Wireless World Circuit Series 7: Power amplifiers-1

Basic power amplifiers

Typical data
Supply: ±15V
Tr: BFY50
R1: 1.2kΩ
R2: 120Ω
R3: 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The classic transformer-coupled class A amplifier has been superseded for most purposes, but may still be applied where good isolation is required between source and load, or where the optimum impedance for maximum undistorted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 R1 and R2 fix the base potential of Tr1 provided the current through them is much greater than the base current. This base current is the required collector quiescent current divided by transistor hFE. These parameters fix the value of R1 + R2 by the approximate relationship R1 + R2 = hFE * Vss/m * Ic. The value of m is 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 Ic) is defined because the p.d. across R4 equals the p.d. defined across R3 minus the Vce of Tr1. 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 R4 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 C2 decouples R3 to prevent negative feedback within the required frequency range. As R4 may be a low resistance, C2 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).

Wireless World Circuit Series 7: Power amplifiers-2

Servo amplifier

Typical performance
Supplies: ±15V, 235mA
Quiescent current: ±1.0mA
A1: 741
Tr1: BF518
Tr2, Tr3: TIP3055
Tr4: BF441
R1: 15kΩ
R2: 47kΩ
R3: 4.7kΩ
R4: 180kΩ
C1: 2nF
C2: 1000pF
C3: 4.7nF
C4: 6.8pF
D1, D2: 1N914
R1: 180Ω
Risetime < 30μs (4.8V
pk-pk at 1kHz)
Vin: 2.07V r.m.s. without clipping

In servo systems a servo amplifier is needed when a high-power load must be driven from a low-power source. Amplifier A1 acts as a see-saw amplifier having its gain determined by R4/R3 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 Tr1 is forward-biased by only about 350mV due to the p.d. across R3. The base-emitter junctions of Tr4 and Tr5 would be forward-biased to a smaller extent unless R3 was greater than R4. However, including D2 and making R3 = R4 produces the desired bias with D2 providing some temperature compensation for the base-emitter voltage of Tr4. 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 R1 in push-pull. Transistors Tr4 and Tr5 form a Darlington pair while Tr1 and Tr3 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, R4 and C2 provides feedback around this operational amplifier and R4 and C3 provide feedback around the power stage. Bandwidth of the amplifier is controlled by C4R4 time constant which can be held fixed when the gain is varied by R3, if C4 is also adjusted. Diodes D1 and D2 protect the output transistors against breakdown when the load is highly inductive.
Supplies $V' = \pm 6V$,
$V'' = \pm 15V$

Tr1: BFR41
Tr2: BFR81
IC1: 741
R1: 470$\Omega$
R2: 150$\Omega$
Output power 0.92W at 64% efficiency
Output voltage swing:
10.5V pk-pk for $\pm 6V$ supply

Class B
The complementary pair of transistors acting as emitter followers comprise the basic class B push-pull stage. Transistor Tr1 conducts during the positive half-cycle and Tr2 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 ($\sim 0.5V$ 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 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 8$\Omega$ 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.

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 $P_{out} = 3.8W$ for $R_L = 25\Omega$. Total harmonic distortion falls as drive increases:
typically 0.45% for $V_{in} = 2.8V$ and 5.3% for $V_{in} = 150mV$ (supplies $\pm 15V$, $R_L = 18\Omega$ and $f = 1kHz$ sinewave).

Circuit modification
- The Tr1-Tr2 and Tr3-Tr4 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 BFR81 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 $R_2$ and $R_1$ and/or placing a diode in series with $R_2$ and $R_1$. The resistors in the emitters can be selected to provide the required quiescent current. (See circuit left.)

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

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

- 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

Cross references
Series 7 cards 1 & 12.
Series 2 card 4.
Series 4 card 8.
Wireless World Circard Series 7: Power amplifiers-3

Pulse buffer amplifier

Typical performance
V1: +14V; V2: +5V
Tr1: TIP355; Tr2: TIP50
IC1: 1N54; IC2: 1N5817
R1: 470Ω; R2: 100Ω
D1: 1N4001; C1: 680µF
Input pulse height: 4V
Duration: 600ns
P.R.F.: 50kHz
Rise time: 20ns

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 greater than the charging current required by the capacitor C1 to achieve a high rate of rise and/or fall. The class B push-pull stage shown has Tr1 driving the capacitor in the positive direction, while Tr2 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 R1 as D2 is allowed to conduct. On the negative-going edge, current through R4 is significantly greater and could cause excessive current flow in Tr2, but the diode is reverse biased and R3 takes the place of limiting action previously provided by R4. It is not possible in a simple circuit of this kind to choose a simple bias network for R1 and R2 which would give the same bias current in both directions.

IC1 is an open-collector high-voltage output device which pulls the potential at the bases of Tr1 and Tr2 to a low value when in conduction, and when out of conduction allows the bias to rise towards V1 via R1.

Wireless World Circard Series 7: Power amplifiers-4

Push-pull class A power amplifier

Typical performance
For supply of 13V, quiescent current of 950mA, max. output for 3% distortion is 12V pk-pk into 1Ω (5.6W). Mean current falls to 20mA 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% i.m.d. Distortion: <1.5%, 1W into 3Ω, 100Hz to 1kHz. 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 R6 remains constant throughout the a.c. wave form while its d.c. value can be adjusted to set the desired quiescent current. Bootstrapping via C1 ensures that any increase in the collector potential of Tr1 is transferred via the emitter follower action of Tr1 to reappear at the junction of R3 and R4. Hence the charge in p.d. across R4 approaches zero except at very low frequencies where the reactance of C1 becomes significant. As there is no change in R3 current, any increase in Tr1 current increases the base current of Tr2 while reducing the base current of Tr1 by substantially the same amount. Accurate current phase-splitting together with matched current gains of Tr1, Tr2 keep the distortion low. Overall negative feedback via R4 defines the output quiescent voltage as a multiple of the base voltage of Tr1 (~1.3V) and the ratio R3/R4 scales this base voltage up to half the supply voltage, i.e. the output transistors operate with equal Vbe as well as equal Ie.
Component changes
- Transistors Tr1 and Tr2 can be replaced by BFR41 & BFR81 or BC125 & BC126 with poorer rise and fall times. Typical comparison rise time (ns) fall time (ns)
  T845/50  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, V1: 14V, V2: +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.
- IC1: 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 Tr1 which brings Tr2 into conduction, clamping the output to the negative supply rail, with very small saturation effects. Conversely, if the output swings negative Tr2 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.

Cross references
Series 6, cards 1, 2 & 8.

Component changes
Tr1, Tr2: Power transistors with closely matched hfe at operating current. Quiescent power (at least twice max. output) determines types and heat sinks.
2N3055 for Pe > 5W, MJE321 for Pe > 1W
BFY50, BFR41, etc., for Pe < 1W.
Tr1: BFY50, BFR41, 2N3053 for most applications.
C1: Reactance < R1 at lowest freq. Typically 200 to 5000μF.
C2: Reactance < R2 at lowest freq. Typically 100 to 500μF.
R1, R2: Set output current \( \frac{V_s}{2(R_1 + R_2)} \approx 2I_o / h_{FE} \). One resistor made variable to adjust mean current. Typical range 100Ω to 1kΩ (higher values for low-power circuits).
R3: Sets voltage gain \( \approx R_3 / R_2 \) and input resistance \( \approx R_3 \).
R4, R5: Set output voltage (quiescent) to \( \approx 2V_s [R_3 / R_2] + 1 \).
Current in R2, R3 to 5 to 20 times base current of Tr1. Typical values R4: 100 to 500Ω. R5: 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 Vgs rather than at the supply centre point, i.e. resistors require readjusting for different supply volts. Adding Tr1 increases open-loop gain, allowing 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 midpoint of R3 ≈ R4. Voltage gain \( \approx (R_3 / R_2) + 1 \). Reactance of C2 < R2 at lowest frequency of interest. Typically R3, R4: 1 to 10kΩ. R5, R6: 20 to 200kΩ. Other values as before.
- For higher input impedance, input potential divider may be bootstrapped. Interchanging locations of R3, C4 allows R4 to be bootstrapped, almost doubling input impedance.
- Quiescent current depends on current gains of Tr1, Tr2. By monitoring circuit mean current and using result to control drive current Tr1, mean current can be made constant, e.g. for Tr2 a germanium transistor, D1, a silicon diode, mean p.d. across R3 is controlled at 0.4V.

Further reading
High-voltage amplifier

Typical performance
Supply: ±100V
Tr1: MJE340
Tr2: 2N3819
R1: 10kΩ; R2: 450Ω
Quiescent voltage: 52V
Vp: 10V
Input signal: 7V pk-pk
Output voltage: 84V pk-pk

Gain constant up to 20kHz
Variation of output with R1 not decoupled shown opposite.
Effective output impedance of transistor con-

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 characteris-
tics are obtained by Tr2 and the high voltage characteristics by Tr1. 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 value. Transistor Tr2 is thus a field-effect transistor whose gain is controlled by the quiescent current, which may be set by R1. The drain of Tr2 feeds into the emitter of Tr1 whose base is maintained at a constant potential, just high enough to ensure that Tr1 has a quiescent voltage that is above its pinch-off value. The bias voltage should be obtained from a low impedance circuit.

Hence Tr1 is operating into a low impedance, while Tr2 is virtually a common-base stage and has thus the highest voltage rating that it could possibly have. The current at the collector of Tr1 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 Tr1 and the gate of Tr2 as the voltage swing at the collector is isolated from the gate of Tr2. The capacitance between Tr1 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 R1 for a.c. and determines the bandwidth of the amplifier. The problem is more severe than in many low-voltage amplifiers because R1 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 VCB breakdown of Tr1.

Component changes
- Decouple R1 with 150μF capacitance to retain gm 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 Vp 8 to 11V – value not critical, with no significant effect on performance.

Class C power amplifier

Typical data
Supply: 12V
Tr1; Tr2: BFR41
R1: 100Ω; R2: 50Ω
(Carbon)
C1: 180pF; C2: 360pF
C3: 47pF; C4: 10nF
C5: 500pF; C6: 190pF
C7: 805pF; L1: 2.7μH
L2: 2.16μH; L3: 2.38μH
L4: 230μH; L5: 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 C1, C2, C3, C4 and C5 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 graph) but taking account of the d.c. drop (3.52V) across the r.f. choke L4 efficiency rises to 66.5%. Hence L4 should have low resistance, but its effect is less noticeable at lower currents. Transistors Tr1 and Tr2 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 L3, L4 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.
Supply may be increased up to 300V with appropriate changes in R1 and R2 to control quiescent voltage. Typically (i) supply: +200V, \( V_G \): 110V, \( R_2 \): 122Ω, \( V_{in} \): 7V pk-pk; \( V_{out} \): 180V pk-pk; \( R_1 \): 10kΩ. (ii) supply: +300V, \( V_G \): 150V, \( R_2 \): 1.5kΩ, \( R_3 \): 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 at positive, \( R_G \) 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

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 \( R = V_{sec}/2P_{out} \). 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 \( C_3 \). Complete design procedures are given in the first three references.

Further reading
Motorola, application note AN-282: Systemizing r.f. power amplifier design, 1967.

Cross references
Circard series 7, card 1.
Wireless World Circuit Series 7: Power amplifiers—7

Bridge output amplifiers

Typical data
ICT1, ICT2: 741
R1: 10kΩ pot
R2: 10kΩ
Supplies: ±15V
R3: 2kΩ
Output voltage: 15V
r.m.s. into 2kΩ (17.5V)
For m.s. o/c for k = 0.1 to 1.0 at 1kHz.

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 crucial performance limitations imposed by transistors arise from the need for an alternative output configuration for increased output voltage swing. When 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 needs to 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 R4 = R1. Equal magnitudes of output are ensured for this condition assuming ideal amplifiers because their two resistors carry equal current while their junctions are virtual earth points. 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.

Wireless World Circuit Series 7: Power amplifiers—8

Class B quasi-complementary output

Typical performance
Supply: ±20V
Tr1: BFR81; Tr2, Tr3: TIP3055
R1: 1.5kΩ; R2: 1kΩ
R4: 470Ω; R5: 330Ω
R6: 1.8kΩ; R7: 8.2kΩ
C1: 100µF; C2: 22µF;
C3: 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
Grads of harmonic distortion and efficiency versus output power for loads of 15Ω and 8Ω shown opposite.

Circuit description
This is a circuit of a class B push-pull amplifier in which transistors Tr2 and Tr3 complement the pair Tr1 and Tr4. To use n-p-n transistors in the output stage for economy, the configurations of the two sections are different, i.e. Tr3 and Tr4 are connected as a Darlington pair and Tr1 and Tr2 as a complementary pair. They receive essentially the same a.c. drive, but with the bases separated by Tr5, Tr6 and Tr7 conduct

for positive-going output signals and Tr8 supplies base current drive to Tr1 and Tr4 for negative-going output signals. Transistor Tr5 is used in the so-called amplified diode configuration in which the potential difference between the bases of Tr6 and Tr7 is set as a multiple of the Vbe of Tr1 by the potential divider R9, R10, i.e. R9 can be adjusted to give the desired quiescent current in transistors Tr2 and Tr3. 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 Tr5 is an inverting amplifier with overall negative feedback through R7, the values of R8 and R7 determining the d.c. output potential
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.

- An alternative to the bridge circuit for increased voltage swing is the principle of supply bootstrapping of which this is one version.

- 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 $R_4$ 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.

Further reading


Del Conso, 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.


in conjunction with $R_4$. Because $R_5$ is decoupled, the a.c. properties of the arrangement are determined by the ratio of $R_5$ to the source resistance. Resistors $R_1$ and $R_5$ are centretapped and this point is taken to the output via $C_1$, which bootstraps $R_5$ 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 $R_5$ between the emitters and the output terminal (left).

- Middle circuit shows an alternative arrangement, adding transistors $TR_1$ and $TR_2$. These are normally non-conducting except under overload conditions, i.e. as the output current increases the voltage drop across $R_5$ or $R_6$ causes $TR_1$ or $TR_2$ to turn on and divert the base current available to $TR_4$ or $TR_3$, limiting the output current to $Vbe/R_3$.

- Alternative configurations for the output stages are shown right (i) requires low and high power p-n-p and n-p-n transistors to make up the Darlington pairs, the minimum p.d.

between input and output circuits being twice the $Vbe$ of a single transistor, (ii) uses complementary Darlington 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.


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 Series 7: Power amplifiers-9

**Broadband amplifier**

![Circuit diagram](image)

**Typical performance**
- Supply: +20V, 118mA
- \( T_1 \): BFR41; \( T_2 \): BFY50
- \( R_1 \): 35Ω; \( R_2 \): 150Ω
- \( R_3 \): 220Ω; \( R_4 \): 22kΩ
- \( R_5 \): 120Ω; \( R_6 \): 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 \( h_{fe} \)arity 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, \( T_1 \) and \( T_5 \) operate in regions where their current gain is much smaller than the normal values and they therefore have relatively high quiescent currents. 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 \( R_5 \). Resistor \( R_3 \) is determined from the required voltage gain (AV) for equal source and load resistances \( R_3 \approx \frac{AV \cdot R_L}{AV + h_{fe}} \). Inductor \( L \) is included to offset the capacitive loading due to \( T_1 \) 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 \( R_4 \).

Wireless World Circard Series 7: Power amplifiers-10

**Class A op-amp power booster**

![Circuit diagram](image)

**Typical performance**
- \( T_1 \): BFR81
- \( T_3 \); \( T_4 \); 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: 1.25mA (set by \( R_1 \))

**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 \( R_4 \) then as the base potential of \( T_1 \) increases the emitter current of \( T_1 \) decreases and with it the collector current.

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 1Ω load.

This fall is fed to the base of \( T_1 \); causing it to conduct less, while the fall in emitter current releases more of the constant current in \( R_3 \) to flow in the base of \( T_1 \). Provided the current gain of \( T_1 \) is reasonably high, the magnitude of the base current charges in \( T_1 \) and \( T_3 \) are equal but the signs are opposite. This represents an approach to ideal current phase-splitting. The constant current in \( R_4 \) is provided by the bootstrap capacitor \( C_2 \), such that any change in the potential at the base of \( T_3 \) is coupled via the follower action to the positive end of \( R_3 \), i.e. with no resulting change of p.d. across \( R_3 \) in the ideal case. Resistors \( R_2 \) or \( R_3 \) require to be variable to set the
Component changes
With \( V_{cc} \) (min) = +5V, \( V_{in} \) (max) \( \approx 140 \text{mV} \) r.m.s., supply current is 30mA, and \( P_{out} \approx 11 \text{mW} \). 

\( T_1 \) and \( T_2 \) can both be BFY50 or BFR41. 

\( T_1 \) 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 on 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 \( T_1 \) may be drastically reduced. Resistor \( R_2 \) 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 \( T_1 \) 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 \( T_1 \) is d.c. or a.c. grounded. The feedback is not decoupled and the voltage gain is determined by the ratio \( R_1/R_2 \). The input resistance is largely that of \( R_1 \) except at high frequencies where the feedback falls and the impedance at \( T_1 \) base must be considered.
- Inclusion of \( R_4 \), as shown right, may be applied to both the previous circuits to allow an output to be taken from the collector of \( T_2 \). To maximize the signal swing in the collector circuit of \( T_2 \) the bias network must be readjusted to leave a small voltage at \( T_2 \) emitter, say by reducing \( R_3 \) and \( R_1 \) in the original circuit. The output resistance is approximately \( R_1 \); this stage is therefore convenient for feeding directly into any other low impedance stage, such as that left, with \( R_4 \) removed. This mismatch can often be of advantage in extending the bandwidth of the amplifier.

Further reading


Lo, A. W. (and others), Transistor for Electronics, Chapter 9, Prentice-Hall, 1955.


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

output current and stability of that current then depends on hve variation in \( T_1 \), \( T_2 \). The base-emitter p.d.s of \( T_1 \), \( T_2 \) 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 \( R_2 \), \( R_4 \) above the supply on positive signal swings. Hence it can drive \( T_1 \) base far enough positive to saturate it 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 \( R_4 \) may replace \( R_2 \).
- 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.
- 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 \( R_2/R_4 \) + 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.

Further reading


Cross references
Series 7, cards 2, 4 & 8.
**Wireless World Circard Series 7: Power amplifiers-11**

**D.C. power amplifier**

![D.C. power amplifier circuit](image)

**Circuit description**

This circuit uses a voltage regulator, i.e., package to supply an output stage $\text{Tr}_1$, where the amplifier is to be used primarily with a unipolar signal, though it can also be interpreted as a unipolar stage. 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 $\text{Tr}_1$ 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 $R_6$, the quiescent current in $\text{Tr}_1$ is then set by the bias resistor $R_2$ in conjunction with this predetermined voltage.

**Typical performance**

- $\text{IC}_1$: LM305 or LM100
- $\text{Tr}_1$: MJE271
- Supplies: $\pm 15\text{V}$
- $R_1$, $R_2$: $150\Omega$; $R_3$: $1.5\text{k}\Omega$
- $R_4$: $1\text{k}\Omega$; $R_5$: $1.8\text{k}\Omega$
- $R_6$: $10\text{k}\Omega$
- $C_1$: $1\mu\text{F}$ (tantalum);
- $C_2$: $10\mu\text{F}$

**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.

An a.c. signal can be superimposed at pin 6 via $C_1$ and $R_4$ to 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 $R_4$ falls and with it the current through $R_4$. The current in $R_4$ for this voltage exclusion can never exceed $R_4$, even when the transistor current falls to zero in the positive direction; however, much greater currents can be provided through $\text{Tr}_1$. The amplifier is thus an inefficient class A amplifier whose effectiveness can be improved by replacing $R_4$ by a constant-current stage which can sustain a given peak current in $R_4$, almost equal to the quiescent value, even for large voltage excursions in the negative direction.

Capacitor $C_1$ is used to suppress f.f. oscillation and a low inductance type must be used.

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**Wireless World Circard Series 7: Power amplifiers-12**

**Class D switching amplifier**

![Class D switching amplifier circuit](image)

**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 $R_5$ and $R_4$. The periodic time of the squarewave fed to $R_4$ depends on the time constant $R_5C_3$ if $R_4$ is much greater than $R_5$. To obtain a realistic switching frequency with

\[ V_{\text{bias}} = -640\text{mV} \] to set

- mean load voltage to zero with $V_{\text{in}} = 0$;
- switching frequency: 27.8kHz, max 40 kHz;
- With $V_{\text{in}} = 0$, supply current is $\pm20\text{mA}$; with $V_{\text{in}} = 3.4\text{V pk-pk}$, current is $\approx 130\text{mA}$; power in 15-\Omega load $\approx 1.66\text{W}$; residual "carrier" $\approx 300\text{mV}$ across 15\-\Omega; overall efficiency $64\%$; output stage efficiency $\approx 76\%$; 3-dB bandwidth $\approx 600\text{Hz}$.

reasonable components and also to obviate the need for large input signals a compromise must be made in the value of $R_4$. Current in $R_4$ flows alternately in $R_5$ and $R_4$ producing p.d.s across these resistors that are sufficient to switch on $\text{Tr}_1$ and $\text{Tr}_2$ respectively. The signal applied to $R_4$ causes the mark-to-space ratio of the output waveform from the astable to vary in sympathy with the instantaneous value of $V_{\text{in}}$, 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 $V_{\text{in}} = 0$, it can be made so by a suitable choice of the bias supply and $R_4$. Diodes $D_1$ and $D_2$ protect $\text{Tr}_1$ and $\text{Tr}_2$ against breakdown when the load impedance is highly inductive.
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 R4 is replaced by the Baxandall constant-current circuit shown left TR1; BFR81, TR2; TIP355, R1; 18Ω, R4; 3.9kΩ. This permits a much greater input signal level before peak clipping occurs. Resistor R1 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 TR2 output current is 200mA pk-pk, then the output current swing in load R1 is twice that for the case when R4 is 15Ω with the same quiescent current. A comparison of instantaneous currents for the two possible circuits between a and b is tabulated above.

- 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.

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

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