Class E Power Amplifiers for QRP

David Cripe NMØS
Mount Vernon IA
May 14, 2009

ABSTRACT
A Class-E power amplifier is an RF switching power amplifier capable of near 100% DC-to-RF conversion efficiency. These are simple, robust circuits which are worth considering of inclusion within QRP transmitters. This presentation covers the theory of operation, a practical, cookbook approach to design, and comparison with more conventional Class-B power amplifiers.

INTRODUCTION
Much of the appeal of the QRP hobby lays in the simplicity and economy of the QRP radio equipment. The popular 'Pixie' and 'Tuna Tin' kits, for example, have exemplified this concept of reducing amateur communication equipment to a minimal state. Both concepts of simplicity and economy are part of a larger notion of *efficiency*. Efficiency of design yields a simple radio, with few, and inexpensive components. Efficiency of power allows the fewest watts to effectively provide communication from one station to another. To take full advantage of the low-power aspect of the QRP hobby, attention should also be paid to maximizing the electrical efficiency of the output amplifier.

The majority of amateur radio transmitters have been designed with final power amplifiers set up to operate in Class-B or AB, which typically yield DC to RF efficiencies of at most 50 or 60%, and often well below that. Even a good Class-C amplifier, optimized for CW operation, is hard pressed to reach 80% efficiency. In contrast, by utilizing a final power amplifier operating in *Class-E*, it is easily possible to obtain DC-to-RF efficiency exceeding 90%.

The term 'Class-E' may not be familiar to most hams, but it is a technology that has been around for over thirty years. It is a type of switching power amplifier invented in the mid 1970s by Nathan Sokal, WA1HQC and his son Alan Sokal(1). It has gradually come to achieve widespread application, and is today used in areas ranging from broadcast transmitters, to cell phones, RF induction heating equipment, and others.

There have been relatively few amateur radio products or projects that have taken advantage of the benefits Class-E power amplifiers have to offer. Perhaps the most well-known amateur radio product utilizing a Class-E power amplifier was the SGC 'Mini-Lini', a cigar-box sized, 500-watt Class-E amplifier (2). One ham who has done much toward popularizing Class-E circuitry is Steve Cloutier WA1QIX, through his website www.classeradio.com (3). Most of the designs published on his site are elaborate multi-hundred watt QRO transmitters, which might not hold much appeal to the QRP enthusiast. The benefits of Class-E power amplifiers are not only limited to high-power applications. The author's NS-40 transmitter (4) is a 5 watt, 40 meter Class-E transmitter, one of few QRP Class-E products currently available.

The purpose of this presentation will be to describe simple, low-powered circuits that can exploit the advantages of Class-E power amplifiers for the QRP enthusiast.

(INSERT FIGURE 1)

THE CLASS-E POWER AMPLIFIER

At first glance the Class-E power amplifier circuit, shown in Figure 1, does not appear to be significantly different from other more-familiar classes of RF power amplifier. It contains a transistor, tuned output circuit, and output harmonic filtering, just as does a Class-B or -C power amplifier. The distinction in the Class-E circuit lies in the specific values of the inductors, capacitors, and load impedance surrounding the transistor, and the specific transistor voltage and current waveforms it produces, allowing it to operate most efficiently.

(INSERT FIGURE 2)

Sokal (1) defines a Class-E power amplifier as possessing waveforms with the following characteristics:

- Saturated transistor conduction occurring over 180 degrees of the RF cycle.
- A resonant waveform at the transistor drain during the interval of transistor non-conduction, that swings to zero volts immediately prior to the commencement of transistor conduction.
- Drain current that begins at zero amperes during the half-cycle of transistor conduction.
- A drain waveform with a zero voltage-rate-of-change immediately prior to the commencement of transistor conduction.

These conditions are fulfilled a circuit that generates the specific set of waveforms that are unique to Class-E power amplifiers, which are depicted in Figure 2. For this reason, it is common to tune and optimize a Class-E power amp by observing its drain waveform on a high-speed oscilloscope, which can be compared to the ideal waveforms. By observing the specific manner in which the sample waveforms depart from the ideal, it is possible to determine the specific circuit components that must be adjusted.

Unlike a linear power amplifier such as Class-B, the output power of switching power amplifier classes such as Class-E is relatively unaffected by amplitude of the drive signal. Instead, if the output power of Class-E amplifiers is to be controlled, it is done so by varying the supply voltage.

EFFICIENCY

The power dissipated in a given power amplifier device is the product of voltage across it, and current passing through it, integrated over an RF cycle. A switch is capable of controlling passage of electrical power very efficiently as it acts either as a very small resistance, or a very large one. In its 'off' state, a switch supports voltage while allowing no current flow. In its 'on' state, a switch permits current flow through a low resistance. The power loss in the switch in either the 'on' or 'off' conditions are minimized. If the power amplifier active device behaves as a switch it has the potential to convert power at near 100% efficiency.

One loss mechanism associated with switching circuits in general is the result of the discharge of any capacitance associated with the switch at the beginning of its conduction cycle. The energy stored in this capacitance C, when charged to voltage V, is $\frac{1}{2} V^2 \cdot C$. If this capacitor is charged and discharged through the switch at frequency F, the power loss in the switch will be $\frac{1}{2} V^2 \cdot C \cdot F$. Consequently, it is essential that the voltage across the switch and its capacitance be zero at the commencement of switch conduction.

Since the active device in a Class-E power amplifier is utilized as a switch, its conduction losses are minimized. The characteristic drain voltage waveform allows zero-voltage switching, which contributes to its efficiency by the reduction of switching losses.

It is not difficult to design Class-E power amplifiers for HF with efficiency surpassing 90%. Per watt of output, a Class-E transmitter will have one-third to one-fifth the transistor dissipation as one using the same amplifier transistor operating in Class-B or Class-C. The implication of this is that the Class-E power amplifier will use less power, run cooler and more reliably, and can be constructed with fewer or smaller output transistors. The results of this are simpler, more economical transmitter circuits which are ideally suited for QRP. For a battery-powered station, the 40% reduction in power consumption-per-watt of transmitted power translates to a significant improvement in battery life.

(INSERT FIGURE 3)

DESIGN OF CLASS-E POWER AMPLIFIERS

It is relatively easy to design a Class-E PA if a cookbook approach is followed. Sokal published the essential relationships to determine component values for a given frequency, power, and input voltage in his original patent and his comprehensive QEX article (6). Simplified equations can suffice, though, for typical cases, with sufficient accuracy for the hobbyist.

Figure 3 depicts a practical QRP transmitter circuit, a modification of Sokal's basic circuit to include an additional circuit at the output for impedance transformation and improved harmonic attenuation. To determine the component values of the circuit, the desired power output P is defined, and the power supply voltage, B, is selected.

The power amplifier MOSFET can be chosen next based on the power output and supply voltage. Bear in mind that in a Class-E power amplifier, the peak drain voltage swings to 3.5 times that of the supply voltage, and higher when operating into non-unity VSWR. For a transmitter operated from a 12 or 13.6 volt power supply, a 60 volt-rated part is adequate if VSWR is kept below 2:1. A MOSFET rated to 100 volts V_{ds} operating from 13.6 VDC will not likely experience VSWR-related voltage breakdown.

There are two readily available, inexpensive MOSFETs that will serve the majority of 12-volt QRP applications. The first is the 2N7000, a plastic TO-92 part, with 2 ohms R_{ds}, and 20 pF C_{oss} output capacitance. The second is the IRF510, a TO-220 with 0.54 ohms R_{ds}, and 81 pF C_{oss} output capacitance. The 2N7000 will serve well up to 1 watt output. A two watt transmitter can be constructed of two, paralleled 2N7000s. The IRF510 will serve well at two to ten watts output. Both parts will work well up to 21 MHz.

Another MOSFET worth considering in QRP service is the 2N6661. It is a MOSFET with similar ratings to a 2N7000, but with an upgraded 90 volt drain breakdown voltage rating, housed in a TO-39 package. Because this metal can package can be effectively heat-sunk, this part will be preferable to the 2N7000 where high reliability is required. These benefits are somewhat offset by its higher price tag, in the $3 range.

Once the desired output power P has been selected, the load resistance R for the PA can be calculated from the formula:

$$R = 0.32 \cdot B^2 / P - 1.5 \cdot R_{ds}$$

For purposes of example, let us design a 5 watt PA, operating from 13.6 volts, for 40 meters. This would require a load resistance of 11.0 ohms.

Then the total PA shunt capacitance C_1 is determined. This value includes the intrinsic C_{oss} capacitance of the MOSFET. This is given by the formula:

$$C_1 = 0.19 / (2 \pi \cdot F \cdot R)$$

Our 5 watt, 40 meter PA would then require 385pF of total drain capacitance. When using an IRF510, which has a C_{oss} (drain-source) capacitance of 81 pF, this yields a value for C1 of 304 pF. We may use two, 150 pF capacitors paralleled. In practice, the physical layout of the circuit requires that this capacitance must be mounted as close as possible to the body of the MOSFET – even directly bridging the drain and source leads, if possible. Excessive inductance between the MOSFET and this capacitor will ring during device switching, leading possible loss of device efficiency or instability.

The series L_2 – C_2 network between the MOSFET drain serves to attenuate the drain waveform harmonics, and pass a largely sinusoidal current. The value of the capacitor C_2 should yield a reactance with a magnitude somewhere between one to two times that of the load resistance R, preferably selected to be a common standard value of capacitor. For our sample circuit, we select C2 to equal 1000pF.

The series L_2 – C_2 network has a further requirement that its net reactance have an inductive component with a magnitude equal to 1.8 · R. This means that the value of L_2 is determined by the formula:

$$L_2 = [1.8 \cdot R + 1 / (2 \pi \cdot F \cdot C_2)] / (2 \pi \cdot F)$$

For our sample circuit, L2 will be 0.94 uH.

From here, it is necessary to match the load impedance R of the power amplifier to 50 ohms. While transformers are the most frequently used solution, if the load impedance R of the PA is between about 10 and 250 ohms, the easiest way to perform an impedance transformation is with a Pi network. This is comprised of C_3, C_5, and L_3. Capacitor C_4 parallels L_3 to resonate at the second harmonic of the operating frequency, providing the extra attenuation required to meet FCC limits on harmonic emission.

The value of C_3 and C_5 are given by:

$$C_3 = C_5 = 1 / (2 \pi \cdot F \cdot (R \cdot 50)^{0.5})$$

This yields a value for C_3 and C_5 of 950 pF. This may be approximated by paralleling two 470 pF capacitors.

The value of L_3 is given by

$$L_3 = 0.75 \cdot (R \cdot 50)^{0.5} / (2 \pi \cdot F \cdot R)$$

This gives a value of 0.38 uH for L_3.

The value of C_4 is given by:

$$C_4 = C_3 / 3$$

This gives a value of C_4 as 317 pF. We may use a 100 pF capacitor paralleled with a 220 pF.

Finally, the value of L_1 is chosen so that its impedance is high compared to the load resistance. It should be greater than ten times the value of L_2, and capable of handling the DC current draw of the power amplifier without saturation.

Wherever possible, the values for C_1, C_2 and L_2 should be within 10% of the ideal values, while C_3, C_4, C_5 and L_3 should be within 5%. Pay attention to the current rating of the components during their selection process. Generally, use of silver mica capacitors throughout the output network will prove adequate. RF- or microwave-rated chip capacitors are also preferred. Check the manufacturers' data sheets for current ratings during the design process.

As the inductors L_2 and L_3 are non-standard values, these should be wound on iron-powder toroids, and the turns spacing adjusted as necessary to attain the desired values. The power output may be fine-tuned to some degree by adjusting L_2.

Once these component values have been determined, they may be used to create a simulation to verify the circuit power output, tuning, and efficiency. Analytic tools such as PSPICE and similar circuit simulation programs have taken much of the cut-and-try iteration out of PA design. One excellent freeware circuit simulation program is SWCAD III, available for download from www.linear.com (5). Intended for use in design and optimization of switch-mode power supply circuitry, this software includes electrical models for many common MOSFETs and bipolar transistors familiar to the QRP community. Incremental changes to the components in the power amplifier may be made to bring the output power and drain waveform shape closer to design goal.

HARMONICS AND OUTPUT FILTERING

All power amplifiers possess some degree of non-linearity which results in some degree of harmonic generation in the process of amplifying a sinusoidal carrier wave. The Class-E drain waveform, being non-sinusoidal, contains a high harmonic content, depicted in Figure 4. Operating courtesy demands that out-of-spectrum emissions be limited, while FCC specifies all harmonics are to be -45 dB with respect to the carrier. Examining the spectral plot, it is apparent that the second harmonic, at -6 dBc, is closest in both frequency and amplitude to the carrier, so is most difficult to attenuate.

Generally, the most effective approach is to add a second harmonic notch in the output filter of the transmitter, and in a circuit with a minimum number of inductors, can be formed by adding a capacitor in parallel with the choke in the output Pi network. The output network described in the cookbook design circuit of Figure 3, consisting of a series L-C bandpass, followed by a Pi-lowpass network with a second harmonic notch will attenuate harmonics sufficiently to meet the FCC requirements. In the absence of a spectrum analyzer, an estimate of the harmonic output of the transmitter can easily done by modeling the entire amplifier using the Fast-Fourier Transform capability of SWCAD III.

THERMAL CONSIDERATIONS

The primary failure mode for MOSFETs operating in a switching mode is from excessive temperature at the semiconductor junction. The worst-case effect of operating a Class-E power amplifier into a poorly-matched antenna impedance is a decrease of efficiency leading to increased heat dissipation in the MOSFET. To be reliable, the power amplifier must be designed to maintain its transistor at a safe temperature.

A very conservative rule of thumb for reliable amplifier design is to keep the semiconductor die to below 100 degrees C. Although most MOSFETs are rated for operation at temperatures up to 150 or 175 degrees C, the 'on' resistance of most MOSFETs will increase 60% between 25 and 100 degrees C. A Class-E power amp operating at 90% efficiency at room temperature will see its efficiency reduced to roughly 84% if the transistor reaches 100 degrees.

It is possible to calculate the temperature rise of the device knowing the heat dissipation, the thermal resistance of the power device, published in that manufacturer's data sheet, and that of the heat sink used, if any. The thermal resistance of the 2N7000 in its plastic TO-92 package is 312 degrees C-per-watt. Allowable heat dissipation for this part would be around 0.25 watts. The thermal resistance of the IRF510 in its TO-220 package, without a heat sink, is 62 degrees C per watt, safely allowing over a watt of dissipation.

Addition of a heat sink to a TO-220 can reduce the total thermal resistance of the part significantly. The common, folded aluminum sink of the type sold at Radio Shack will reduce the thermal resistance of a TO-220 transistor to about 20 degrees C per watt, further increasing its allowable heat dissipation.

DRIVE REQUIREMENTS

As a MOSFET is a voltage-controlled device, driving a Class-E power amplifier is a matter of applying enough RF voltage to the MOSFET gate to drive the device fully into saturation. What complicates this it that the gate driver circuit must charge and discharge the large MOSFET gate-source capacitance with each RF cycle. The upper frequency at which a particular driver circuit may be used will be inversely proportional to the product of driver output impedance and the MOSFET gate capacitance. There quite a number of commercial ICs intended for driving large MOSFETs in switching power applications, most of which do not function well at HF.

Fortunately, there is a simple solution that lends itself well to this application. Common 74HC-family logic gates can supply sufficient current to function effectively in this application. A single 74HC gate can directly drive a single 2N7000 at up to 14 MHz, or two paralleled 2N7000 MOSFETs at up to 7 MHz. Alternately, two paralleled 74HC gates can be used to drive a single IRF510 at up to 7 MHz, and so on. Another less common logic family is the 74AC family which has approximately one-third of the output impedance of the 74HC family. Although not available from every parts supplier, they are worthy of consideration.

While a zero-to-five volt logic signal is usually adequate to drive most MOSFETs, there is an additional trick that can further improve the effectiveness of the digital logic gate as a drive amplifier. Since the turn-on threshold of most MOSFETs is in the 3 to 4 volt range, the MOSFET can be driven harder into conduction by shifting the 0-to-5 output voltage of the logic gate upward by 1.5 volts to provide a 1.5-to-6.5 volt swing. This can be done by capacitively coupling the output of the logic gate to the gate of the MOSFET, and then resistively biasing the MOSFET gate up to 4 volts. Figure 5 depicts a circuit that serves this function. The 'T/r' input signal is a TTL-level signal that goes to 5 volts during CW transmission, providing this bias to the MOSFET gate. By switching this bias voltage at the beginning of each CW element, and back down just prior to its end, the CW element can be shaped to soften the rise and fall times, reducing spectrum bandwidth of the transmitted signal.

(INSERT FIGURE 5)

COMPARISON WITH CLASS-B POWER AMPLIFIER

With our power amplifier designed, it is worthwhile to compare it with a Class-B power amplifier also using the IRF510 MOSFET, at the same power output. With a circuit simulation program such as SWCAD III, it is easy to perform a head-to-head comparison quickly, and obtain quantitative data for evaluation of their respective merits.

A sample Class-B circuit was designed for comparison with a Class-E circuit. As with the sample design produced from the cookbook equations, which also used an IRF510, and delivered 5 watts from 13.6 volts. It used a resonant drain circuit with a Q of 5 to generate a sinusoidal output waveform, and to attenuate output harmonics.

Although both circuits are set up to deliver equal output power, the most significant difference between the two circuits was the power dissipated in the MOSFET. The MOSFET in the Class-E power amplifier dissipates 0.34 watts. The Class-B power amplifier transistor, in contrast, dissipates 3.4 watts. This significant different in the transistor dissipation between the two circuits results in higher efficiency, cooler junction temperatures, reduced heat sink size, and improved reliability of the Class-E power amplifier.

(INSERT FIGURE 6)

It is also worth examining the performance of the two power amplifier circuits when operating into non-ideal loads. The load impedances of the SWCAD III circuits were modified to present 2:1 VSWRs, and varied at 45 degree intervals around the Smith chart VSWR circle. The resulting variations in output power of the Class-B and -E circuits are plotted in Figure 6. It is notable that the power varies more significantly with load changes with the Class-E circuit than with the Class-B circuit. As the Class-E power amplifier can be considered a low-impedance RF voltage source, its output current will vary inversely with the load resistance. The Class-B circuit, on the other hand, exhibits less variation under load mismatch, as its drain current is largely defined by the gate drive voltage.

(INSERT FIGURE 7)

Next, an examination of transistor dissipation as a function of phase angle under 2:1 VSWR conditions is performed. The results are compiled in Figure 7. Within the 2:1 VSWR circle, the transistor dissipation of the Class-B amplifier varies from a low of 2.8 watts to a high of 5.5 watts. In contrast, the Class-E power amplifier transistor has transistor dissipation that ranges between a low of 0.15 w and a high of 0.87 w for all 2:1 VSWR loads. It can be seen that for 2:1 VSWR, the worst case transistor dissipation for the Class-E PA is still well less than the best case dissipation for the Class-B power amplifier, and at all times the Class-B dissipation is at least four times higher than that of the Class-E transistor dissipation.

The roughly three-fold increase in transistor dissipation that the Class-E MOSFET experiences at worst-case 2:1 VSWR conditions should be taken into account by the designer. While the junction temperature of a non-heatsunk IRF510 will remain below 100 degrees C at this dissipation, a heat sink should be added if it is anticipated that more challenging antenna impedances might be presented.

(INSERT FIGURE 8)

A comparison of DC-to-RF conversion efficiencies of the Class-B and -E power amplifiers is plotted in Figure 8. Despite the varying VSWR angles, the efficiency of the Class-E power amplifier remains near 90%, while the Class-B power amplifier efficiency varies from 38 to 55 percent.

(INSERT FIGURE 9)

The designer of the QRP transmitter should take into account the effect that non-unity VSWR can have on the peak drain voltage of the MOSFET. Figure 9 depicts the peak MOSFET drain-source voltage occurring on a Class-E amplifier operating from 13.6 volts, over all 2:1 VSWR angles. Normally, the peak voltage will be approximately 46 volts, 3.4 times the supply voltage. The worst-case peak voltage at 2:1 VSWR will be 54 volts. While this is a safe voltage for a 100 volt-rated IRF510, this is close to the 60 volt rating of the 2N7000.

FURTHER TOPICS

It is possible to modify the Class-E power amplifier to operate in a push-pull mode. The benefit of this is that the second harmonic is largely canceled, eliminating the need for a second harmonic notch in the output filter. Additionally, the value of the drain choke L1 may be reduced by a factor of four compared to that of the single-switch circuit. Design equations for a push-pull amplifier are contained in the reference (7).

As noted previously, the Class-E power amplifier can not by itself be used to linearly amplify an RF signal. They can and do find frequent application in amplitude-modulated transmitters. The Class-E ham transmitters described at the WA1QIX's www.classeradio.com website use switch-mode, Class-S modulators to provide an audio-modulated voltage supply to produce an AM signal. To utilize a Class-E power amplifier to amplify SSB signals, a technique of Envelope-Elimination-and-Restoration (EER) is used. Invented by Leonard Kahn (8), EER is a technique where a switching power amplifier us used to amplify the instantaneous phase information contained in the SSB signal. A separate audio-frequency amplifier is used to amplify the envelope waveform, which is then supplied to the RF power amplifier as a modulated drain voltage supply. The SSB waveform is recreated by the multiplicative action that takes place in the switching power amplifier. With the technique of EER, it is possible to amplify a complex, SSB signal at high efficiency, usually exceeding 80%. The SGC Mini-Lini (2) 'linear' amplifier utilized EER in conjunction with a Class-E RF power amplifier.

REFERENCES

(1) Sokal, Nathan and Sokal, Alan US3919656, 'High Efficiency Tuned Switching Power Amplifier'
(2) The Mini-Lini, http://www.sgcworld.com/500watt.html
(3) Class-E Radio, www.classeradio.com
(4) NS-40, http://www.wa0dx.org/wa0itp/ns40.html
(5) SwitcherCAD III, Linear Technology Corp, www.linear.com
(6) Sokal, Nathan 'Class-E RF Power Amplifiers', *QEX*, Jan/Feb 2001, pp 9-20
(7) Cripe, David US5327337, 'Resonant Push-Pull Switching Power Amplifier'
(8) Kahn, Leonard 'Single-Sideband Transmission by Envelope-Elimination and Restoration', *Proc. I.R.E*, Vol. 40, pp 803-806, July 1952.

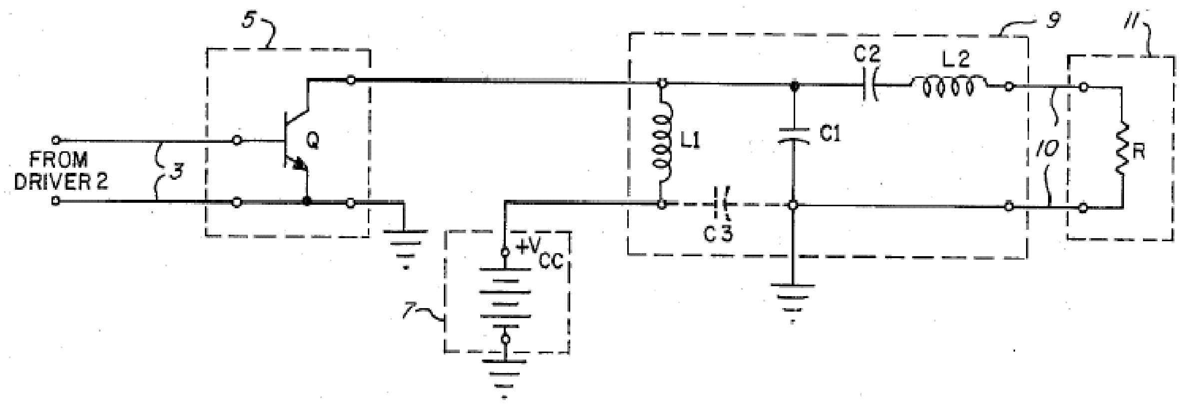

Figure 1: Basic Class-E Circuit (1)

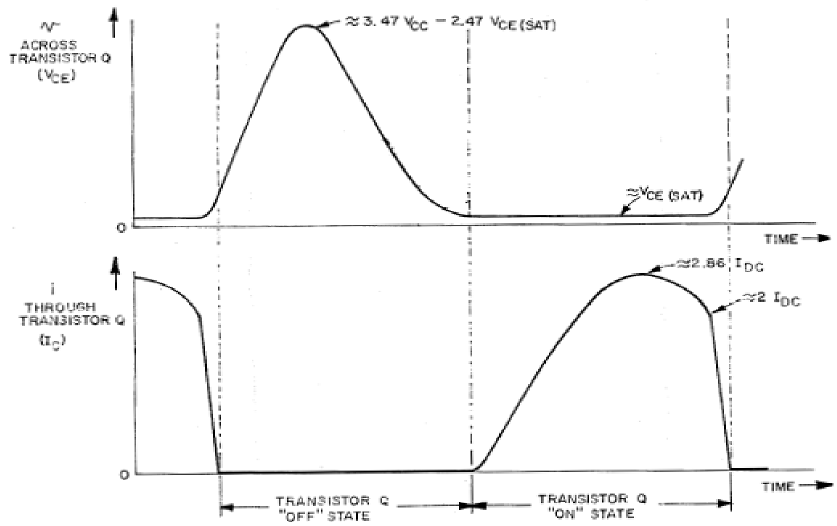

Figure 2: Characteristic Class-E Waveforms (1)

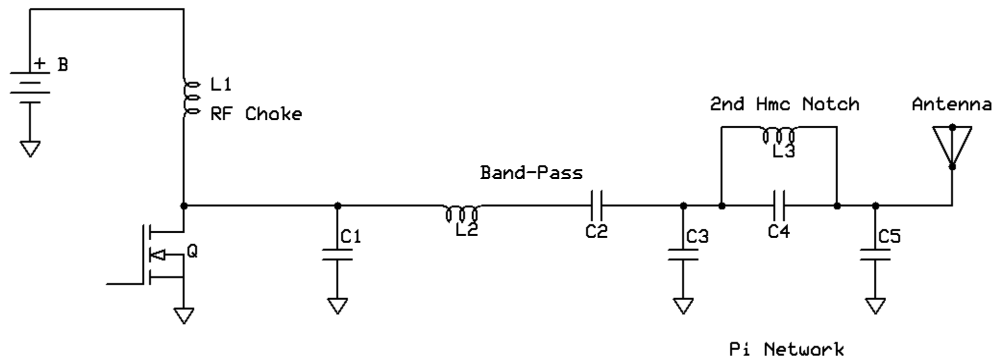

Figure 3: Practical Class-E Power Amplifier

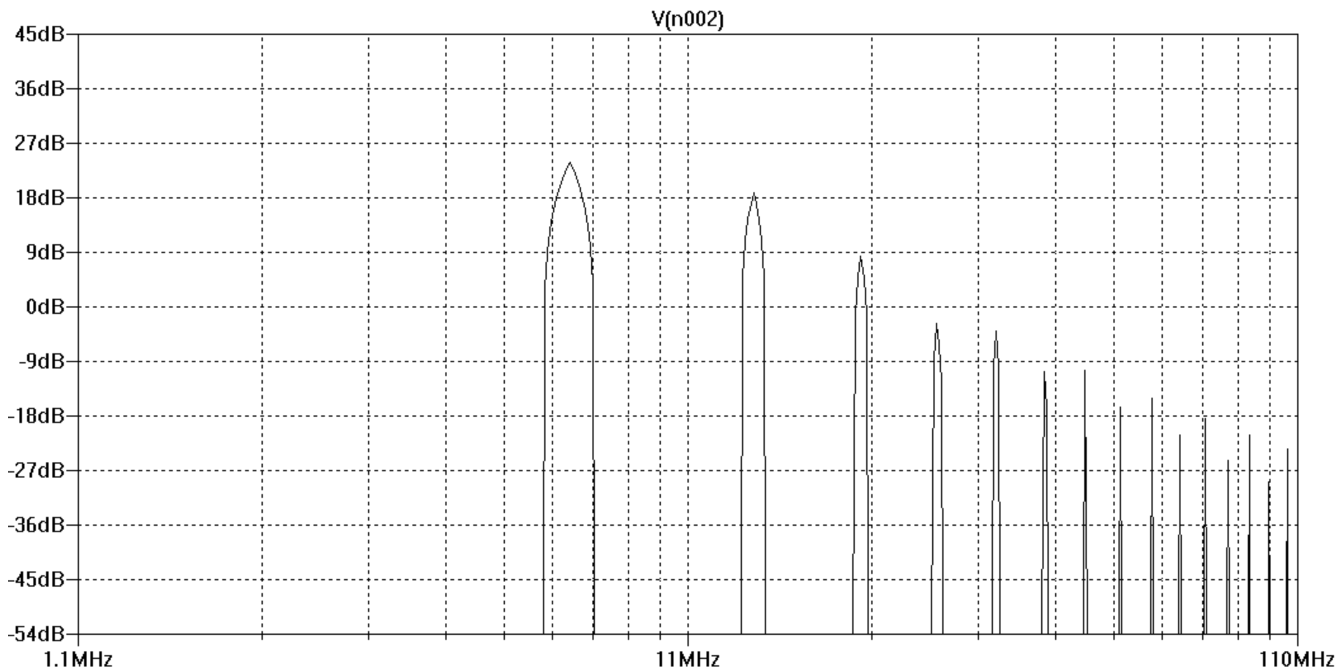

Figure 4: Harmonic Content of the Class-E Drain Waveform

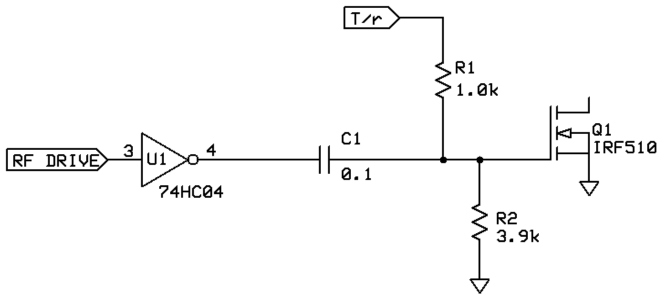

Figure 5: Drive Circuit for Class-E PA

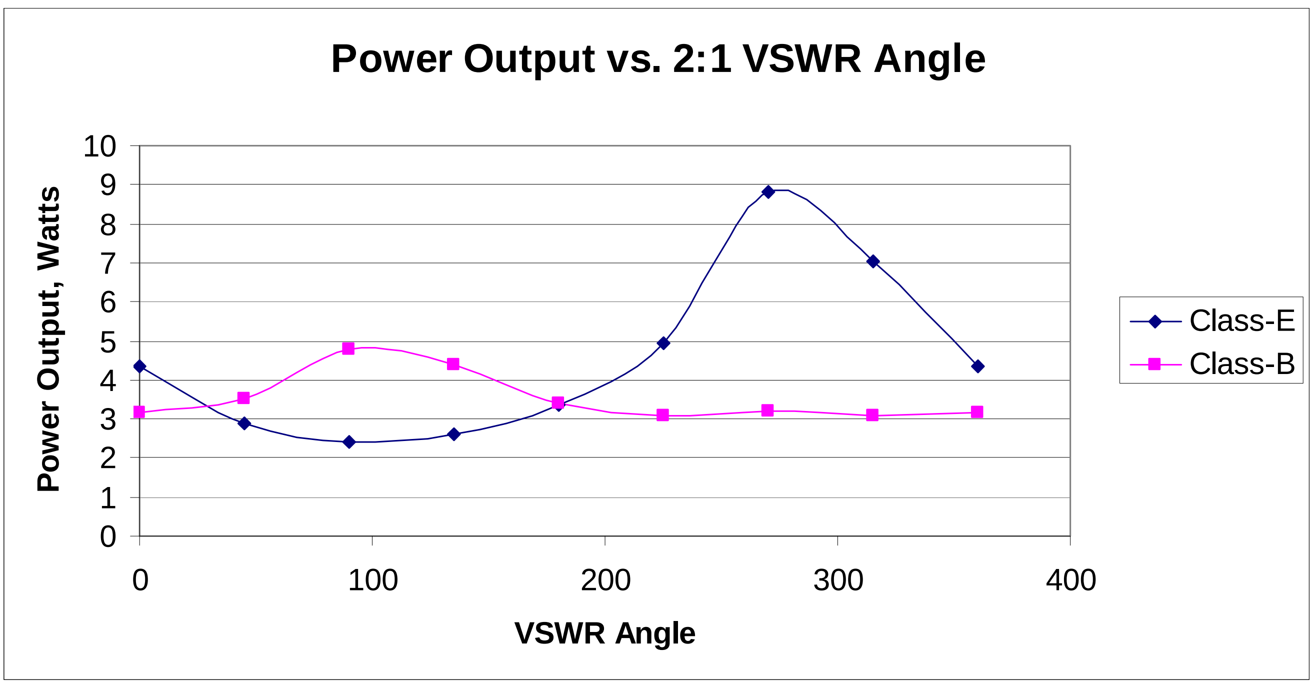

Figure 6: Power Output of Class-B and E PA at 2:1 VSWR

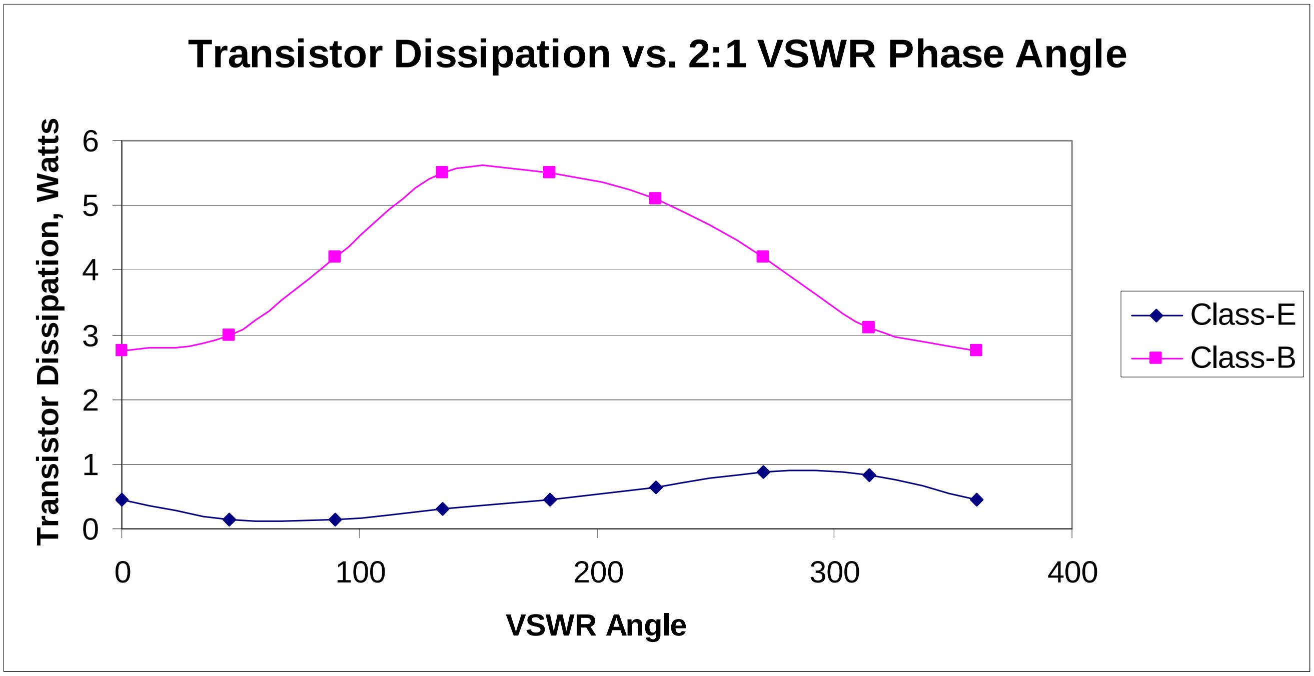

Figure 7: Transistor Dissipation of Class-B and -E PA at 2:1 VSWR

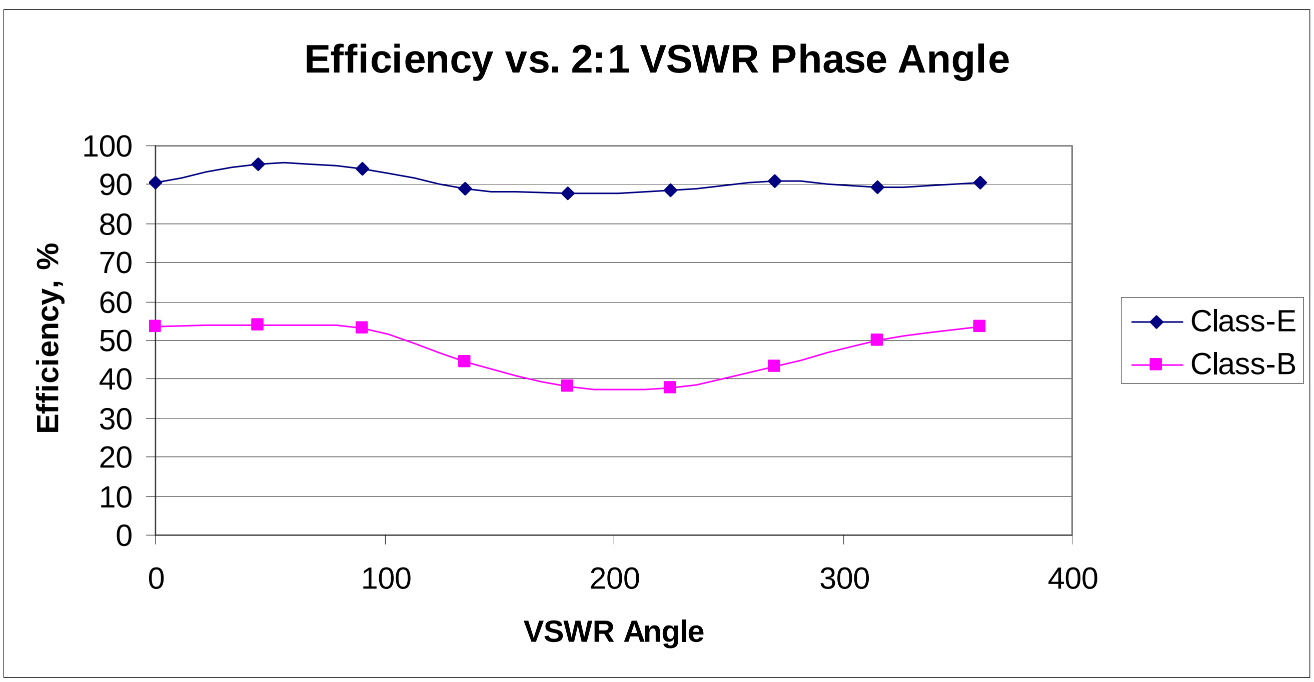

Figure 8: Efficiency of Class-B and -E PA at 2:1 VSWR

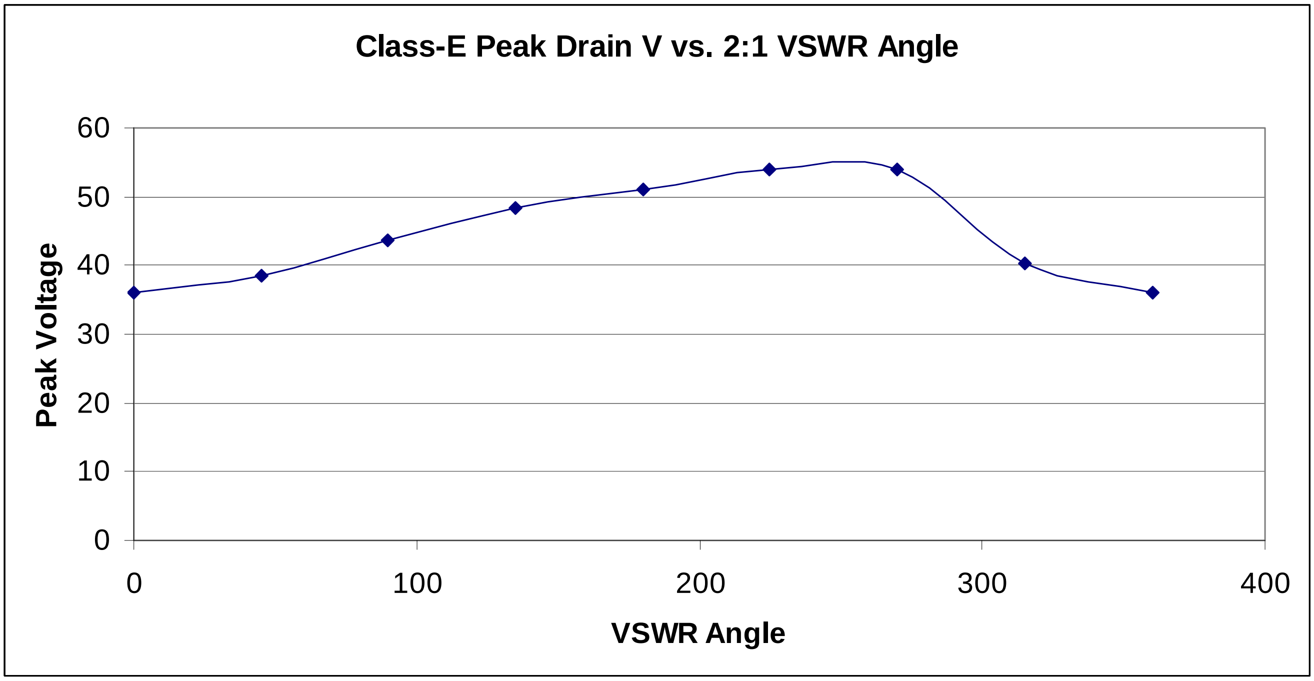

Figure 9: Peak Drain Voltage for Class-E Power Amplifier, 13.6V input