, e.g. a sine wave. Thus the transistor must be biased to some quiescent voltage/current setting which will allow positive and negative-going output swings. If distortion is to be avoided the transistor must remain conducting throughout the cycle, i.e. neither the current nor the p.d. across the transistor shall fall to zero. This mode of operation, class A, may be defined in terms of the "angle of conduction", being the full 360° of the cycle. In class B each device conducts for precisely 180° or half the cycle and in class C conduction is for $<180°$.

In Fig. 3 (a) if the direct current is permitted to flow in the load, equal positive and negative excursions occur for a voltage across the transistor equal to half the supply voltage (assuming an ideal transistor). Thus the peak of the a.c. component of load voltage is $\hat{V} = V/2$. Hence the a.c. power in the load is $(V/2)^2/2R_L = V^2/8R_L$. In the absence of signal the current drawn from the supply is $V/2R_L$ giving a supply power $V^2/2R_L$. This shows an efficiency of 0.25. The quiescent power splits 50/50 between transistor and load resistance.



$P_o \propto 1/R_L$

Figs. 1 and 2. In a voltage amplifier and a current amplifier, load voltage or current is defined for a given input signal and load power is proportional to the conductance of R_L (Fig. 1 above) or the resistance of R_L (Fig. 2 below), suggesting that power amplifiers using shunt-derived feedback could be looked on as voltage amplifiers, but capable of handling large powers.



$P_o \propto R_L$

It is possible to do still worse. In Fig. 3 (b) the load is capacitively coupled to the amplifier to eliminate the direct current in the load. A collector resistor is still required to allow the flow of current in the transistor, establishing the quiescent conditions for class A operation. Now the total a.c. power is split between R and R_L and the maximum efficiency is reduced to 0.125.

The situation can be retrieved if the collector resistor can be replaced by some constant-current stage as in Fig. 3 (c). The positive peak current in the load can then equal the quiescent current even when the collector approaches the supply voltage (assuming a constant-current stage that can function with a p.d. falling towards zero). Hence the load can have a maximum current swing simultaneously with a maximum voltage swing. In Fig. 3 (b) when the transistor current falls to zero, R and R_L are effectively in series and the p.d. across R limits that across R_L.

The constant-current stage may consist of an inductor whose reactance is high compared to the resistance of the load at all frequencies of interest; an ideal transformer that also allows the use of arbitrary load resistance by proper choice of turns ratio; a transistor biased to deliver a constant current. The disadvantage of (d) and (e) is that the amplitude-frequency response is limited unless bulky and expensive inductor/transformers are available. They do offer the possibility of higher efficiency than any of the other circuits. For example, Fig. 3 (f) allows the peak current in the load to equal the quiescent current, and the peak voltage to equal the transistor quiescent voltage, i.e. half the supply voltage for the best operating conditions. Thus load power is $(I_{dc}V/2)/2$ while supply power is $V.I_{dc}$, giving an efficiency of 0.25 bringing the efficiency back to the level of Fig. 3 (a) but with the d.c. component removed from the load. In Fig. 3 (d), the peak current in the load is still equal to the quiescent current, but the inductance allows the collector voltage to swing positive with respect to the supply line as the transistor current falls—a load peak voltage equal to the supply voltage being available. The a.c. power in the load then becomes $VI_{dc}/2$ for the same supply power VI_{dc}, and maximum efficiency is 0.5. This is the highest efficiency possible in class A and the transformer-coupled circuit of Fig. 3 (e) has the same capabilities. It is common for practical circuits using small transformers to have efficiencies in the region of 0.25 to 0.4.

The low efficiencies attained by these single-device circuits lead to the investigation of multiple transistor circuits. Simply operating transistors in parallel may increase the quiescent power they can dissipate and hence the available output power, but the method offers only second-order improvements in efficiency by reduction of saturation voltage etc. Before turning to other classes of amplifier, consider the natural extensions that are possible of the circuit of Fig. 3 (f). Replace the transistor used as a current source by one receiving a signal as in Figs 4 (a) to (d). The signals to the two transistors depend on the configuration used, but the aim in each case is to cause one transistor to decrease its current by the same amount as the other increases it, still assuring class A operation for each transistor individually. In this way the peak current in the load may equal twice the circuit quiescent current: at the point when the current in one transistor reaches zero the other has doubled. The peak load voltage remains at half the supply voltage when the output is biased to the supply mid-point for maximum undistorted output.

So far the term "matching" has been avoided. For low-level amplifiers, input impedance is frequently matched to that of the source; this is the condition for minimum noise. To maximize the power gains of following stages the output and input impedances may be matched, i.e. made equal. This remains the practice in r.f. amplifiers, but at lower frequencies deliberate mismatch is more common as it allows for better control over the gain.

A fallacy that is based on the experience with these low-level stages, and derives from the maximum power-transfer theorem is often extended to power amplifiers. For each of the class A stages described earlier there is a value of load resistance that maximizes the output power without clipping the waveform peaks. As shown for each individual case there are separate limits for both peak voltage and peak current, and the optimum load will have a resistance given by the ratio of these peaks. This load resistance has nothing to do with the output resistance of the transistor.

Consider Fig. 5 which represents the operation of Fig. 3 (d). Draw the load line representing the load resistance through the quiescent point; the maximum output power without distortion is achieved for the slope giving symmetrical voltage and current swings in the positive and negative directions. Such a line intersects the V, I axes at $2V_c$ and $2I_c$ respectively and the slope of the line is the same as that joining the V_c, I_c points on the axes. This optimum load is thus confirmed as depending on the quiescent conditions with the slope of the transistor characteristics (the true output resistance) playing no part. Life is rarely that simple in practice, and the results are modified by saturation effects as well as by the various non-linearities, but not sufficiently to disturb the principle, which applies equally to the circuits of Fig. 4.

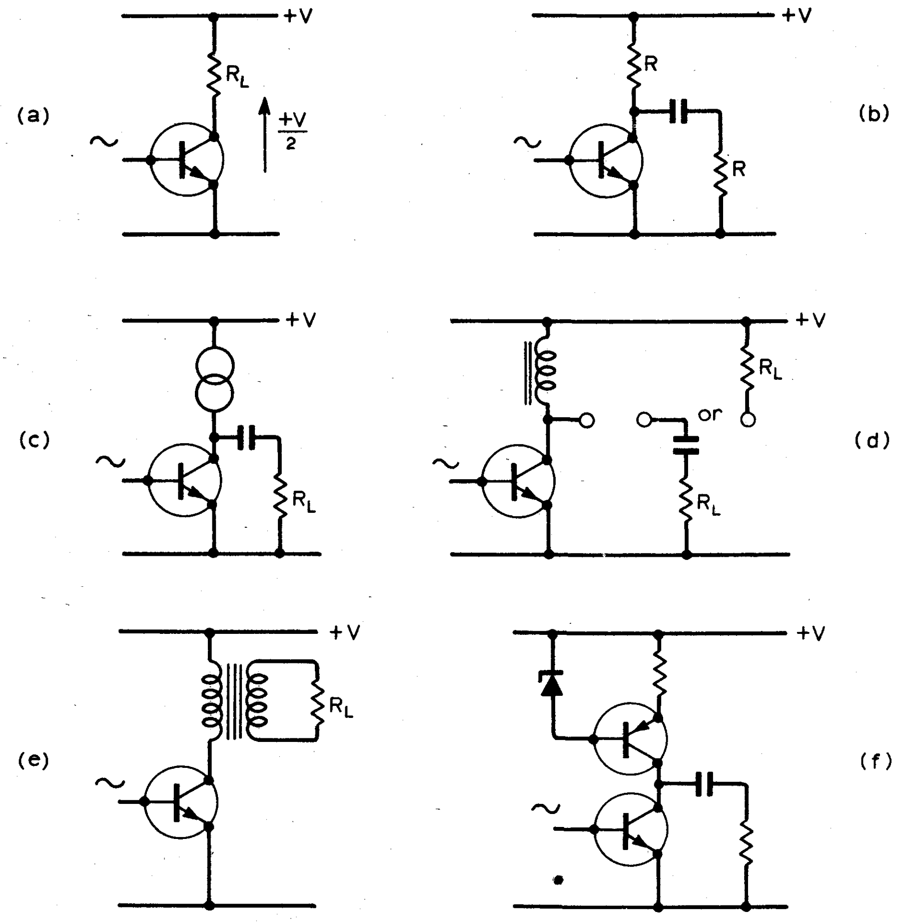


Fig. 3. In a resistive-load class A amplifier (a) efficiency is only 25%; with a capacitatively-coupled load efficiency could be 12½% for equal values of resistor shown (b). Replacing collector resistor by a constant-current circuit (c) means the peak load current can equal the quiescent current. Examples of constant-current circuits are a simple inductor (d) or transformer (e) both giving maximum possible efficiencies of 50%, or at (f) using an additional transistor, efficiency 25%.

Hence for Fig. 4 (c) when used as a class A amplifier the quiescent current (I_s) may be calculated from the supply voltage (V_s) and the intended load resistance. Peak voltage in load is $V_s/2$; peak current in load is $2I_s$; therefore optimum load resistance is $V_s/4I_s$. Resulting mean load power is $(V_s/2)2I_s/2 = V_sI_s/2$, corresponding to an efficiency of 0.5 for ideal transistors.

These circuits are not restricted to class A operation. Indeed they are more commonly used as push-pull class-B amplifiers in which the bias network (not shown) is adjusted to bring both transistors to the edge of conduction. Each transistor then conducts during one half-cycle, there being no quiescent current. There is no comparable limit to the peak current that may be provided; class B simply demands that conduction takes place in a device for 180° in the cycle. A limit will be imposed in any particular design by the current/power limitations of the transistors/power supply. In principle any basic design for a class B power amplifier using configurations such

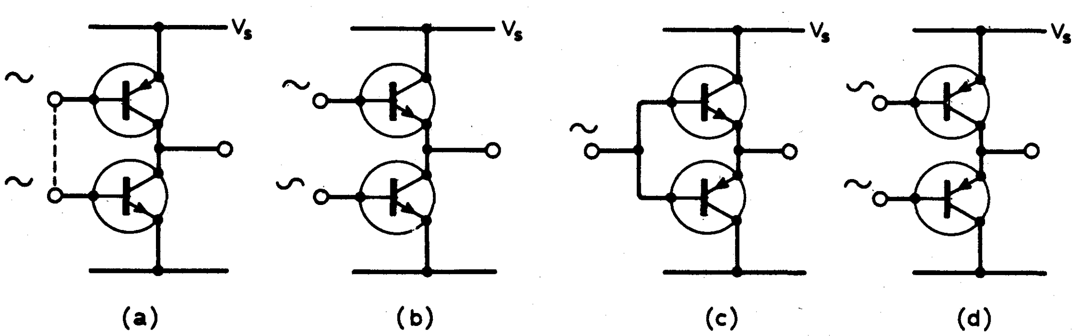


Fig. 4. Using a second for signal handling enables the peak load current to equal twice the quiescent current, with efficiencies of up to 50%.

as those of Fig. 3 (c) and (d) may be extended to higher current levels by replacing the output transistors with Darlington pairs, complementary pairs etc. Thus 100W and 100mW amplifiers may be surprisingly similar in configuration. At high power levels the importance of protection and of maintaining stable bias leads to the addition of circuits monitoring and/or controlling the current in the output stage.

To minimize the distortion due to device non-linearity at low currents (cross-over distortion) the bias networks are set to provide some quiescent current, setting the operation intermediate between true class B and class A—often called class AB and further subdivided into AB_1, AB_2 according to the fraction of the cycle for which each device is non-conducting. The design of low-distortion power amplifiers is a highly specialized subject that will warrant separate treatment in a later series though outline design procedures and practical examples of simple and economical circuits are given in this series of Circards.

Quiescent power in class B is zero. Maximum output power with ideal transistors (Fig. 6) is $\hat{V}^2/2R_L$ where $\hat{V}$ is $V_s/2$. Therefore $P_{L\,max}$ is $V_s^2/8R_L$.

Under these conditions, the mean current drawn from the supply is $V_s/2\pi R_L$. This is because the current is drawn from the supply only during the positive half-cycle; the negative half-cycle results in charge being withdrawn from the large coupling capacitor, which charge is restored during the next positive half-cycle. The mean power drawn from the supply is $V_s I_{dc} = V_s^2/2\pi R_L$ and the corresponding efficiency is

$$\frac{P_L}{P_s} = \frac{V_s^2}{8R_L}\cdot\frac{2\pi R_L}{V_s^2} = \frac{\pi}{4}$$

or 78.5%.

As the load power is proportional to the square of the output voltage while the supply power is proportional to the voltage it follows that efficiency is proportional to output voltage. It is also true that at some intermediate level of output, the load power having fallen faster than the supply power, the power in the transistors passes through a maximum. For sine wave drive the maximum dissipation in each transistor is at an output voltage where $\hat{V}$ is V_s/π, and the dissipation is then about one fifth of the maximum output power derived above, i.e. a 10W amplifier could theoretically be constructed using a pair of transistors with power ratings of only 2W each.

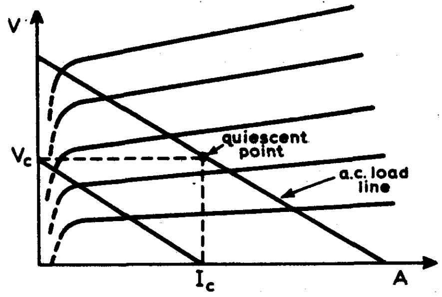


Fig. 5. Optimum load for maximum signal handling is represented by load line cutting axes at $2V_C$, $2I_C$, for Fig. 3 (d) or (e) and depends on the quiescent condition and not transistor output resistance. In general optimum load can be calculated from peak voltage/peak current.

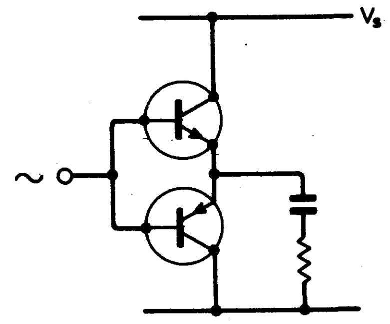


Fig. 6. In this typical class B stage, text shows maximum efficiency to be 78.5%.

Class C amplifiers are normally restricted to tuned amplifier systems, conduction taking place for only a small part of a cycle, and the recovery of a sinusoidal waveform then demands a high-Q circuit. The exception is the power-control circuits such as controlled rectifier and triac circuits in which only the mean value of voltage/current/power is of interest and waveform shape is non-critical. These differ from conventional class C circuits in that the conduction has a controlled starting angle but always finishes near the end of a half-cycle; class C r.f. amplifiers are biased such that the conduction angle is symmetrical about the peak of the drive waveform.

A further difference is that the power control devices are operated as nearly as possible as perfect switches, while at the high frequencies normally associated with class C stages, a very detailed design procedure is required to cope with transistor parameters. This will normally include complex conjugate matching to source and and load, to optimise performance. Efficiency can exceed that for class B, though power losses in the passive components involved in the tuning/matching processes are inevitable. A further application of class C power amplifiers is in frequency multiplication where the output circuit is tuned to a harmonic of the input frequency. These aspects are germane to more detailed later studies of r.f. circuits.

Class D is the generic term for switching circuits in which the active devices are switched multiply in and out of conduction during a single cycle of the input signal. They are also power realizations of pulse modulation systems, the theory for which can be used to determine the spectral components of the output. As one example, circuits such as those of Fig. 4 may have their signal drive replaced by high-speed pulse waveforms whose widths are modulated by the received signal. If the load is fed via an LC filter, the pulse frequency terms can be removed and the output transferred to the load is proportional to the signal.

For ideal transistors the switching process allows for zero power dissipation; at all times either the transistor p.d. or its current tend to zero. If the unwanted terms are to be well outside the band of frequencies that it is required to amplify, then the pulse frequency may have to be so high that serious power losses occur during the instants of switching, while charge storage in either of the transistor base-emitter junctions can lead to excessive current spikes through the series path then provided by the two transistors. The principle is more readily applicable to small servo-motor systems than to audio amplifiers as the electromechanical properties of the load do not require very high switching frequencies. In some cases efficiencies may exceed 90% with 100% as the theoretical upper limit.

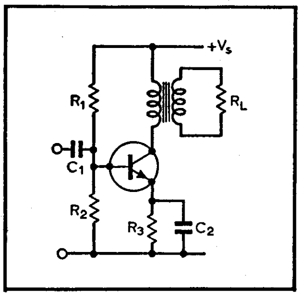
In all the above circuits an ever-present problem is that of protecting both the circuit and its load from excessive current flow. Much time and energy is expended on systems for protection against faults, but inevitably accidents happen, so often after some improvement or embellishment has been added. One can only wonder if such thoughts may have been in the mind of William Wordsworth when he wrote

I have submitted to a new control;
A power is gone which nothing can restore;
A deep distress hath humanized my soul.

A cry from the heart that will speak to all designers of power amplifiers.

Wireless World Circard

Basic power amplifiers



Typical data
Supply: +15V
Tr: BFY50
R_1: 1.2kΩ
R_2: 120Ω
R_3: 10Ω
T: 3.25:1 turns ratio
Quiescent current: 70mA
Output power into 25Ω load: ~400mW for 10% distortion

Class A

The classic transformer-coupled class A amplifier has been superseded for most purposes, but may still be applied where good isolation is required between source and load, or where the optimum impedance for maximum undistorted output is very different from the load impedance. Resistors R_1 and R_2 fix the base potential of Tr_1 provided the current through them is much greater than the base current. This base current is the required collector quiescent current divided by transistor h_{FE}. These parameters fix the value of $R_1 + R_2$ by the approximate relationship $R_1 + R_2 = h_{FE}\, V_S/m\, I_C$. The value of m the ratio of divider current to base current, is a compromise between stability and wasted power. Typically $m = 5$ to 20.

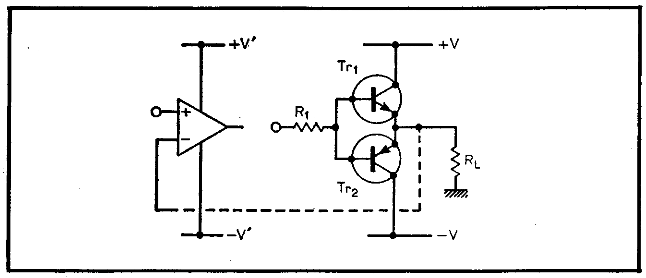
Emitter current (and hence I_C) is defined because the p.d. across R_3 equals the p.d. defined across R_2 minus the V_{be} of Tr_1. For silicon transistors this is 0.6V and is stable to within 10 or 20% for most transistors under most operating conditions. The resulting p.d. across R_3 is again a compromise between high values for better stability and low values for minimum wasted power – not less than 0.5V and not greater than say 20% of supply voltage as a guide for power stages. Capacitor C_2 decouples R_3 to prevent negative feedback within the required frequency range. As R_3 may be a low resistance, C_2 must then have high capacitance.

Class C

The basic principle behind class C amplifiers is simple, the efficient realization difficult. The transistor conducts only on positive peaks of the input signal with the RC time constant determining the angle in the cycle for which conduction continues, the base-emitter of the transistor acting as a diode and allowing C to charge during the peak. The current in the output circuit is then in the form of pulses of current of which the fundamental term flows in the load if the LC circuit resonates at the fundamental frequency. A high-Q circuit ensures that the harmonics are sharply attenuated giving good output waveform simultaneously with high efficiency. A wide range of load and source impedances can be accommodated by introducing suitable LC networks at input and output (see card 6).

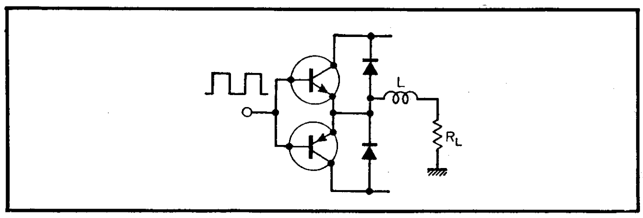
Class B

The complementary pair of transistors acting as emitter followers comprise the basic class B push-pull stage. Transistor Tr_1 conducts during the positive half-cycle and Tr_2 during the negative half cycle. For input voltages close to zero neither transistor conducts as each requires a finite base-emitter voltage for conduction to commence (~0.5V for silicon devices). Non-linearities at low-levels make direct voltage drive at the bases unattractive, with the resulting cross-over distortion being very apparent in badly designed amplifiers of this time. If the output stage is included within



Supplies $V = \pm 6$V, $V' = \pm 15$V
Tr_1: BFR41
Tr_2: BFR81
IC_1: 741
R_1: 470Ω
R_L: 15Ω
Output power 0.92W at 64% efficiency
Output voltage swing: 10.5V pk-pk for ±6V supply

the feedback loop of a high-gain amplifier the negative feedback reduces the distortion very considerably. At high frequencies the falling gain of the op-amp prevents the feedback from being fully effective and the crossover reappears. Voltage-gain as shown is unity, but standard feedback networks may be used to obtain any desired voltage gain. Output may be increased to 1.75W into 8Ω but heat-sinking is then advisable. If the objectionable audio effects of crossover are to be minimized biasing networks are inserted between the transistor bases.



Class D

In the class D amplifier, one or more transistors act as switches, connecting the drive point of an LR series circuit to the supply lines. This delivers a square wave to the LR circuit and provided the reactance of the inductor is high at the switching frequency there is little output. If the duty-cycle of the input waveform is altered the output will have a mean level which is a function of the duty cycle. A frequency lower than that of the basic switching frequency is used to modulate the pulse-width/position of the square wave generator and the low voltage is then a function of that signal voltage. For ideal transistors there is no power lost at the switching frequency and the overall efficiency can approach 100%. Diodes clamp the output voltage to the supply lines. The drive voltage must be large enough to saturate the transistors.

Further reading

Oxborne, M. R., Design of tuned transistor power amplifiers, *Electronic Engineering*, 1968, pp.436-43.
Stewart, H. E., Engineering Electronics, Allyn & Bacon 1969, pp.589-642.
Birt, D. R., Modulated Pulse Amplifiers, *Wireless World*, 1963, pp.76-83. (Also subsequent articles and letters.)

Cross references

Series 7, cards 4, 5, 9, 10, 11 (class A), 2, 3, 7, 8 (class B), 6 (class C), 12 (class D).

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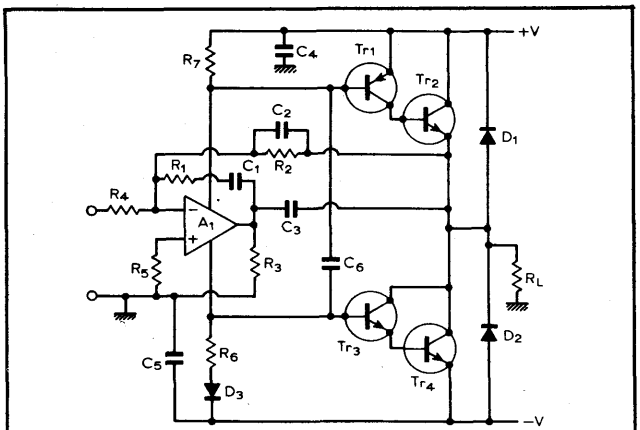
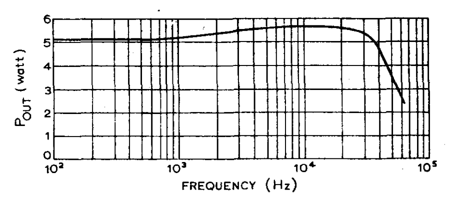
Servo amplifier





Typical performance

Supplies: $\pm$15V, 235mA
Quiescent current: $\pm$1.8mA
A_1: 741
Tr_1: BFR81
Tr_2, Tr_4: TIP3055
Tr_3: BFR41
R_1, R_2: 15kΩ
R_3: 47Ω; R_4, R_5: 4.7kΩ
R_6, R_7: 180Ω
C_1: 2nF; C_2: 100pF
C_3: 4.7nF; C_4, C_5, C_6: 470nF
D_1, D_2: SP2; D_3: 1N914
R_L: 18Ω
Risetime $\approx 30\mu s$ (4.8V pk-pk at 1kHz)
V_{in}: 2.07V r.m.s. without clipping

In servo systems a servoamplifier is needed when a high-power load must be driven from a low-power source. Amplifier A_1 acts as a see-saw amplifier having its gain determined by R_2/R_4 which can be adjusted to accommodate a wide range of input signal levels from a transducer. With no input signal, the output power transistors are virtually cut off, the only drain from the supply being the quiescent current of the operational amplifier (around 2mA). Hence the base-emitter junction of Tr_1 is forward-biased by only about 350mV due to the p.d. across R_7. The base-emitter junctions of Tr_3 and Tr_4 would be forward-biased to a smaller extent unless R_6 was greater than R_7. However, including D_3 and making $R_6 = R_7$ produces the desired bias with D_3 providing some temperature compensation for the base-emitter voltage of Tr_4. The amplifier has a class B push-pull output stage so that a bipolar input signal produces class B currents in its supply leads. These currents are used to provide the base drive to the compound power transistors which supply the load currents to R_1 in push-pull. Transistors Tr_3 and Tr_4 form a Darlington pair while Tr_1 and Tr_2 are its complementary equivalent. The Darlington configuration is used to provide high current gain to ensure that the load current is much larger than the amplifier's quiescent current. To guard against instability, R_1 and C_1 provide feedback around the operational amplifier and R_3 and C_3 provide feedback around the power stage. Bandwidth of the amplifier is controlled by C_2R_2 time constant which can be held fixed when the gain is varied by R_2, if C_2 is also adjusted. Diodes D_1 and D_2 protect the output transistors against breakdown when the load is highly inductive.

Component changes

Useful range of supplies: $\pm$6 to $\pm$18V.
Output power and efficiency fall as supply voltage is reduced: typically P_{out} is 0.8W and efficiency is 65% with $\pm$6V at 1kHz. With maximum drive, P_{out} falls as R_L increases: for supplies of $\pm$15V, typically, P_{out} is 12.6W for $R_L = 6.8\Omega$ and $P_{out} = 3.8$W for $R_L = 25\Omega$. Total harmonic distortion falls as drive increases: typically 0.45% for $V_{in} = 2.8$V and 5.3% for $V_{in} = 150$mV (supplies $\pm$15V, R_L: 18Ω and $f = 1$kHz sinewave).

Circuit modification

- The Tr_1-Tr_2 and Tr_3-Tr_4 Darlington pairs in the output stage may be made single n-p-n and p-n-p transistors. Ideally, these transistors should have high current gains to provide a peak load current that is significantly in excess of the quiescent current in the amplifier. They also need to have a higher power rating and the combination of high power, high current gain and wide bandwidth is not an easy specification to meet at low cost. The use of single BRF81 and BRF41 transistors provides a reasonable compromise.
- A modification which can improve stability while allowing some quiescent current in the output stage, i.e. biasing in class AB, is obtained by including resistors in the equivalent emitters of the drive transistors, increasing the p.d. across R_6 and R_7 and/or placing a diode in series with R_6 and R_7. The resistors in the emitters can be selected to provide the required quiescent current. (See circuit left.)
- In principle, any other feedback configuration may be used; for example taking the input signal to the non-inverting input of the operational amplifier and grounding the input end of R_4 converts the feedback to a series-applied form with the accompanying increase in input impedance. (See circuit right.) The operational amplifier may be supplied with differential input signals if desired.

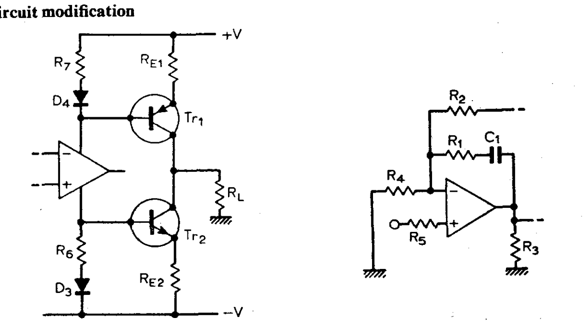
Further reading

Campbell, D. L. & Westlake, R. T., Build a high-current servoamplifier with i.cs, *Control Engineering*, December 1969, pp.91-4.
Garza, P. P., Getting power and gain out of the 741-type op-amps, *Electronics*, 1 Feb., 1973, p.99.

Cross references

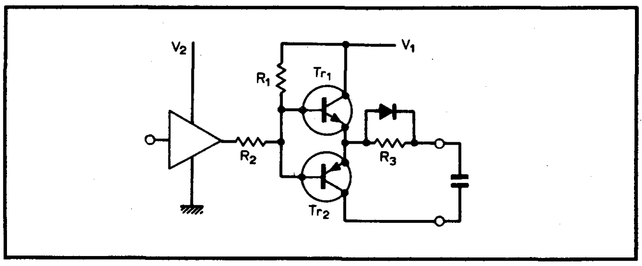
Series 7 cards 1 & 12.
Series 2 card 4.
Series 4 card 8.

Circuit modification



Wireless World Circard

Pulse buffer amplifier

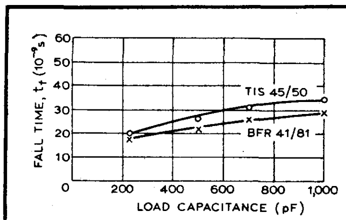


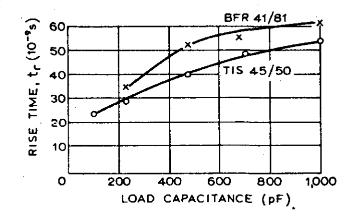
Typical performance

V_1: +14V; V_2: +5V
Tr_1: TIS45; Tr_2: TIS50
IC_1: 1/6 SN7406
R_1: 470; R_2: 100Ω
R_3: 10Ω
D_1: PS101; C_1: 680pF
Input pulse height: 4V
Duration: 600ns
P.R.F.: 50kHz
Rise time: 20ns

Output pulse: rise time 49 ns; fall time 32ns; pulse height: ≈ V_1 (Rise and fall times measured between 10% and 90% levels). Variation of rise and fall time with several capacitive loads shown right

Some small distortion effects on input drive pulse were not apparent on the output pulse.





Circuit description

The complementary symmetry output stage commonly used in class B amplifiers is equally applicable to pulse outputs. The problem here is that using only a single transistor in the output will only allow any capacitive load to have either a fast rise time or a fast fall time, but not both. Or if the output stage is operated in class A, it needs a quiescent current greatly in excess of the charging current required by the capacitor to achieve a high rate of rise and/or fall. The class B push-pull stage shown has Tr_1 driving the capacitor in the positive direction when a positive-going edge is applied at the base connection, while Tr_2 drives the capacitor in the negative direction. Rise and fall times are now determined by the current flow in the capacitor, which on the positive-going edge is limited by the base current that can be supplied by R_1 as D_1 is allowed to conduct. On the negative-going edge, current through R_2 is significantly greater and could cause excessive current flow in Tr_2 but the diode is reverse biased and R_3 takes the place of limiting action previously provided by R_1. It is not possible in a simple circuit of this kind to choose a simple bias network for R_1 and R_2 which would give the same bias drive current in both directions.

IC_1 is a open-collector high-voltage output device which pulls the potential at the bases of Tr_1 and Tr_2 to a low value when in conduction, and when out of conduction allows the bias to rise towards V_1 via R_1.

Component changes

- Transistors Tr_1 and Tr_2 can be replaced by BFR41 & BFR81 or BC125 & BC126 with poorer rise and fall times. Typical comparison

	rise time (ns)	fall time (ns)
TIS45/50	12	12
BC125/126	28	14
BFR41/81	38	15

- For each capacitor value, overshoot on leading and trailing edges of output pulse is approximately 25% of pulse level.
- Resistive load: 100Ω, V_1: 14V, V_2: +5V; output pulse excursion is from 1.6 to 12V.

Pulse width: 6μs. Useful frequency range 3 to 100kHz. Corresponding mean current from supply 1.5 to 30mA d.c.

- IC_1: SN75451A or SN7407 for greater output voltage levels and faster rise times.

Circuit modification

- Rise and fall times for the circuit left are given centre. The lower level of drive pulse from the i.c. is approximately zero and hence pulse rise times will be slightly larger than in the original circuit.
- An alternative arrangement is shown right. If the drive voltage goes positive, the Zener diode transfers current to the base of Tr_2 which brings Tr_2 into conduction, clamping the output to the negative supply rail, with very small saturation effects. Conversely, if the output swings negative Tr_1 conducts and clamps the output to the positive rail, i.e. the peak-to-peak output swing into the load is almost equal to the supply rail values.

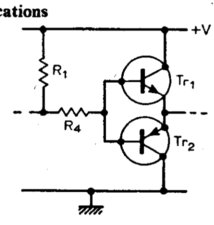
Further reading

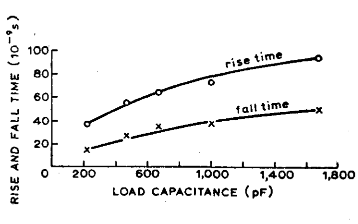
Texas Instruments Technical Seminar 1972, m.o.s. memory drivers.
SGS-Fairchild, Industrial Circuit Handbook, 1967, p.38.
Williams, P., Voltage following, *Wireless World*, vol. 74, 1968, pp.296.

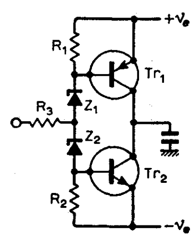
Cross references

Series 6, cards 1, 2 & 8.

Circuit modifications

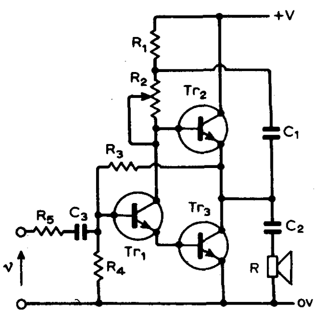






Wireless World Circard

Push-pull class A power amplifier



Tr_2, Tr_3: TIP3055
Tr_1: BFR41
C_2: 2,000μF
C_1: 470μF
C_3: 100μF
R_1, R_5: 220Ω
R_2: 250Ω
R_3: 470Ω
R_4: 120Ω
R: 3Ω
Supply: 12 to 14V

Typical performance

For supply of 13V, quiescent current of 950mA, max. output for 3% distortion is 12V pk-pk into 5Ω (3.6W). Mean current falls to 820mA at max. output. Full power bandwidth: 20Hz to 100kHz. Hum and noise: 80dB below full output. Quiescent current: 1.25A @ 13V. Output power: 5W into 3Ω @ 5% t.h.d. Distortion: <1.1%, 1W into 3Ω, 100Hz to 10kHz. Voltage gain ~ −2. Input impedance ~250Ω.

Circuit description

Class A push-pull amplifiers have at least two active devices in the output stage, and each device should operate under the same quiescent conditions. A drive circuit using one or more devices provides antiphase signals to the output pair which should have matched parameters. Thus a minimum of three transistors is called for and more are commonly required. By using current phase-splitting, a simple circuit results which still gives adequate efficiency and distortion figures. The key feature of the circuit is that the current in R_2 remains constant throughout the a.c. wave form while its d.c. value can be adjusted to set the desired quiescent current. Bootstrapping via C_1 ensures that any increase in the collector potential of Tr_1 is transferred via the emitter follower action of Tr_2 to reappear at the junction of R_1 and R_2. Hence the change in p.d. across R_2 approaches zero except at very low frequencies where the reactance of C_1 becomes significant. As there is no change in R_2 current, any increase in Tr_1 current increases the base current of Tr_3 while reducing the base current of Tr_2 by substantially the same amount. Accurate current phase-splitting together with matched current gains of Tr_2, Tr_3 keep the distortion low. Overall negative feedback via R_3 defines the output quiescent voltage as a multiple of the base voltage of Tr_1 (~1.3V) and the ratio R_3/R_4 scales this base voltage up to half the supply voltage, i.e. the output transistors operate with equal V_{ce} as well as equal I_c.

Component changes

Tr_2, Tr_3: Power transistors with closely matched h_{FE} at operating current. Quiescent power (at least twice max. output) determines types and heat sinks.
2N3055 for $P_o > 5W$. MJE521 for $P_o > 1W$.
BFY50, BFR41, etc., for $P_o < 1W$.
Tr_1: BFY50, BFR41, 2N3053 for most applications.
C_2: Reactance $< R_L$ at lowest freq. Typically 200 to 5000μF.
C_1: Reactance $\ll R_1$ at lowest freq. Typically 100 to 500μF.
R_1, R_2: Set output current $V_s/2(R_1 + R_2) \approx 2I_s/h_{FE}$. One resistor made variable to adjust mean current. Typical range 100Ω to 1kΩ (higher values for low-power circuits).
R_5: Sets voltage gain $\sim -R_3/R_5$ and input resistance $\sim R_5$.
R_3, R_4: Set output voltage (quiescent) to $\sim 2V_{be}[(R_3/R_4) + 1]$. Current in R_3, R_4 to 5 to 20 times base current of Tr_1. Typical values R_4: 100 to 500Ω. R_3: 300Ω to 3kΩ.

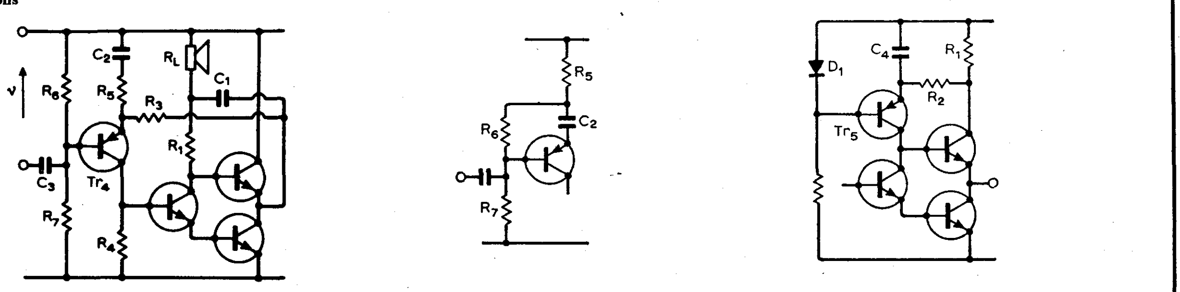
Circuit modification

- Open-loop gain of the original circuit is low and feedback that can be used may not reduce distortion sufficiently. Simple bias circuit leaves the output at a fixed multiple of V_{be} rather than at the supply centre point, i.e. resistors require readjusting for different supply volts. Adding Tr_4 increases open-loop gain, allows 100% d.c. series-applied feedback and has input feedback and load all referred to same supply line. This eliminates bootstrap capacitor provided speaker can tolerate direct quiescent current of driver stage. For output at mid-point of supply $R_6 \approx R_7$. Voltage gain $\approx (R_3/R_5) + 1$. Reactance of $C_2 \ll R_5$ at lowest frequency of interest. Typically R_4, R_5: 1 to 10kΩ, R_6, R_7: 20 to 200kΩ. (circuit left)
- For higher input impedance, input potential divider may be bootstrapped. Interchanging locations of R_5, C_2 allows R_6 to be bootstrapped, almost doubling input impedance. (centre)
- Quiescent current depends on current gains of Tr_2, Tr_3. By monitoring circuit mean current and using result to control drive current Tr_1, mean current can be made constant, e.g. for Tr_5 a germanium transistor, D_1 a silicon diode, mean p.d. across R_1 is controlled at 0.4V. (right)

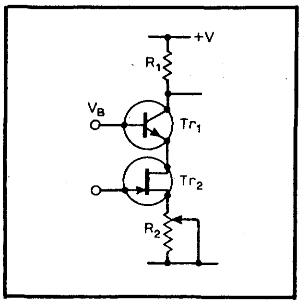
Further reading

Linsley Hood, J. L., Simple class A amplifier, *Wireless World*, vol. 75, 1969, pp.148-53.
Markus, J. (ed.), Improving signal transfer in Electronics Circuits Manual, 1971, p.19.
Allison, W., Self-biasing class A power amplifier, *Wireless World*, vol. 78, 1972, p.577.

Circuit modifications



High-voltage amplifier



Typical performance
Supply: +100V
Tr_1: MJE340
Tr_2: 2N3810
R_1: 10kΩ; R_2: 456Ω
Quiescent voltage: 52V
V_B: 10V
Input signal: 7V pk-pk
Output voltage: 84V pk-pk
Gain constant up to 20kHz
Variation of output with R_2 not decoupled shown opposite.
Effective output impedance of transistor configuration 5MΩ.

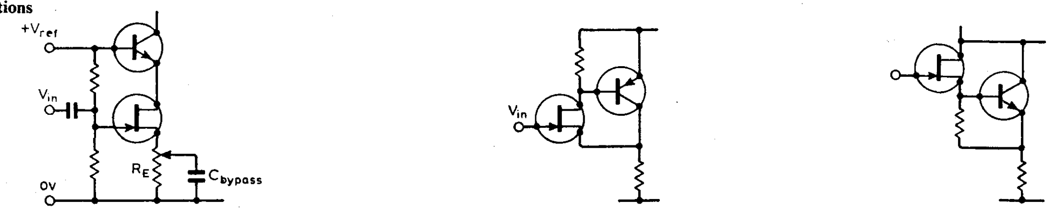
Circuit description
The characteristics required by an amplifier may include high voltage gain and in some applications the ability to withstand high output voltages simultaneously. Such a combination is not available within a single device, but the circuit shown arranges that the necessary input impedance gain characteristics are obtained by Tr_2 and the high voltage characteristics by Tr_1. The input characteristics aimed at were that the device should behave with a defined gain, so that the whole system could be considered equivalent to a valve. Transistor Tr_2 is thus a field-effect transistor whose gain is controlled by the quiescent current, which may be set by R_2. The drain of Tr_2 feeds into the emitter of Tr_1 whose base is maintained at a constant potential, just high enough to ensure that Tr_2 has a quiescent voltage that is above its pinch-off value. The bias voltage should be obtained from a low impedance circuit. Hence Tr_2 is operating into a low impedance, while Tr_1 is virtually a common-base stage and has thus the highest voltage rating that it could possibly have. The current at the collector of Tr_1 is essentially the same as the emitter current as the current gain from emitter to collector is nearly unity. There is no significant Miller/Blumlein effect between the collector of Tr_1 and the gate of Tr_2 as the voltage swing at the collector is isolated from the gate of Tr_2. The capacitance between Tr_1 collector and base is now effectively a capacitance to ground rather than to the input of the amplifier. However this capacitance still affects the output characteristics, as it is in parallel with R_1 for a.c. and determines the bandwidth of the amplifier. The problem is more severe than in many low-voltage amplifiers because R_1 will have a much higher value for a given quiescent current because the p.d. across it may be in excess of 100V. This is the usual penalty to be paid for a high-voltage gain, i.e. the associated high load impedance will have a longer time constant for a given capacitance. The voltage rating is close to the V_{CE} breakdown of Tr_1.

Component changes
- Decouple R_2 with 150μF capacitance to retain g_m of the combined transistors. Output 82V pk-pk for an input signal of 2.4V pk-pk. Low frequency cut-off then 5Hz.
- Range of V_B 8 to 11V – value not critical, with no significant effect on performance.
- Supply may be increased up to 300V with appropriate changes in R_1 and R_2 to control quiescent voltage. Typically (*i*) supply: +200V, V_Q: 110V, R_2: 122Ω, V_{in}: 7V pk-pk; V_{out}: 180V pk-pk; R_1: 10kΩ. (*ii*) supply: +300V, V_Q: 150V, R_2: 1.5kΩ, R_1: 68kΩ, V_{out}: 275V pk-pk.
- Increase of +V from 100 to 200V, maintaining circuit resistors constant reduces h.f. cut-off by approximately 20% indicating that this is dependent more on external components rather than operating conditions.

Circuit modifications
- F.E.T. g_m is controllable by varying negative feedback. A wide range of control is possible with the circuit shown left. Because the gate is positive, R_E can be large for chosen quiescent value of drain current, the feedback being varied via C without then altering the d.c. state of the circuit.
- F.E.T. g_m can be boosted by adding a p-n-p bipolar transistor to achieve a complementary pair (centre), or an n-p-n transistor for a Darlington pair (right), as the output impedance may be considered to be approximately $1/g_m$. The effective output impedance is less than that of the f.e.t alone.

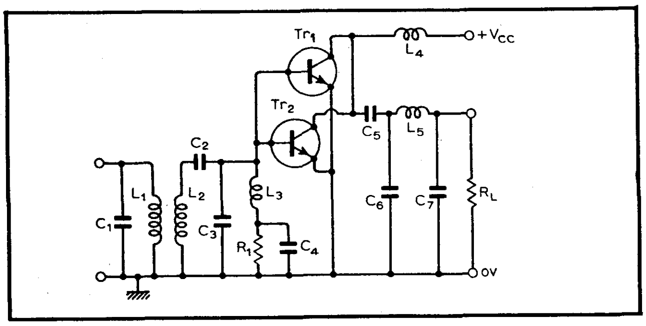
Further reading
Designers casebook. *Electronics*, 1 Feb., 1973, p.99.
Greiter, O., Transistor amplifier output stages, *Wireless World*, vol. 69 1963, pp.310-3.
High voltages switched with a single transistor, 400 Ideas for Design, Vol. 2, 1971, Hayden.

Circuit modifications



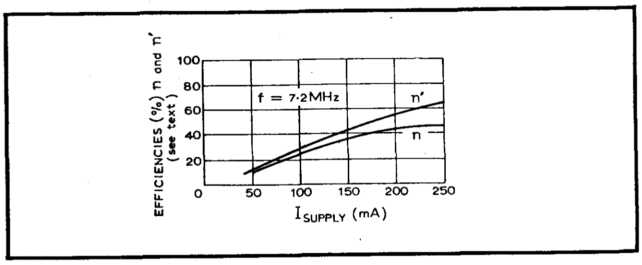
Wireless World Circard

Class C power amplifier



Typical data

Supply: 12V
Tr_1, Tr_2: BFR41
R_1: 100Ω; R_2: 50Ω (carbon)
C_1: 180pF; C_2: 360pF
C_3: 47pF; C_4: 10nF
C_5: 500pF; C_6: 190pF
C_7: 805pF; L_1: 2.7μH
L_2: 2.16μH; L_3: 2.38mH
L_4: 230μH; L_5: 1.51μH



Circuit description

Many class C amplifiers find application in the v.h.f. and u.h.f. bands, special transistor fabrication techniques being used to optimize their performance. For correct design it is necessary to establish a suitable model for the transistor behaviour under class C conditions, some manufacturers providing the appropriate data. In general, this data is not available for class C designs operating at frequencies lower than about 10MHz, so that a successful circuit normally results from a breadboard version using variable capacitors. The circuit shown above was produced on this basis where C_2, C_3, C_5, C_6, C_7 were originally fixed capacitors 'padded' out with variables. Source and load resistance were 50Ω and the output power obtained at 7.2MHz was 1.41W with a drive signal producing 250mA supply current. Overall efficiency was only 47% (see graphs) but taking account of the d.c. drop (3.52V) across the r.f. choke L_4 efficiency rises to 66.5%. Hence L_4 should have low resistance, but its effect is less noticeable at lower currents. Transistors Tr_1 and Tr_2 were general-purpose transistors connected in parallel to reduce dissipation problems. The tuned networks in the input and output circuits should match the source to the transistors and the transistors to the load for maximum power transfer. Careful layout is essential and the circuit can easily oscillate as L_3, L_4 and the collector-base capacitance of the transistors form the basic arrangement of a Hartley-type oscillator.

Component changes

The circuit can operate over a limited frequency range and a wide range of supply voltages and power levels provided the input and output networks are re-adjusted to cater for the changing values of transistor input and output resistance and capacitance.

Alternative general-purpose transistors can be used, such as BFY50.
Single transistor can be used when reduced power is acceptable.
Input transformer can be dispensed with if alternative input and output networks used (see over).

Circuit modifications

Correct design procedures for class C r.f. power amplifiers tend to be highly analytical due to the need to consider the correct choice of input and output coupling networks, their working Q-factors, degree of harmonic rejection, possible causes of spurious oscillation and the d.c. operating conditions. For a successful design the impedances at the transistor input and output terminals must be known under the desired operating conditions. Use of small-signal parameters leads to considerable errors in a class C design as the voltage and current swings are so large in such a power amplifier. When class C transistor data is available it is normally provided in the form of equivalent parallel input resistance and reactance and parallel output capacitance as a function of frequency and power output. The equivalent parallel output resistance is given approximately by $R = V^2_{cc}/2.P_{out}$. Even with this data available a choice must be made from the large number of possible input and output coupling networks. Often a T-configuration is suitable for both networks as shown left. These networks complex-conjugate match the source to the transistor and the transistor to the loads. Both networks introduce losses due to component imperfections. Choice of the working Q-factors is a compromise between losses in the coupling networks, their selectivity and realizable component values. If the loaded-Q is high the capacitors will be small, the selectivity will be high but the losses will be large. A low working Q-factor implies the opposite. When the available data is correctly interpreted it will normally still be necessary to tune the amplifier for optimum performance, for example by adjustment of C_1 to C_4. Complete design procedures are given in the first three references.

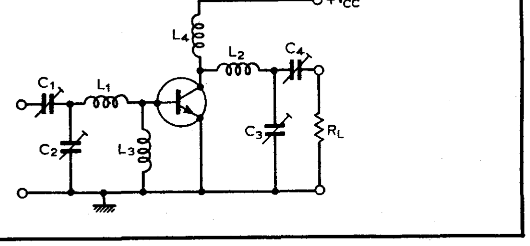
Further reading

Motorola, application note AN-282: Systemizing r.f. power amplifier design, 1967.
Hilbers, A. H., On the input and load impedance and gain of r.f. power transistors, *Electronics Applications*, vol. 27, 1967, pp.53-60.
Mulder, J., On the design of transistor r.f. power amplifiers, *Electronic Applications*, vol. 27, No. 4, pp. 1967, 155-171.
Markus, J. (ed.), 8-MHz, 3-W amplifier, in Electronic Circuits Manual, 1971, p.15.

Cross references

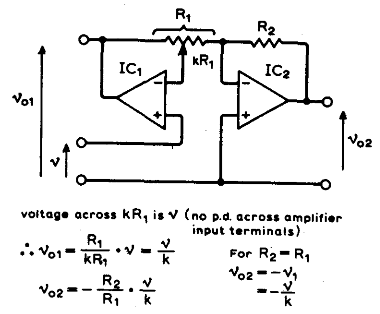
Circard series 7, card 1.

Circuit modifications



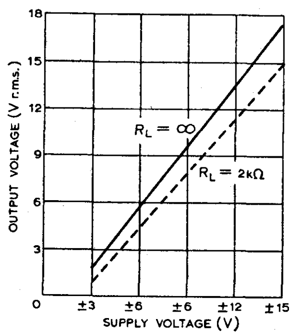
Wireless World Circard

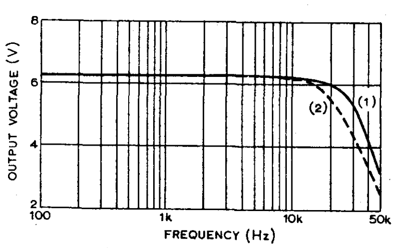
Bridge output amplifiers



Typical data

IC_1, IC_2: 741
R_1: 10kΩ pot
R_2: 10kΩ
Supplies: ±15V
R_L: 2kΩ
Output voltage: 15V r.m.s. into 2kΩ (17.5V r.m.s. o/c) for $k = 0.1$ to 1.0 at 1kHz.





(1) Max output including increased drive at [illegible].
(2) Max output for distortion <5%.

Circuit description

Most power amplifiers have a single-ended output, delivering to the load a voltage whose peak-to-peak value is at most equal to the total supply voltage. If transformers/inductors are allowed such single-ended stages may produce peak-to-peak output voltage swings of up to double the supply voltage, but only if the transistor breakdown voltages are equally high. The economic and performance limitations imposed by transformers point to the need for an alternative output configuration for increased output voltage swing. If the load is taken between the outputs of two amplifiers delivering inverted outputs of equal magnitude, then the load voltage being the difference between the two has twice the magnitude of each separately. The method is illustrated using standard operational amplifiers, but is applicable to amplifiers at all power levels, where the constraint of a grounded load need not be met. This particular configuration offers the advantage that a single potentiometer controls the gain of both channels. The exact balance is adjusted if required by setting $R_2 = R_1$. Equal magnitudes of output are ensured for this condition assuming ideal amplifiers because the two resistors carry equal current while their junction is a virtual earth point. A further advantage of this circuit is the high input impedance. As only one amplifier has a common-mode signal, the amplitude response differs somewhat, but the difference is only significant at those frequencies where the characteristic of each amplifier has departed significantly from the ideal. Slew-rate limiting, an output circuit phenomenon, determines the highest frequency at which large output voltages are obtainable with low distortion.

Component changes

- Replace amplifiers by any compensated type (307, etc.); alternatively use uncompensated types (748, 301, etc.) with appropriate compensation capacitor (reduced compensation possible with increased gain leading to higher slew rate).
- Resistor values non-critical but $R_1 = R_2$ gives push-pull output (circuit usable as phase-splitter for succeeding stages). Resistor R_2 may be made adjustable to take up tolerances if outputs are required to be given ratio, leaving tapping point on potentiometer to vary total gain. Typical values for R_1, R_2; 1kΩ to 250kΩ. Higher values lead to offset, drift and additional h.f. limitations; lower values absorb too much of the available output current.
- If unity gain is sufficient, IC_1 may be replaced by voltage follower, R_1 replaced by fixed resistor.

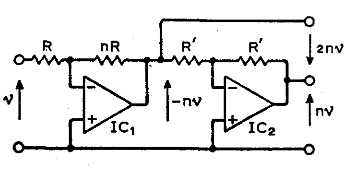
Circuit modifications

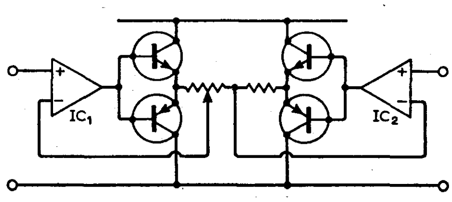
- Using two separate inverting amplifiers, with second set for a gain of −1, control over both outputs is obtained by varying the gain of the first. As both are used as virtual-earth stages feed-forward compensation may be used to obtain stable performance with considerable increase in slew-rate and cut-off frequency.
- Current capability of the output stages can be increased by any of the ways suggested on the cards describing class B/class A amplifiers. The simplest addition is a pair of complementary emitter-follower combinations. Output current capability may be increased by one or two orders of magnitude, but the output voltage swing is slightly reduced because of the base-emitter p.d. of the transistors. Crossover distortion may be minimized by the addition of diode/transistor biasing networks to the transistor base circuits. (middle)
- An alternative to the bridge circuit for increased voltage swing is the principle of supply bootstrapping of which this is one version. (right)

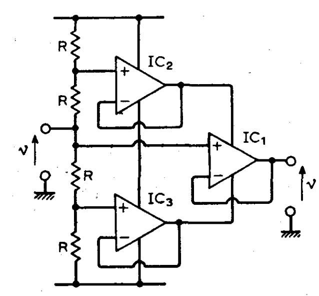
Further reading

Greiter, O., Transistor amplifier output stages, part 1, bridge circuits, *Wireless World*, vol. 69, 1963, pp.17-20.
Del Corso, D. & Giordana, M., Simple circuit to double the output-voltage swing of an operational amplifier with increased slew rate, *Electronics Letters*, vol. 8, pp.151/2.

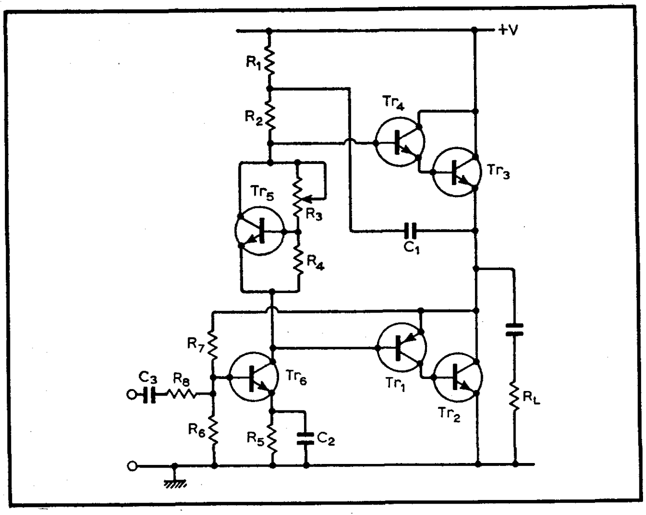
Circuit modifications





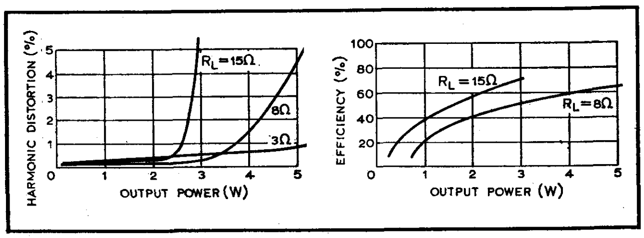


Class B quasi-complementary output



Typical performance

Supply: +20V
Tr_1: BFR81; Tr_2, Tr_3: TIP3055
Tr_4, Tr_5, Tr_6: BFR41
R_1, R_2: 1.5kΩ; R_3: 1kΩ
R_4: 470Ω; R_5: 330Ω
R_6: 1.8kΩ; R_7: 8.2kΩ
R_8: 1kΩ; R_L: 8Ω
C_1: 100μF; C_2: 22μF;
C_3: 10μF
Main d.c. output: 10V
Input signal: 2.6V pk-pk
Output signal: 6.7V pk-pk
Output power: 5.4 watts
Harmonic distortion: 5.8%
Quiescent current: 0.41A
Graphs of harmonic distortion and efficiency versus output power for loads of 15Ω and 8Ω shown below



Circuit description

This is a circuit of a class B push-pull amplifier in which transistors Tr_3 and Tr_4 complement the pair Tr_2 and Tr_1. To use n-p-n transistors in the output stage for economy, the configurations of the two sections are different, i.e. Tr_3 and Tr_4 are connected as a Darlington pair and Tr_1 and Tr_2 as a complementary pair. They receive essentially the same a.c. drive, but with the bases separated by Tr_5. Tr_3 and Tr_4 conduct for positive-going output signals and Tr_6 supplies base current drive to Tr_1 and Tr_2 for negative-going output signals. Transistor Tr_5 is used in the so-called amplified diode configuration in which the potential difference between the bases of Tr_1 and Tr_4 is set as a multiple of the V_{be} of Tr_5 by the potential divider R_3, R_4, i.e. R_3 can be adjusted to give the desired quiescent current in transistors Tr_2 and Tr_3. A forward bias is available which may allow the transistors to conduct to a small extent, just sufficient to minimize the crossover distortion that can never be entirely absent. Transistor Tr_6 is an inverting amplifier with overall negative feedback through R_7, the values of R_6 and R_7 determining the d.c. output potential in conjunction with R_5. Because R_5 is decoupled, the a.c. properties of the arrangement are determined by the ratio of R_7 to the source resistance. Resistors R_1 and R_2 are centre-tapped and this point is taken to the output via C_1, which bootstraps R_2 so that the current through it remains constant throughout the cycle of output voltage swing.

Circuit modifications

- To avoid dangerous overcurrent in either of the output stage transistors, the current may be limited by adding series resistors R_e between the emitters and the output terminal (left).
- Middle circuit shows an alternative arrangement, adding transistors Tr_7 and Tr_8. These are normally non-conducting except under overload conditions, i.e. as the output current increases the voltage drop across R_{e1} or R_{e2} causes Tr_7 or Tr_8 to turn on and divert the base current available to Tr_4 or Tr_1, limiting the output current to V_{be}/R_e.
- Alternative configurations for the output stages are shown right (*i*) requires low and high power n-p-n and p-n-p transistors to make up the Darlington pairs, the minimum p.d. between input and output circuits being twice the V_{be} of a single transistor, (*ii*) uses complementary equivalent pairs with only one base – emitter path between input and output. Each pair comprises two inverting stages with 100% series - applied negative feedback giving unit gain.

Component changes

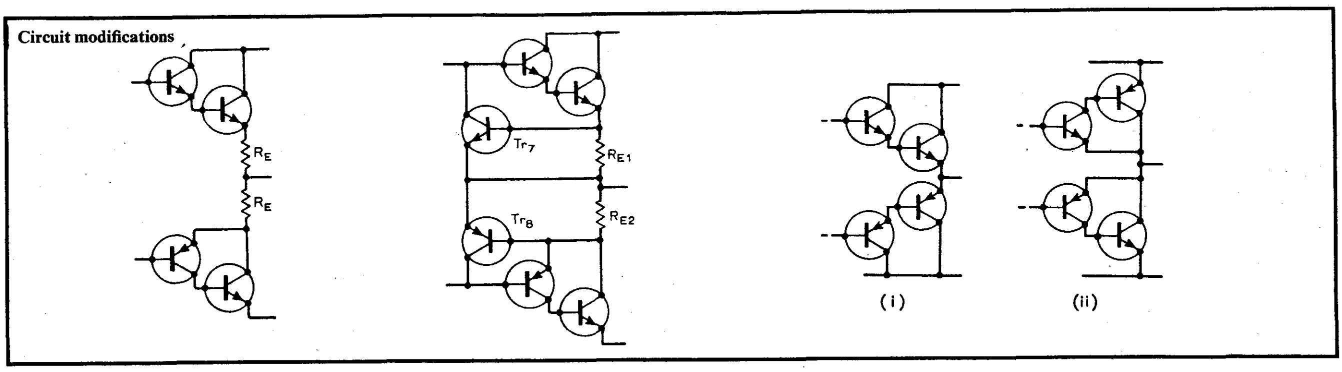
Adjustment of R_3 to avoid just visible crossover distortion gives a quiescent current of 7mA.

Further reading

New uses for the LM100 regulator, National Semiconductor application note AN8-7.
Grebene, B., Analog integrated circuit design, Van Nostrand 1972, pp.163-7.
Amplifier efficiency (Letters), *Wireless World*, vol. 75, 1969, p.381.
Hartz, R. S. & Kamp, F. S., Power output and dissipation in class B transistor amplifiers, RCA publication AN-3576. (Also in publication SSD-204A, p.594.)

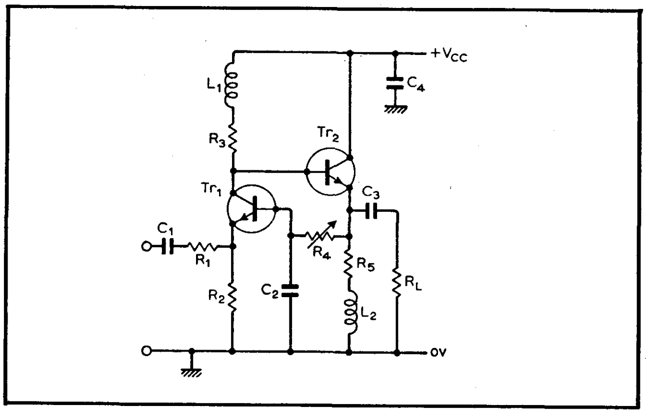
Cross references

Series 7, cards 1, 2 & 3.



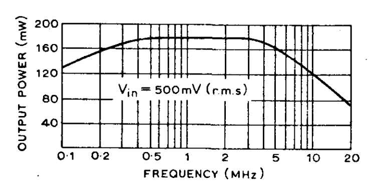
Wireless World Circard

Broadband amplifier



Typical performance

Supply: +20V, 118mA
Tr_1: BFR41; Tr_2: BFY50
R_1: 33Ω; R_2: 150Ω
R_3: 220Ω; R_4: 22kΩ
R_5: 120Ω; R_L: 50Ω
(carbon)
L_1: 1.7μH; L_2: 220μH
Power gain ≈ 14dB
3dB bandwidth 64kHz to 16mHz



Circuit description

In many applications the transfer of power to a load at maximum efficiency is not the primary consideration. Often, power gain is required for small input signals over a wide frequency range without introducing significant intermodulation and harmonic distortion. The common-base stage offers the best linearity of voltage gain against collector current, the latter changing in sympathy with this input signal. The emitter follower, while not providing voltage gain, gives a current gain of the same order as a common-emitter stage and is therefore very useful for transferring power to a load. To obtain this transfer with little distortion, it is necessary to operate the emitter follower at a relatively high quiescent current even for quite small input signals. The circuit uses a common-base stage feeding the load via an emitter follower. To maximise the gain-bandwidth product, Tr_1 and Tr_2 operate in regions where their current gain is much smaller have relatively high quiescent currents and thus have lower-than-normal current gains. The input resistance of the common base stage is inverse to its quiescent current, so that a high current allows the amplifiers input resistance to be matched to that of the source by a suitable choice of R_1. Resistor R_3 is determined from the required voltage gain (A_V) for equal source and load resistances $R_3 \approx A_V h_{fe2} R_L/(A_V + h_{fe2})$. Inductor L_1 is included to offset the capacitive loading due to Tr_2 and strays to maintain the gain at high frequencies. To deliver as much output current to R_L as possible at high frequencies choke L_2 is included in series with R_5.

Component changes

With V_{cc} (min) = +5V, V_{in} (max) ≈ 140mV r.m.s., supply current is 30mA, and P_{out} ≈ 11mW.
Tr_1 and Tr_2 can both be BFY50 or BFR41.
Tr_1 can carry a much smaller quiescent current, using for example an ME4103, with increased values of R_2, R_3 and R_4.
R_1 can be increased or decreased to allow matching to source resistances greater or less than 50Ω respectively.

Circuit modifications

If the input signals are very small, output powers of around half a watt can still be obtained over a wide bandwidth by cascading a pair of amplifiers of the type described. When the gain-bandwidth product of the amplifier is not the most critical requirement and a higher efficiency is needed, the quiescent current in Tr_1 may be drastically reduced. Resistors R_2 and R_3 would then need to be increased, with a corresponding increase in R_4, if this is to be the means of controlling the quiescent operating conditions. The lower Tr_1 current may be chosen to make the natural input resistance of the stage, in the absence of R_1, the value required to match the source.

- Input resistance may be defined using shunt-applied feedback, as shown left, where the emitter of Tr_1 is d.c. or a.c. grounded, the feedback is not decoupled and the voltage gain is determined by the ratio R_A/R_B. The input resistance is largely that of R_A except at high frequencies where the feedback falls and the impedance at Tr_1 base must be considered.
- Inclusion of R_6, as shown right, may be applied to both the previous circuits to allow an output to be taken from the collector of Tr_2. To maximize the signal swing in the collector circuit of Tr_2 the bias network must be readjusted to leave a small voltage at Tr_2 emitter, say by reducing R_4 and R_2 in the original circuit. The output resistance is approximately R_6; this stage is therefore convenient for feeding directly into any other low impedance stage, such as that left, with R_A removed. This mismatch can often be of advantage in extending the bandwidth of the amplifier.

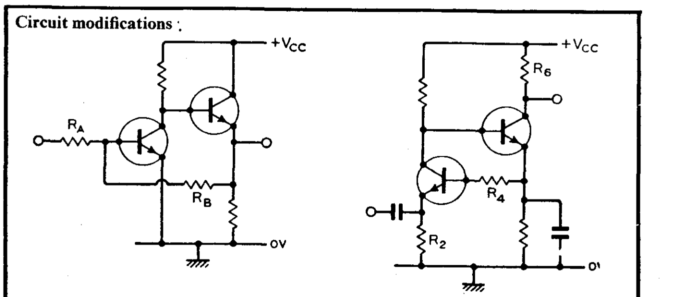
Further reading

Hirst, R., Wideband linear amplifier, *Wireless World*, vol. 75, 1969, pp.168-70.
Meindl, J. D. & Hudson, P. H., Low-power linear circuits, *IEEE Journal of Solid-State Circuits*, vol. 1, 1966, pp.100-11.
Lo, A. W. (and others), Transistor for Electronics, Chapter 9, Prentice-Hall, 1955.
Griffiths, H. N., Simple wideband amplifier, *Wireless World*, vol. 75, 1969, p.478.

Cross references

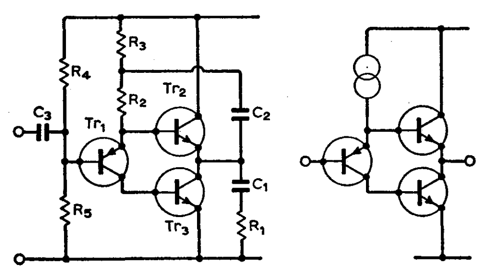
Series 7, cards 1, 4, 5 & 10.

Circuit modifications:



Wireless World Circard

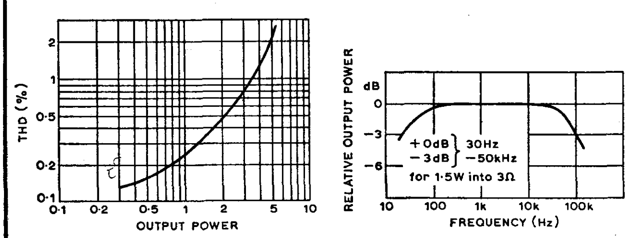
Class A op-amp power booster



Typical performance

Tr$_1$: BFR81
Tr$_2$, Tr$_3$: TIP3055
R$_1$: 3Ω; R$_2$: 250Ω pot.
R$_3$: 220Ω; R$_4$, R$_5$: 10kΩ
C$_1$: 2,000μF; C$_2$: 470μF
C$_3$: 10μF
Supply voltage: 12V
Quiescent current:
1.25A (set by R$_1$)

Output power for 1% t.h.d. into 3Ω load: 4.2W (supply current falls to 1.05A at full output). Output voltage swing to within about 0.7V of supply lines for 3Ω load and about 0.15V for 15Ω load.



Circuit description

Available operational amplifiers have limited output currents, but may have a voltage swing approaching supply values. The circuit shown is class A buffer amplifier of unity voltage gain which may be added to such amplifiers to increase their output current to 1A or more. In addition the circuit is a very simple version of the voltage follower, having a low d.c. offset between input and output, a voltage gain very close to unity and a high input impedance. With the bootstrap technique applied the amplifier is capable of driving low load resistances to within <1V of each supply line.

If a constant current flows in R$_2$, then as the base potential of Tr$_1$ increases the emitter current of Tr$_1$ decreases and with it the collector current.
This fall is fed to the base of Tr$_3$ causing it to conduct less, while the fall in emitter current releases more of the constant current in R$_2$ to flow in the base of Tr$_3$. Provided the current gain of Tr$_1$ is reasonably high, the magnitudes of the base current charges in Tr$_2$, Tr$_3$ are equal but the signs are opposite. This represents an approach to ideal current phase-splitting. The constant current in R$_2$ is provided by the bootstrap capacitor C$_2$, such that any change in the potential at the base of Tr$_2$ is coupled via the follower action to the positive end of R$_2$, i.e. with no resulting change of p.d. across R$_2$ in the ideal case. Resistors R$_2$ or R$_3$ require to be variable to set the output current and stability of that current then depends on h_{FE} variation in Tr$_2$, Tr$_3$. The base-emitter p.ds of Tr$_1$, Tr$_2$ substantially cancel, as they can readily be chosen for junction area ratios matching the quiescent current ratios. As a class A amplifier, maximum theoretical efficiency is 50%. At full output the load power may approach 40% of supply power in practice, but the quiescent power is somewhat higher than the supply power at full load.

Circuit modifications

- The good d.c. offset characteristics allow the amplifier to be used as a voltage follower with d.c. coupling to the load. Bootstrapping should be retained unless the amplitude response is required to extend to d.c., as it swings the junction of R$_2$, R$_3$ above the supply on positive signal swings. Hence it can drive Tr$_2$ base far enough positive to saturate Tr$_2$ hard making maximum use of available supply voltage. If the load is to be a.c. coupled but may carry a small quiescent current, the load resistance R$_1$ may replace R$_3$. (left)
- Any other constant-current circuit may replace the bootstrap arrangement, e.g. a f.e.t. either with gate strapped to source as shown or with a resistor in the source lead to define some lower value of current. (middle)
- Although the distortion of the buffer stage above is low, the addition of a high voltage gain amplifier such as an op-amp can increase the voltage gain to $(R_B/R_A)+1$ while providing sufficient overall feedback to make distortion very low. The wide bandwidth of the buffer stage together with its unity gain minimizes the risk of instability at high frequencies. Should this be troublesome an op-amp with external compensation may be used with increased compensation capacitor. (right)

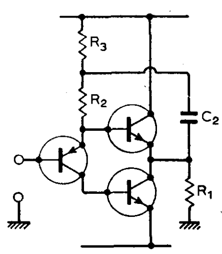
Further reading

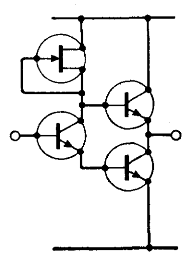
Belcher, D. K., Inexpensive circuit boosts op-amp output current, 400 Ideas for Design, vol. 2, Hayden, pp.1-2.
Bloodworth, G. G., D.C. amplifier with unity voltage gain, *Electronic Engineering*, 1965, pp.112-4.
Electronic Circuit Design Handbook, Current boosters for i.c. op-amps, Tab, 1971, p.161.

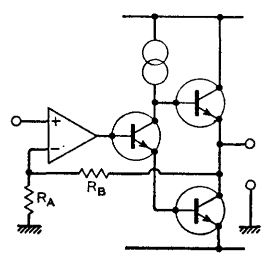
Cross references

Series 7, cards 2, 4 & 8.

Circuit modifications

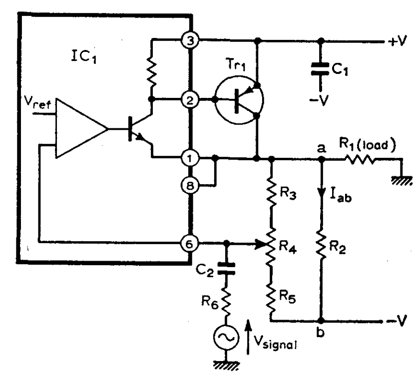






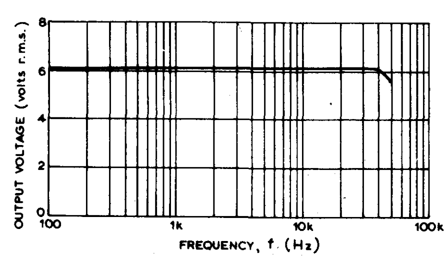
Wireless World Circard

D.C. power amplifier



Typical performance
IC_1: LM305 or LM100
Tr_1: MJE271
Supplies: ±15V
R_1, R_2: 150Ω; R_3: 15kΩ
R_4: 1kΩ; R_5: 1.8kΩ
R_6: 10kΩ
C_1: 1μF (tantalum);
C_2: 10μF

Input signal: 2.8V r.m.s. at 100Hz
Maximum output voltage before symmetrical clipping 6.2V r.m.s.
Output power: 250mW
Harmonic distortion: 0.35% at 1kHz, and 0.32% at 20kHz.



Circuit description
This circuit uses a voltage regulator i.c. package to supply an output stage Tr_1 where the amplifier is to be used primarily with a unipolar signal, though it can also be interpreted as a class A output stage which can be a.c. coupled to a load. The i.c. regulator contains its own reference voltage and a separate feedback point (terminal 6) which allows the potential at the collector of Tr_1 to be set to some stable value which is a multiple of the internal reference voltage, that multiple being set by R_3, R_4 and R_5 the quiescent current in Tr_1 is then set by the bias resistor R_2 in conjunction with this predetermined voltage.

An a.c. signal can be superimposed at pin 6 via C_2 and R_6 the circuit then behaving as a see-saw amplifier, as the reference voltage leaves the feedback terminal 6 as an a.c. virtual earth. For d.c. purposes, the circuit may be treated as series applied feedback. The peak current in the load is limited to a fraction of the quiescent current for negative excursions; as the voltage goes negative the p.d. across R_2 falls and with it the current through R_2. The current in R_1 for this voltage exclusion can never exceed R_2 even when the transistor current falls to zero in the positive direction; however, much greater currents can be provided through Tr_1. The amplifier is thus an inefficient class A amplifier whose effectiveness can be improved by replacing R_2 by a constant-current stage which can sustain a given peak current in R_1 almost equal to the quiescent value, even for large voltage excursions in the negative direction. Capacitor C_1 is used to suppress h.f. oscillation and a low inductance type must be used.

Onset of slew-rate limitation occurs at 70kHz for an output signal level of 16V pk-pk when the signal level is reduced to 3 to 5V pk-pk by reducing the input signal. Voltage gain is flat up to 100kHz, with 3dB fall-off occurring about 250kHz.

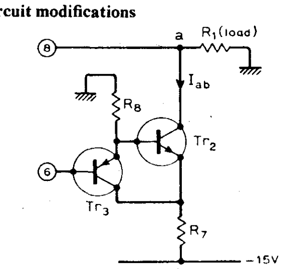
Circuit modifications
- Resistor R_2 is replaced by the Baxandall constant-current circuit shown left Tr_1: BFR81, Tr_2: TIP3055, R_7: 18Ω, R_8: 3.9kΩ. This permits a much greater input signal level before peak clipping occurs. Resistor R_7 is chosen for approximately a 100mA constant quiescent current in the path a-b (about 1.8V is available at terminal 6). If the transistor Tr_1 output current is 200mA pk-pk, then the output current swing in load R_1 is twice that for the case when R_2 is 150Ω with the same quiescent current. A comparison of instantaneous currents for the two possible circuits between a and b is tabled below.

- The regulator may be replaced by the operational amplifier emitter-follower circuit, shown right. To maintain the d.c. stability of the output, the non-inverting terminal must be connected to a suitable stable reference voltage. If the d.c. power supply is stabilized, then this may be a tapping on a potential divider connected across the supply. For minimum drift, the effective resistance seen at both input terminals of the op-amp should be comparable.

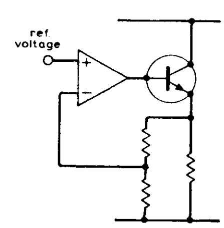
Further reading
New uses for the LM100 regulator, National Semiconductor application note AN-8, 1968.
Amplifier efficiency (Letter), *Wireless World*, vol. 75, 1969, p.535.

Cross references
Series 3, card 8.
Series 7, card 12.

Circuit modifications

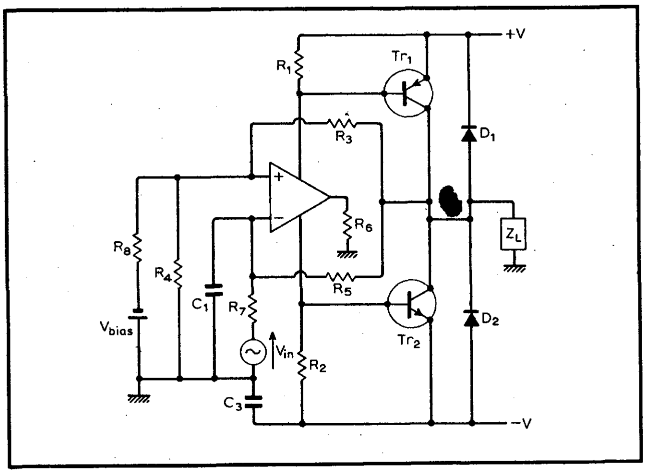


current through Tr_1 (mA)	I_{ab} (mA)	load current (mA)	circuit element in a–b
100 200 0	100 150 50	0 +50 −50	R_2
100 200 0	100 100 100	0 +100 −100	constant current source



Wireless World Circard

Class D switching amplifier



Typical performance

A_1: 301
Tr_1: 301
Tr_2: BFR41
D_1, D_2: SD2

Supply: ±10V (4 to 15V)
R_1, R_2: 180Ω, R_3: 10kΩ
R_4: 100Ω, R_5: 4.7kΩ
R_6: 470Ω; R_7, R_8: 1kΩ
C_1, C_2, C_3: 100nF
Z_L: 1mH ($r = 0.9\Omega$) +15Ω

V_{bias}: −640mV to set mean load voltage to zero with $V_{in} = 0$.

Switching frequency: 27.8kHz, max 40 kHz. With $V_{in} = 0$, supply current is ±20mA; with $V_{in} = 3.4$V pk-pk 100Hz; current is ≈130mA; power in 15-Ω load ≈1.66W; residual "carrier" ≈300mV across 15Ω; overall efficiency 64%; output stage efficiency ≈76%; 3-dB bandwidth ≈600Hz. With rectangular input at 100Hz, output rise and fall times ≈600μs.

Circuit description

Basically, the circuit is an astable oscillator, generating a squarewave that is used to drive a complementary pair of output transistors into conduction on alternate half-cycles of the squarewave. The output transistors thus switch the voltage to the load at a frequency that is much higher than that of the signals to be amplified. The squarewave generator is designed around the operational amplifier A_1 which uses positive feedback via R_3 and R_4. The periodic time of the squarewave fed to R_6 depends on the time constant R_5C_1 if R_7 is much greater than R_5. To obtain a realistic switching frequency with reasonable components and also to obviate the need for large input signals a compromise must be made in the value of R_7. Current in R_6 flows alternately in R_1 and R_2 producing p.ds across these resistors that are sufficient to switch on Tr_1 and Tr_2 respectively. The signal applied to R_7 causes the mark-to-space ratio of the output waveform from the astable to vary in sympathy with the instantaneous value of V_{in}, so that the mean value of the voltage applied to the load also varies directly with the input signal. If the load impedance has an external filter, or is by its nature self-filtering such as with a motor, then the power drawn from the amplifier at the switching frequency is low and the useful signal power in the load will be high.

If the mark-to-space ratio of the squarewave generated by the astable is not unity with $V_{in} = 0$, it can be made so by a suitable choice of the bias supply and R_8. Diodes D_1 and D_2 protect Tr_1 and Tr_2 against breakdown when the load impedance is highly inductive.

Circuit modifications

- The bias source to set the mark-to-space ratio of the squarewave to zero can be obtained by a potentiometer connected between ground and the appropriate supply line.
- While an inductor is normally used in series with a resistive load to filter out the h.f. squarewave, any suitable low-pass filter can in principle be connected between the junction of Tr_1 and Tr_2 collectors and R_1. Another possible method is to connect a capacitor in parallel with the inductive smoothing choke so that it is resonant at the switching frequency. For example, with a choke of 1mH and $r = 1\Omega$, a parallel capacitance of 16nF would be resonant at switching frequency of 40kHz. At signal frequencies less than about 500Hz, the impedance of this tuned network is inductive, having a maximum impedance of about 3Ω.
- The complementary pair of transistors forming the output stage can be replaced by a bridge-type network as shown left. The four transistors are fed with complementary pulse-width-modulated squarewaves which cause the transistors to be switched on and off in pairs. With Tr_1 and Tr_4 on current flows in the load in one direction and is reversed when Tr_2 and Tr_3 are switched on.
- Another practical form of bridge output stage is shown right using a pair of voltage comparators to generate the complementary pulse-width-modulated switching waveforms. The bridge of power transistors is connected across a single-ended supply. Component details are given in the first reference.

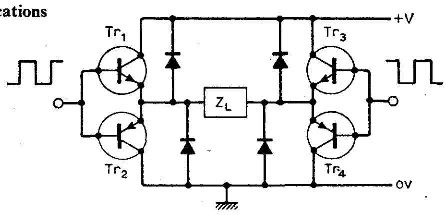
Further reading

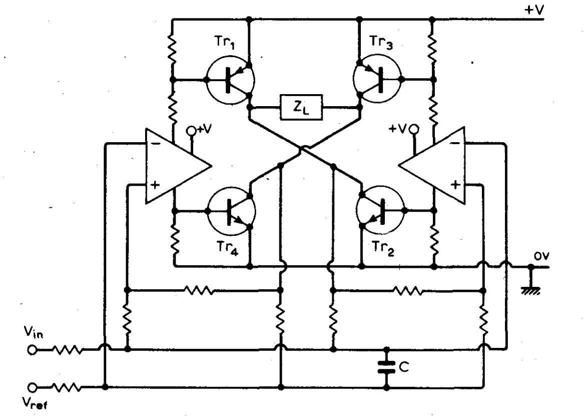
National Semiconductor, data sheet and application notes on the LM311 voltage comparator, 1970.
Camenzind, H. R., Modulated pulse audio and servo power amplifiers, International Solid-State Circuits Conference, University of Pennsylvania, Philadelphia, 1966, pp.90/1.
Meidl, J. D., Micropower circuits, Wiley, 1969, pp.61, 64 & 65.

Cross references

Series 7, cards 1 & 2. Series 2, card 4. Series 3, card 1. Series 4, card 8.

Circuit modifications



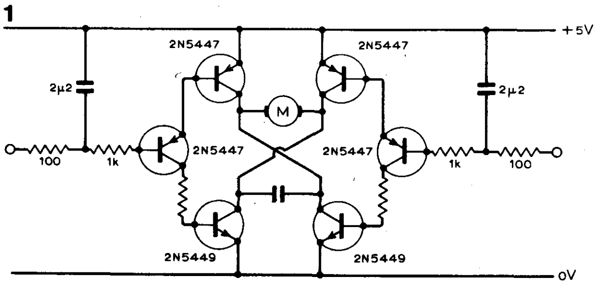


Wireless World Circard

Power amplifiers

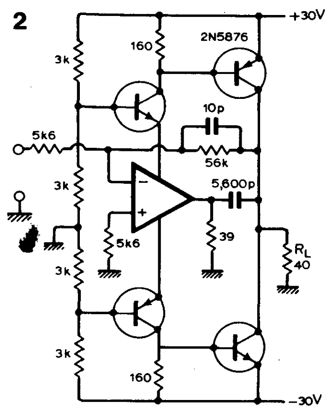
1. A bridge output configuration is very convenient for driving small motors in servo systems. It avoids the need for a centre tapped supply and with it the possibility of unbalance in the supplies which can cause unequal response in the two directions. The inputs can be driven from t.t.l. open-collector gates, though decoupling between the motor and logic-circuit supplies is desirable. The servo system is of the pulse proportional type with the transistors in the switched mode. The dissipation in each transistor is low and small plastic encapsulated types are adequate. For higher motor supply voltages, special open-collector devices are available with voltage ratings to 15V (e.g. SN7401A).
Bessant, M. F. Multi-channel proportional remote control, *Wireless World*, vol. 79, 1973, pp.479–82.

1



2. The use of the supply leads to an operational amplifier as signal output connections allows anti-phase current drives to a complementary pair of power transistors. By adding cascode-connected transistors the voltage limits of the op-amp can be adhered to while allowing any desired supply voltage. The cascode transistors experience the difference between supply and op-amp voltages while the output stages in this circuit have to cope with the full 60V. As shown, the peak power in the load can be as high as 22W and the output can swing to within 0.5V of either supply line. Amplitude response is flat to 30kHz.
Garza, P. P. Getting power and gain out of the 741-type op-amp. *Electronics*, vol. 46, 1973, p.99, (Feb. 1)

2



3. Recent monolithic audio power amplifiers have more and more of the required bias circuitry incorporated within the i.c. In some cases the only external components are a gain control and one or two components to ensure stability with particular loads (low load resistances in this circuit). In addition supply decoupling (with 0.1μF disc ceramic capacitor close to i.c.) may be needed if the supply is more than 2 to 3in from amplifier. The LM380 input is d.c. coupled, has internal feedback to set the gain and requires only a single polarity supply with the signal ground-referred. The bridge condition as drawn permits the output power to be doubled (limited by thermal ratings) and needs no coupling capacitor to load provided 1MΩ offset control sets outputs to same d.c. level.
Byerly, J. E. and Kooi, M. V. LM380 power audio amplifier, National Semiconductor application note AN-69, 1972, p.6.

3