# 10

# Power Amplifiers

Class-A amplifiers produce outputs with little distortion because the transistors are biased and driven so that they are always active. However, when a transistor is active, the voltage and current are large at the same time, so that the dissipated power is substantial and the efficiency is poor, in the range of 35%. In addition, the amplifier dissipates power even when there is no output. These are severe limitations for even modest output power levels; consequently, few power amplifiers run Class A. To eliminate the power drain when there is no signal, we can leave the transistor unbiased, so that it does not dissipate power when it is off. In addition, if we drive the transistor clear to saturation, using the transistor as a switch, the dissipated power can be greatly reduced because the saturation voltage is low. This is Class-C amplification, which achieves excellent efficiencies, in the range of 75%. We will also see variations of Class C, the Class D, E, and F amplifiers, that achieve even higher efficiencies. The disadvantage of operating Class C is that the output amplitude no longer follows the input level. There is significant distortion at both low and high levels. We say the amplifier is *nonlinear*, and this presents challenges in amplifying signals that vary in frequency and amplitude at the same time, such as music in stereo amplifiers. However, Class C is quite suitable for signals that simply turn on and off, such as Morse Code in the NorCal 40A, or signals that only vary in frequency, such as FM transmissions. Class C also works for signals that vary in amplitude alone, such as AM broadcasts, because the amplitude can be controlled by the supply voltage.

Class-B amplifiers are active for half of a cycle. They usually have a small bias to reduce low-level distortion, and they operate at output levels below saturation to stay linear. They represent a compromise between the poor efficiency of Class A and the distortion of Class C. The bias is usually small enough that the power drain is not a problem. They achieve efficiencies that are substantially better than Class A, but not as good as Class C, in the range of 60%. The output stage of the Audio Amplifier in the NorCal 40A is a Class-B amplifier. Figure 10.1 shows the relationships among the different amplifier classes.

Even efficient transistor amplifiers are often limited in their output power by heat. Transistors are small, and this makes it difficult to get the heat out. In addition, transistors are limited to modest operating temperatures, and they are made of semiconductors that are only fair conductors of heat. This is in contrast

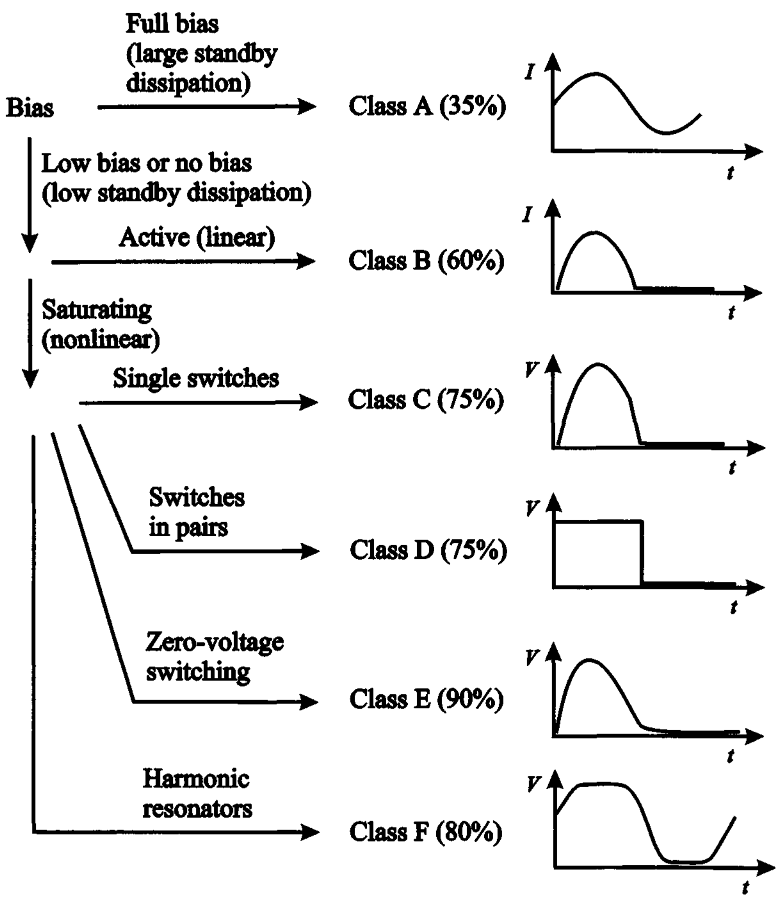



**Figure 10.1.** Classes of transistor power amplifiers with characteristic waveforms. Representative efficiencies at radio frequencies are given in parentheses. The classification can be quite confusing, partly because transistor and vacuum-tube amplifiers operate differently, and partly because usage has changed over time. Transistor amplifiers achieve high efficiency by saturating, whereas vacuum-tube amplifiers achieve high efficiency by being active only over a small part of the cycle. Traditionally Class C refers to amplifiers that conduct over less than half of the cycle. This is appropriate for tube amplifiers, but difficult to apply to the transistor Power Amplifier in the NorCal 40A, where the transistor conducts for a full half cycle. We will call saturating amplifiers Class C even if they conduct over more than half a cycle. We will also call amplifiers Class C if they conduct for less than half a cycle, even if they never saturate. We will see that the oscillators we study in the next chapter operate in this mode. Class D uses transistors as switches in pairs to increase both efficiency and power. Class E uses a network that allows switching when the voltage is low, achieving spectacular efficiencies, as high as 90%. Class-F amplifiers add harmonic resonators to shape the voltage waveforms to reduce the peak voltage.

to vacuum tubes, which are large structures of metal, glass, and ceramic that can operate at high temperatures and dissipate large amounts of heat. Transistor limitations make it important to understand their thermal behavior. We can make a thermal model for a power amplifier that is the analog of an RC circuit.

## 10.1 Class-C Amplifiers

Figure 10.2 shows the Class-C Power Amplifier in the NorCal 40A. It is a common-emitter amplifier, with no emitter resistor at all. The collector voltage is supplied through a large inductor called an *RF choke*. The choke gives the supply a large RF impedance, so that it effectively acts as a DC current source. The load current is taken through a large capacitor called a *DC block*. The blocking capacitor has a small RF impedance that does not affect AC currents, but it prevents DC current from the supply from getting to the antenna. Many antennas have transformers that make them DC short circuits. The Harmonic Filter removes the harmonic components (Problem 13).

We represent the transistor by a switch that opens and closes at the operating frequency (Figure 10.3a). When the switch is open, the transistor is off. When it is closed, the transistor is on. A voltage source $V_{on}$ is included to take the on-voltage into account. This is an approximation, because the on-voltage varies with the current. When the switch turns off, there is a large ringing voltage caused by the

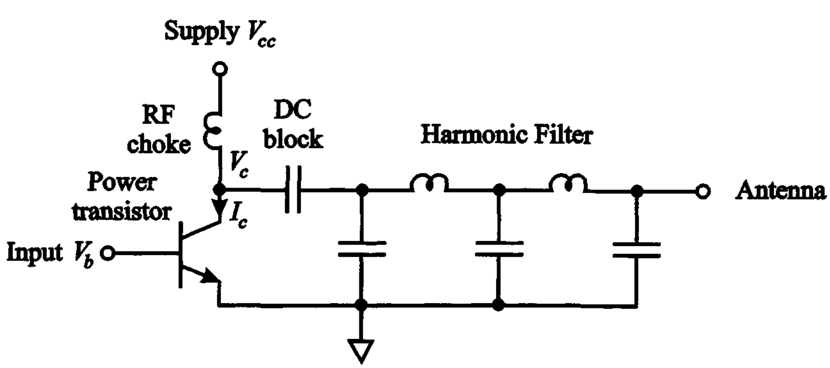



**Figure 10.2.** Class-C Power Amplifier in the NorCal 40A.

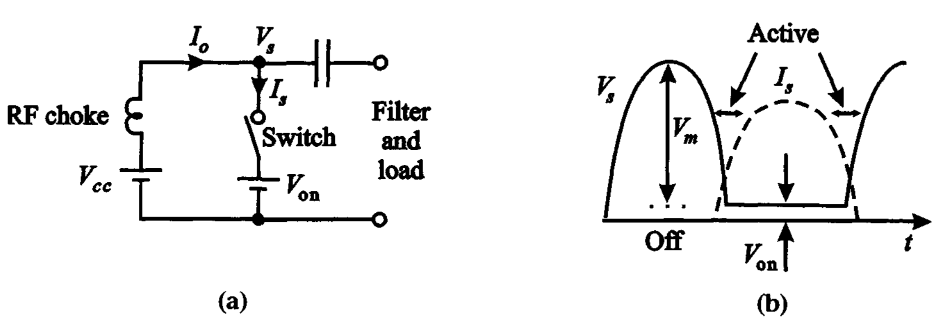



**Figure 10.3.** Switch model for the Class-C amplifier (a), and switch voltage $V_s$ and current $I_s$ waveforms (b).

Harmonic Filter. We saw a voltage like this in the transistor switch in Problem 5. As an approximation, the switch voltage $V_s$ is a rectified cosine wave superimposed on $V_{\text{on}}$ (Figure 10.3b). The switch current $I_s$ is zero when the switch is off. When the switch turns on, current flows. There is some overlap with the cosine voltage, because the transistor is active during the transition from off to on. There is another active period when the switch turns off.

First we can relate the supply voltage $V_{cc}$ to the switch voltage $V_s$. We can write $V_s$ in the form

$$V_s(t) = \begin{cases} V_{\text{on}} + V_m \cos(\omega t), & \text{switch off,} \\ V_{\text{on}}, & \text{switch on,} \end{cases} \tag{10.1}$$

where $V_m$ is the peak value of the rectified cosine wave. Because the choke has no DC resistance, the DC or average value of $V_s$ must be the same as the supply voltage. The average value of the rectified cosine over a full cycle is $V_m/\pi$. Therefore we can write

$$V_{cc} = V_{\text{on}} + V_m/\pi. \tag{10.2}$$

We can write $V_m$ in terms of the supply voltage as

$$V_m = \pi(V_{cc} - V_{\text{on}}). \tag{10.3}$$

This formula indicates that we can interpret $V_{cc} - V_{\text{on}}$ as the effective supply voltage.

We can write the power from the supply $P_o$ as

$$P_o = V_{cc} I_o, \tag{10.4}$$

where $I_o$ is the DC supply current. We can also write an expression for the switch loss. Because the blocking capacitor passes no DC current, the DC switch current must also be $I_o$. For now, we assume that the time that the transistor is active is small, and we neglect it. Therefore we can write the power dissipated in the switch $P_d$ as

$$P_d = V_{\text{on}} I_o. \tag{10.5}$$

Now that we have accounted for the power lost in the transistor, the remaining power must be the output $P$. We can write

$$P = P_o - P_d = (V_{cc} - V_{\text{on}}) I_o. \tag{10.6}$$

The efficiency is given by

$$\eta = P/P_o = (V_{cc} - V_{\text{on}})/V_{cc}. \tag{10.7}$$

The efficiency is the ratio of the effective supply voltage to the actual supply voltage. This formula tells us that to make efficient amplifiers, we need to make the on-voltage small and the supply voltage large. We can reduce the on-voltage by reducing the DC current, but this also reduces the output power. Moreover, we have to be careful not to exceed the manufacturer's limit for peak voltage.

To go further with the analysis, we need to interpret the voltage waveform as a sum of harmonic frequency components. This is called a Fourier series. In

Appendix B, Section 3, we derive a formula for the Fourier coefficients. They are given by

$$V_s(t) = V_{cc} + \frac{V_m}{2}\cos(\omega t) + \frac{2V_m}{\pi}\left(\frac{\cos(2\omega t)}{3} - \frac{\cos(4\omega t)}{15} + \frac{\cos(6\omega t)}{35} - \cdots\right). \tag{10.8}$$

The first term in the sum is just the DC voltage. The next term is the component at the transmitter frequency. This is called the *fundamental*. The peak value of the fundamental component is $V_m/2$. This makes sense, because we have only half a cosine. The other components are even harmonics, the second, the fourth, and so on. Because the filter greatly reduces the harmonics at the load, we will not consider the power in them. We will assume that the input impedance of the filter at the fundamental is real and given by $R$. Since the peak voltage of the fundamental component is $V_m/2$, we can write the output power as

$$P = \frac{V_m^2}{8R} = \frac{\pi^2(V_{cc} - V_{\mathrm{on}})^2}{8R}. \tag{10.9}$$

We can increase the power by increasing the supply voltage, as long as we do not exceed the maximum voltages, currents, and temperatures specified for the transistor. Since the output power varies inversely with the input impedance of the filter, if we want to increase the power, we should reduce the impedance. We considered this in Problem 13. There are limits to how much the power can be increased, however, because lower impedances raise the current, and this increases $V_{\mathrm{on}}$. This reduces both the effective supply voltage and the efficiency. In addition, at high current levels in bipolar transistors, $I_c$ is proportional to $\sqrt{I_b}$ rather than $I_b$. This is shown in Figure 10.4a. This limits the current that can be driven through the transistor.

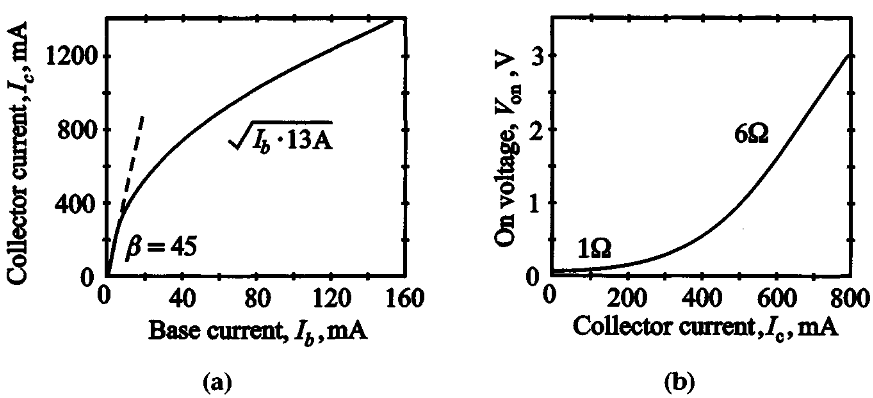



**Figure 10.4.** Effects of large currents in the 2N3553 bipolar transistor used in the Nor-Cal 40A Power Amplifier. (a) Collector current $I_c$ as a function of base current $I_b$. The collector-emitter voltage is 2.5 V. The relationship at low currents is linear, with a $\beta$ of 45, but for collector currents above 200 mA, the measurements follow the curve $I_c = \sqrt{I_b \cdot 13\,\mathrm{A}}$ quite closely. (b) On-voltage $V_{\mathrm{on}}$ as a function of the collector current $I_c$. The base current is 20 mA. There is a kink in the curve, where the slope changes from 1 Ω to 6 Ω.

For Class-A amplifiers, we derived an amplifier output impedance, which is useful for writing a Thevenin equivalent circuit and for matching for maximum power transfer. For transistor Class-C amplifiers, the output impedance is usually small, and it is harder to pin down. In a Class-A amplifier, the output impedance tells us how much the output voltage will drop as current is drawn. In a Class-C amplifier, $V_{\text{on}}$ plays this role, because it increases with current, and this reduces the effective supply voltage. At the large currents in a power amplifier, the collector voltage makes a gradual transition from saturation to active, so that there is not a simple relation between $V_{\text{on}}$ and $I_c$ (Figure 10.4b).

## 10.2 NorCal 40A Power Amplifier

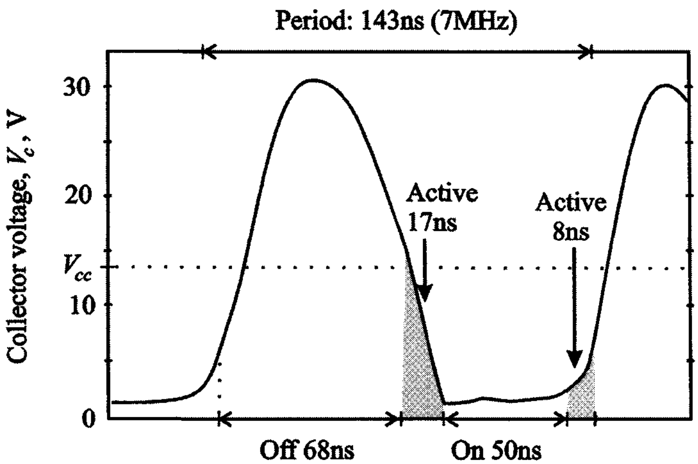
We can use the formulas we developed for Class-C amplifiers to make a quick but optimistic prediction of the performance of the NorCal 40A Power Amplifier and to compare these predictions with measured values. Figure 10.5 shows the measured collector and base voltage waveforms. The supply voltage is $V_{cc} = 12.8$ V, and $V_{\text{on}} \approx 2$ V. The on-voltage of 2 V is much larger than we have seen in the low-level switches in Problems 19 and 20. This large voltage is caused by the large current that flows in the power transistor when it is on. The input impedance of the filter $R$ is close to 50 Ω. From Equations 10.9 and 10.7, we predict a power of 2.9 W and an efficiency of 84%, which are somewhat higher than the 2.5 W and the 78% that was measured.

In both Class-C and Class-D amplifiers, there is additional loss when the transistor is active. If you look carefully at Figure 10.5a, you can see a slope change where the active regions begin. The active transition from off to on is the easier one to see, and it is 17 ns long. The transition from on to off is not as distinct and lasts only 8 ns. For comparison, the on time is 50 ns, and the off time is 68 ns. The off time is actually somewhat less than half of the cycle, which is 143 ns long. The off time corresponds closely to the time that the base voltage in Figure 10.5b is below 0.6 V. It seems surprising that the base–emitter diode conducts through more than half the cycle, because the base is driven with a sine wave. The explanation is that a saturated transistor requires additional time to turn off, because there are a large number of electrons in the base that must recombine or leave first. This is called a *charge-storage* delay. It also occurs in pn diodes, but not in Schottky diodes, which have a metal anode instead of a silicon one.

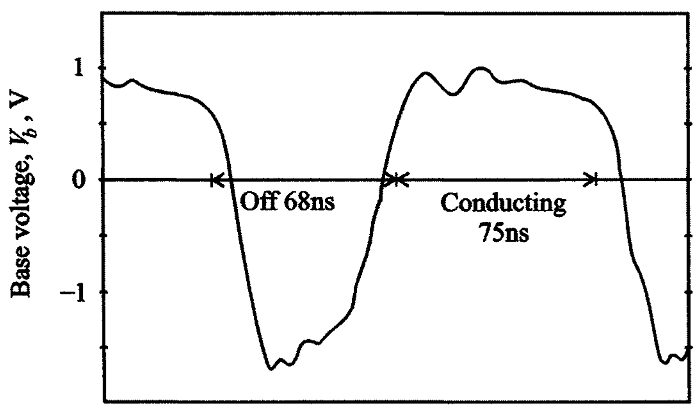
It is interesting to account for the transistor loss $P_d$. We can calculate it as

$$P_d = P_o - P = 3.2\,\text{W} - 2.5\,\text{W} = 700\,\text{mW}. \tag{10.10}$$

We start with the active period loss. The active transition from off to on has a larger loss than the one from on to off, because it is longer and the voltages are higher. Physically we account for the loss as a capacitive discharge through the transistor (Figure 10.6a). The input capacitor in the Harmonic Filter C45 (330 pF) is connected through a DC block to the collector of the power transistor Q7. The transistor itself, the zener diode D12, and the RF Filter also have capacitance, but




(a)




(b)

**Figure 10.5.** Voltages for the NorCal 40A Power Amplifier at 7 MHz. The output power is 2.5 W, with a supply voltage of 12.8 V and a supply current of 250 mA. The efficiency is 78%. Collector voltage (a), and base voltage (b).

all are small, and we will neglect them. The capacitor is charged to 15 V at the beginning of the active period. We can write the capacitive energy $E$ as

$$E = CV^2/2 = 37\,\text{nJ}, \tag{10.11}$$

where $C$ is the capacitance of C45, 330 pF, and $V$ is 15 V. During the active period, this energy dissipates in the transistor. At the end of the active period, the voltage has dropped to 2 V, and the capacitive energy is less than a nanojoule. We can

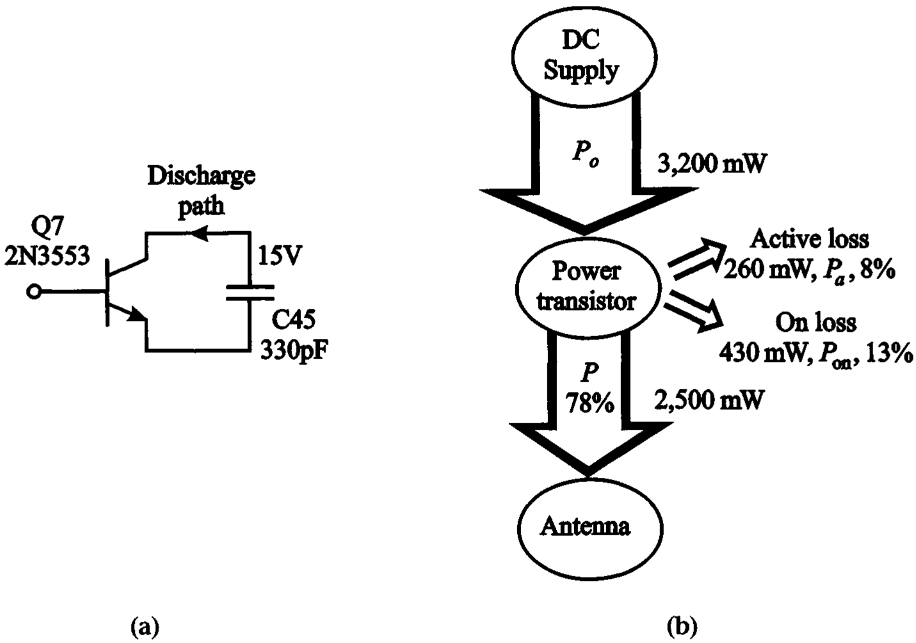



**Figure 10.6.** (a) Capacitive discharge through the collector of the power transistor. The DC block (C44) is not shown. C45 is charged to 15 V when the active transition begins. (b) The power flow in the NorCal 40A Power Amplifier when the output is 2.5 W.

convert this dissipation to a power $P_a$ by multiplying by the frequency to get

$$P_a = Ef = 260\,\text{mW}, \tag{10.12}$$

where $f$ is the frequency, 7 MHz.

We can find the current $I_c$ associated with the capacitive discharge by multiplying the capacitor charge Q by $f$. We write

$$Q = CV = 5.0\,\text{nC} \tag{10.13}$$

so that

$$I_c = Qf = 35\,\text{mA}. \tag{10.14}$$

We can calculate the on-period current $I_{\text{on}}$ by subtracting $I_c$ from the DC current $I_o = 250$ mA. This gives us

$$I_{\text{on}} = I_o - I_c = 215\,\text{mA}. \tag{10.15}$$

We can find the loss during the on period $P_{\text{on}}$ by multiplying $I_{\text{on}}$ by $V_{\text{on}} = 2$ V:

$$P_{\text{on}} = V_{\text{on}} I_{\text{on}} = 430\,\text{mW}. \tag{10.16}$$

We get the total dissipated power $P_d$ by adding $P_a$ and $P_{\text{on}}$. This gives us

$$P_d = P_a + P_{\text{on}} = 690\,\text{mW}, \tag{10.17}$$

which is close to the measured value. Figure 10.6b summarizes the power flow in the amplifier.

## 10.3 Class D

We can extend the idea of a switching amplifier by employing a pair of transistor switches that alternately connect a load to a voltage source and to ground. This is a Class-D amplifier, and a simplified circuit is shown in Figure 10.7a. The switch pair is called a *push–pull* circuit, in contrast to the Class-C amplifier, which is said to be *single-ended*. The supply voltage $V_{cc}$ is connected directly to a switch without a choke. There is a band-pass filter to prevent DC and harmonic currents from reaching the load. We can simplify the circuit by representing the pair of switches by a single double-throw switch at the operating frequency (Figure 10.7b). This gives us a square-wave voltage superimposed on the on-voltage of the transistor (Figure 10.7c). We write the switch voltage $V_s$ as

$$V_s(t) = \begin{cases} V_{cc} - V_{\text{on}}, & S_1 \text{ on}, S_2 \text{ off}, \\ V_{\text{on}}, & S_2 \text{ on}, S_1 \text{ off}. \end{cases} \tag{10.18}$$

The difference between the maximum and minimum voltages is given by

$$V_m = V_{cc} - 2V_{\text{on}}. \tag{10.19}$$

For a Class-C amplifier, we interpreted the rectified cosine in terms of its frequency components. We can also write a square wave this way. The coefficients are derived in Appendix B, in Section 2. The voltage $V_s(t)$ is given by

$$V_s(t) = \frac{V_{cc}}{2} + \frac{2V_m}{\pi}\left(\cos(\omega t) - \frac{\cos(3\omega t)}{3} + \frac{\cos(5\omega t)}{5} - \cdots\right). \tag{10.20}$$

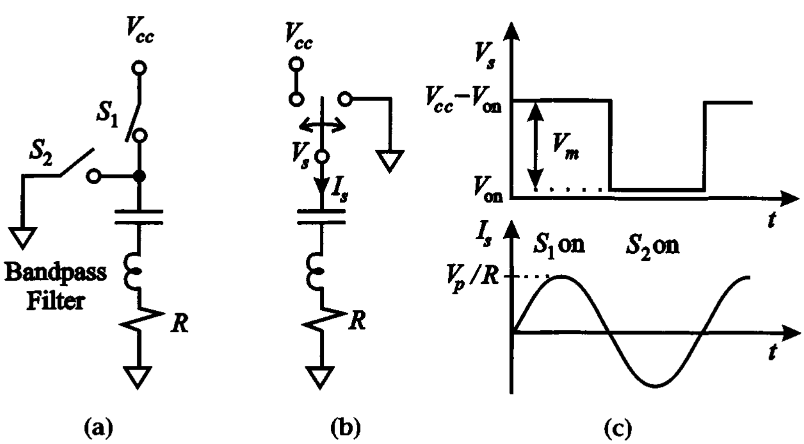



**Figure 10.7.** Class-D amplifier with a pair of transistor switches (a), and a simplified circuit with a single double-throw switch (b). The switch voltage and current waveforms (c).

There are only odd harmonics. The peak voltage of the fundamental component $V_p$ is given by

$$V_p = 2V_m/\pi. \tag{10.21}$$

This means that we can interpret $V_m = V_{cc} - 2V_{\text{on}}$ as the effective supply voltage. Since the filter stops all of the current components except the fundamental, the output power $P$ is given by

$$P = \frac{V_p^2}{2R} = \frac{2V_m^2}{\pi^2 R}. \tag{10.22}$$

The output power is inversely proportional to the load resistance. The input power $P_o$ is given by

$$P_o = V_{cc} I_o, \tag{10.23}$$

where $I_o$ is the average supply current. The supply current is a rectified cosine with a peak value $I_p$ of

$$I_p = V_p/R, \tag{10.24}$$

so that we can write

$$I_o = \frac{I_p}{\pi}. \tag{10.25}$$

We can rewrite this in terms of the effective voltage $V_m$ by substituting from Equation 10.21. This gives us

$$I_o = \frac{2V_m}{\pi^2 R}, \tag{10.26}$$

and the output power $P$ is

$$P = V_m I_o. \tag{10.27}$$

We get the efficiency by dividing Equation 10.27 by Equation 10.23 to get

$$\eta = P/P_o = V_m/V_{cc}. \tag{10.28}$$

The efficiency is the ratio of the effective supply voltage to the actual supply voltage. In practice, the efficiencies of Class-D amplifiers are comparable to Class-C amplifiers. The square-wave voltage of a Class-D amplifier is a major advantage in many situations because the maximum voltage is low, the same as the supply voltage. By comparison, the Class-C amplifier peak voltage is $\pi$ times larger than the DC voltage. However, the switches must be carefully synchronized, and this limits Class D to lower frequencies. They are common in AM transmitters and power supplies.

## 10.4 Class E

In the NorCal 40A Power Amplifier, 40% of the loss comes from capacitive discharge when the transistor turns on. One thing that you might notice in Figure 10.5a is that the voltage is already coming down when the transistor goes active. If we substitute a series resonant circuit for a series inductor in the harmonic filter (Figure 10.8a), the voltage can be made to come all the way down to zero before the transistor turns on (Figure 10.8b). This is called zero-voltage switching, and it eliminates capacitive discharge loss. Gerald Ewing demonstrated this idea in 1964, and it is the basis for the Class-E amplifier defined by Nathan and Alan Sokal in 1975. In the Class-E amplifier, both the voltage and its slope are zero when the transistor goes active. This means that even if the switching point is mistimed, or the switching is slow, the loss is still low. Class-E amplifiers are the most efficient amplifiers known, with efficiencies in the 90% range. The disadvantage is that the peak voltages for Class-E amplifiers are even higher than those for Class-C amplifiers.

You might wonder why it is important to work to raise the efficiency from 75% to 90%. There would be a modest improvement in the electric bill for a transmitter or battery life for a portable transceiver. The real answer, however, is heat – the loss is two and a half times lower for 90% efficiency than for 75% efficiency. If an amplifier is limited by the heat it can dissipate, we can calculate the maximum power output as a function of efficiency. We can write the output power $P$ in terms of the supply power $P_o$ as

$$P = \eta P_o \tag{10.29}$$

and the dissipated power $P_d$ as

$$P_d = (1 - \eta) P_o. \tag{10.30}$$

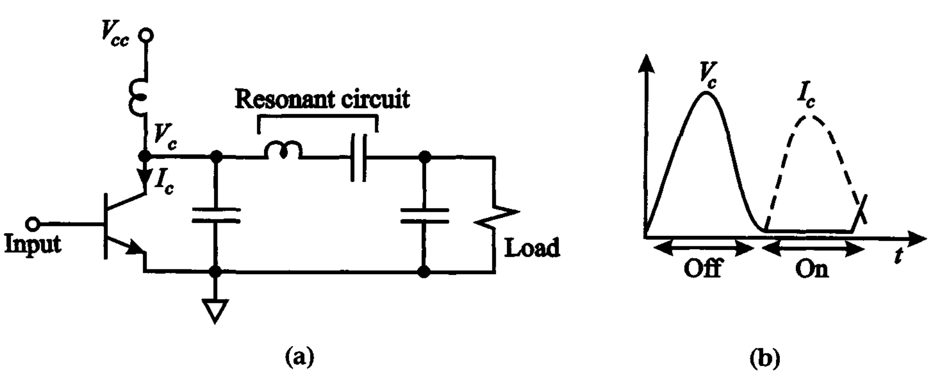



**Figure 10.8.** Class-E amplifier (a), and collector voltage and current waveforms (b). The components in the resonant circuit are adjusted so that the collector voltage comes all the way down to zero before the transistor turns on.

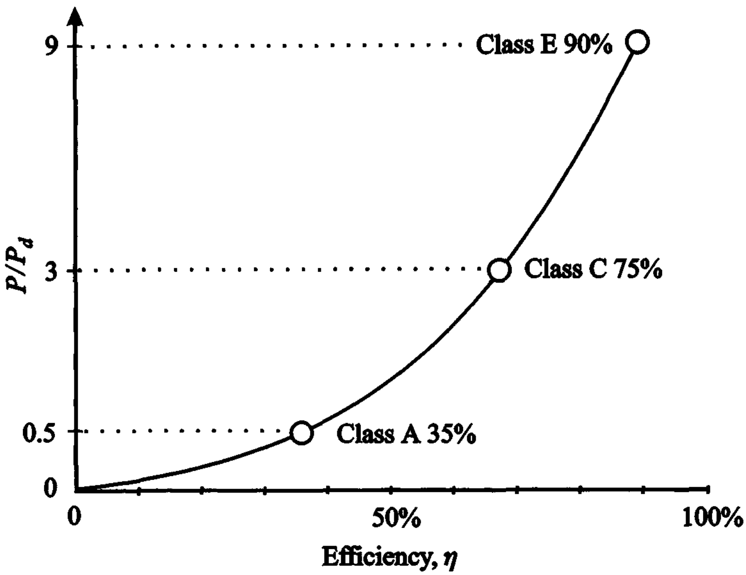



**Figure 10.9.** Output power versus efficiency for amplifiers that are limited by heat.

We can eliminate the input power between these two formulas and get

$$\frac{P}{P_d} = \frac{\eta}{1-\eta}. \tag{10.31}$$

The maximum output power increases greatly at high efficiencies (Figure 10.9). As an example, a large power transistor might be able to dissipate 50 W safely. The maximum power that we would be likely to get out of a Class-A amplifier with this transistor would be 27 W. For a Class-C amplifier, we would expect 150 W, and for a Class E, we could get 450 W.

An interesting Class-E amplifier was developed by Caltech undergraduate students that uses the NorCal 40A as a driver. This amplifier allows the output power to be increased from the few watts available from the NorCal 40A to 500 W. The circuit is shown in Figure 10.10a. It uses a field-effect transistor called a MOSFET. MOS is short for *metal-oxide-semiconductor*. In the circuit, $C_d$ sets the current. For higher current, we would increase $C_d$. $L_s$ and $C_s$ form the resonant circuit. The resonant frequency for $L_s$ and $C_s$ is set somewhat below the operating frequency. $L_2$ and $C_2$ transform the antenna impedance of 50 Ω to about 10 Ω, which is appropriate for this amplifier. In addition, $L_2$ and $C_2$ have a resonance at the second harmonic. This circuit is called a *trap*, and it acts as a band-stop filter to prevent the second harmonic from getting to the output.

The drain voltage is shown in Figure 10.10b. The peak voltage is 400 V, more than ten times larger than the peak voltage in the NorCal 40A. The efficiency is extremely high, about 90%, and this makes it possible to dissipate the power in an aluminum heat sink without a fan.

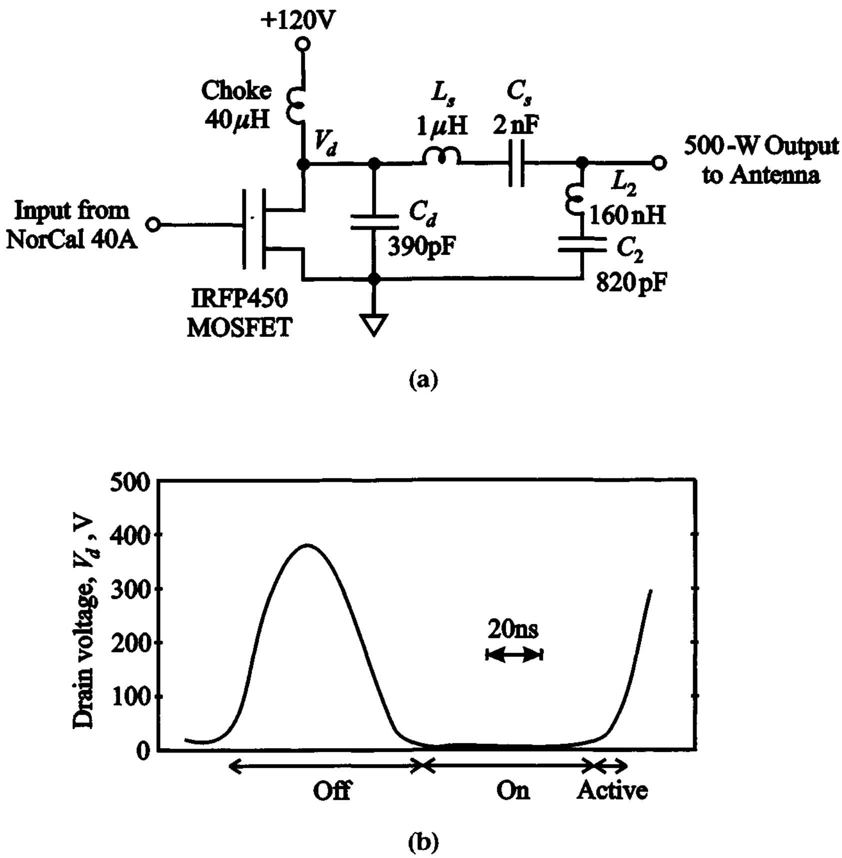



**Figure 10.10.** 500-W Class-E amplifier developed at Caltech. Circuit diagram (a), and drain voltage (b). For more information, see "High-Efficiency Class-E Power Amplifiers," by Eileen Lau, Kai-Wai Chiu, Jeff Qin, John Davis, Kent Potter, and David Rutledge, in *QST* magazine, Part 1, May 1997, pp. 39–42, and Part 2, June 1997, pp. 39–42.

## 10.5 Class F

Class-D amplifiers give a square-wave voltage that is attractive because it keeps the peak voltage low. However, Class-C amplifiers are simpler, because they need only one transistor switch. The Class-F amplifier uses a single switch like the Class C, but it adds a third-harmonic trap to produce a flattened voltage as in Class D. We can think of it as a single-ended Class D. The circuit is shown in Figure 10.11a. $L_3$ and $C_3$ have a resonance at the third harmonic that causes a third-harmonic component to be added to the collector voltage to flatten it (Figure 10.11b).

The efficiency of Class-F amplifiers is typically higher than that of Class C but lower than that of Class E. Which design is appropriate depends on whether the transistor output is limited by collector voltage or heat. It is not easy to generalize about this, because the answer depends on the frequency and power levels. Also, the amplifier may only operate a fraction of the time. The fraction of the time

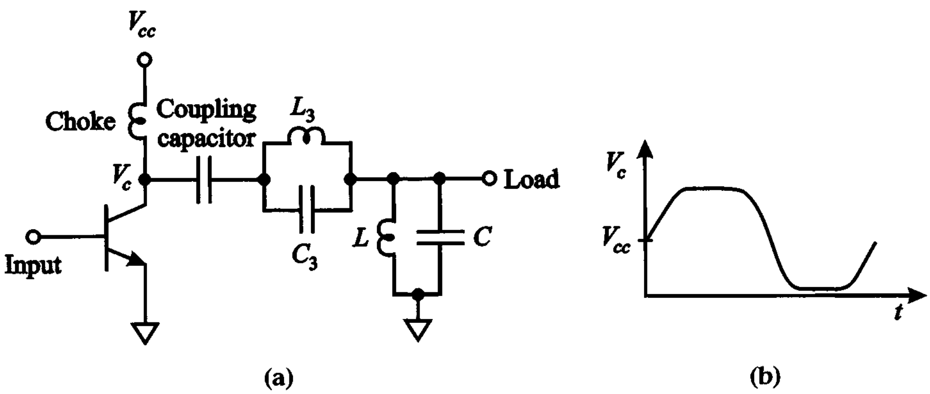



**Figure 10.11.** Class-F amplifier (a), and collector voltage waveform (b).

that an amplifier is on is called the *duty cycle*. A small duty cycle will reduce the heat load, but the peak voltages are not affected. At higher frequencies in the GHz range, the maximum voltages for transistors may only be a few volts, and a Class-F design may be better. However, for higher-power transmitters at lower frequencies, Class E may be appropriate.

## 10.6 Class B

The high-efficiency classes C, D, E, and F all are nonlinear, and this means that they are not suitable for amplifying signals that have both frequency and amplitude changes. In mathematical terms, an amplifier is *linear* if we can relate the output voltage $V$ and the input voltage $V_i$ by a scalar multiplier:

$$V = \alpha V_i. \tag{10.32}$$

Notice that this is different from the definition we use in mathematics, where

$$y = ax + b \tag{10.33}$$

would be considered linear. In Figure 10.12, I plot output voltage against input voltage for the NorCal 40A Power Amplifier in Class C. The relationship is not at all linear, and you can see distortion at both low and high levels.

The curve shows a *threshold*, with very little output when the input is less than 1.2 Vpp. Lower voltages are not large enough to turn on the base–emitter diode. We can eliminate threshold distortion by biasing for Class A. However, Class A is inefficient and there is large DC power consumption even when there is no input signal. However, there is a way to reduce the threshold distortion, and to keep high efficiency, if we bias enough to shift the threshold near zero (Figure 10.12). This is Class B, and it combines efficiency with low threshold distortion. You may see it called Class AB to indicate that there is a bias. There is a tradeoff between the amount of bias power and the reduction in distortion.

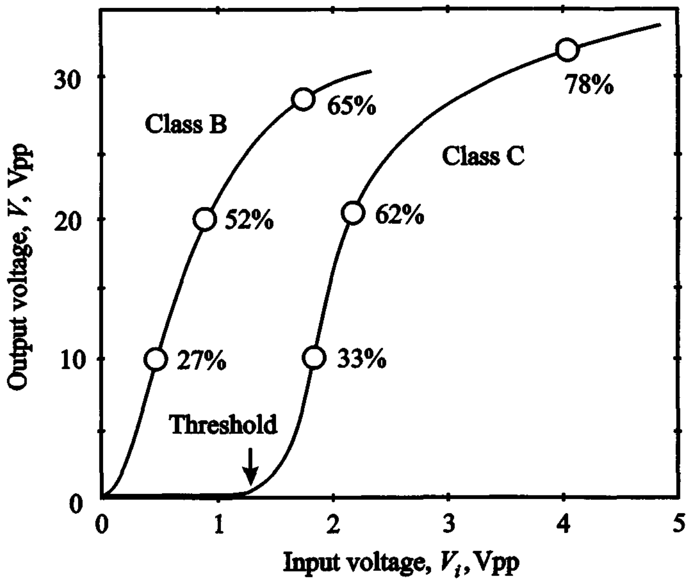



**Figure 10.12.** Input and output voltages for the NorCal 40A Power Amplifier, with the efficiencies noted on the curves. The curve marked Class C is normal operation, while the curve marked Class B is for a base bias of 500 mV (calibrated for 50-Ω source and load). The base bias greatly reduces the threshold so that the amplifier can be used for linear operation. At a given output, the efficiency for Class-B operation is lower than that for Class C.

Class-B amplifiers are more complicated that either Class-A or Class-C amplifiers, but for many applications they represent the best of both worlds. There is high-level distortion in both Class B and Class C where the transistors saturate. The efficiency rises as we push further into saturation, and this tells us that we must trade off distortion for power and efficiency.

Figure 10.13 shows the collector-voltage waveform during Class-B operation. The transistor is active for only half the cycle, but the voltage is approximately sinusoidal because of the ringing of the harmonic filter. Residual threshold distortion can be seen, as well as the onset of saturation.

We can derive a formula for the maximum possible efficiency for a Class-B amplifier by analyzing idealized waveforms (Figure 10.14a). The voltage is a cosine with a maximum of twice the supply voltage and a minimum of zero. The current is a rectified cosine. The DC current $I_o$ can be written as

$$I_o = I_m/\pi, \tag{10.34}$$

where $I_m$ is the maximum value of the rectified cosine. The supply power is

$$P_o = V_{cc}I_o = V_{cc}I_m/\pi. \tag{10.35}$$

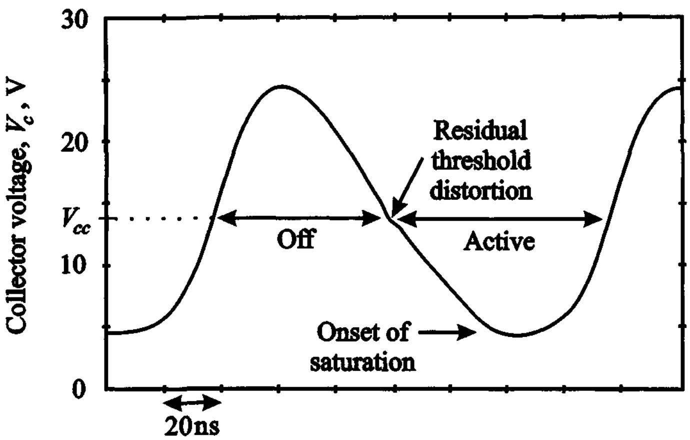



**Figure 10.13.** Collector voltage for the NorCal 40A Power Amplifier in Class-B operation at 7 MHz. The output is 1 W into a 50-Ω load. The base RF voltage is set for 880 mVpp, with a DC offset of 500 mV, calibrated for a 50-Ω source and load.

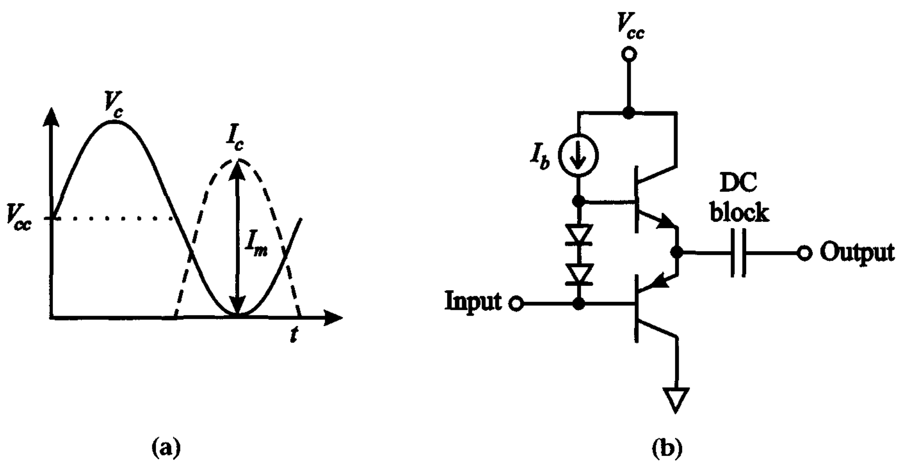



**Figure 10.14.** Idealized collector voltage and current waveforms for a Class-B amplifier (a). A push-pull Class-B audio amplifier (b).

Section B.2 shows that for a rectified cosine, the peak value $I_p$ of the fundamental component of the current is given by

$$I_p = I_m/2 \tag{10.36}$$

so that the output power is

$$P = V_{cc}I_p/2 = V_{cc}I_m/4 \tag{10.37}$$

and the maximum efficiency is given by

$$P/P_o = \pi/4 = 79\%. \tag{10.38}$$

This is the ideal efficiency, and 60% would be typical of practical Class-B amplifiers. This is much better than for Class A, but falls short of Class-C efficiencies.

One problem with a single-ended Class-B amplifier is that the harmonic filter limits the bandwidth. We cannot amplify a signal at a frequency and its harmonic at the same time, because the harmonic will be taken out by the filter. This poses a problem in an audio amplifier, where we would like to amplify over a very wide frequency range. To solve this problem, the Audio Amplifier in the NorCal 40A uses a push–pull Class-B amplifier. In a push–pull amplifier, there are two transistors, and each transistor provides current for half the cycle. This amplifier comprises the final stage of the LM386N-1 integrated circuit. There are two stages of Class-A amplification, followed by a final Class-B power amplifier. A simplified version of the power amplifier is shown in Figure 10.14b. It is an emitter follower. When the input voltage is high, the top npn transistor is active, and the bottom pnp transistor is not. When the input voltage is low, the situation is reversed, and the bottom pnp transistor is active. The two diodes set a voltage difference between the two bases that minimizes threshold distortion.

## 10.7 Thermal Modeling

Controlling the temperature is critical for high-power amplifiers. Manufacturers specify a maximum operating temperature for their devices. This is typically between 150° and 200°C. Because the failure rate for transistors increases dramatically as the temperature increases, it is a good idea to be well below this limit. We will use an electrical analogy to make a mathematical model that relates the dissipated power to the temperature rise. In the analogy, the temperature $T$ corresponds to voltage, and the dissipated power $P_d$ corresponds to current:

$$T \Longleftrightarrow V, \tag{10.39}$$

$$P_d \Longleftrightarrow I. \tag{10.40}$$

We call this a *dual* relationship. In our power amplifier, we clip a piece of metal called a *heat sink* onto the transistor that allows the heat to escape into the air. The power that is lost to the air is proportional to $T - T_0$, where $T$ is the temperature of the heat sink and $T_0$ is the ambient temperature. This means that we can characterize the heat sink by the ratio of $T - T_0$ to the dissipated power $P_d$. This ratio is analogous to resistance, and it is called the *thermal resistance* $R_t$. We write it as

$$R_t = (T - T_0)/P_d. \tag{10.41}$$

The units of thermal resistance are °C/W. In addition to the heat lost to the air, thermal energy is stored in materials as they heat up. The temperature rise is proportional to the stored energy, which is the integral of the power. The integral of power is dual to charge, and thus the ratio of heat energy to temperature rise is like capacitance. It is called the thermal capacitance $C_t$, and we write

$$C_t T' = P_d, \tag{10.42}$$

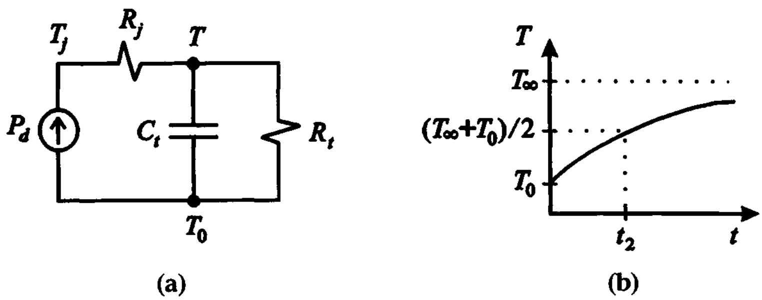



**Figure 10.15.** Thermal model for the transistor and heat sink (a), and plot of heat-sink temperature $T$ versus time (b).

where the prime denotes a time derivative. The units of thermal capacitance are J/°C.

Figure 10.15a shows a thermal circuit model for the transistor and heat sink. We draw a current source for the power, and include the thermal resistance to air and the thermal capacitance. We need to consider an additional effect. There is a large thermal resistance between the transistor and its package, because the transistor is small, and this makes it difficult to get the heat out. It is traditional to write this resistance as $R_j$, where $j$ stands for junction. We can write

$$R_j = (T_j - T)/P_d, \tag{10.43}$$

where $T_j$ is the transistor temperature. $R_j$ is in series with the other components. We can understand this if we consider that the thermal power from the transistor must pass to the package and heat sink before we can consider the effect of thermal capacitance and thermal resistance to air. This means that the transistor will have a greater temperature than the sink.

The equations for this circuit are like the ones for RC circuits we studied in Chapter 2. We start with the first-order differential equation

$$f(t) + \tau f'(t) = 0, \tag{10.44}$$

where $\tau$ is a time constant. This equation has a decaying solution

$$f(t) = f_0 \exp(-t/\tau), \tag{10.45}$$

where $f_0$ is the initial value of $f$. If the right side in Equation 10.44 is zero, the equation is *homogeneous*. We want to solve the *inhomogeneous* equation, where the right side is equal to some value $x$:

$$f(t) + \tau f'(t) = x. \tag{10.46}$$

In the long term, the derivative will approach zero, and this means that $x$ is just the final value for $f(t)$. We relabel $x$ as $f_\infty$ to show this, obtaining

$$f(t) + \tau f'(t) = f_\infty. \tag{10.47}$$

To solve this equation, define a new variable $g$, which is the difference between $f(t)$ and the final value $f_\infty$:

$$g(t) = f(t) - f_\infty. \tag{10.48}$$

The derivative of $g$ is the same as the derivative of $f$:

$$g'(t) = f'(t). \tag{10.49}$$

The equation for $g$ is homogeneous, and it looks the same as Equation 10.44, with $g$s replacing $f$s:

$$g(t) + \tau g'(t) = 0. \tag{10.50}$$

This gives us the solution

$$g(t) = g_0 \exp(-t/\tau), \tag{10.51}$$

or in terms of $f$,

$$f(t) = f_\infty - (f_\infty - f_0) \exp(-t/\tau). \tag{10.52}$$

Now let us apply this approach to find the transistor temperature from the thermal circuit. We start by finding the heat-sink temperature $T$. We can write the dissipated power $P_d$ as a sum of two parts: a resistive component $(T - T_0)/R_t$ and a capacitive component $C_t T'$. Notice that $R_j$ does not affect the heat-sink temperature. We have

$$P_d = \frac{T(t) - T_0}{R_t} + C_t T'(t). \tag{10.53}$$

We can multiply by $R_t$ and rearrange to find

$$T(t) + R_t C_t T'(t) = P_d R_t + T_0. \tag{10.54}$$

This is in the same form as our inhomogeneous equation (Equation 10.46), and we can rewrite it as

$$T(t) + \tau T'(t) = T_\infty, \tag{10.55}$$

where

$$\tau = R_t C_t \tag{10.56}$$

and

$$T_\infty = P_d R_t + T_0. \tag{10.57}$$

The heat-sink temperature is given by

$$T(t) = T_\infty - P_d R_t \exp(-t/\tau). \tag{10.58}$$

The transistor temperature $T_j$ is given by

$$T_j = T(t) + R_j P_d. \tag{10.59}$$

## FURTHER READING

Power amplifiers raise entirely different issues from small-signal amplifiers, and a good place to start is *Radio Frequency Transistors, Principles and Practical Applications*, by Norm Dye and Helge Granberg, published by Butterworth and Heinemann. This book is full of the details of construction and measurements. *Solid State Radio Engineering*, by Herbert Krauss, Charles Bostian, and Frederick Raab, published by Wiley, is an excellent reference. The chapters on power amplifiers were written by Frederick Raab, a pioneer in Class-F power-amplifier development. Raab also gives a list of all the letter classes of amplifiers as they have been defined and redefined in the literature.

*Puff* is a linear circuit simulator, and therefore it cannot handle nonlinear circuits such as power amplifiers and oscillators. The standard program for simulating nonlinear circuits is called SPICE. This program was developed at the University of California at Berkeley and there are many commercial versions. A popular one is PSPICE, and it is available free from the OrCad Company in a demonstration version that is fine for many problems. Consult the company web site at http://www.orcad.com for more information.

## PROBLEM 24 – POWER AMPLIFIER

The Power Amplifier is shown in Figure 10.16. The transistor is the 2N3553, and the manufacturer's data sheet in Appendix D specifies the maximum collector voltage as 40 V.

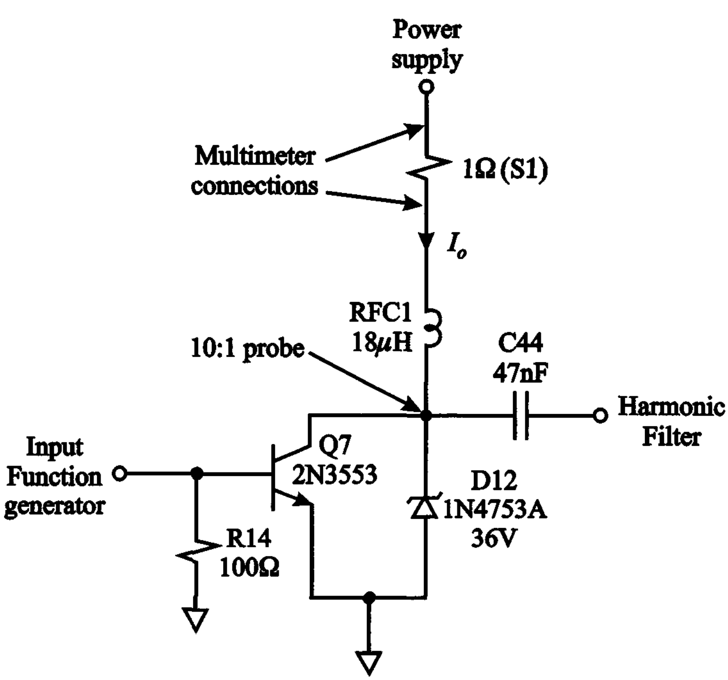



**Figure 10.16.** Power Amplifier.

D12 is a zener diode across the output that conducts at 36 V to prevent excessive collector voltages. R14 is a 100-Ω resistor across the input. This limits the reverse voltage on the base–emitter diode, which should not exceed 4 V.

To start, solder in the transistor Q7. You should slip a plastic spacer over the leads to keep the can from touching the board. The can is connected to the collector, and it will short out the circuit if it touches the base or emitter solder pads. Then install C44, D12, and RFC1 ("RFC" stands for radio-frequency choke). Leave the cathode of the diode high enough above the board that you can get a probe under it. Slip a heat sink onto the transistor for cooling.

We will use the 1-Ω resistor across S1 for monitoring the current. We will need to measure its resistance accurately. There is a problem in measuring small resistance, however, because the test leads add appreciable resistance. Clip the test leads from the multimeter together to measure the lead resistance. Now measure the resistance of the 1-Ω resistor, and subtract the lead resistance. Record this corrected resistance for future reference.

Connect the scope to the Antenna jack J1 and use a 50-Ω termination. The function generator should be connected across R14. It should be set for a 1-Vpp, 7-MHz sine wave, with no offset. In addition, a 10:1 probe should be connected to the cathode of D12 to monitor the collector voltage.

**A.** Calculate the peak-to-peak voltage across the 50-Ω load that is required for an output power of 2 W. Gradually increase the function-generator voltage until the output power is 2 W. Sketch the collector voltage. What is the available power from the function generator? Calculate the gain $G$ in dB. Use the multimeter to measure $V_{CC}$ and record this value. A good place to make a connection for measuring $V_{CC}$ is the choke side of your 1-Ω current-sensing resistor.

**B.** Next make a series of measurements for peak-to-peak output voltages of 5, 10, 15, 20, 25, and 30 V. For each output voltage, use the multimeter to measure the DC voltage across the 1-Ω current-sensing resistor. Calculate the DC supply current $I_o$, subtracting 2 mA for regulator current. Now using your previous measurement of $V_{CC}$, calculate the supply power $P_o$.

**C.** Calculate the output power for each output voltage. Plot the efficiency $\eta$ given by

$$\eta = P/P_o, \tag{10.60}$$

with output power $P$ on the $x$ axis. Now plot the power that is dissipated in the circuit $P_d$, given by

$$P_d = P_o - P, \tag{10.61}$$

with output power $P$ on the $x$ axis.

**D.** At sufficiently high output voltages the efficiency begins to drop. Extend your plots until you can see this drop off.

## PROBLEM 25 – THERMAL MODELING

We will use the thermal model in Figure 10.15a to characterize our amplifier. $R_j$ can be difficult to measure, because the transistor is inside the can. However, since $R_j$ depends only on the characteristics of the transistor and the package, and not on the heat sink, a manufacturer can do this measurement for us. Motorola specifies that for the 2N3553 transistor, $R_j = 25°C/W$.

Thermal time constants are much longer than electrical time constants, and this means that you can take the data by hand as the temperature rises. However, this requires planning, because you cannot hurry things. If something goes wrong during the measurements, you need to wait for things to cool down before you start over. You will be making measurements over a twenty-minute period.

Coat the end of a thermometer with heat-sink compound. This helps to keep the thermal resistance between the thermometer and the heat sink low. Try not to get the compound on anything except the thermometer and the heat sink, because it is difficult to get off.

**A.** Place the thermometer bulb on the heat sink. Some heat sinks have clips with enough tension to hold the thermometer upright. Make sure that the thermometer is oriented so that you can read the temperature easily. Also make sure that you take each reading with your head in the same place to avoid parallax error. Take an initial temperature reading with the power off. This is the ambient temperature $T_0$.

**B.** The connections are the same as in the previous problem, except that you do not need the 10:1 probe. Turn up the function-generator voltage until the output across the 50-Ω scope termination is 30 Vpp. This is an output power of 2.25 W. Take a temperature reading each minute for the first ten minutes. You may need to adjust the function generator from time to time if the output voltage changes.

**C.** At the end of ten minutes, use the multimeter to measure the voltage across the 1-Ω sense resistor and to measure $V_{CC}$. From the multimeter measurements, deduce the power being dissipated by the transistor. You should allow for a 2-mA regulator current.

**D.** Take a final temperature reading at twenty minutes to get $T_\infty$. Use this measurement to calculate $R_t$ and the final value of $T_j$.

**E.** Make a plot of heat-sink temperature versus time, and from the plot calculate the time $t_2$ that it takes for the temperature to come a factor of two closer to the final temperature. A good starting point for this calculation is the temperature after one minute. Often the first minute of a thermal measurement is complicated by the time that it takes to adjust the equipment. Use the measurement of $t_2$ to calculate $C_t$. You should wipe the thermal compound off the thermometer after you are done with it.

Now we consider the entire transmitter amplifier chain from the Buffer Amplifier through the Driver Amplifier and the Power Amplifier. Install C48 (10 nF). This should complete the circuits shown on Figure 10.17. The capacitor C48 is next to the Key jack J3. It has a low impedance at the 7-MHz operating frequency, only about 2 Ω, and its purpose is to keep RF voltages on the Keyline low, so that they do not cause mischief.

We can also understand the effect of the Schottky diode D10. This diode only lets current leave the Driver Amplifier circuit; it does not allow it to flow back in. This prevents other circuits from charging up C56.

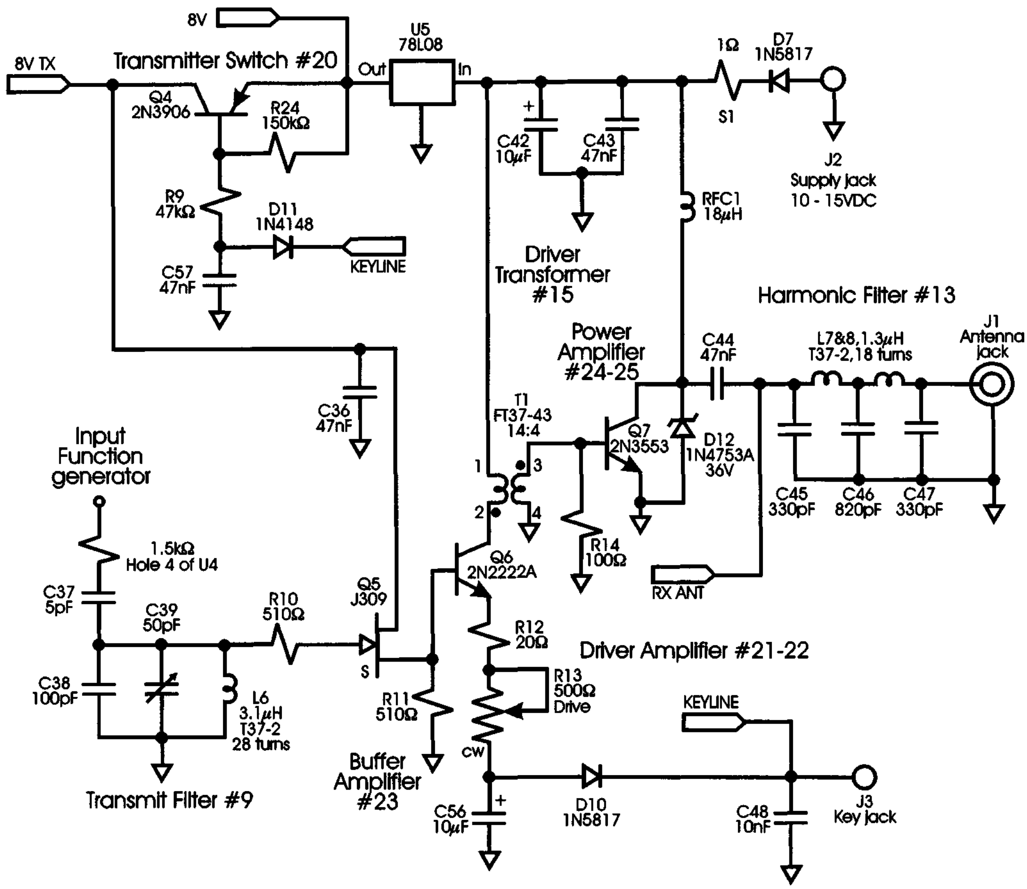



**Figure 10.17.** The transmitter amplifier chain.

So far, when we have tested the Power Amplifier, we have adjusted the function-generator voltage to set the output power. In the actual transmitter, however, the input signal that comes from the Transmit Mixer is at a fixed level, and the output is set with the Drive pot R13, which controls the gain of the Driver Amplifier.

To start, plug in the power supply, attach leads for the function generator, and connect the Antenna output to the scope with a 50-Ω termination. Insert a shorting plug into the Key jack J3. Set the function generator to a frequency of 7 MHz and an amplitude of 600 mVpp. Check that C39 is tuned for maximum output, and set the Drive pot R13 to give an output of 2 W. You should leave the pot in this position for the rest of the lab.

Now we investigate the bias for the Drive Amplifier. You will need to turn off the signal from the function generator. The bias circuit is simple and elegant. It comes directly from the JFET Buffer Amplifier. The resistor R11 (510 Ω) controls the bias for both the JFET Buffer and the BJT Driver. The value of R11 is chosen by balancing off the bias currents of the Buffer and the Driver. Making the resistance larger reduces the JFET bias current but increases the BJT current.

**F.** Find the emitter resistance (R12 and R13 together). In making an ohmmeter reading, you will need to pull out the shorting plug. Make sure you put the shorting plug back after you take the reading, or else the amplifiers will have no bias. Use a multimeter to find the following: the base–emitter diode drop for Q6, the drop across R12 and R13 together, and the diode drop for the Schottky diode D10. What is the emitter bias current $I_e$?

**G.** You should notice that the Schottky diode has a lower forward voltage than a pn diode. A Schottky diode is made as a contact between metal and silicon. Usually the silicon will be n-type, and many metals can be used. Gold and platinum are common. What would $I_e$ become if the designer had used a pn diode for D10 instead of a Schottky? You may assume that a pn diode has a forward voltage of 0.6 V. What problem would this new emitter current cause?

The Driver capacitor C56 is important in turning the transmitter signal off. Transmitters should not be turned off suddenly, because the sudden change creates spurious frequency components that are heard by operators on other frequencies as annoying clicks. Typically transmitters are designed to turn off gradually over a period of several milliseconds. The same consideration applies in turning a transmitter on. This timing is controlled by the Transmit Mixer, and we will look at it later.

When the Key jack J3 is short-circuited (key down), the emitter bias current for the Driver flows through the emitter resistors, and out through the Schottky diode D10. The Driver capacitor C56 is charged to the forward voltage of the diode. When the Key jack is open-circuited (key up), the current through D10 stops. This leaves C56 charging through the emitter resistance. The emitter bias current decays exponentially. As the emitter current drops, this reduces the peak-to-peak current that the amplifier can produce, giving the gradual reduction in output power that we need.

**H.** Measure the time $t_2$ that it takes the emitter current to drop by a factor of 2, and compare it with theory. To do this measurement, you will need to cycle repeatedly through key up and key down. Use a keying relay cable for this. You should set the frequency to 20 Hz.

**I.** Finally, we will look at the gain of the entire amplifier and filter chain from the Bandpass Filter through the Buffer, Driver, Power Amplifier, and Harmonic Filter. To start, the function generator should be set to an amplitude of 250 mV and a frequency of 7.03 MHz. This is a little higher than the frequency in previous measurements, and it is the actual frequency that we will use for the transceiver tests. The gain of the Driver Amplifier should be set to its maximum value, with the R13 pot set fully clockwise. Adjust the filter capacitor C39 for maximum output voltage. Vary the frequency a few kilohertz on each side to confirm that the peak is within 10 kHz of 7,030 kHz. Now adjust the amplitude setting on the function generator until the output power $P$ is 2 W. Calculate the power available $P_+$ from the function generator with the 1.5-kΩ resistor. Calculate the gain in dB of the entire chain. Find the 3-dB bandwidth. You can remove the 1.5-kΩ resistor from hole 4 of U4 when the measurements are completed.