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Development of a Radar Pulse Modulator

Eckard Friedrich Natter
University of Central Florida

DEVELOPMENT OF A RADAR PULSE MODULATOR

BY

ECKARD F. NATTER

RESEARCH REPORT

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ABSTRACT

This report summarizes the design of a 25kW peak power pulse modulator for an airborne medium-range weather radar.

It is the purpose of the modulator to collect and store energy over a certain time period and to form this energy into a short, high-power pulse.

The modulator is required to drive a coaxial magnetron with a pulse of 5kV at 5A. System considerations make a pulse width of 3.5us and a repetition rate of 99Hz necessary.

The pulse is generated in a line-type pulse circuit which utilizes an SCR as a switching device. It is shown that a solid-state modulator can use a commercial grade SCR for the pulse generation. Although currents of 100A are switched, the instantaneous power dissipation in the SCR is reduced significantly through the use of a saturable delay reactor.

A pulse transformer is used to achieve maximum power transfer from the modulator to the magnetron. The pulse transformer is insulated with a semi-rigid epoxy. Corona generation is avoided by limiting the voltage gradients in the insulation to 80V/mil.

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DEVELOPMENT OF A RADAR PULSE MODULATOR

1.1 PROBLEM PRESENTATION

The task is to develop a radar modulator for an airborne X-Band weather radar with a 200-mile range. This requires a nominal RF-output pulse of 10kW and 3.5µs duration at a repetition rate of 99Hz. The block diagram of the radar is shown below.
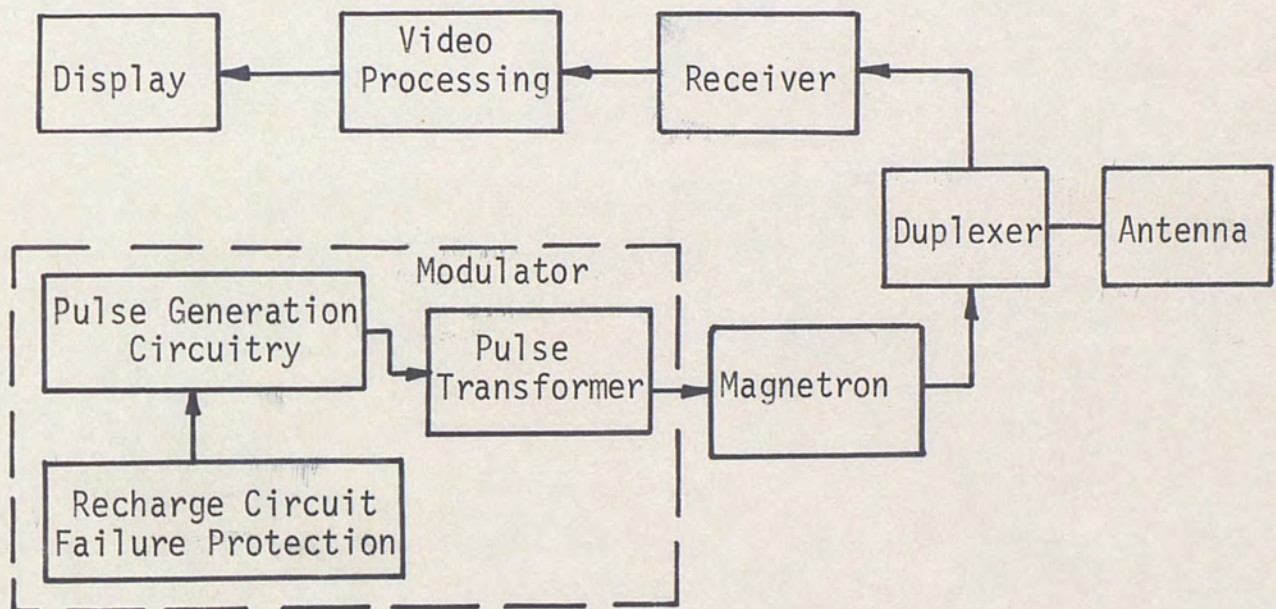
Figure 1. Block Diagram of Radar System

Reliability considerations make a coaxial magnetron necessary. The chosen magnetron (L5362)* requires a drive of 5kV at 5A. The modulator under discussion consists of the following three areas:

  a) Impedance matching to magnetron by means
      of a pulse transformer.

*Litton, Coaxial Magnetron L5362

b) Pulse generation circuitry.

c) Charging circuit and failure protection
of modulator from arcing and misfire of
magnetron.

1.2 PREDESIGN CONSIDERATIONS

The following guidelines are to be considered during the
design phase of the modulator.

1. Solid-state design.

2. Low manufacturing costs and easy assembly
through use of printed circuit board layout.

3. Light-weight construction.

4. Oil is not permitted as insulation material
since possible leakage will reduce the
reliability.

5. The temperature range of operation is from -55°C
to +100°C with storage from -55°C to +125°C.
Maximum operating altitude is 55,000 feet.

The modulator will drive a coaxial magnetron as shown in
Figure 1. A magnetron is basically a magnetically biased vacuum
diode that in its V-I characteristic is similar to a zener diode.
The magnitude of the bias field influences the "knee voltage" of
the magnetron, thus, changing the effective impedance. Figure 2
shows the V-I characteristic of the magnetron used.
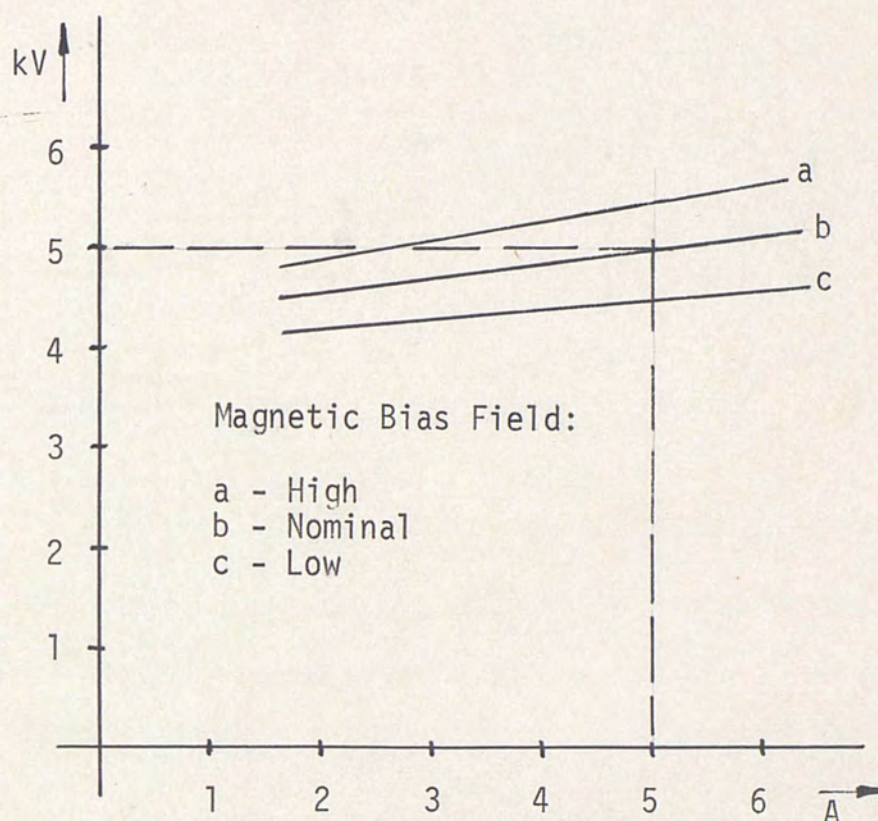
Figure 2. Magnetron Characteristic

The following problem areas are common under magnetrons and have to be considered to protect the modulator and assure reliable operation.

1. The cathode and filament are connected together, and since the anode of the magnetron is grounded, it has to be driven with a negative pulse at the cathode. Thus, the heater voltage "rides" on the pulse. Therefore, special attention must be given to the way the heater voltage is applied.

2. With increasing life, the tendency to misfire in- creases strongly. Misfire occurs when the pulse drive reaches its nominal value and the magnetron fails to conduct.

3. Arcing between anode and cathode can occur when
    the magnetron is misfiring or if the magnetron
    is subjected to a large mismatch in its micro-
    wave output.

1.3 INITIAL DESIGN

It is the purpose of the modulator to collect and store
energy over a certain time period and to form this energy into a
short and high-powered periodic pulse.

The three sections of the modulator are:

a) The pulse transformer for impedance
    matching.

b) The pulse generation circuitry also
    called pulse-forming network.

c) The charging circuit that charges the
    pulse-forming network and provides fail-
    ure protection.

The input impedance of the magnetron has to be matched to a
suitable modulator output impedance to assure maximum power
transfer. A pulse transformer is a solution for this problem.
A dual secondary winding, if connected as in Figure 3, can be
used to apply the heater voltage and as a series resistor for the
heater of the magnetron to limit a high switch-on current. This
switch-on current is large since the cold filament of the heater
has only approximately 1/8 of the resistance it has in operating
condition.
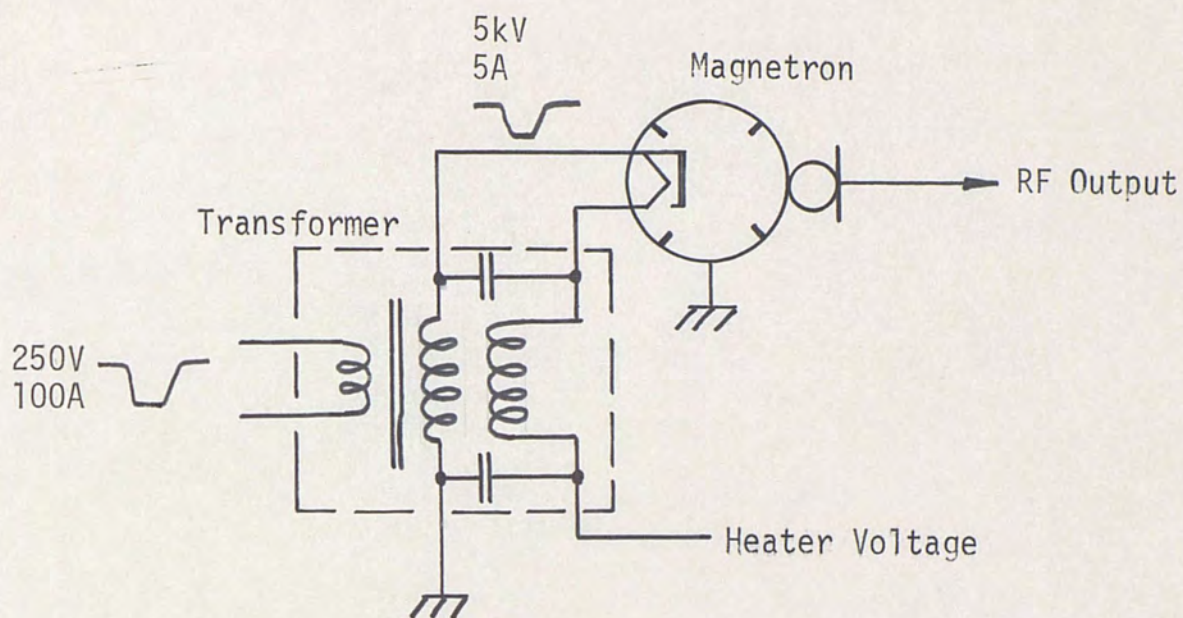
Figure 3.   Pulse Transformer and Magnetron

Since there are no solid-state devices available that can switch 5kV, the pulse transformer is also used as a voltage transformer to reduce the pulse voltage. A 20 to 1 step-up ratio is an optimum. This ratio will bring the primary pulse down to 250V and 100A. Neglecting losses, the primary pulse requirements of the transformer are then 250V and 100A. This is a primary impedance of

$$\frac{250V}{100A} = 2.5\ \Omega$$

For the pulse-forming circuitry, two different concepts are considered--a magnetic pulse-shaping circuit versus a line-type pulse generator. The two circuits differ in the way the pulse energy is stored. The magnetic modulator stores the energy in the magnetic field of an inductor. The line-type modulator stores

the energy in the electric field of a capacitor. Weight and
volume considerations eliminated the magnetic modulator.

A line-type pulse generator is shown in Figure 4.
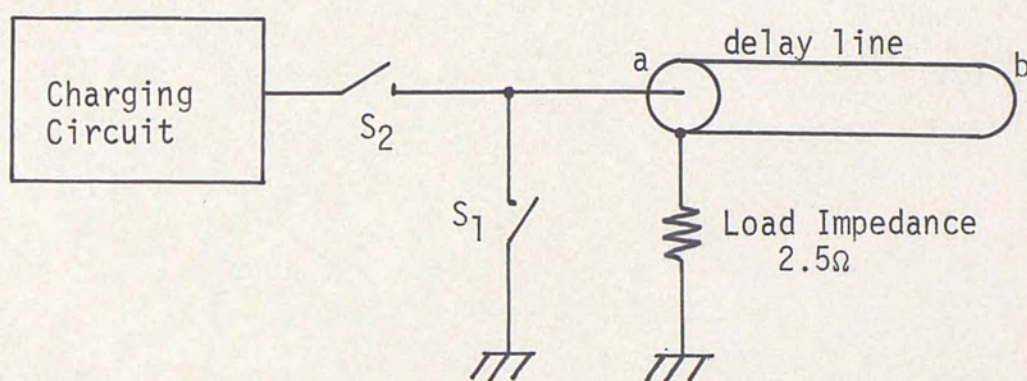
Figure 4. Line-type Pulse Generator

When the switch S_1 closes, the voltage in the charged delay
line will divide between the load resistor and the characteristic
impedance of the line. If the load impedance equals the charac-
teristic impedance of the delay line, the voltage will divide
equally. The pulse width is equal to twice the delay time of the
line. This can be understood by imagining that when the switch
S_1 closes, a step function is introduced at the input of the
line (point a). This step function travels down the line, is
reflected at the open end (point b), and travels back to the
input (point a) where no reflection occurs. Thus, the step-
function needs twice the delay time to deplete the line of its
charge.

Distributed delay lines can only be used for relative short
pulses because of their large volume per unit delay. A lumped

element delay line has a much lower volume per unit delay, however, the number of sections used is critical to rise and fall time of the pulse. The more sections used, the steeper the rise and fall times, and the lower is the ripple on the pulse; thus, volume, weight, and cost are increased.

The charging voltage of the line is estimated (not considering losses) to be

$$2 \times 250V = 500V$$

which is easily switched by an SCR. The SCR used is a 2N4444 with a 600V forward breakdown rating.

Failure considerations cover three different phenomena: latch-up of switching SCR, misfire, and arcing of magnetron.

Latch-up of the modulator occurs when the main switching SCR fails to turn off after the pulse-forming network is discharged. This problem is avoided if the recharge of the delay line occurs after the pulse loop has settled and the main switching SCR has turned off. The recharge is delayed through switch S_2 in Figure 4. To prevent a latch-up, S_1 and S_2 are not closed simultaneously. Switch S_2 is the recharge delay.

Misfire of the magnetron (failure to conduct) results in at least twice the voltage being generated by the pulse-forming network. This problem is commonly avoided by placing a gas filled spark gap across the magnetron. The firing voltage of the spark gap has to be considerably larger than the nominal "knee voltage" of the magnetron. Rather than using a spark gap, the pulse transformer was designed to withstand 2.5 times the nominal voltage, 13kV.

In case the magnetron misfires, the pulse transformer should saturate as quickly as possible but without affecting the pulse shape. The saturated transformer, being nearly a short, will then reverse charge the pulse-forming network. At this point, a reverse voltage limiter will absorb the charge before the pulse-forming network is recharged. This is, of course, only possible with the above described recharge delay.

Arcing inside the magnetron is, for the modulator, the same as a short and covered by the above circuitry.

Switch S_2 is realized with another SCR; however, since S_2 is riding on the supply voltage, the SCR is turned on through a trigger transformer.

The delay between "firing" the main SCR (S_1) and the control SCR (S_2) is realized through an R-C network which differentiates the trigger pulse to the modulator. The leading edge of the trigger pulse turns the main SCR on and the trailing edge turns the control SCR on.

The applied trigger pulse is 250μs long. This determines the delay between the conduction of both SCR's.

There still exists the possibility that S_1 will conduct if a transient voltage is induced on the trigger input when the current through S_2 is above the holding current of SCR-S_1. This is a latch-up condition. Both SCR's are conducting and can only be turned off if the current path is interrupted. A relay serves this purpose. The relay coil serves as a current sensor and opens the contacts S_3 of Figure 5 if the current drawn from the power supply exceeds .15A.

Figure 5 shows a simplified schematic with the appropriate voltage wave forms.

Figure 5.   Simplified Schematic and Voltage Waveforms of Modulator

2.1 EFFECTS OF PULSE TRANSFORMER PARAMETERS ON PULSE SHAPE

The effect of the pulse transformer (with a biased diode or magnetron as load) on pulse shape has to be viewed in three different time segments. This is necessary since the load is nonlinear. The three time intervals are as follows:

a) Rise time of pulse until load conducts.

b) Top of pulse with low dynamic load impedance.

c) Tail and backswing of pulse due to the energy stored in the magnetic field of the core, L_e.

An electromagnetic transformer is reasonably well approximated in the equivalent circuit of Figure 6. Losses of primary and secondary windings (copper losses) are neglected.
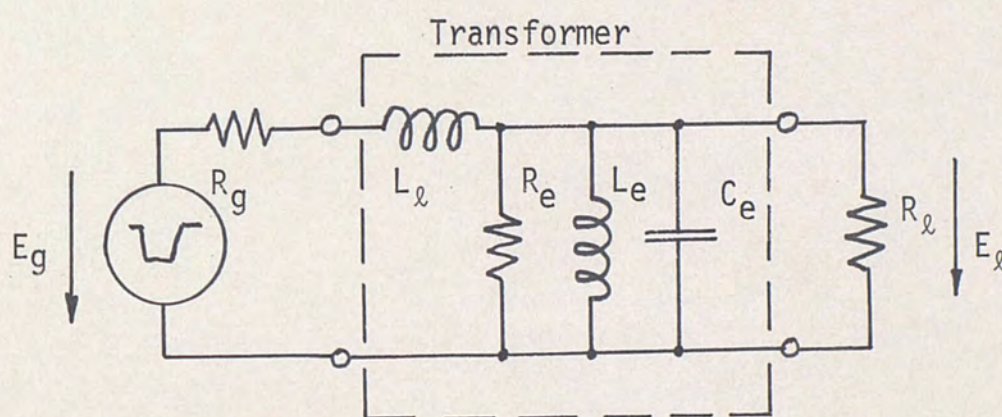

Figure 6. Equivalent Circuit for Pulse Transformer

E_g -- Ideal Negative Pulse Source

R_g -- Internal Impedance of Source

L_ℓ -- Lumped Leakage Inductance

R_e -- Equivalent Resistor Representing

Losses in Primary and Secondary

Winding and Core

L_e -- Primary Inductance

C_e -- Lumped Equivalent of Distributed

Capacitance Between Windings.

R_ℓ -- Load Impedance

E_e -- Output Voltage

The rise time of the pulse on a magnetron or biased diode load is considered for a transformer in which the effect of R_e is negligible compared with R_g and where the rise time is so short compared with the pulse length that L_e is also negligible. This leaves the equivalent circuit of Figure 7.
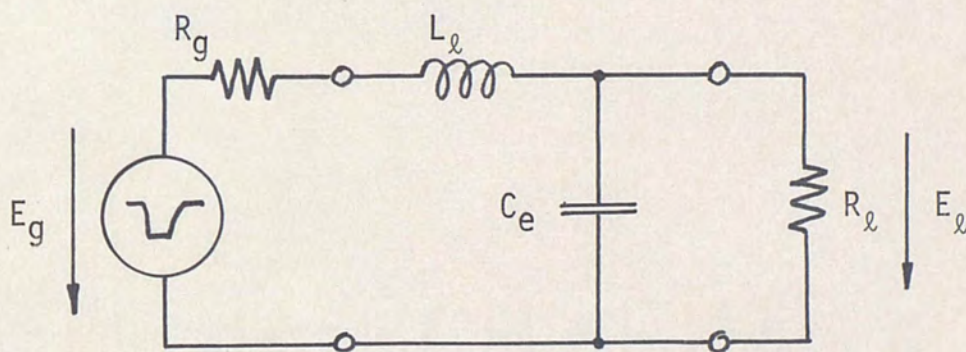

Figure 7. Equivalent Circuit For Rise Time

Since R_ℓ is usually much larger than R_g, the output voltage will rise quickly until

$$\frac{E_\ell}{E_g} = 1$$

Up to that point, only R_g, L_ℓ, and C_e determine the thransfer function because of the non-conducting magnetron with its high dynamic impedance.   Figure 8 shows the transfer function.

$$\frac{E_\ell}{E_g} = \frac{R_\ell}{R_g+R_\ell}\left[1-e^{-\frac{t}{\tau}}\cos\left(\frac{t}{\sqrt{L_\ell C_e}}\right)\right]\Bigg|_{o<\frac{E_\ell}{E_g}<1}$$

$$\frac{1}{\tau} = \frac{R_g}{2L_\ell} + \frac{1}{2C_e R_\ell}$$
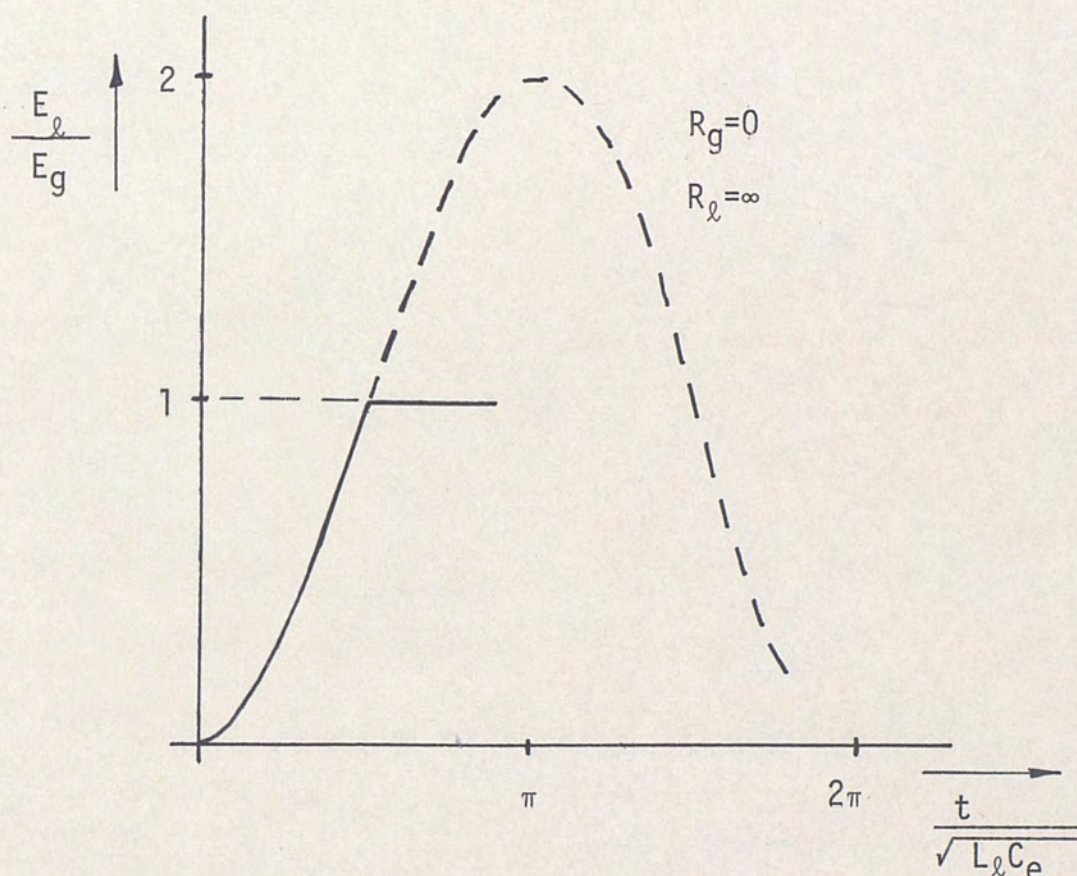

Figure 8.   Voltage Rise of Pulse

<u>The top of the pulse</u> starts when the transfer function becomes unity. At that point, the biased diode or magnetron conducts and R_e becomes less than R_g. Now a new equivalent diagram is valid, Figure 9.
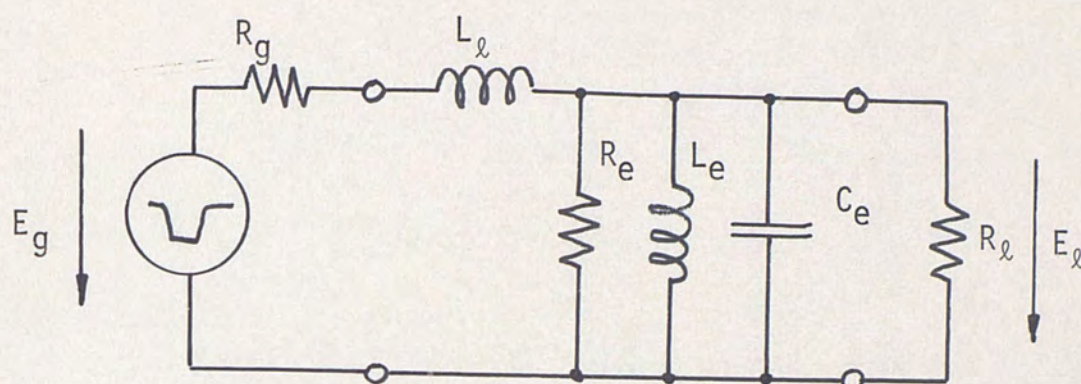
Figure 9. Equivalent Circuit for Top of Pulse

Since R_g is larger than zero, L_e will introduce a pulse decay with a time constant of

$$\tau = \frac{L_e \left(R_g + R_\ell \right)}{R_g \quad R_\ell}$$

The nonlinear load will make this decay much more evident in the current pulse through the load than in the voltage across it. The current pulse is important since it is equivalent in shape to the envelope of the RF pulse emitted by the magnetron.

C_e and L_e will introduce some ringing to the pulse, again much more visible on the current pulse, but the low value of R_ℓ will make this ringing decay with an exponential function of time constant τ_1.

$$\tau_1 = 2 \frac{L_e}{R_\ell}$$

The tail and backswing of the pulse is determined by the amount of energy stored in L_e. The magnetic field in L_e is represented by the flux density B in Figure 10. The energy stored in the magnetic field of the transformer is equal to

$$E = \frac{1}{2} L_e I_{max}^2$$

where I_{max} is the current through L_e at the end of the pulse. The magnetic field that is building up during the pulse has to return from its maximum value, B_{max}, to a minimum value of B_{min} during the interpulse period. This change in flux will induce a voltage inverse to the pulse polarity. The energy stored in the magnetic field has to be dissipated during the pulse interval making the stored energy important to the efficiency of the pulse transformer. The equivalent circuit for this time period is shown in Figure 10c.
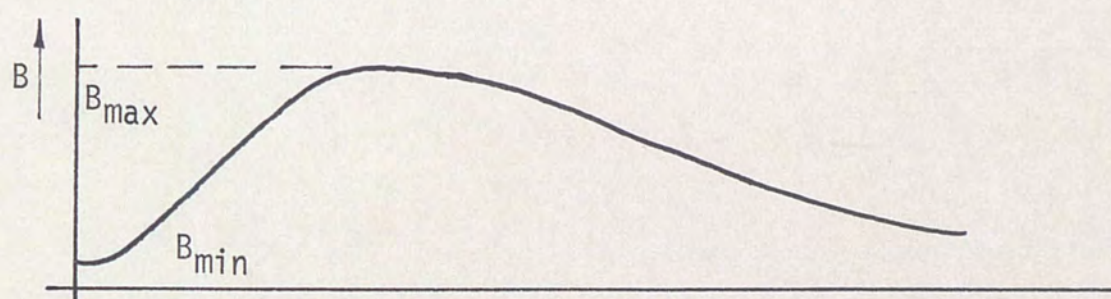

Figure 10a. Magnetic Field in Transformer Core
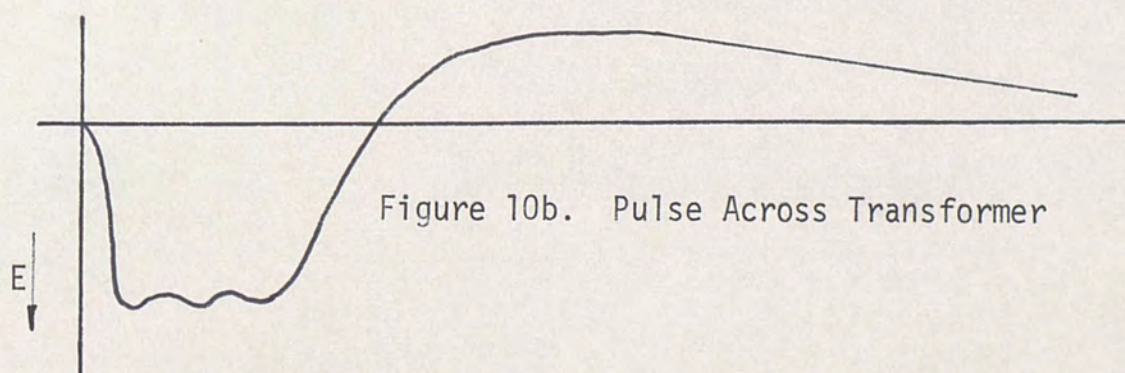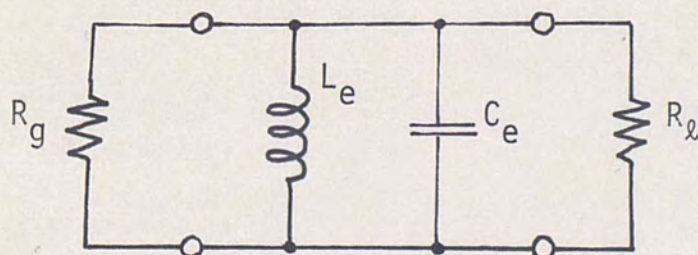

Figure 10b. Pulse Across Transformer

Figure 10c. Equivalent Circuit for Fail and Back Swing of Pulse.

L_e and C_e form a ringing circuit that is damped by R_g. R_ℓ can again be neglected since it is large compared with R_g.

2.2 PULSE TRANSFORMER

In Chapter 1, it is mentioned that a step-up ratio of 1 to 20 is considered the optimum value for this transformer using a hypersil* core. The core that is used was chosen after approximately a dozen different design tries using different cores. The following parameters had to be met.

1. Secondary maximum voltage, 13kV under fault condition.

2. Maximum voltage gradient in insulation, 80V/mil for nominal operation.

3. Rise time with resistive load, 100-300 ns.

4. Fall time with resistive load less than 1us.

5. Overshoot, less than 15%.

6. Secondary copper resistance total of 1 ohm at 25°C, each winding .5 ohms.

7. Primary copper resistance less than 30m ohms.

8. Core must saturate with 500V applied after approximately 5us.

9. Magnetization current after 3.5us with 250V applied must be less than 20A.

In the first step in the design of the transformer only tape wound C-core are considered. They are the easiest to assemble. Torroidal cores will give the highest permeability, however, they are very difficult to wind especially if large wire sizes are involved. Further, the uniform application of insulation material on torroidal cores is nearly impossible.

*Hypersil--trade name for grain oriented transformer lamination.

From the Massachusetts Institute of Technology Handbook,
Pulse Generators[1], a certain winding or interleaving pattern is
chosen to give the strongest coupling between primary and secon-
dary winding. Figure 11 shows this pattern.

Primary
Windings

Secondary
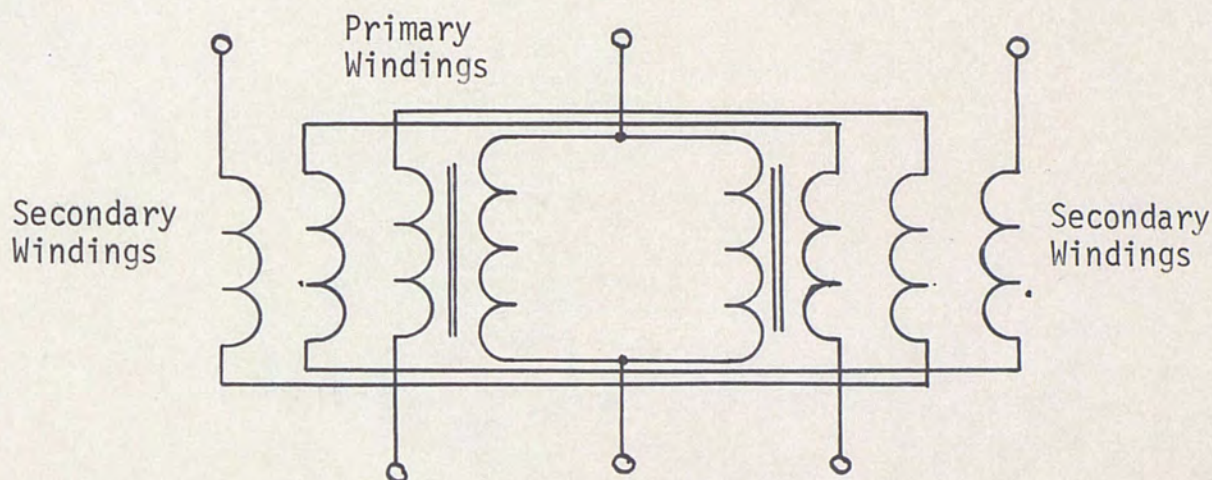Windings

Secondary
Windings

Figure 11. Winding Pattern

This winding configuration gives some information about the
core to be used. Each secondary layer has approximately

$$\frac{1\Omega}{6 \text{ layers}} = .18\Omega/\text{layer}$$

of copper resistance since, as mentioned in Chapter 1, both secon-
dary windings are used as series resistors to the magnetron fila-
ment.

The transformer will be placed on each of the two legs of
the C-core. This makes it necessary to have, as closely as possi-
ble, the same number of turns on both primary windings. An experi-
ment showed that the closest tolerance possible for production was

[1]G. Norris Glascoe and J. V. Lebacqz, Pulse Generators,
Massachusetts Institute of Technology Radiation Laboratory Series
(Lexington, Mass.: Boston Technical Publishers, Inc., 1964), p. 518.

±.25 turns. Therefore, six turns are chosen as minimum for the primary winding.

The maximum magnetization current is then used to figure the minimum inductance six turns should give on a certain core:

$$L_{min} = \frac{250V \times 3.5\mu s}{20A} = 43\mu H$$

An air gap has to be considered in this calculation. The air gap is necessary to increase the uniformity between different batches of cores. Since the transformer is to be used in a single-sided pulse application, the residual magnetism after each pulse recurrence (B_{min}) is of importance. A reduction of inductance through an air gap will reduce the remanance by approximately the same ratio.

The maximum voltage between two secondary layers of the nominal pulse is

$$\frac{2}{3} \times 5kV = 3.3kV$$

After several experiments, 5mil Kraft paper is chosen as insulation material because of its porosity and its ability to absorb impregnating compounds. From the voltage gradient and the maximum voltage between layers, the required insulation space is found:

$$\frac{3.3kV}{80V/mil} = 41mil \text{ per layer}$$

This requires eight wraps of 5mil Kraft paper between layers.

A "dry transformer" is pictured in Figure 12. The term "dry transformer" implies a wound specimen that is not yet impreg-

nated and potted. A single step impregnating and potting proced-
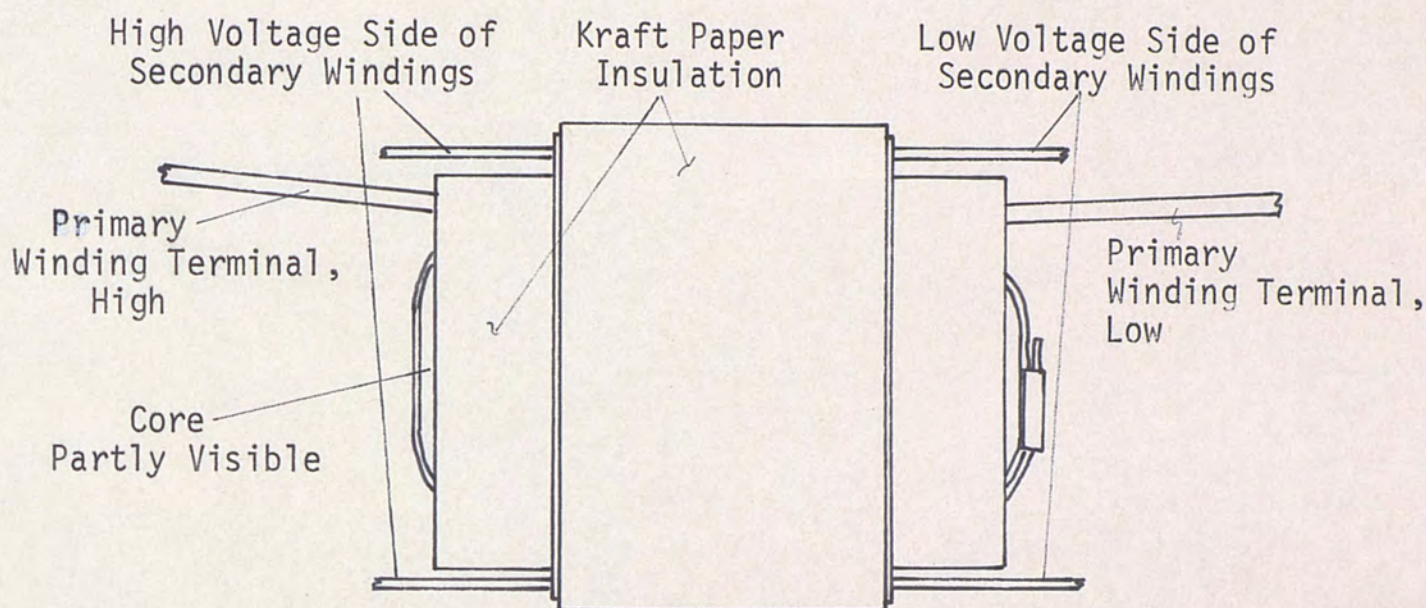ure was developed utilizing a semi-rigid epoxy.*

High Voltage Side of   Kraft Paper   Low Voltage Side of
Secondary Windings   Insulation   Secondary Windings

Primary
Winding Terminal,   Primary
High   Winding Terminal,
                                                         Low

Core
Partly Visible

Figure 12. Pulse Transformer (Actual Size)

Corona is a most important factor with any high voltage
component. Corona,[2] if present, will slowly destroy any type of
solid insulation through several different mechanisms. Briefly,
corona constitutes a periodic electrical discharge. Commonly,
corona occurs if microscopic gas enclosures are present in the
insulation material.

Some frequencies of the corona frequency spectrum extend
into the HF and VHF range. It is, therefore, convenient to use
the high frequency range of the corona spectrum to indicate the
presence of corona. To test each unit for corona, the following
test circuit was chosen. See Figure 13.

*Scotch Cast 431, 3-M Company.

[2] E. Kuffel and M. Abdullah, High-Voltage Engineering (New York:
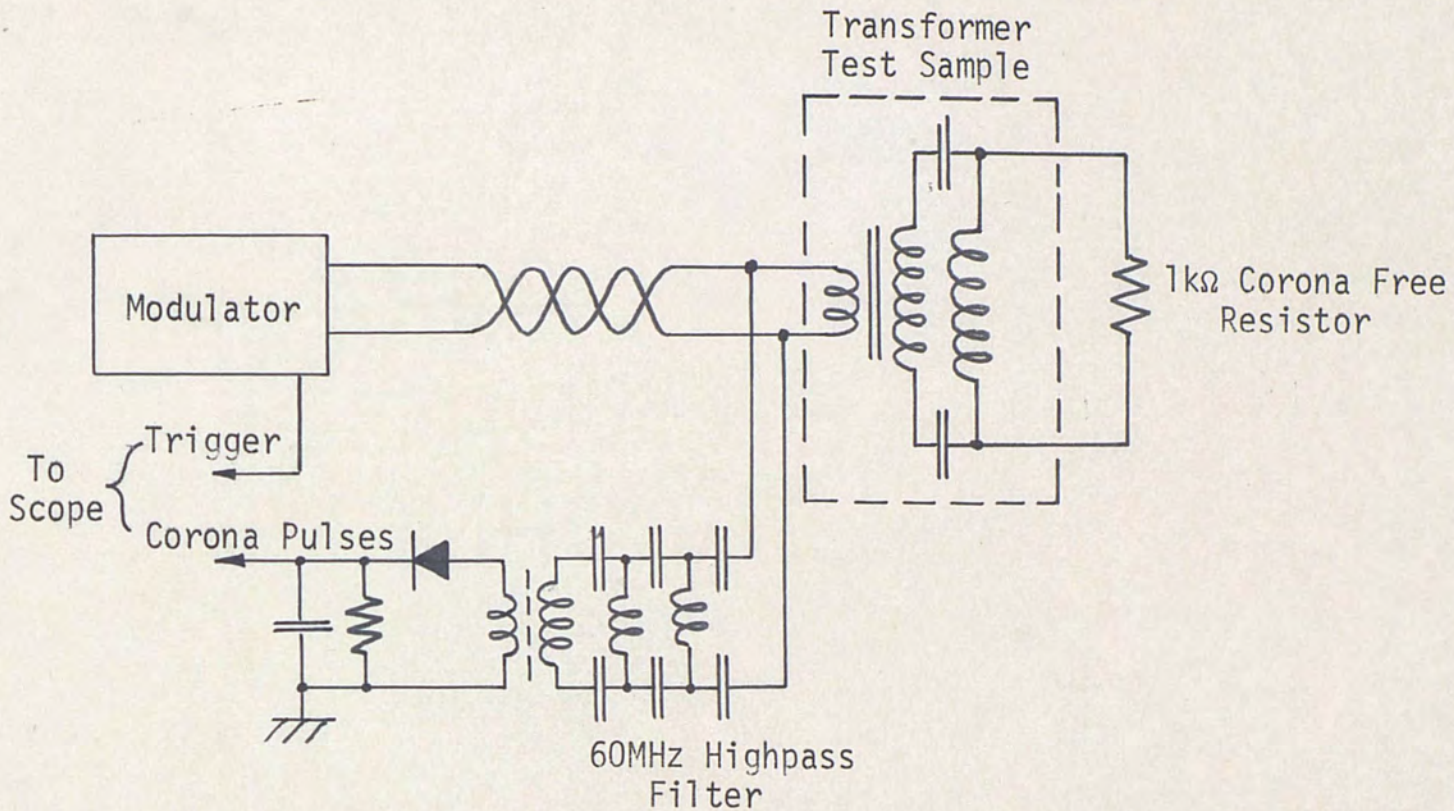Pergamon Press, Ltd., 1966), Chapters 2 and 3.

Figure 13. Corona Test Circuit

The corona test circuit is not considered to be an accurate calibrated indicator of corona but merely an indicator of the presence of corona. It consists of a symmetrical high-pass filter with a cutoff frequency of 60MHz that is connected across the primary winding of the pulse transformer.

Several transformers were tested with this circuit and showed a definite "corona onset" point at 9 to 12kV of pulse voltage across the secondary winding. This is well above the operating voltage of 5kV.

2.3 THE PULSE-FORMING NETWORK

As discussed in Chapter 1, a lumped-element delay line is used to generate the pulse. A lumped line is made up of a cascade of symmetrical networks such as the T-section of Figure 14.
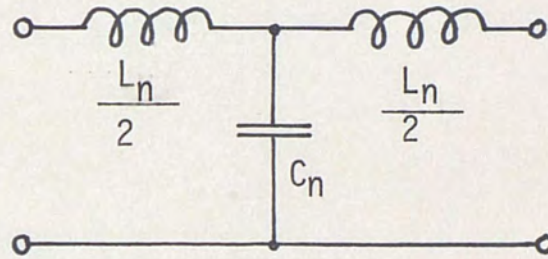

Figure 14. T-Section of Delay Line

The quantity $\sqrt{L_n C_n}$ represents the delay time per section; therefore, an n-section line, when used as a pulse-forming network, will produce a pulse of duration τ if all sections are equal.

$$\tau = 2n \sqrt{L_n C_n}$$

The cutoff frequency of one section is approximated by

$$f_c = \frac{1}{2\pi \sqrt{L_n C_n}}$$

The rise time of the pulse is determined mainly by the first L-C-section. For an equal section delay line, the first inductor is equal to $L_n/2$. This gives an expression for the rise time T_r.

$$T_r \approx \frac{1}{10 f_c}$$

The number of sections necessary to achieve a rise time of 150ns

is then given by

$$n = \frac{\pi \tau}{10\ T_r} = \frac{\pi \times 3.5\mu s}{1.5\mu s} = 7.3$$

The total capacitance of the network has to store the energy required for each pulse. This energy is

$$E = V \times I \times \tau = 250V \times 100A \times 3.5\mu s = .0875Ws$$

The stored energy in a capacitor is

$$E = \frac{1}{2}\ CV^2$$

this determines the total capacitance of the network

$$C = \frac{2E}{V^2} = \frac{.175}{.25}\ 10^{-6}F = .68\mu F$$

$$C_n = \frac{.68}{7}\ \mu F = .097\mu F$$

The capacitor C_n of each section is .097μF. The inductance L_n between each capacitor is determined with

$$L_n = \left(\frac{\tau}{2n}\right)^2 \frac{1}{C_n} = .52\mu H$$

This does not include losses. The transformer is assumed to have an efficiency of 95%. The pulse-forming network is estimated with 85% efficiency making the total efficiency approximately 80%. This power loss has to be compensated with an increased storage of energy. The capacitance of each line section is

chosen with

$$C_n = \frac{.097\mu F}{.8} = .12\mu F$$

In the previous section, the primary impedance of the pulse transformer was determined with 2.5Ω. The characteristic impedance of the line will be approximately of the same value. Slight mismatches will not affect the power transfer to any great extent as Figure 15 shows. η is the ratio of power into any load resistance R_ℓ to the power into a matched load Z_o.

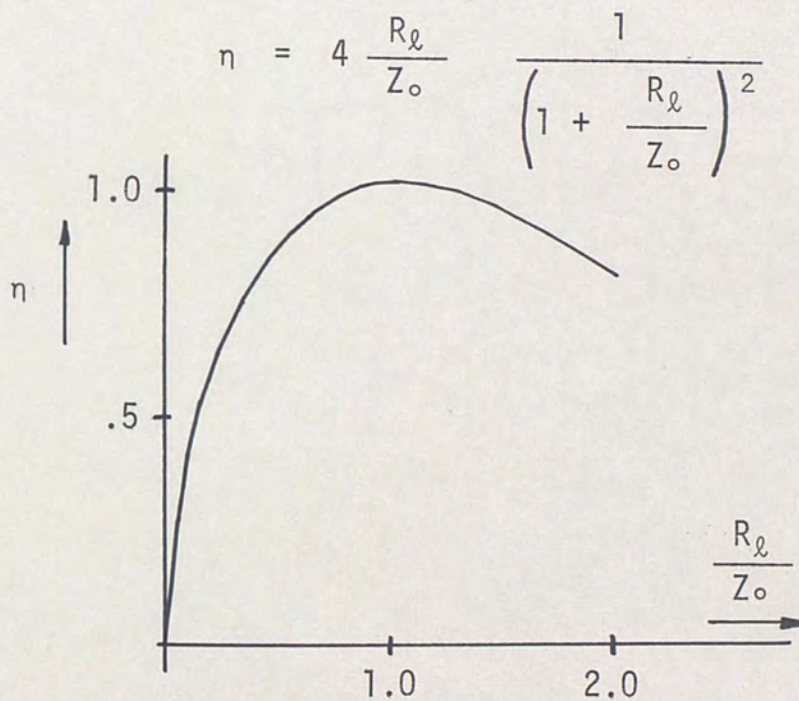$$\eta = 4\frac{R_\ell}{Z_o} \frac{1}{\left(1 + \frac{R_\ell}{Z_o}\right)^2}$$

Figure 15. Power Transfer Efficiency

The capacitors have to withstand large current surges; therefore, foil wrapped capacitors are chosen for the pulseforming network. The current in each one of the capacitors are approximated in Figure 16 for a four-section network. It is shown there that the last capacitor in the line has the largest pulse, the highest stress, and would be the first one to fail.
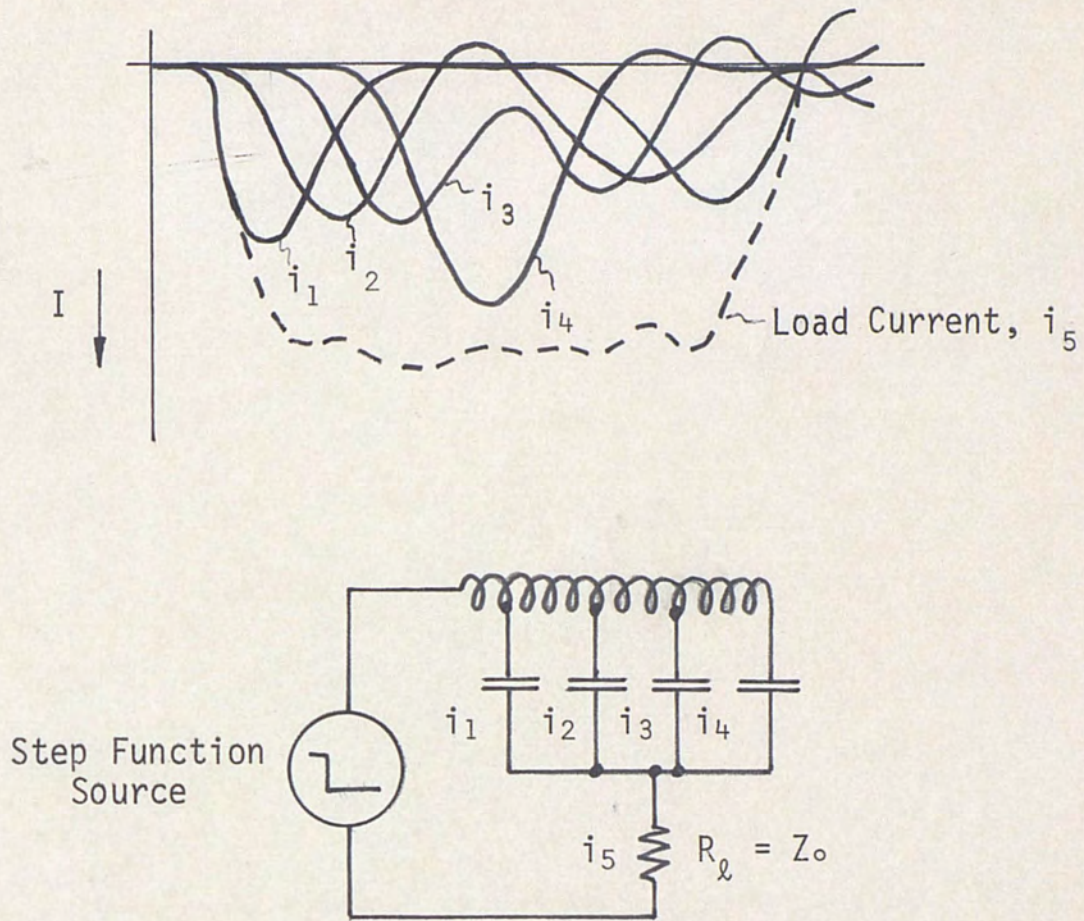
Figure 16. Capacitor Currents in Four-Section
Pulse-Forming Network

2.4 SWITCHING CIRCUIT

As previously mentioned, an SCR was chosen as a switching device. However, the SCR is too slow to achieve the desired rise time. Furthermore, the instantaneous dissipation is barely in the limits of the device. The efficiency is low since approximately 30 percent of the energy available from the pulse-forming network is dissipated in the SCR.

This makes a secondary switching device necessary. A saturable choke is used for this purpose. From Figure 17, it is concluded that the choke should saturate at approximately 1.5μs after the SCR initially starts to conduct. 1.5μs is a sufficient delay to make the SCR conduct. Furthermore, the choke had to saturate at a current level of approximately 5A so that minimum energy would leak out of the pulse-forming network. The initial current through the SCR before the choke saturates is called the priming current.

Using a torroidal core for the saturable choke and assuming a uniform current sheet around that core, the saturated inductance of the core will be

$$L_s = \frac{\mu_0 N^2}{\ell} A (\varepsilon \mu_s - \varepsilon + 1)$$

L_s = Saturated Inductance

A = An Area Enclosed by a Turn

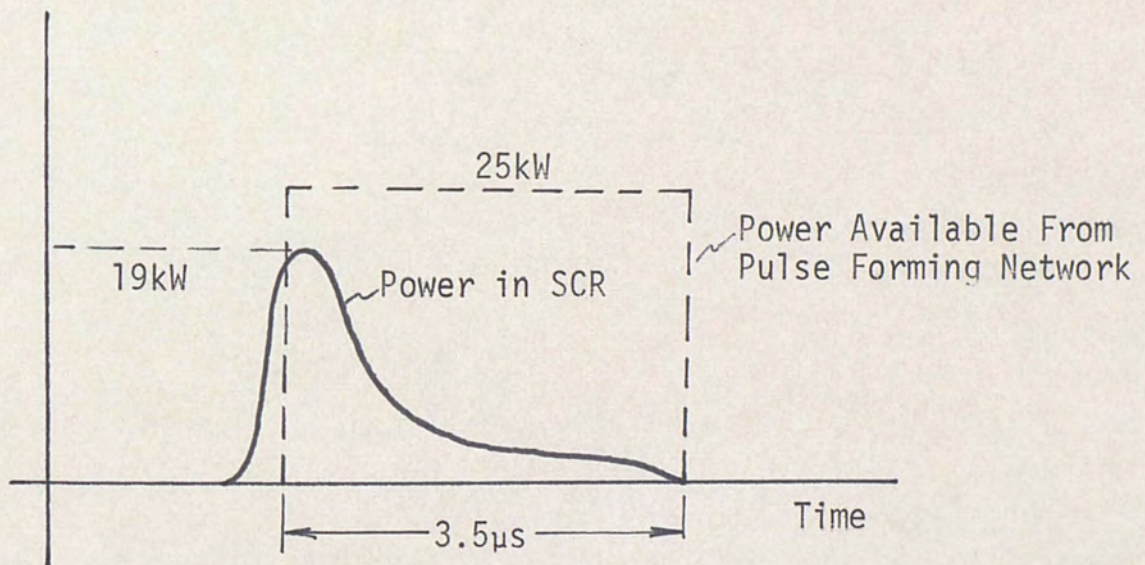1 = Mean Magnetic Path Length
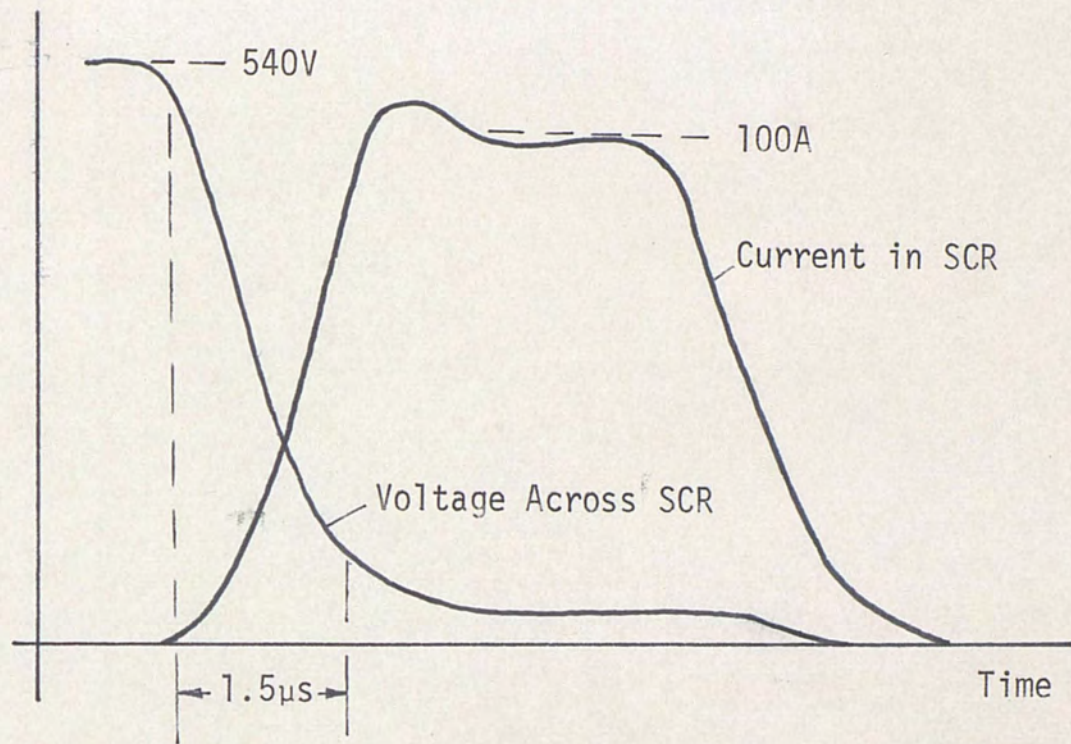
N = Number of Turns

Figure 17. Power Dissipation in Main Switching SCR, No Delay Reactor

The time to saturation will be governed by the following law:

$$E \times t_p = \Delta B \times \alpha \times N$$

E = Applied Voltage

t_p = Time to Saturation, Priming Time

α = Effective Core Area, Iron Area

Assuming the rise time of the modulator pulse to be

$$t_r = 2.2 \; \frac{L_s'}{R_\ell + Z_n}$$

where

$R_\ell = Z_n$

L_s' = Total Inductance in Pulse Loop

R_ℓ = Load Resistance

Z_n = Pulse-Forming Network Impedance

Since t_r had to be in range from 100 to 300μs, a formula for t_r had to be found.[3]

$$t_r = 1.1\mu_o \left(\mu_s - 1 + \frac{1}{\epsilon}\right) \; \frac{W}{\alpha \; \ell} \; \left(\frac{t_p}{\Delta B}\right)^2$$

$$t_r = \frac{(\text{composite permeability}) \; (\text{peak power in load}) \; (\text{priming time})^2}{(\text{core volume}) \; (\text{available flux density charge})^2}$$

The priming choke chosen has 19 turns of AWG #18 on Magnetics, Inc. 50002-1A core. The delay obtained is 1μs with 540V on the pulse forming network.

[3]T. H. Robinson, "Circuit Configuration and Thyristor Rating for Solid-State Modulators," Proceedings of the Tenth Modulatory Symposium, 1968.
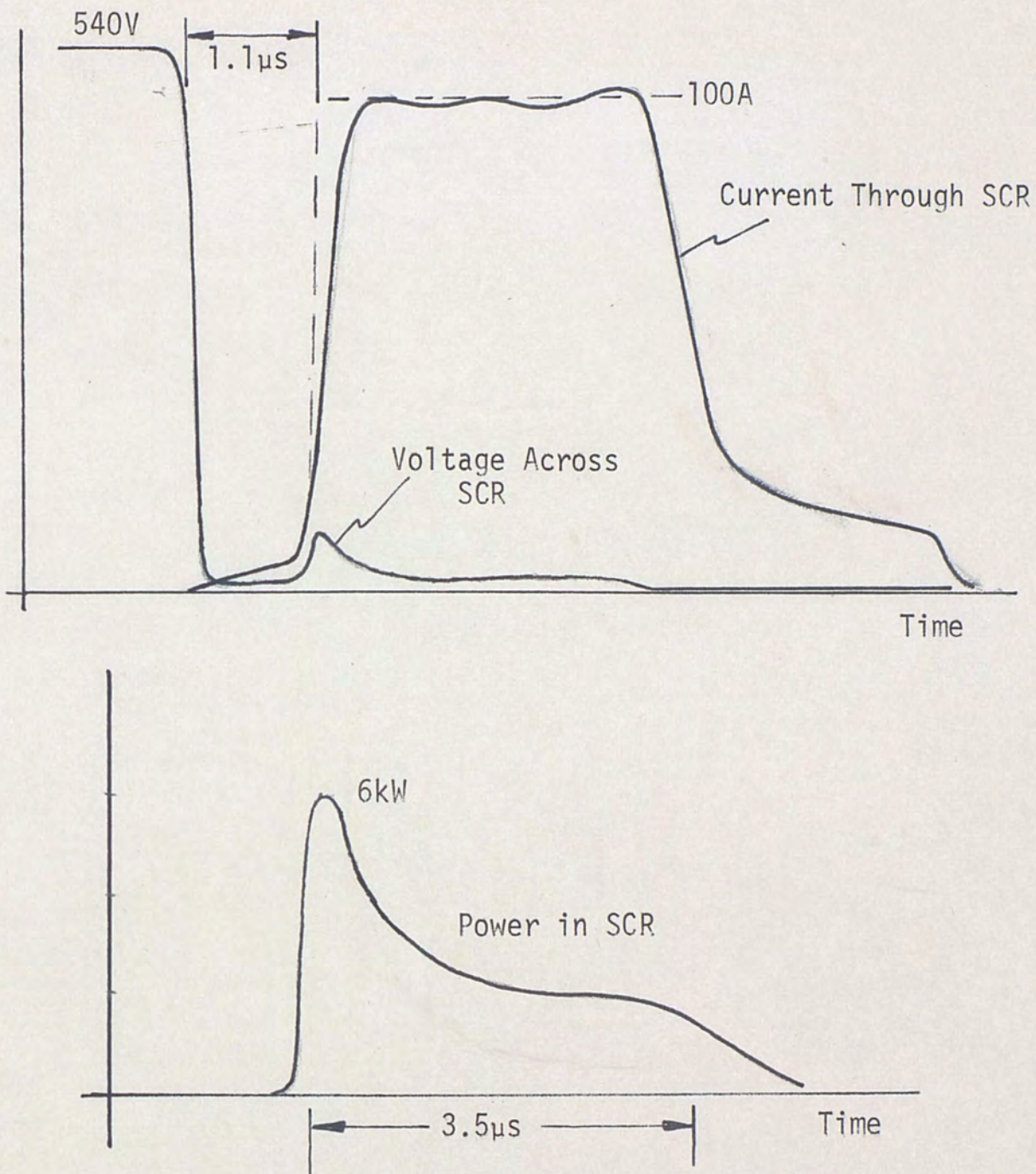
Figure 18. Power Dissipation in Main SCR with Delay Reactor

The introduction of the delay reactor in series with the main switching SCR reduces the power dissipated in the device by a factor of 10. Figure 18 shows the instantaneous dissipation in the SCR during each pulse.

2.5   CHARGING CIRCUIT

Although the charging circuitry has very little effect on
the output pulse of the modulator, the design of the circuitry
and the choice of the components is important to the over-all
efficiency.

The charging circuit must isolate the power supply from
the main switching SCR during and shortly after the pulse.
Furthermore, the charging circuit must spread the recharge of
the pulse-forming network over as much time as possible, and
still fully recharge the network before the next pulse is
developed.

Resonant charging is used for this modulator. Figure 19
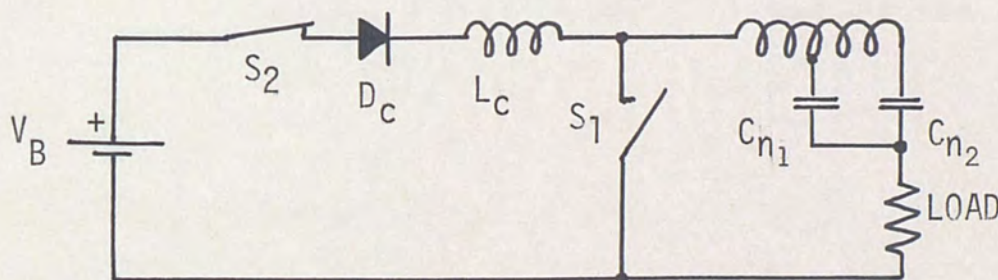shows the simplified diagram.

Figure 19.   Resonant Charging Circuit

The inductance of the pulse-forming network is neglected
since its value is small compared to the inductance of the re-
charge coil, L_c. When switch S_2 closes, L_c and C_n will combine,
forming a resonant circuit. The voltage rise on C_n will be as
high as $2V_B$ depending on the value of the total series resist-
ance of the charging loop, R_c. Since the hold-off diode D_c pre-
vents a reversal of the charging current, the maximum voltage
will remain on the pulse-forming network until S_1 closes.

Since the pulse repetition rate of the modulator is approximately 100 pulses per second, the value of L_c is defined by

$$L_c \leqq \frac{T^2}{4\pi^2 \, C_n} = 4.7H$$

where

T = Pulse Interval

C_n = Total Pulse-Forming Network Capacitance

L_c is chosen with 3H.

3.1 SUMMARY AND CONCLUSIONS

The preceding chapters show how the modulator was designed. By the time this report was written, production quantities of modulators were built and incorporated into the radar systems.

The modulators, up to this time, performed well and reliably. However, there is an area that can be improved: The relay that is used to prevent a latch-up should be replaced by a solid-state circuit.

A new circuit is under investigation that combines the functions of the relay and the control SCR (S_2) in one transistor switching stage. Photo optical couplers will be used to drive the transistor switch.
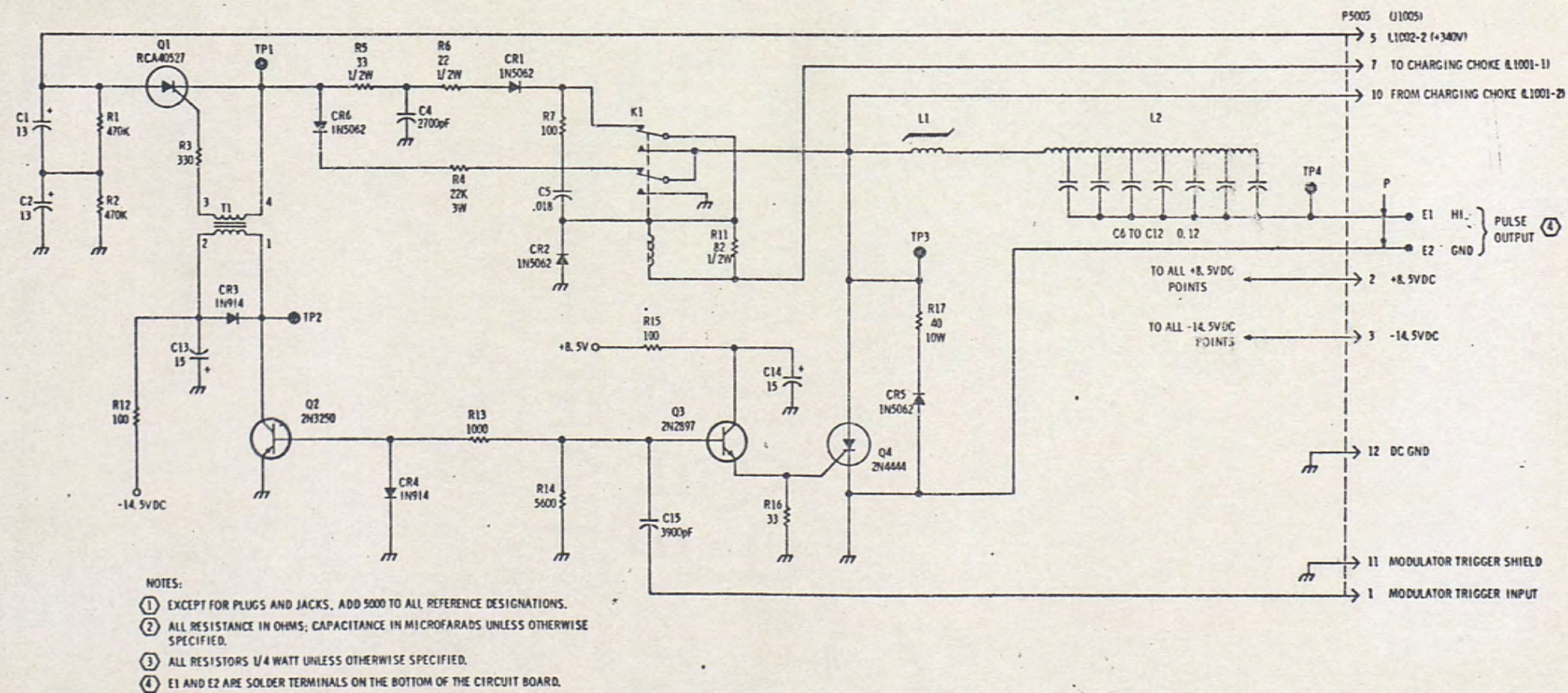
Figure A-1. Schematic of Modulator Printed Circuit Board
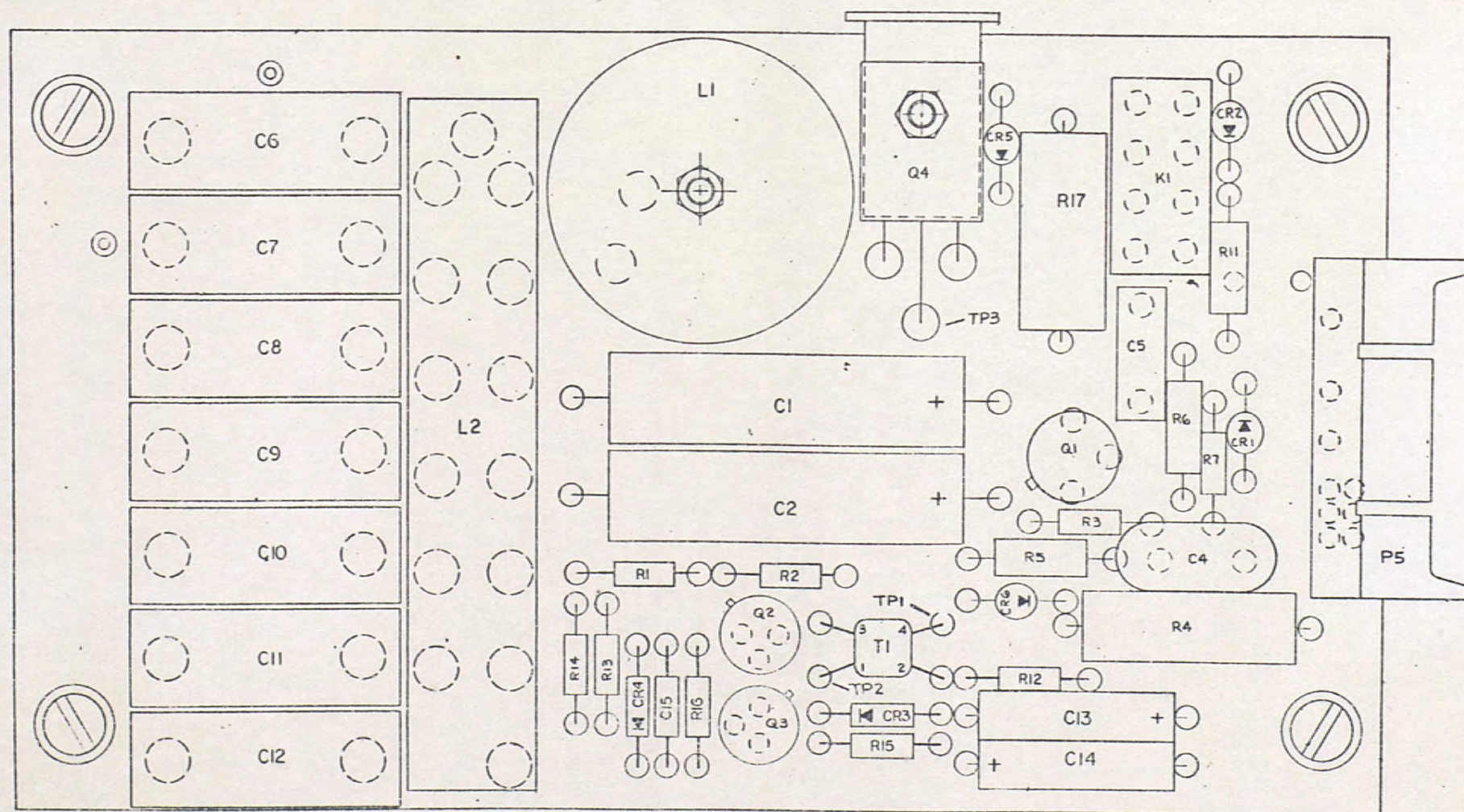
APPENDIX

Figure A-2.   Printed Circuit Board Layout

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