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This dissertation describes a new radio-frequency power-amplifier circuit and mode of operation that exceeds the efficiency of the conventional class C amplifier in the low and medium h-f range. It will be found useful in applications where linearity between the input and output are not required, as in all applications where a conventional class C amplifier would be used, and where the highest efficiency with regard to r-f power output for a given d-c input power is desired. The reasons for this high efficiency are the transistor operates in a pure switching mode, and the voltage across the transistor and the current flowing through it can both be made exactly equal to zero during the interval of the switch-on transient.

A complete design procedure is given along with tabulated design parameters to simplify the numerical calculations. To illustrate this, a 20-watt transistor r-f power amplifier was designed, built and tested in the laboratory. This amplifier was found to be so

efficient that an external heat radiator on the power transistor was unnecessary. The author suggests that such a design could be used to advantage in a transmitter for a remote radio-navigation beacon, as an emergency marine life-boat transmitter or in similar applications where input power is limited and expensive to provide.

HIGH-EFFICIENCY RADIO-FREQUENCY POWER AMPLIFIERS

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HIGH-EFFICIENCY RADIO-FREQUENCY POWER AMPLIFIERS

INTRODUCTION

Radio-frequency power amplifiers may be classified into two major groups; those that preserve the amplitude of the input signal and those that do not. This dissertation will be primarily concerned with the latter group which is generally classified as Class C amplifiers. High efficiency with regard to the r-f output for a given d-c input power is of primary importance in this type of power amplifier.

In the following pages a new, very efficient class C amplifier will be developed along with the design equations, which will be illustrated by a practical power-amplifier design. Since the active device will be used as a high-speed switch the transister will be more efficient than the vacuum tube but a very high-perveance tube will also give good results.

CLASSIFICATION OF AMPLIFIERS

Before the new mode of class C r-f amplification operation is introduced it seems advisable to review the well known Class A, B, and C amplifiers with particular attention given to maximum attainable efficiency. Efficiency will be defined as the ratio of the r-f power output to the sum of the d-c, and r-f power inputs to the amplifier stage.

Class A Amplifier

The class A amplifier is characterized by a continuous flow of bias current through the active device. This bias current which is modulated by the input r-f signal flows through the load impedance giving rise to an output voltage which is proportional to the input. At no time during the r-f cycle does the bias current cease to flow except in the limiting case when it just drops to zero on the negative peak of the r-f cycle. Figure 1 shows a class A transistor, r-f amplifier with appropriate d-c biasing. Figure 2 illustrates the theoretical limiting case of maximum output for a class A amplifier. In order to approach this condition in practice the on or saturation resistance of the transister must be very small to make the voltage v_c approach zero while full peak current is flowing through it.

The maximum possible collector efficiency for a class A

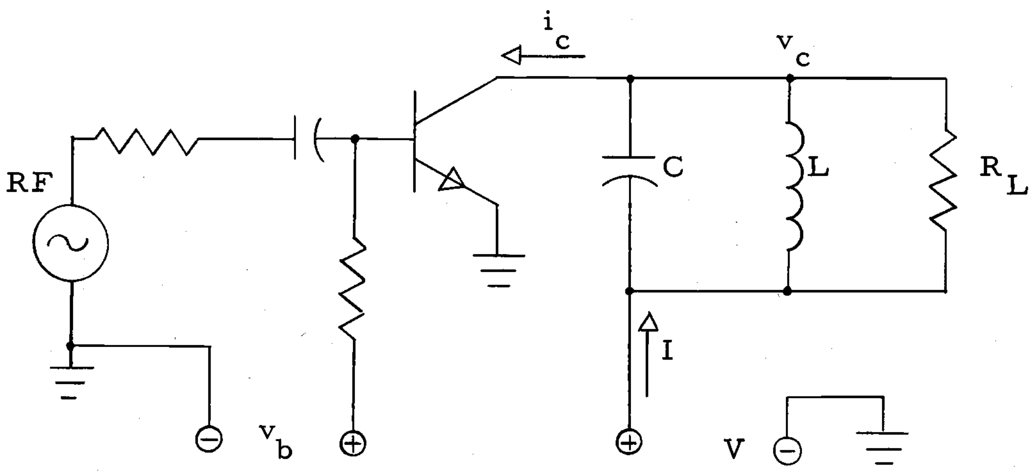


Figure 1. Class A R-F Amplifier

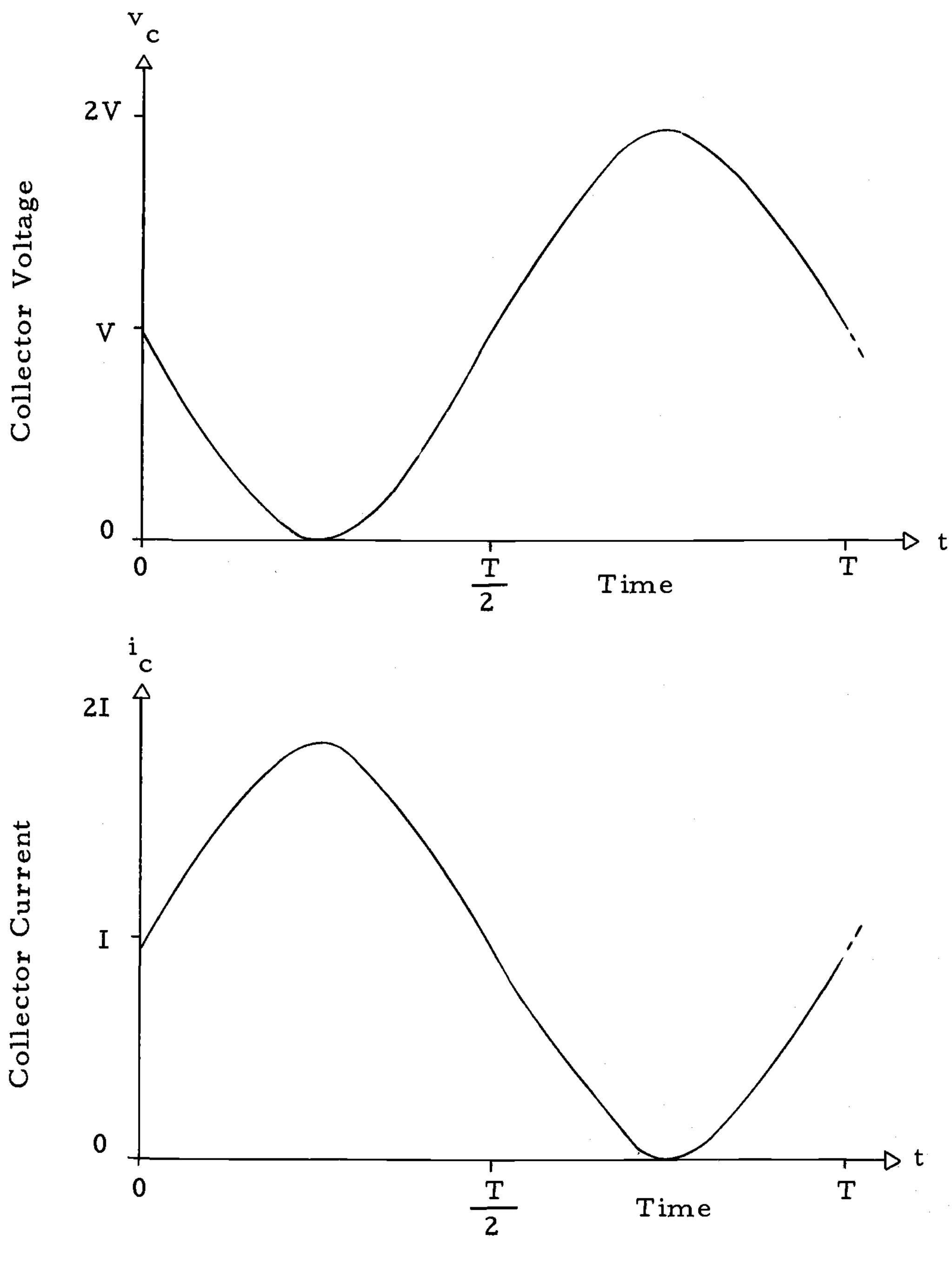


Figure 2. Class A R-F Amplifier Collector Voltage and Current vs Time

amplifier may be computed by solving for P_d the average power dissipated in the transistor.

$$P_d = \frac{1}{T}\int_o^T v_c i_c \, dt \qquad (1)$$

$$P_d = \frac{1}{T}\int_o^T V(1 - \sin\frac{2\pi}{T}t)\, I\,(1 + \sin\frac{2\pi}{T}t)\, dt$$

$$P_d = \frac{VI}{T}\int_o^T (1 - \sin^2\frac{2\pi}{T}t)\, dt$$

$$P_d = \frac{VI}{T}\int_o^T (\frac{1}{2} + \frac{1}{2}\cos\frac{4\pi}{T}t)\, dt$$

$$P_d = \frac{VI}{T}\left[\frac{1}{2}t + \frac{T}{8\pi}\sin\frac{4\pi}{T}t\right]_o^T$$

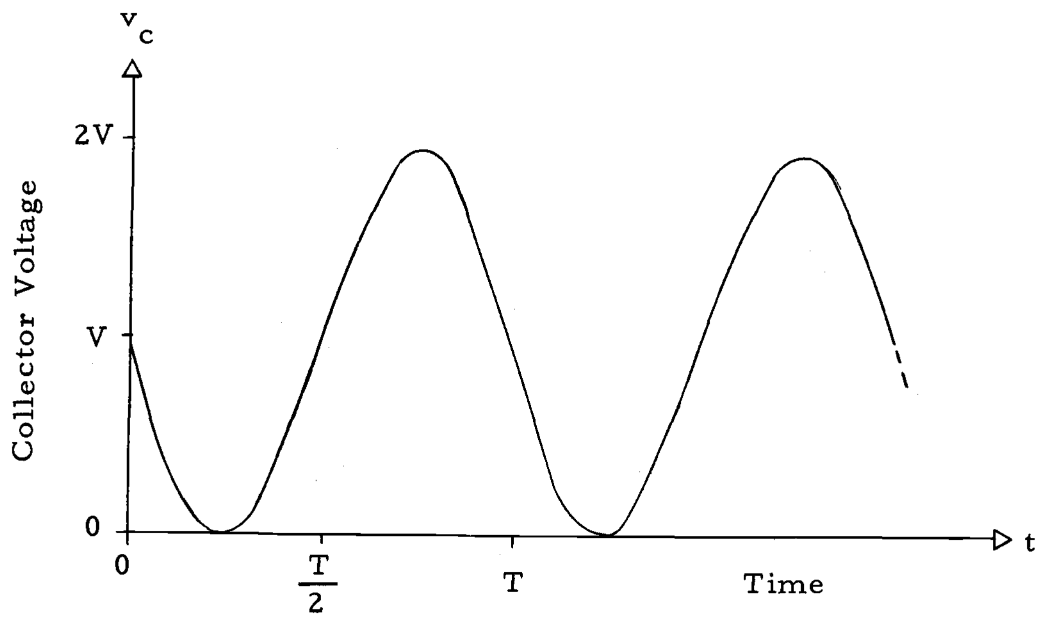
$$P_d = \frac{VI}{2}$$

Since the d-c collector-input power is VI, the maximum theoretical collector efficiency is 50 percent for the pure class A amplifier. In practice the overall efficiency is less than 50 percent since the transister always has a finite on resistance and requires additional input power to supply base current.

The class A amplifier is usually used in low-power level applications where low-output amplitude distortion is more important than high efficiency.

Class B Amplifier

Pure class B operation is achieved when the operating point is selected in such a way that collector current i_c flows for one-half of the r-f cycle. Figure 3 shows the theoretical maximum-output



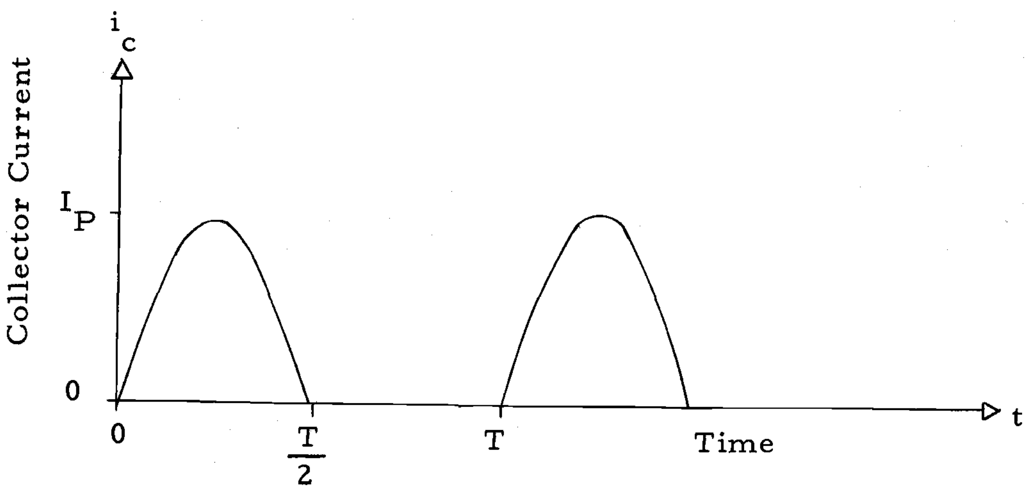


Figure 3. Class B R-F Amplifier Collector Voltage and Current vs Time

condition for the class B r-f amplifier. The parallel-resonant tuned circuit in the collector lead will supply a good sinusoidal output voltage provided the Q of the circuit is reasonably high, $Q \simeq 10$.

As in the case of the class A amplifier the class B amplifiers maximum collector efficiency may be calculated by computing the average power dissipated in the transistor for the limiting case of zero saturation voltage occurring at the instant of peak collector-current flow. Again from equation (1),

$$P_d = \frac{1}{T}\int_o^T v_c i_c \, dt.$$

It is necessary to establish the peak current I_p flowing through the transistor as a function of the average d-c current supplied to the collector circuit in order to evaluate the theoretical maximum collector efficiency. This is easy to do since the collector current flows in sinusoidal pulses as shown in Figure 3.

$$I = \frac{1}{T}\int_o^T i_c dt \qquad (2)$$

$$I = \frac{1}{T}\int_o^{T/2} I_p \sin \frac{2\pi}{T} t \, dt$$

$$I = \frac{I_p}{T}\left[-\frac{T}{2\pi} \cos \frac{2\pi}{T} t\right]_o^{T/2} = \frac{I_p}{\pi} \text{ or } I_p = \pi I.$$

Now the expression for P_d may be evaluated.

$$P_d = \frac{1}{T}\int_o^{T/2} V\,(1 - \sin\frac{2\pi}{T}t)\,\pi I \sin\frac{2\pi}{T}\,t\,dt$$

$$P_d = \frac{\pi VI}{T}\int_o^{T/2} (\sin\frac{2\pi}{T}\,t - \frac{1}{2} + \frac{1}{2}\cos\frac{4\pi}{T}t)\,dt$$

$$P_d = \frac{\pi VI}{T}\left[-\frac{T}{2\pi}\cos\frac{2\pi t}{T} - \frac{1}{2}t + \frac{1}{8\pi}\sin\frac{4\pi t}{T}\right]_o^{T/2}$$

$$P_d = \frac{\pi VI}{T}\left(\frac{T}{2\pi} + \frac{T}{2\pi} - \frac{T}{4}\right) = \pi VI\left(\frac{1}{\pi} - \frac{1}{4}\right)$$

$$P_d = VI \quad (1 - \frac{\pi}{4}) \,. \tag{3}$$

Since the d-c input power to the collector circuit is VI, the power available at the load is $\frac{\pi}{4}$ VI as evident from equation (3). The maximum collector efficiency of the class B amplifier is therefore $\frac{\pi}{4}$ or 78.5 percent. Again, as in the case of class A amplifiers the efficiency attained in practice is much lower than this value due to finite saturation resistance (5, p. 357-358).

The tuned class B amplifier affords the highest overall efficiency of the linear class of amplifiers. It is used in the high-level stages of r-f transmitters where reasonable linearity and efficiency are prerequisite.

Class C Amplifier

In a large class of r-f transmitters, amplitude linearity in the final r-f power amplifier is not required. Such is the case in CW pulse or code transmitters, frequency-modulated or high level collector or plate amplitude-modulated transmitters. Here high efficiency is of primary concern, particularly in portable and remote stations where the basic input power is limited and expensive to supply. In these cases the class C amplifier is used to advantage.

Good analysis and design procedures have been worked out for the vacuum tube class C amplifier (2, p. 629-652; 4, p. 432-440). One can expect plate efficiencies between 65 percent and 85 percent in practical circuits. For the transistor class C r-f amplifier insufficient published material seems to be available to guide the prospective designer.

As it was for the class A and B transistor r-f amplifier, the class C mode of operation will be defined in terms of the collector current conduction angle or time. For collector-current conduction angles less than 180°, the transistor will be operating class C by definition. With reference to Figure 1, the base-bias voltage V_b would have to be negative if the r-f excitation voltage was sinusoidal to achieve class C operation. However, if the excitation was applied to the base of the transistor in the form of a series of current pulses with a time duration of less than T/2, class C operation would be achieved without the negative-base power supply. With this type of current pulse excitation, the amplitude and period is easy to adjust and hold within design limits. This would not be the case for a sinusoidal source since the conduction time would depend very critically on the amplitude of the sine wave and the value of the negative-bias voltage V_b.

Figure 4 shows the collector-voltage and current wave forms of a pulse-excited transistor r-f amplifier. To find an expression for the collector efficiency, equation (1) is applied as before.

$$P_d = \frac{1}{T}\int_{-\frac{t_1}{2}}^{\frac{t_1}{2}} [V - (V-V_s) \cos \frac{2\pi}{T} t] I_p dt$$

$$P_d = \frac{I_p}{T} [Vt - \frac{T}{2\pi} (V-V_s) \sin \frac{2\pi}{T} t)]_{-\frac{t_1}{2}}^{\frac{t_1}{2}}$$

$$P_d = \frac{I_p}{T} [Vt_1 - \frac{T}{\pi} (V-V_s) \sin \frac{\pi t_1}{T}]$$

where V_s will be defined as the saturation voltage of the transistor under the given circuit conditions with I_p flowing through it. Since the saturation resistance R_s ($R_s \simeq \frac{V_s}{I_p}$) of the transistor is dependent on voltage, current and temperature, the following equations are to this extent approximations.

Applying equation (2) results in

$$I = \frac{1}{T}\int_{-\frac{t_1}{2}}^{\frac{t_1}{2}} I_p dt = I_p \frac{t_1}{T} . \quad (4)$$

Now solving for I_p and substituting into the expression for P_d yields,

$$P_d = I [V - \frac{T}{\pi t_1} (V-V_s) \sin \frac{\pi t_1}{T}] \text{ or}$$

$$P_d = VI [1 - \frac{T}{\pi t_1} (1 - \frac{V_s}{V}) \sin \frac{\pi t_1}{T} .$$

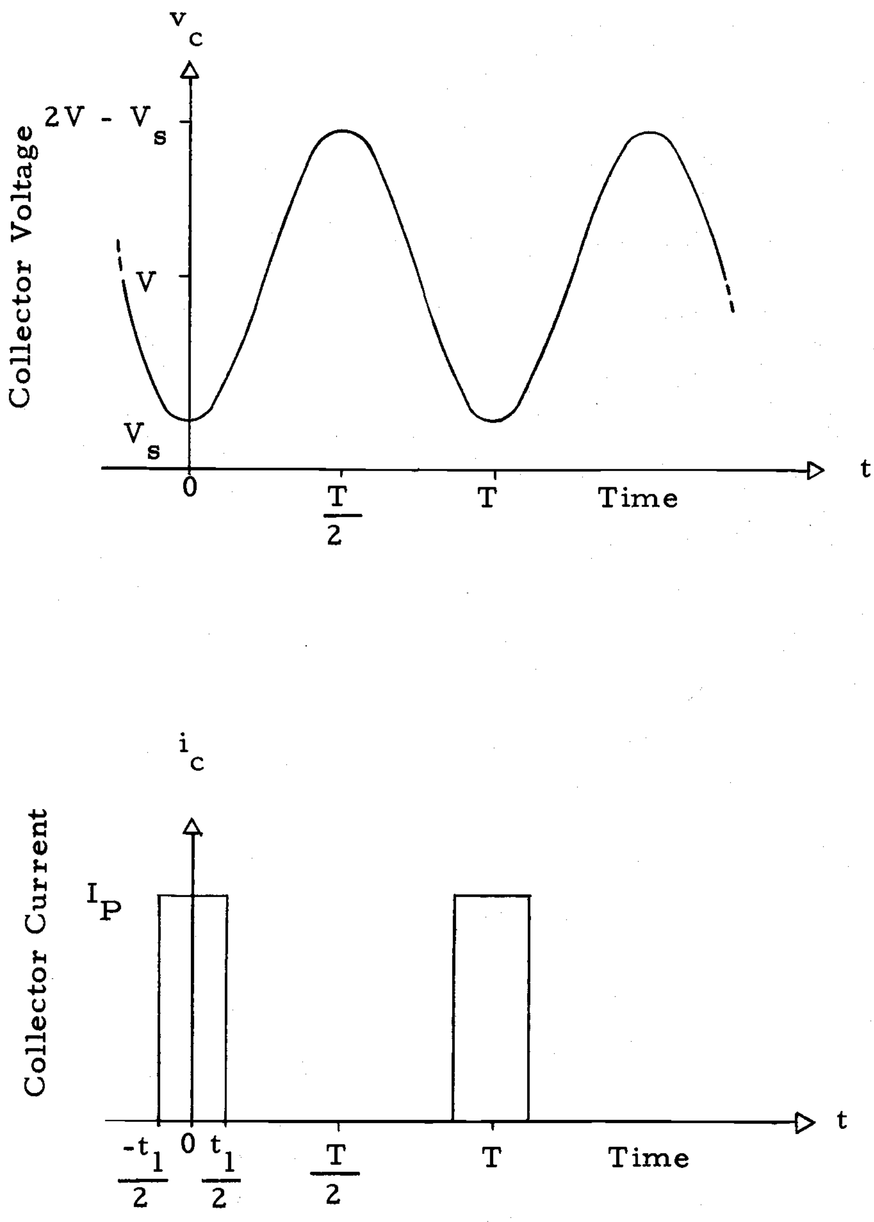


Figure 4. Class C R-F Amplifier Collector Voltage and Current vs Time

Since the collector efficiency is equal to power out divided by the power in to the collector circuit,

$$\frac{VI - P_d}{VI} = 1 - \frac{P_d}{VI} \quad \text{or}$$

$$\text{eff.} = \frac{T}{\pi t_1} (1 - \frac{V_s}{V}) \sin \frac{\pi t_1}{T} \ . \qquad (5)$$

From equation (5), it is evident that the collector efficiency of a pulse-excited class C r-f amplifier is dependent upon the conduction angle $\frac{\pi t_1}{T}$ and the ratio of the saturation voltage to the d-c supply voltage.

Since the saturation voltage V_s is dependent to a first approximation linearly on the collector-pulse current I_p, equation (5) may be written

$$\text{eff.} = \frac{T}{\pi t_1} (1 - \frac{R_s I_p}{V}) \sin \frac{\pi t_1}{T} \ . \qquad (6)$$

For any given transistor the saturation resistance R_s and the peak collector current I_p that can be established is fixed and limited respectively. The collector current in a transistor is related to the base-excitation current by the current-transfer ratio, beta, which is first an increasing and then a decreasing function of the magnitude of the collector current. Beta also decreases with operating frequency which places another well-known constraint on the switching speed of the transister (3, p. 4-25, 29). With these limitations in mind the designer, with a knowledge of the characteristics of the selected transister can determine the maximum value of I_p that can be obtained with a reasonable base-current excitation.

To arrive at a typical numerical value of the collector efficiency of a class C transistor r-f amplifier an example will be given. Say an efficient r-f power source of about 2.0 watts output in the low megacycle frequency range is needed. A silicon-mesa 2N697 transistor with the following pertinent parameters will be used.

Saturation resistance $R_s \doteq 5\Omega$ (maximum limit reported by manufacturer) at $I_p = 0.4$ amp.

Maximum collector voltage $\doteq$ 50 volts.

Let the supply voltage V = 24 volts and $\frac{t_1}{T} = \frac{1}{4}$ then;

$$I = I_p \frac{t_1}{T} = 0.4 \times \frac{1}{4} = 0.1 \text{ amp},$$

input-collector power $VI = 24 \times 0.1 = 2.4$ watts,

saturation voltage $V_s = R_s I_p = 5 \times 0.4 = 2$ volts

peak-collector voltage (see Figure 4),

$2V - V_s = 2 \times 24 - 2 = 46$ volts which is less than the specified maximum of 50 volts,

and the collector efficiency

$$\text{eff.} = \frac{4}{\pi}\left(1 - \frac{2}{24}\right) \sin \frac{\pi}{4} = 0.826 \text{ or}$$

$$\text{eff.} \doteq 83 \text{ percent}$$

The preceeding calculations show that the practical efficiency of a class C transistor r-f power amplifier is significantly higher than even the theoretical maximum efficiency of a class B amplifier. The principle disadvantage of the class C amplifier as compared to the class B, aside from linearity, is the necessary circuitry and subsequent power-drain requirements to generate excitation base-current pulses of duration much less than one-half of the r-f cycle. This disadvantage can be removed and at the same time achieve an increase in efficiency by using the new r-f power amplifier to be developed.

HIGH-EFFICIENCY R-F POWER AMPLIFIER DEVELOPMENT

During the course of the author's research in the area of switching mode transistor r-f power amplifiers, it was discovered that very high efficiencies were obtained with a series resonant R, L, C circuit connected about the transistor as shown in Figure 5. A careful analysis of this circuit reveals that the proper choice of transistor and circuit parameters yields an extremely efficient r-f power source. As will be shown, the reasons for this high efficiency are that the transistor operates in a pure switching mode and that the voltage across the transistor v_c and the current flowing through it i_c can both be made equal to zero during the interval of the switch-on transient. This latter condition is seldom, if ever, achieved in power-switching applications, as an example, even in the very efficient transistor d-c to a-c inverters (6, p. 447-461).

Circuit Analysis

In the analysis of the circuit shown in Figure 5, the Q of the series-tuned circuit will be made high enough to make valid the assumption that the load current i_L is sinusoidal. That is,

$$Q = \frac{X}{R_L} \tag{7}$$

where X is the reactance of L or C at the resonant operating

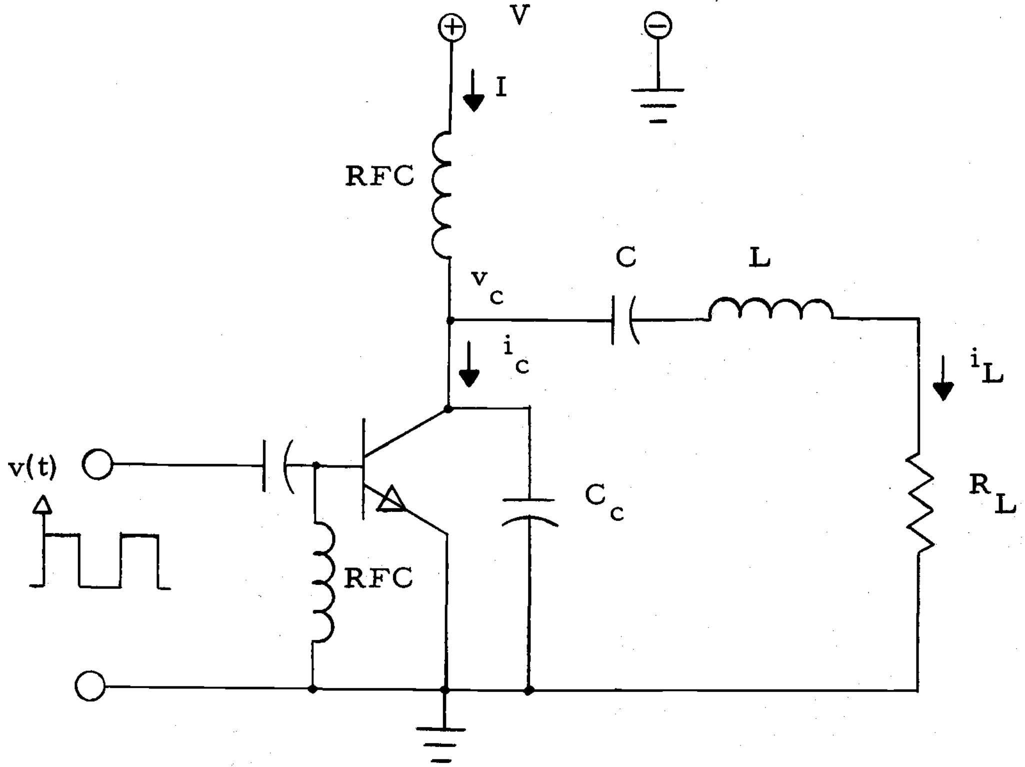


Figure 5. High Efficiency R-F Power Amplifier

frequently f and R_L is the sum of the load resistance and the finite series loss resistance of the inductor and capacitor. In practice, a Q of about ten will give a good sinusoidal shape to the load current. It is also assumed that all the capacitance at the collector node seen at v_c, including the intrinsic collector to emitter capacitance of the transistor, will be contained in C_c and that the saturation resistance R_s of the transistor will be much less than R_L.

Under these assumptions the load current may be expressed as follows:

$$i_L = I_L \sin \frac{2\pi}{T} (t - t_1) \tag{8}$$

where I_L is the peak value of the load current and $2\pi t_1/T$ is the phase angle between the base-current excitation pulses and i_L. As usual $T = 1/f$, the period at the operating frequency. The current i_c, which is the current flowing to the transistor and C_c, can be expressed as the difference between the d-c current I and the load current i_L.

$$i_c = I - I_L \sin \frac{2\pi}{T} (t - t_1) \tag{9}$$

The voltage v_c at the collector of the transistor may not be expressed analytically as a function of time during the non-conducting or open-circuit interval t_2, since

$$v_c = \frac{1}{C_c} \int_o^T i_c dt$$

where $0 \leq t \leq t_2$. Figure 6 shows plots of i_c and v_c as a function of

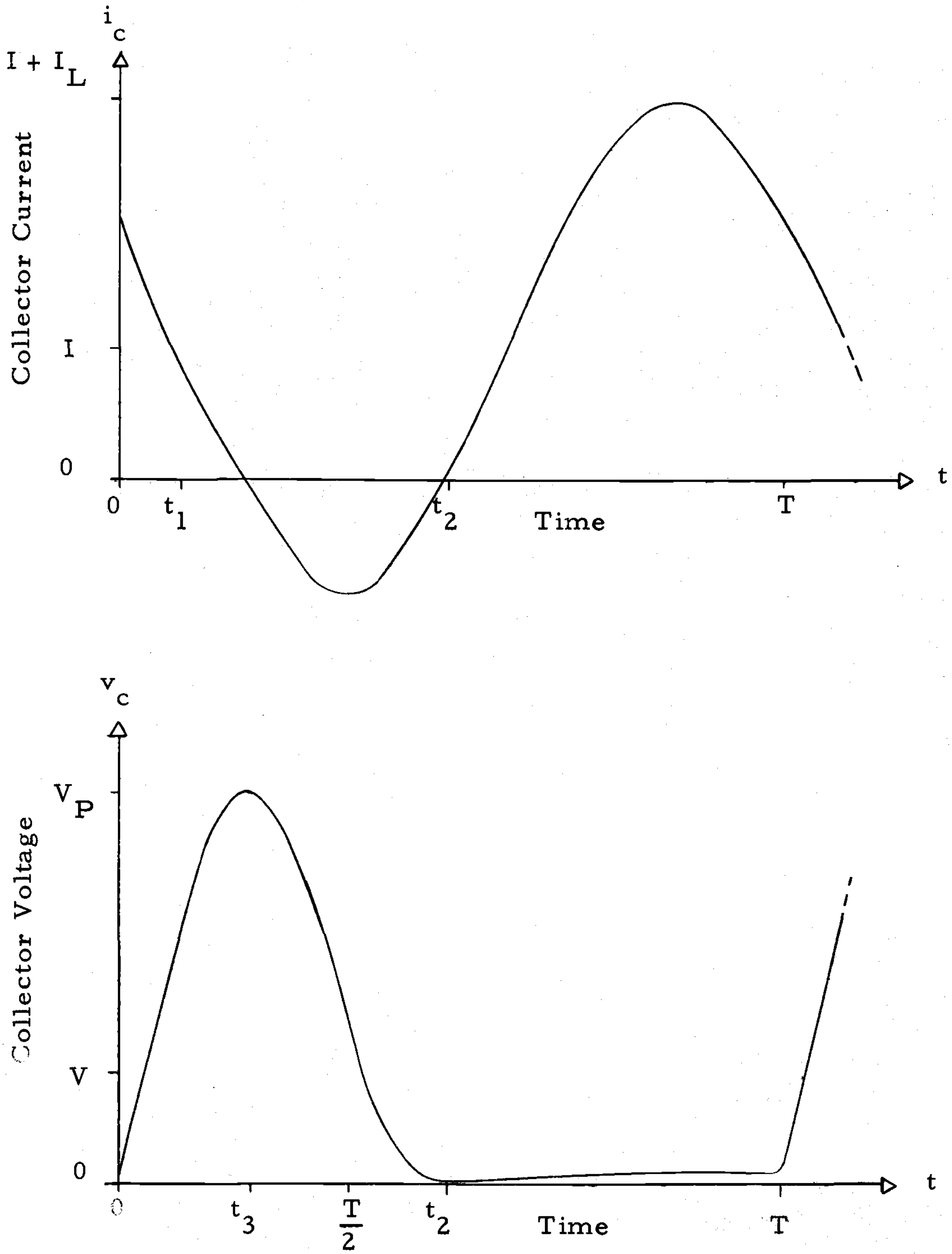


Figure 6. High Efficiency R-F Amplifier Collector Current and Voltage vs Time

time for a very unique condition, that is, i_c and v_c are both zero at time t equal to t_2. This condition insures high efficiency since the transistor is switched on and brought into saturation with zero voltage across it and zero current flowing through it. During the time interval $T-t_2$, the transistor is in saturation carrying the current i_c.

Applying the condition that at $t = t_2$, $i_c = 0$ results in

$$K = \frac{I}{I_L} = \sin \frac{2\pi}{T} (t_2 - t_1). \tag{11}$$

In order to apply the condition that at $t = t_2$, $v_c = 0$, the analytical expression for v_c must be obtained by substituting equation (9) into (10) and integrating

$$v_c = \frac{1}{C_c}\int_0^t [I - I_L \sin \frac{2\pi}{T} (t - t_1)]\, dt.$$

$$v_c = \frac{1}{C_c} \{It + I_L \frac{T}{2\pi} [\cos \frac{2\pi}{T} (t - t_1) - \cos \frac{2\pi}{T} t_1]\} . \tag{12}$$

Setting $v_c = 0$ at $t = t_2$ yields another independent equation for the ratio I/I_L .

$$\frac{1}{C} \{It_2 + I_L \frac{T}{2\pi} [\cos \frac{2\pi}{T} (t_2 - t_1) - \cos \frac{2\pi t_1}{T}]\} = 0$$

$$K = \frac{I}{I_L} = \frac{T}{2\pi t_2} [\cos \frac{2\pi t_1}{T} - \cos \frac{2\pi}{T} (t_2 - t_1)] . \tag{13}$$

By eliminating K between equations (11) and (13) an expression relating the phase angle $2\pi \frac{t_1}{T}$ to the nonconduction angle $2\pi \frac{t_2}{T}$ is obtained. For convenience let

$$\phi_1 = 2\pi \frac{t_1}{T} \tag{14}$$

and $$\phi_2 = 2\pi \frac{t_2}{T} \tag{15}$$

then

$$\sin(\phi_2 - \phi_1) = \frac{1}{\phi_2}\left[\cos\phi_1 - \cos(\phi_2 - \phi_1)\right]$$

or

$$\phi_2 = \frac{\cos\phi_1 - \cos(\phi_2 - \phi_1)}{\sin(\phi_2 - \phi_1)}$$

Now by using the trigonometric identities for the sine and cosine of the difference of two angles

$$\phi_2 = \frac{\cos\phi_1 - \cos\phi_2 \cos\phi_1 - \sin\phi_2 \sin\phi_1}{\sin\phi_2 \cos\phi_1 - \cos\phi_2 \sin\phi_1} \; .$$

By dividing the numerator and the denominator by $\cos\phi_1$

$$\phi_2 = \frac{1 - \cos\phi_2 - \sin\phi_2 \tan\phi_1}{\sin\phi_2 - \cos\phi_2 \tan\phi_1}$$

where

$$\tan\phi_1 = \frac{\sin\phi_1}{\cos\phi_1} \; .$$

Now solving for ϕ_1 as a function of ϕ_2

$$\tan\phi_1 = \frac{\phi_2 \sin\phi_2 + \cos\phi_2 - 1}{\phi_2 \cos\phi_2 - \sin\phi_2}$$

or

$$\phi_1 = \arctan\left(\frac{\phi_2 \sin\phi_2 + \cos\phi_2 - 1}{\phi_2 \cos\phi_2 - \sin\phi_2}\right) \tag{16}$$

Again with reference to Figures 5 and 6, the average voltage at the collector of the transistor must be equal to the d-c supply voltage less any d-c voltage drop across the r-f choke, therefore;

$$V = \frac{1}{T}\int_{o}^{t_2} v_c dt + \frac{1}{T}\int_{t_2}^{T} R_s i_c dt. \quad (17)$$

By multiplying V by I, an expression for the average power input is obtained.

$$VI = \frac{I}{T}\int_{o}^{t_2} v_c dt + \frac{I}{T}\int_{t_2}^{T} R_s i_c dt \ . \quad (18)$$

Another independent expression for the average input power is

$$VI = P_L + P_d \quad (19)$$

where P_L is the power output into the load resistance R_L and P_d is the average power dissipated in the transistor. Expressions for P_L and P_d are given as follows:

$$P_L = \frac{1}{2} R_L I_L^2, \quad (20)$$

$$P_d = \frac{1}{T}\int_{t_2}^{T} R_s i_c^2 dt \ . \quad (21)$$

By equating the two independent expressions for the average input power to the collector-load circuit an equation relating the operating frequency to the network parameters and the nonconduction angle may be obtained as follows:

$$\frac{I}{T}\int_{o}^{t_2} v_c dt + \frac{I}{T}\int_{t_2}^{T} R_s i_c dt = \frac{1}{2} R_L I_L^2 + \frac{1}{T}\int_{t_2}^{T} R_s i_c^2 dt ,$$

or

$$\frac{R_L I_L^2}{2} + \frac{R_s}{T}\int_{t_2}^{T} (i_c^2 - I i_c) dt = \frac{I}{T}\int_{o}^{t_2} v_c dt \ . \quad (22)$$

Substituting equations (9) and (12) into (22) and replacing I_L with I/K yields

$$\frac{R_L I^2}{2K^2} + \frac{I^2 R_s}{T} \int_{t_2}^{T} \{[1 - \frac{1}{K} \sin \frac{2\pi}{T} (t-t_1)]^2 - 1 + \frac{1}{K} \sin \frac{2\pi}{T}(t-t_1)\} \, dt$$

$$= \frac{I^2}{TC_c} \int_{o}^{t_2} \{t + \frac{T}{2\pi K} [\cos \frac{2\pi}{T} (t-t_1) - \cos \frac{2\pi}{T} t_1]\} \, dt .$$

Dividing out I^2 and expanding the first integral

$$\frac{R_L}{2K^2} + \frac{R_s}{T} \int_{t_2}^{T} [\frac{1}{K^2} \sin^2 \frac{2\pi}{T} (t-t_1) - \frac{1}{K} \sin \frac{2\pi}{T} (t-t_1)] \, dt$$

$$= \frac{1}{TC_c} \int_{o}^{t_2} \{t + \frac{T}{2\pi K} [\cos \frac{2\pi}{T} (t-t_1) - \cos \frac{2\pi}{T} t_1]\} \, dt .$$

Solving this equation for the operating frequency f which is equal to $1/T$

$$f = \frac{R_L C_c}{A - B R_s C_c} \tag{23}$$

where

$$A = 2K^2 \int_{o}^{t_2} \{t + \frac{T}{2\pi K} [\cos \frac{2\pi}{T} (t-t_1) - \cos \frac{2\pi t_1}{T}]\} \, dt \tag{24}$$

and

$$B = 2 \int_{t_2}^{T} [\sin^2 \frac{2\pi}{T} (t-t_1) - K \sin \frac{2\pi}{T} (t-t_1)] \, dt. \tag{25}$$

Equation (23) is an implicit function of frequency since A and B are both functions of the period; therefore, integral equations (24) and (25) must be solved and f extracted. From equation (24)

$$A = 2K^2 \left\{ \frac{t^2}{2} + \frac{T}{2\pi K} \left[\frac{T}{2\pi} \sin \frac{2\pi}{T}(t-t_1) - (\cos \frac{2\pi}{T} t_1)t \right] \right\}_o^{t_2}$$

$$A = 2K^2 \left\{ \frac{t_2^2}{2} + \frac{T}{2\pi K} \left[\frac{T}{2\pi} \sin \frac{2\pi}{T}(t_2 - t_1) + \frac{T}{2\pi} \sin \frac{2\pi t_1}{T} - (\cos \frac{2\pi t_1}{T}) t_2 \right] \right\} .$$

Now using equations (14) and (15) to write A in terms of the radian angles ϕ_1 and ϕ_2

$$A = \frac{K^2 T^2}{4\pi^2} \left\{ \phi_2^2 + \frac{2}{K} \left[\sin(\phi_2 - \phi_1) + \sin \phi_1 - \phi_2 \cos \phi_1 \right] \right\} .$$

Let

$$a = f^2 A \tag{26}$$

then

$$a = \frac{K^2}{4\pi^2} \left\{ \phi_2^2 + \frac{2}{K} \left[\sin(\phi_2 - \phi_1) + \sin \phi_1 - \phi_2 \cos \phi_1 \right] \right\} . \tag{27}$$

For equation (25) after substituting in the trigonometric identity for the sine squared

$$B = 2 \int_{t_2}^{T} \left[\frac{1}{2} - \frac{1}{2} \cos \frac{4\pi}{T}(t-t_1) - K \sin \frac{2\pi}{T}(t-t_1) \right] dt,$$

$$B = 2 \left[\frac{t}{2} - \frac{T}{8\pi} \sin \frac{4\pi}{T}(t-t_1) + \frac{KT}{2\pi} \cos \frac{2\pi}{T}(t-t_1) \right]_{t_2}^{T} ,$$

$$B = T - t_2 - \frac{T}{4\pi} \sin \frac{4\pi}{T}(T-t_1) + \frac{T}{4\pi} \sin \frac{4\pi}{T}(t_2-t_1) + \frac{KT}{\pi} \cos \frac{2\pi}{T}(T-t_1) - \frac{KT}{\pi} \cos \frac{2\pi}{T}(t_2-t_1) .$$

Again, using equations (14) and (15)

$$B = T - \frac{T\phi_2}{2\pi} - \frac{T}{4\pi} \sin(4\pi - 2\phi_1) + \frac{T}{4\pi} \sin 2(\phi_2 - \phi_1) +$$

$$\frac{KT}{\pi} \cos(2\pi - \phi_1) - \frac{KT}{\pi} \cos(\phi_2 - \phi_1).$$

Factoring out $T/4\pi$ and simplifying

$$B = \frac{T}{4\pi} \{2(2\pi - \phi_2) + \sin 2(\phi_2 - \phi_1) + \sin 2\phi_1 +$$

$$4K[\cos\phi_1 - \cos(\phi_2 - \phi_1)]\} .$$

Let

$$b = fB \tag{28}$$

then

$$b = \frac{1}{4\pi} \{2(2\pi - \phi_2) + \sin 2(\phi_2 - \phi_1) + \sin 2\phi_1 +$$

$$4K[\cos\phi_1 - \cos(\phi_2 - \phi_1)]\} . \tag{29}$$

Now substituting equations (26) and (28) into (23)

$$f = \frac{R_L C_c}{\frac{a}{f^2} - \frac{b}{f} R_s C_c}$$

Solving for f the desired result is obtained

$$f = \frac{a}{R_L C_c + bR_s C_c} \tag{30}$$

where a and b are dimensionless quantities which are functions of the nonconduction angle ϕ_2.

Average Power Dissipated in the Switching Device

The average power dissipated in the collector of the transistor is expressed by the integral

$$P_d = \frac{1}{T}\int_{t_2}^{T} R_s i_c^2 \, dt.$$

Substituting into equation (9) and (11),

$$P_d = \frac{R_s}{T}\int_{t_2}^{T} [I - \frac{I}{K}\sin\frac{2\pi}{T}(t-t_1)]^2 \, dt;$$

$$P_d = \frac{R_s I^2}{T}\int_{t_2}^{T} [1 - \frac{2}{K}\sin\frac{2\pi}{T}(t-t_1) + \frac{1}{K^2}\sin^2\frac{2\pi}{T}(t-t_1)] \, dt.$$

Using the trigonometric identity for the sine squared

$$P_d = \frac{R_s I^2}{T}\int_{t_2}^{T} [1 - \frac{2}{K}\sin\frac{2\pi}{T}(t-t_1) + \frac{1}{2K^2} - \frac{1}{2K^2}\cos\frac{4\pi}{T}(t-t_1)] \, dt$$

$$P_d = \frac{R_s I^2}{T}[t + \frac{T}{\pi K}\cos\frac{2\pi}{T}(t-t_1) + \frac{t}{2K^2} - \frac{T}{8\pi K^2}\sin\frac{4\pi}{T}(t-t_1)]_{t_2}^{T}$$

$$P_d = \frac{R_s I^2}{T}[(1 + \frac{1}{2K^2})(T-t_2) + \frac{T}{\pi K}\cos\frac{2\pi}{T}(T-t_1) - \frac{T}{\pi K}\cos\frac{2\pi}{T}(t_2-t_1) - \frac{T}{8\pi K^2}\sin\frac{4\pi}{T}(T-t_1) + \frac{T}{8\pi K^2}\sin\frac{4\pi}{T}(t_2-t_1)].$$

Now, writing the power dissipation as a function of ϕ_1 and ϕ_2

$$P_d = R_s I^2 \left[\left(1 + \frac{1}{2K^2}\right)\left(T - \frac{\phi_2 T}{2\pi}\right) + \frac{T}{\pi K} \cos (2\pi - \phi_1) - \right.$$

$$\left. \frac{T}{\pi K} \cos (\phi_2 - \phi_1) - \frac{T}{8\pi K^2} \sin 2 (2\pi - \phi_1) + \frac{T}{8\pi K^2} \sin 2(\phi_2 - \phi_1) \right].$$

Simplifying and factoring

$$P_d = \frac{R_s I^2}{8\pi K^2} [2(2K^2 + 1)(2\pi - \phi_2) + 8K \cos \phi_1 - 8K \cos (\phi_2 - \phi_1)$$

$$+ \sin 2\phi_1 + \sin 2(\phi_2 - \phi_1)]. \tag{31}$$

Let

$$D = \frac{1}{8\pi K^2} [2(2K^2 + 1)(2\pi - \phi_2) + 8K \cos \phi_1 - 8K \cos (\phi_2 - \phi_1)$$

$$+ \sin 2\phi_1 + \sin 2(\phi_2 - \phi_1)] . \tag{32}$$

then

$$P_d = D R_s I^2 . \tag{33}$$

D-C Input Resistance of the High-Efficiency R-F Amplifier

The d-c resistance of the collector circuit seen by the power supply may be expressed as a function of the circuit parameters and the nonconduction angle ϕ_2. A knowledge of this resistance is essential for the design of the amplifier. It is also the load resistance presented to a high-level modulator which would be used to secure an amplitude-modulated r-f signal output.

By dividing equation (19), the expression of the average input power by the d-c input current squared given the input resistance.

$$\frac{V}{I} = \frac{P_L}{I^2} + \frac{P_d}{I^2} . \tag{34}$$

Substituting equations (20) and (33) into (34) yields the desired result,

$$\frac{V}{I} = \frac{R_L}{2K^2} + D\,R_s , \tag{35}$$

since,

$$I_L = \frac{I}{K} .$$

Collector Efficiency

The collector efficiency of the high-efficiency r-f power amplifier may now be calculated by substituting equation (33) into equation (19) which yields

$$VI = P_L + DR_s I^2 . \tag{36}$$

Now by definition the collector efficiency is equal to P_L/VI or after substituting in equation (36)

$$\text{eff.} = \frac{P_L}{P_L + DR_s I^2} .$$

Dividing the numerator and denominator by P_L and applying equation (20)

$$\text{eff.} = \frac{1}{1 + \dfrac{2DR_s I^2}{R_L I_L 2}} \quad \text{or}$$

$$\text{eff.} = \frac{1}{1 + 2K^2 D \dfrac{R_s}{R_L}} \tag{37}$$

since

$$I_L = \frac{I}{K} \, .$$

Under the conditions set down earlier the collector efficiency as shown in equation (37) is a function of the nonconduction angle ϕ_2 and the ratio of the saturation resistance to the load resistance. For a given conduction angle the smaller this ratio the higher efficiency will be.

Peak Voltage Across the Switching Device

When a transistor is used as the switching device in the high-efficiency r-f amplifier circuit, the peak voltage that occurs across it will usually limit the maximum attainable power output when it is carrying maximum peak current. It will be shown later by actual measured performance data that the efficiency of this circuit is so high that the power dissipated in a transistor with a reasonably high beta and low saturation resistance will usually be much less than it is capable of dissipating.

Let V_p equal the peak voltage across the transistor and at its occurrence, let t equal t_3. From equation (10) then

$$\left.\frac{dv_c}{dt}\right|_{t=t_3} \left.\frac{1}{C_c} i_c\right|_{t=t_3} = 0 \, .$$

As shown by this equation and as it is illustrated in Figure 5, the current i_c passes through zero at the instant v_c reaches V_p; therefore, from equation (9)

$$I - I_L \sin \frac{2\pi}{T}(t_3 - t_1) = 0$$

or

$$\frac{I}{I_L} = \sin \frac{2\pi}{T}(t_3 - t_1) .$$

Since by previous definition $K = I/I_L$;

$$\sin(\phi_3 - \phi_1) = K \tag{38}$$

where

$$\phi_3 = \frac{2\pi t_3}{T} . \tag{39}$$

Solving equation (38) for ϕ_3 yields

$$\phi_3 = \phi_1 + \text{arc} \sin K. \tag{40}$$

Now substituting equations (39) and (40) into (12) and factoring

$$V_p = \frac{IT}{2\pi C_c} \{\phi_3 + \frac{1}{K}[\cos(\phi_3 - \phi_1) - \cos \phi_1]\} .$$

Let

$$V_p = \frac{IE}{C_c f} \tag{41}$$

where $f = 1/T$ and

$$E = \frac{1}{2\pi} \{\phi_3 + \frac{1}{K}[\cos(\phi_3 - \phi_1) - \cos \phi_1]\} . \tag{42}$$

Conditions for Maximum Power Output

An expression for the load resistance as a function of the non-conduction angle ϕ_2 and the specified peak voltage V_p and current I_p will now be derived. As might be suspected this relationship will play an important part in the design procedure to be developed.

Solving equation (40) for frequency

$$f = \frac{EI}{C_c V_p} .$$

Equating this expression for frequency to that given by equation (30)

$$\frac{EI}{C_c V_p} = \frac{a}{C_c(R_L + bR_s)}$$

or after simplifying and rearranging terms

$$R_L = \frac{aV_p}{EI} - bR_s . \tag{43}$$

Now from equation (9) or as shown in Figure 6, the peak current flowing through the transistor is given by

$$I_p = I + I_L$$

or

$$I_p = I(1 + \frac{1}{K}) . \tag{44}$$

Solving equation (44) for I and substituting into equation (43) yields the desired result

$$R_L = \frac{(1 + \frac{1}{K})a \; V_p}{E \quad I_p} - bR_s . \tag{45}$$

If maximum power output from a given switching device, with a specified maximum V_p, I_p and R_s, and operating at a given non-conduction angle ϕ_2, is desired, then the load resistance given by equation (45) is used. Under these conditions the power output is given by

$$P_L = \frac{R_L I_p^2}{2(1 + K)^2} \qquad (46)$$

which is obtained by substituting equations (11) and (44) into equation (20).

DESIGN PROCEDURE

The following design procedure will assume that the operating frequency f, the power output P_L, and the nonconduction angle ϕ_2 will initially be specified. A detailed discussion on the choice of the nonconduction angle will be given later.

For the given value of ϕ_2 the following dimensionless design parameters are calculated from their respective equations. From equation (16),

$$\phi_1 = \arc \tan \left(\frac{\phi_2 \sin \phi_2 + \cos \phi_2 - 1}{\phi_2 \cos \phi_2 - \sin \phi_2} \right) .$$

From equation (11) with equations (14) and (15) applied,

$$K = \sin (\phi_2 - \phi_1) .$$

From equation (40),

$$\phi_3 = \phi_1 + \arc \sin K .$$

From equation (27),

$$a = \frac{K^2}{4\pi^2} \{ \frac{\phi_2^2}{2} + \frac{2}{K} [\sin (\phi_2 - \phi_1) + \sin \phi_1 - \phi_2 \cos \phi_1] \} .$$

From equation (29),

$$b = \frac{1}{4\pi} \{ 2 (2\pi - \phi_2) + \sin 2 (\phi_2 - \phi_1) + \sin 2\phi_1$$

$$+ 4K [\cos \phi_1 - \cos (\phi_2 - \phi_1)] \} .$$

From equation (32),

$$D = \frac{1}{8\pi K^2} [2(2K^2 + 1)(2\pi - \phi_2) + 8K \cos \phi_1 - 8K \cos (\phi_2 - \phi_1)$$

$$+ \sin 2\phi_1 + \sin 2(\phi_2 - \phi_1)] .$$

From equation (42),

$$E = \frac{1}{2\pi} \{\phi_3 + \frac{1}{K} [\cos (\phi_3 - \phi_1) - \cos \phi_1]\} .$$

The preceding parameters have been computed on an electronic digital computer for five degree steps in the nonconduction angle ϕ_2. The results of these calculations are given in the appendix in tabular form.

It is now necessary to derive expressions that will relate V_p, I_p, and C_c to the desired output power and operating frequency. Once these expressions are obtained and evaluated, a switching device can be specified that will meet the requirements.

By substituting equation (45) into equation (46) the output power can be expressed as

$$P_L = \frac{\left[\frac{(1 + \frac{1}{K})a \ V_p}{E \ I_p} - bR_s \right] I_p^2}{2(1 + K)^2}$$

Since

$$bR_s \ll \frac{(1 + \frac{1}{K}) \ a \ V_p}{E \ I_p}$$

for any reasonable set of conditions, the following approximation may be made

$$P_L \approx \frac{aV_pI_p}{2EK(1 + K)} \quad . \tag{47}$$

From equation (47), it is seen that the peak voltage and current requirements of the switch must be equal to or exceed

$$V_pI_p \approx \frac{2EK}{a} (1 + K) P_L \, . \tag{48}$$

Now solving equation (44) for I and substituting into equation (41) gives

$$V_p = \frac{E\,I_p}{(1 + \frac{1}{K})\, fC_c} \quad .$$

From this equation it is seen that the total collector-circuit node capacitance must be equal to

$$C_c = \frac{E\,I_p}{(1 + \frac{1}{K})\, fV_p} \tag{49}$$

The active switching device selected to meet the volt-ampere requirements set down in equation (48) must also have an intrinsic capacitance that is less than C_c as given by equation (49). The sum of the circuit-stray capacitance and the capacitance of the switching device forms the minimum value of C_c that will meet the conditions of the high-efficiency amplifier.

At this point in the design a switching device is selected that meets the conditions specified in equations (48) and (49). Thus, the values of V_p, I_p, and C_c are known along with the saturation resistance R_s. The required load resistance may now be calculated from equation (43)

$$R_L = \frac{(1 + \frac{1}{K})a\ V_p}{E\ I_p} - bR_s \qquad (50)$$

This value of amplifier load resistance will not in general be the correct value required, for example, the resistance of a radio antenna, but a simple L-C network will accommodate the match and also provide additional harmonic suppression over that already achieved with the series-tuned circuit (1, p. 118-122).

The current drawn from the d-c power supply is calculated from equation (44).

$$I = \frac{I_p}{I + \frac{I}{K}} \quad . \qquad (51)$$

Now the d-c power-supply voltage required is calculated by substituting the values of I and R_L into equation (35).

$$V = \left(\frac{R_L}{2K^2} + DR_s \right) I \qquad (52)$$

The value of the tuned-circuit inductance L or capacitance C will be calculated from the reactance X at the operating frequency f. Solving equation (7) for X gives

$$X = QR_L \qquad (53)$$

where Q is made equal to about ten or greater to insure a reasonably sinusoidal load current. Either the value of L or C should be made adjustable to permit adjustment of the circuit to achieve the proper load-current phase angle ϕ_1. From Figure 6, it is seen that the load current lags the collector voltage by ϕ_1; therefore, the L-C circuit must appear inductive at the operating frequency. To achieve this condition by adjusting C, (L might just as well have been chosen to be the adjustable element) its value must be made larger than that which would give a reactance exactly equal to X. From simple circuit analysis of a series-RLC circuit the phase angle between the voltage and current is given by

$$\tan \phi_1 = \frac{\omega L - \frac{1}{\omega C}}{R_L}$$

where $\omega = 2\pi f$.

Now solving this equation for C, and evaluating, yields the required value necessary to give a phase shift ϕ_1.

$$C = \frac{1}{\omega(\omega L - R_L \tan \phi_1)}$$

or

$$C = \frac{1}{2\pi f\,(X - R_L \tan \phi_1)} \; . \tag{54}$$

The fixed value of L is now given by

$$L = \frac{X}{2\pi f} \; . \tag{55}$$

At this point in the design the only remaining elements in the amplifier to evaluate are the two r-f chokes in the collector and base circuits. The only criteria in choosing their values are that their reactance at the operating frequency should be large compared to the a-c circuit impedance they supply and that their d-c resistance should be reasonably small to insure low loss and high efficiency.

EXCITATION REQUIREMENTS

The pulse-excitation requirements of the switching device depends on the type of active device used, vacuum tube or transistor, the operating frequency, and the over-all efficiency of the exciter and amplifier combination. All of these requirements can be related to the parameters of the switching device itself and the nonconduction angle ϕ_2.

From the tabular results shown in the appendix, K decreases, E increases, and the parameter a first increases and then decreases with increasing nonconduction angle ϕ_2. These parameters are related to the peak voltage and current switched by the active device by equation (48). A plot of the product V_pI_p for unit power output as a function of ϕ_2 is shown in Figure 7. For maximum device utilization a minimum value of this function should be chosen consistent with peak current requirements. From the normalized plot of V_pI_p it is seen that the smallest values occur at large nonconduction angles, but it is also apparent from equation (51) that the peak current the switching device must carry is very large with respect to the d-c input current. This is not a desirable situation since it requires large peak-excitation pulses to insure saturation of the switching device for a given power output.

Since the normalized plot of V_pI_p shows a broad minimum at

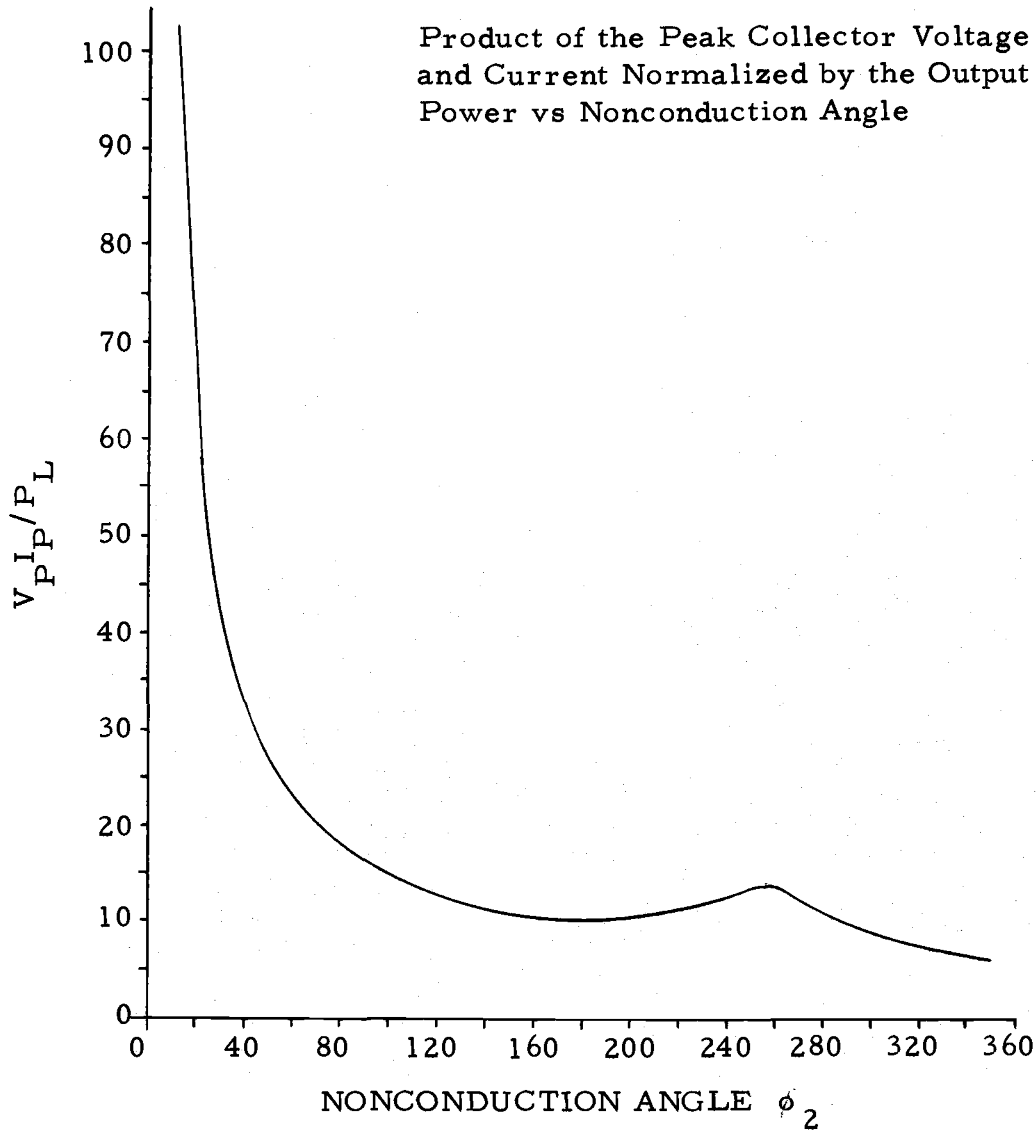

Product of the Peak Collector Voltage
and Current Normalized by the Output
Power vs Nonconduction Angle
100
90
80
70
60
50
40
30
20
10
0
$V_P I_P / P_L$
0
40
80
120
160
200
240
280
320
360
NONCONDUCTION ANGLE ϕ_2

Figure 7

about 180° and it is also seen that the peak current to d-c ratio is reasonable, this would appear to be a good operating point. This is very fortunate from a practical point of view since it is much easier to generate square wave excitation from a frequency-stable sine-wave source than it is to generate a highly asymmetrical pulse train, particularly at radio frequencies.

As is the case for a conventional class C amplifier the collector efficiency improves with increasing nonconduction angle as shown in Figure 8. This plot was obtained from equation (37) for the rather large value of R_s/R_L of one-tenth and for a more typical value of one-fiftieth. If the amplifier design requires the highest possible efficiency it may be beneficial in some cases to operate at angles greater than 180°, but the increased excitation requirements may offset the slight advantage gained in collector efficiency.

If the highest operating frequency from a given switching device with a given load resistance is required, then a nonconduction angle of about 235° should be used. This is evident from an inspection of equation (30) and the tabulation of the parameter a, given in the appendix, since a goes through a maximum at about 235°. The increase in maximum operating frequency gained over operation at 180° is not large though; it is about a 38 percent increase. It is found in practice that the high-efficiency mode of operation is usually limited to the low and medium h-f range due to the limitations

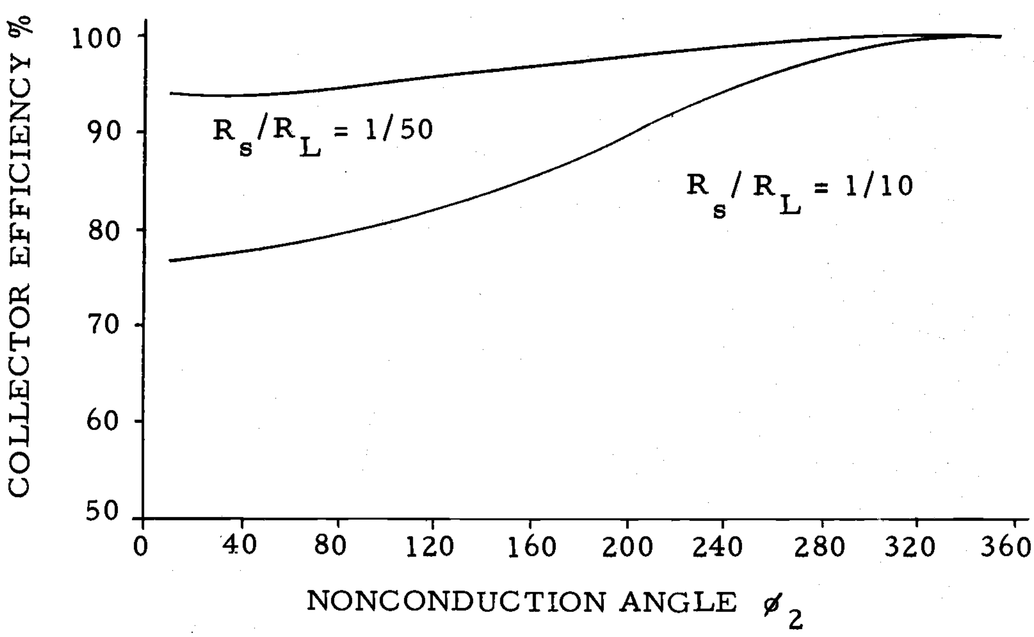


Figure 8. Collector Efficiency vs Nonconduction Angle

imposed by the active switching device. Due to the lower saturation resistance of transistors as compared to vacuum tubes, the load resistance that can be used efficiently is lower for a given minimum value of C_c, consequently higher operating frequencies are attainable with transistors.

EXAMPLE CALCULATION OF COLLECTOR EFFICIENCY

An example calculation of the collector efficiency of a high efficiency r-f power amplifier using a silicon-mesa 2N697 transistor will now be made. The purpose of this computation is to illustrate the high efficiency attainable and to compare the results with that attained earlier using the same transistor and approximate power levels used in the conventional class C circuit.

Selecting a nonconduction angle of 180° for ease of excitation and high efficiency and a power output of about two watts equation 48 yields

$$V_p I_p \simeq \frac{2 \times 0.181 \times 0.537 \times 1.537 \times 2}{2.92 \times 10^{-2}} \simeq 20$$

Let I_p = 0.4 amperes, the same value as was chosen for the conventional class C amplifier, then

$$V_p = 50 \text{ volts.}$$

Substituting the value of R_s equal to five ohms and the known ratio of V_p/I_p in equation (50) the value of R_L is computed.

$$R_L = \frac{(1 + \frac{1}{0.537})\, 2.92 \times 10^{-2} \times 50}{0.181 \times 0.4} - 0.788 \times 5 = 54 \text{ ohms.}$$

Now substituting the known and computed values into equation (37) gives the value of the collector efficiency.

$$\text{eff.} = \frac{1}{1 + 2(0.537)^2\, 2.37 \times \frac{5}{54}} = 0.89$$

It is seen that the collector efficiency for this amplifier is about six percent higher than the conventional class C amplifier operating at approximately the same power output. The improvement in efficiency is even greater if the saturation resistance of the transistor is lower than the value used. This can be easily verified by equations (6) and (37).

DESIGN EXAMPLE

To illustrate the design procedure previously developed, an r-f power amplifier was designed, built and tested in the laboratory. The amplifier uses a Clevite 3TX002 NPN silicon-power transistor driven by a small 2N697 NPN silicon-switching transistor. Figure 9 shows the schematic diagram and the component values for an operating frequency of about 500 kilocycles. Such a design could be used to advantage in a transmitter for a remote radio-navigation beacon or as an emergency-marine life-boat transmitter. In such applications high efficiency is of great importance since input power is usually limied and expensive to provide.

To start the design, let P_L equal about 20 watts and I_p equal to three amperes, which is well within the five ampere rating of the 3TX002. For a nonconduction angle equal to 180˜, which is provided very conveniently by the 2N697 driver transistor when it is excited by a low power sinusoidal source, the peak voltage across the power transistor is calculated from equation (48).

$$V_p = \frac{2EK(1 + K)\, P_L}{a\, I_p} = 10.2 \times \frac{20}{3} = 68 \text{ volts} .$$

The maximum collector to emitter voltage of the 3TX002 is reported as 80 volts, therefore the design value of 68 volts is a reasonable value.

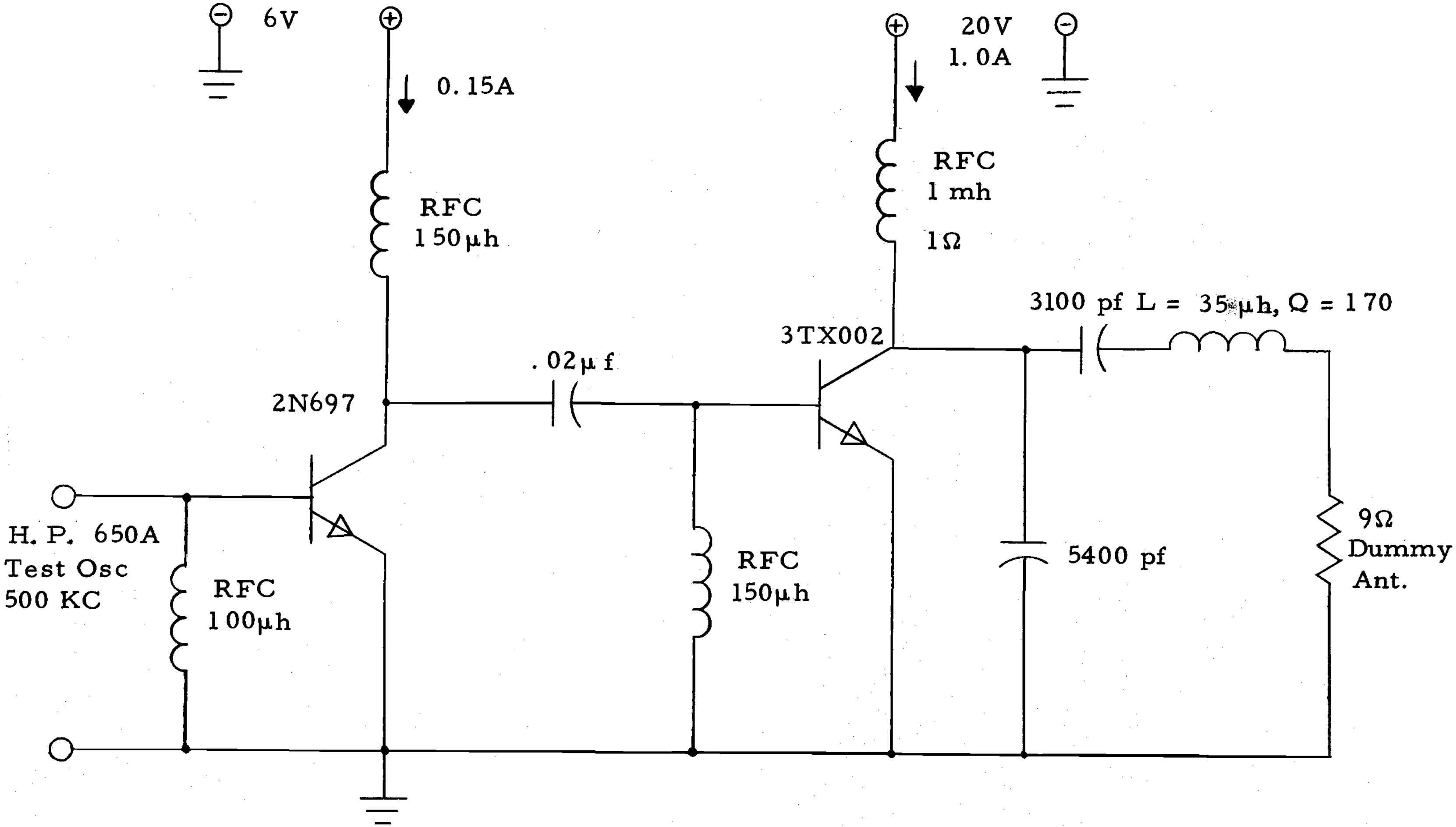


Figure 9. 500 KC R-F Power Amplifier and Exciter

The load resistance, which will include the series-loss resistance of the tuned-circuit inductor, is computed by applying equation (50).

$$R_L = \frac{(1 + \frac{1}{K})\, a\, V_p}{E\, I_p} - bR_s$$

$$R_L = \frac{(1 + \frac{1}{0.537})\, 2.92 \times 10^{-2} \times 68}{0.181 \times 3} - 0.788 \times 0.2$$

$$R_L = 10 \text{ ohms}$$

The saturation resistance of the 3TX002 was measured at about 0.2 ohms at three amperes of collector current. The collector efficiency of the power transistor may now be computed from equation (37).

$$\text{eff.} = \frac{1}{1 + 2K^2 D \frac{R_s}{R_L}} = \frac{1}{1 + 2(0.537)^2\, 2.37 \times \frac{0.2}{10}}$$

$$\text{eff.} = 0.97 \text{ or } 97 \text{ percent} .$$

The required d-c power-supply voltage and current will now be calculated. The current is given by solving equation (44) for I.

$$I = \frac{I_p}{1 + \frac{1}{K}} = \frac{3}{1 + \frac{1}{0.537}} = 1.05 \text{ amperes}$$

The voltage is given by solving equation (52).

$$V = \left(\frac{R_L}{2K^2} + DR_s\right) I = \left[\frac{10}{2(0.537)^2} + 2.37 \times 0.2\right] 1.05$$

$$V = 19 \text{ volts} .$$

It should be remembered that the value of V does not include the finite d-c voltage drop across the collector-circuit r-f choke. The choke used in the amplifier tested had a d-c resistance of about one ohm with an inductance of about one millihenry. The actual power-supply voltage used was 20 volts. This allows for about a one volt drop across the r-f choke.

The value of C_c is calculated by the use of equation (49).

$$C_c = \frac{E\, I_p}{(1 + \frac{1}{K})\, f\, V_p} = \frac{0.181 \times 3}{(1 + \frac{1}{0.537})5 \times 10^5 \times 68}$$

$$C_c = 5.6 \times 10^{-9} \text{ farads}$$

A fixed-mica capacitor with a value of about 5.4×10^{-9} farads was used to supplement the capacitance of the power transister and the circuit strays.

It is easy to see that the losses in the tuned-circuit inductor can easily exceed the power dissipation in the transistor unless a high-Q inductor is used. With this in mind, the author chose an Arnold-Company type A4-1570-570C ($\mu = 20$) Toroidal core which will give a high-Q inductor that is very compact in size. The finished inductor occupies a space less than one by two inches and yields an inductance of 35 microhenrys with a measured Q of 170 at 500 KC.

Solving equation 55 for the reactance X and computing

$$X = 2\pi f L = 2\pi 5 \times 10^5 \times 35 \times 10^{-6} = 110 \text{ ohms} .$$

From equation (53) the loaded-circuit Q is given by

$$Q = \frac{X}{R_L} = \frac{110}{10} = 11 .$$

This value of loaded Q will insure a good sinusoidal shape to the load current as the following oscillograms will show.

The value of the tuned-circuit capacitor C is obtained from equation (54).

$$C = \frac{1}{2\pi f(X - R_L \tan \phi_1)} \quad \frac{1}{2\pi 5 \times 10^5 (110 - 10 \tan 32.50°)}$$

$$C = \frac{1}{2\pi 5 \times 10^5 (110 - 6.36)} = 3.1 \times 10^{-9} \text{ farads.}$$

Figure 10 shows an oscillogram taken from a Tektronix type 545 oscilloscope equipped with a type CA dual-trace preamplifier and 10 megohm - 8 picofarad probes. The sinusoidal wave form displayed is of the inverted-load voltage as measured across an Ohmite-Company 9 ohm dummy-antenna resistor. The vertical scale factor for this wave is 10 volts per division, therefore, the peak-load voltage V_L is 17 volts. Also present in this same picture is the wave form of the collector to emitter voltage across the power transistor shown in proper relative phase to the inverted load voltage. The vertical-scale factor used in this second channel is 20 volts per division, thus the peak voltage across the transister V_p is 68 volts.

The measured direct current from the 20 volt power supply was 1.0 ampere, therefore, the total d-c input power is 20 watts. The input power to the collector circuit excluding the power lost in the one ohm r-f choke is thus 19 watts. This latter figure will be

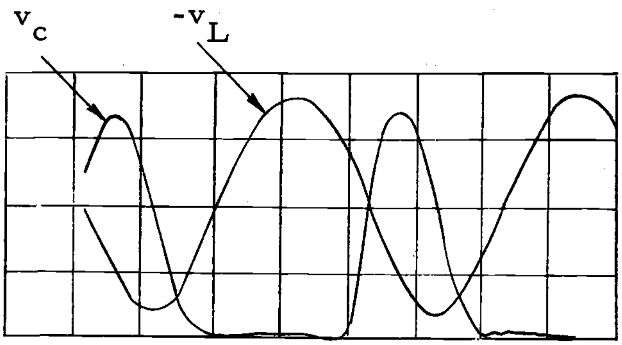


Figure 10

$-v_L$: Sinusoidal Load-Voltage Trace: 10 volts/division

v_c : Collector-Voltage Trace: 20 volts/division

Amplifier Adjusted for Maximum Efficiency used to conpute the collector efficiency.

The peak-load current I_L is given by the ratio of the measured peak-load voltage V_L to the measured load resistance of 9.0 ohms.

$$I_L = \frac{17}{9} = 1.9 \text{ amperes .}$$

The actual power delivered to the 9 ohm load resistance is therefore

$$\frac{V_L I_L}{2} = \frac{17 \times 1.9}{2} = 16 \text{ watts.}$$

The power lost in the series resistance of the inductor is given by

$$\frac{X}{2Q} I_L^2 = \frac{110}{2 \times 170} (1.9)^2 = 1.2 \text{ watts .}$$

The power lost in the 3100 pf capacitor for a measured dissipation factor Δ of 3.3×10^{-3} is given by

$$\frac{\Delta X I_L^2}{2} = \frac{3.3 \times 10^{-3} \times 10^2 (1.9)^2}{2} = 0.6 \text{ watts.}$$

The collector efficiency of the amplifier is now computed from preceding measured quantities.

$$\text{eff.} \simeq \frac{16 + 1.2 + 0.6}{19} = 0.94 \text{ or } 94 \text{ percent .}$$

The accuracy of the efficiency computed from the measured output and input powers is poor since it is the ratio of two numbers which are close to the same value. The author is convinced that the amplifier is operating very close to the theoretically determined efficiency since the case temperature of the power transistor was only 7.5° C above an ambient of 23.4° C with no attached heat radiator.

DETUNING EFFECTS

Figure 11 shows the effects of a four-percent increase in the excitation frequency on the load voltage and collector voltage of the previously described power amplifier. The scale factors are identical to those of Figure 10. In this detuned condition the d-c drawn from the power supply was measured at 0.5 amperes. The supply was held at 20 volts as before along with all circuit parameters. The total input power to the collector circuit is therefore 10 watts. The load voltage measured across the 9 ohm dummy-antenna resistor is 11 volts peak. The power delivered to this load is

$$\frac{(11)^2}{2 \times 9} = 6.7 \text{ watts} .$$

From these two power figures, it is obvious that the efficiency of the amplifier in this detuned condition has decreased markedly.

This decrease in efficiency can be explained by the fact that the collector-base junction of the power transistor is driven into a forward conduction condition following the nonconduction interval. For a short time during each cycle, the transistor carries a reverse current while a negative voltage appears across it. This negative voltage is shown in Figure 11 as the excursion extending below the bottom grid line on the oscillogram. The average time integral of the product of this negative voltage and the reverse current

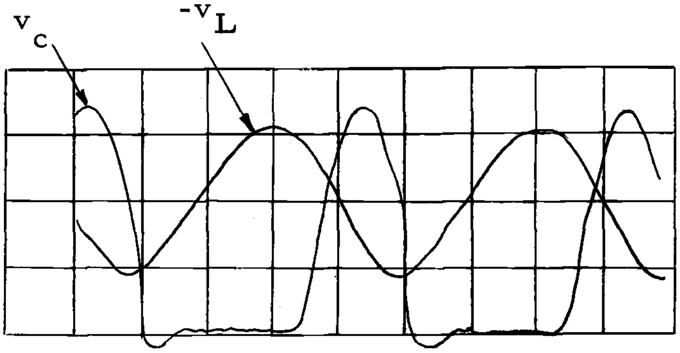


Figure 11

$-v_L$: Sinusoidal Load-Voltage Trace: 10 volts/division

v_c: Collector-Voltage Trace: 20 volts/division

Amplifier Excitation Frequency 4% High

represents additional power dissipated in the transistor.

It is very useful to know that tuning the amplifier for maximum power output does not correspond to the condition of maximum efficiency. This can be shown by analyzing Figure 12, the oscillogram of the same voltages to the same scale factors as previously defined. Maximum power output occurs at a frequency about four percent lower than that which gives maximum efficiency. Under this condition, as it is shown in Figure 12, the voltage across the transistor is positive and finite when it is switched on. The energy stored in C_c is proportional to this voltage squared, but since the transistor switches on and discharges C_c through itself each cycle this energy is lost to the transistor, consequently increasing the average power dissipation.

Under this condition of maximum output the measured d-c drawn from the power supply was 1.25 amperes. Again the power-supply voltage was held at 20 volts which gives a total input power of 12 x 1.25 or 25 watts. The peak-load voltage across the 9 ohm dummy-antenna resistor was 18 volts, thus the power delivered was

$$\frac{(18)^2}{2 \times 9} = 18 \text{ watts} .$$

Again, it is observed that the efficiency of the amplifier in this detuned condition is noticeably reduced. The collector-circuit efficiency, which includes the losses in the r-f choke and the L-C

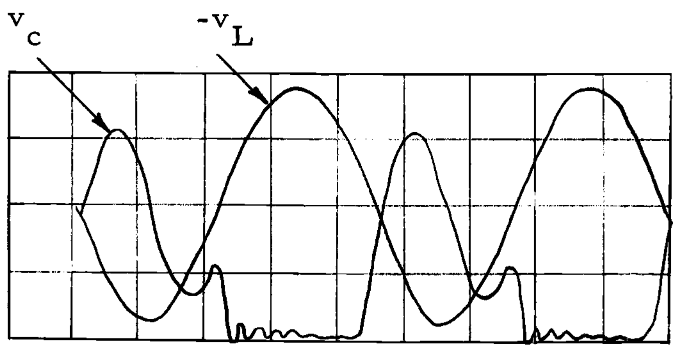


Figure 12

$-v_L$: Sinusoidal Load Voltage Trace: 10 volts/division

v_c: Collector Voltage Trace: 20 volts/division

Amplifier Excitation Frequency: 4% Low

series-tuned circuit, in this condition is 72 percent as compared to 80 percent for the maximum efficiency case. Figure 13 shows in more detail how the collector-circuit efficiency changes with the excitation frequency.

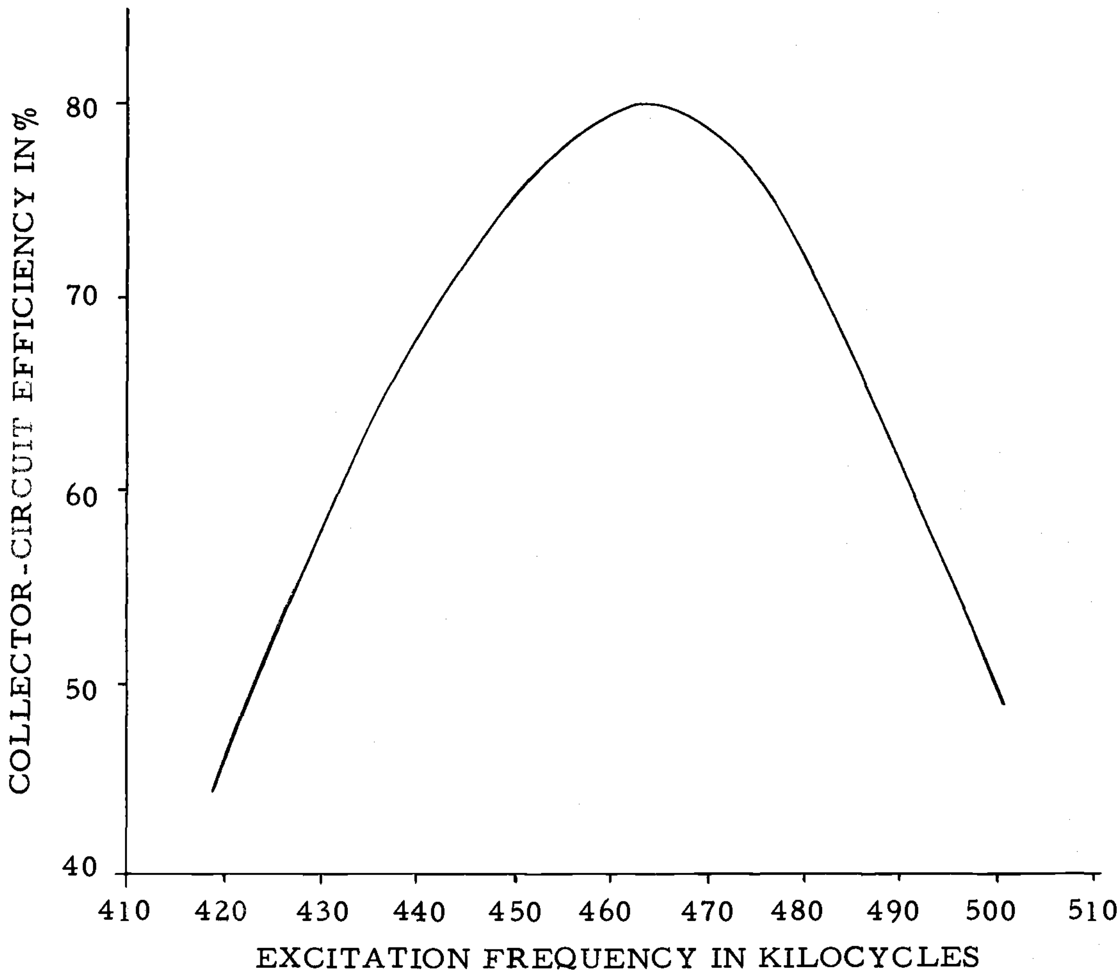


Figure 13. Collector-Circuit Efficiency vs Excitation Frequency

SUMMARY

The most outstanding feature of the high-efficiency r-f power amplifier is the high overall efficiency that can be obtained with a collector-current conduction angle of 180°. As compared to the class C amplifier, which requires a conduction angle of typically less than 90° to reach comparable collector efficiencies, the base-current excitation requirements and peak-collector current capability of the switching transistor can be reduced for a given power output.

As was demonstrated earlier the collector efficiency of the new high-efficiency amplifier using a small silicon-mesa 2N697 switching transistor exceeds the value obtained with the same transistor and power output used in a conventional class C circuit. In this particular example the high-efficiency amplifier gave an 89 percent collector efficiency with a conduction angle of 180° and the class C gave a value of 83 percent with a conduction angle of 90°. It is interesting to note that the improvement in efficiency is even greater if the saturation resistance of the transistor is lower than the value used. The value used in this calculation was the maximum limit reported by the manufacturer.

The 20-watt high-efficiency r-f amplifier which was built and tested in the laboratory gave a measured performance that was within the accuracy of the measuring instruments. The design collector efficiency with a R_s/R_L ratio of one-fiftieth and a conduction angle of 180° was 97 percent. The value obtained from the laboratory measured parameters was 94 percent.

The effects of detuning on this amplifier are given in the previous chapter. A four percent increase in the excitation frequency from that which gave maximum efficiency resulted in a decrease in collector circuit efficiency from 80 percent to 67 percent. A four percent decrease in the excitation frequency resulted in maximum output power in the load, but under this condition the collector efficiency decreased to 72 percent. Proper tuning of the amplifier should result in the collector voltage just reaching zero following the non-conduction period. This condition does not in general result in maximum power output but it does assure maximum efficiency.

Another important advantage of this power amplifier that has not been previously mentioned is the input power decreases with increasing tuning error, except for the modest peak that occurs for a small low frequency error. Although the output power decreases faster, thus the efficiency decreases, the remaining power the switching device must dissipate is much less than would be the case for a conventional class C amplifier of equivalent output power. This is true since the input power to a fully-loaded conventional class C amplifier is approximately constant whether it's in or out of tune, but the collector efficiency varies about like that of the high-efficiency amplifier just described.

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APPENDIX

ϕ_2 (degrees)	ϕ_1 (radians)	K	ϕ_3 (radians)	a	b	D	E
5	-1.513	.99958	.029	.40783E-07*	.10000E+01	.15004E+01	.26123E-05
10	-1.454	.99831	.058	.65136E-06	.10000E+01	.15017E+01	.20914E-04
15	-1.396	.99620	.087	.32877E-05	.10000E+01	.15038E+01	.70675E-04
20	-1.338	.99324	.116	.10348E-04	.99999E+00	.15068E+01	.16782E-03
25	-1.280	.98946	.146	.25128E-04	.99998E+00	.15107E+01	.32854E-03
30	-1.222	.98484	.175	.51766E-04	.99995E+00	.15155E+01	.56931E-03
35	-1.163	.97940	.204	.95164E-04	.99990E+00	.15212E+01	.90706E-03
40	-1.105	.97314	.234	.16090E-03	.99981E+00	.15279E+01	.13592E-02
45	-1.047	.96608	.263	.25512E-03	.99966E+00	.15355E+01	.19438E-02
50	-.988	.95823	.293	.38443E-03	.99943E+00	.15442E+01	.26796E-02
55	-.930	.94959	.322	.55578E-03	.99908E+00	.15540E+01	.35861E-02
60	-.871	.94018	.352	.77630E-03	.99860E+00	.15649E+01	.46839E-02
65	-.813	.93002	.382	.10532E-02	.99793E+00	.15769E+01	.59946E-02
70	-.754	.91912	.412	.13935E-02	.99703E+00	.15901E+01	.75410E-02
75	-.695	.90750	.442	.18041E-02	.99586E+00	.16046E+01	.93472E-02
80	-.637	.89517	.472	.22915E-02	.99435E+00	.16204E+01	.11439E-01
85	-.578	.88216	.503	.28616E-02	.99245E+00	.16377E+01	.13844E-01
90	-.519	.86848	.533	.35195E-02	.99009E+00	.16563E+01	.16593E-01
95	-.460	.85416	.564	.42697E-02	.98720E+00	.16766E+01	.19715E-01
100	-.400	.83921	.595	.51156E-02	.98372E+00	.16984E+01	.23247E-01
105	-.341	.82367	.627	.60593E-02	.97956E+00	.17219E+01	.27224E-01
110	-.282	.80755	.658	.71020E-02	.97466E+00	.17473E+00	.31688E-01
115	-.222	.79087	.690	.82435E-02	.96892E+00	.17745E+01	.36682E-01
120	-.163	.77368	.722	.94819E-02	.96228E+00	.18038E+01	.42254E-01
125	-.103	.75599	.754	.10814E-01	.95466E+00	.18352E+01	.48457E-01
130	-.043	.73783	.787	.12236E-01	.94599E+00	.18689E+01	.55349E-01
135	.017	.71922	.820	.13740E-01	.93618E+00	.19049E+01	.62994E-01

*Note: E-07 = 10^{-7}

ϕ_2 (degrees)	ϕ_1 (radians)	K	ϕ_3 (radians)	a	b	D	E
140	.078	.70021	.853	.15320E-01	.92518E+00	.19435E+01	.71461E-01
145	.138	.68081	.887	.16965E-01	.91292E+00	.19848E+01	.80830E-01
150	.199	.66107	.921	.18665E-01*	.89935E+00	.20290E+01	.91186E-01
155	.259	.64100	.955	.20407E-01	.88442E+00	.20763E+01	.10263E+00
160	.320	.62065	.990	.22177E-01	.86809E+00	.21268E+01	.11527E+00
165	.382	.60004	1.025	.23960E-01	.85034E+00	.21809E+01	.12922E+00
170	.443	.57921	1.061	.25741E-01	.83114E+00	.22387E+01	.14463E+00
175	.505	.5582	1.097	.27500E-01	.81049E+00	.23006E+01	.16165E+00
180	.567	.53703	1.134	.29221E-01	.78840E+00	.23669E+01	.18045E+00
185	.629	.51575	1.171	.30884E-01	.76489E+00	.24378E+01	.20125E+00
190	.692	.49438	1.209	.32470E-01	.74000E+00	.25138E+01	.22427E+00
195	.754	.47297	1.247	.33959E-01	.71378E+00	.25954E+01	.24978E+00
200	.818	.45156	1.286	.35333E-01	.68629E+00	.26829E+01	.27809E+00
205	.881	.43017	1.326	.36570E-01	.65762E+00	.27769E+01	.30955E+00
210	.945	.40885	1.366	.37654E-01	.62787E+00	.28781E+01	.34458E+00
215	1.009	.38763	1.407	.38565E-01	.59714E+00	.29870E+01	.38366E+00
220	1.073	.36656	1.449	.39288E-01	.56557E+00	.31046E+01	.42739E+00
225	1.138	.34567	1.491	.39808E-01	.53330E+00	.32316E+01	.47643E+00
230	1.204	.32500	1.535	.40110E-01	.50048E+00	.33691E+01	.53161E+00
235	1.269	.30459	1.579	.40184E-01	.46728E+00	.35183E+01	.59390E+00
240	1.336	.28449	1.624	.40021E-01	.43389E+00	.36806E+01	.66449E+00
245	1.402	.26472	1.670	.39615E-01	.40050E+00	.38576E+01	.74483E+00
250	1.470	.24533	1.717	.38961E-01	.36729E+00	.40512E+01	.83670E+00
255	1.537	.22637	1.766	.38062E-01	.33449E+00	.42637E+01	.94233E+00
260	-1.536	.20787	-1.745	.36919E-01	.30229E+00	.44978E+01	.10000E+01
265	-1.467	.18988	-1.658	.35539E-01	.27090E+00	.47568E+01	.10000E+01
270	-1.397	.17244	-1.571	.33935E-01	.24054E+00	.50448E+01	.10000E+01

*Note: E-01 = 10^{-1}

(degrees)	(radians)	K	(radians)	a	b	D	E
275	-1.327	.15558	-1.484	.32121E-01	.21139E+00	.53666E+01	.10000E+01
280	-1.256	.13936	-1.396	.30115E-01	.18367E+00	.57286E+01	.10000E+01
285	-1.185	.12382	-1.309	.27941E-01	.15755E+00	.61385E+01	.10000E+01
290	-1.113	.10899	-1.222	.25625E-01	.13319E+00	.66065E+01	.10000E+01
295	-1.039	.09493	-1.134	.23200E-01	.11086E+00	.71457E+01	.10000E+01
300	-.965	.08167	-1.047	.20699E-01	.90354E-01	.77738E+01	.10000E+01
305	-.891	.06926	-.960	.18160E-01	.72087E-01	.85147E+01	.10000E+01
310	-.815	.05774	-.873	.15625E-01	.56017E-01	.94020E+01	.10000E+01
315	-.738	.04715	-.785	.13136E-01	.42171E-01	.10484E+02	.10000E+01
320	-.661	.03754	-.698	.10739E-01	.30535E-01	.11834E+02	.10000E+01
325	-.582	.02894	-.611	.84792E-02	.21052E-01	.13566E+02	.10000E+01
330	-.502	.02140	-.524	.64025E-02	.13617E-01	.15870E+02	.10000E+01
335	-.421	.01494	-.436	.45532E-02	.80779E-02	.19089E+02	.10000E+01
340	-.339	.00961	-.349	.29729E-02	.42302E-02	.23910E+02	.10000E+01
345	-.256	.00543	-.262	.16992E-02	.18210E-02	.31931E+02	.10000E+01
350	-.172	.00242	-.175	.76412E-03	.54916E-03	.47953E+02	.10000E+01
355	-.087	.00061	-.087	.19242E-03	.69674E-04	.95975E+02	.10000E+01

*Note: E-01 = 10^{-1}