The Gas-Discharge Transmit-Receive Switch

By A. L. SAMUEL, J. W. CLARK and W. W. MUMFORD

THE gas-discharge transmit-receive switch has become an accepted part of every modern radar set. Indeed, without such a device, an efficient single-antenna micro-wave radar would be nearly impossible. Many of the early radar sets made in this country employed separate antennae for the transmitter and receiver. The advantages of single antenna operation are so apparent as hardly to require discussion. The saving in space or, if the same space is to be occupied, the increase in gain and directivity of a large single antenna is, of course, apparent. But even more important, perhaps, is the tremendous simplification in tracking offered by a single antenna, particularly where a very rapid complex scanning motion is desired.

The fact that the receiver needs to be operative only during periods between the transmitting pulses makes single antenna operation possible if four conditions are satisfied. These are: (1) the receiver must not absorb too large a fraction of the transmitter power during the transmitting period, (2) the receiver must not be permanently damaged by that portion of the transmitter power which it does absorb, (3) the receiver must recover its sensitivity after any possible overload during the transmitting pulse in a time interval shorter than the interval required by the reflected pulse to arrive back to the receiver from the nearest target, and (4) the transmitter must not absorb too large a fraction of the received power. At frequencies of the order of 700 megacycles and at low power levels these conditions are not impossible of attainment without recourse to any special switching mechanism other than that provided automatically by the usual circuit components. Conditions (1) and (2) can be met by designing the receiver in such a way that the change in input impedance as a result of overload will cause most of the available input power to the receiver to be reflected. Condition (3) requires careful attention to the time constants of all those receiver circuits which are subject to overload. Condition (4) fortunately is automatically satisfied by most transmitters, again as a result of the large mismatch reflections which occur at the connections to the transmitter's "tank" circuit when the transmitter is not operating. The United States Navy Mark 1 radar was operated on this basis.

The speed with which the transmit-receive switch must operate rules out all consideration of mechanical devices, at least for all but the longest range "early warning" equipment. For example, the go and return time to a target at 500 feet distance requires approximately one microsecond. Switching times must, therefore, be measured in microseconds. Since these short time intervals would at first sight seem to be too small to permit the use of gaseous discharge devices, some work was done on the use of specially designed vacuum diodes. It is possible to employ balancing circuits (sometimes called hybrid circuits) to achieve single antenna operation, but such circuits require critical balancing adjustments and they dissipate a large part of the available power in non-useful balancing loads. The need for a still different approach to the duplexing problem was clearly indicated.

Spark discharges either in air or in enclosed gaps bridged across parallel wire transmission lines were used in some of the early experimental long-wave radar sets. Dr. Robert M. Page of the Naval Research Laboratory was one of the pioneers in this work. These devices were only moderately satisfactory because of their erratic behavior and because of electrode wear. However, it was observed that the recovery time of such discharges was not as long as might be expected on the basis of a simple ionization and deionization explanation of their operation. This led to the investigation of the use of low-pressure gas discharges. These very early gas-discharge "switches" were actually much more in the nature of "lightning protectors", their principal function being to limit the power delivered to the receiver during the transmitting pulse in a gross sort of way, with considerable reliance on impedance changes at the receiver and on the rugged overload capabilities of the first tube in the receiver.

The trend toward shorter wavelengths and the desire for better protection led to the development of a partially evacuated gas-discharge tube located in a relatively high Q resonant cavity. In England, cavity type duplex tubes were made by inserting gas in a then current type (Sutton Tube) of local oscillator tube. These devices were called TR boxes (abbreviation for transmit-receive) by the English, a designation which has continued. It is a curious coincidence that some of the earliest cavity type duplex tubes made in this country at the Bell Telephone Laboratories were also constructed by inserting gas in an American type local oscillator tube (the 712A vacuum tube). This tube (later coded the 709A vacuum tube) was tested in an operative system which was subsequently demonstrated to the Army with such satisfactory results that the tube was adopted without change for several radar systems. The 709A vacuum tube and its associated cavity are shown in Fig. 1.

A similar structure, known as the 702A and shown in Fig. 2 (together with the 709A tube) was used for longer wavelengths. The need for these tubes was so very great that no time was allowed for their improvement before production was undertaken by the Western Electric Company.

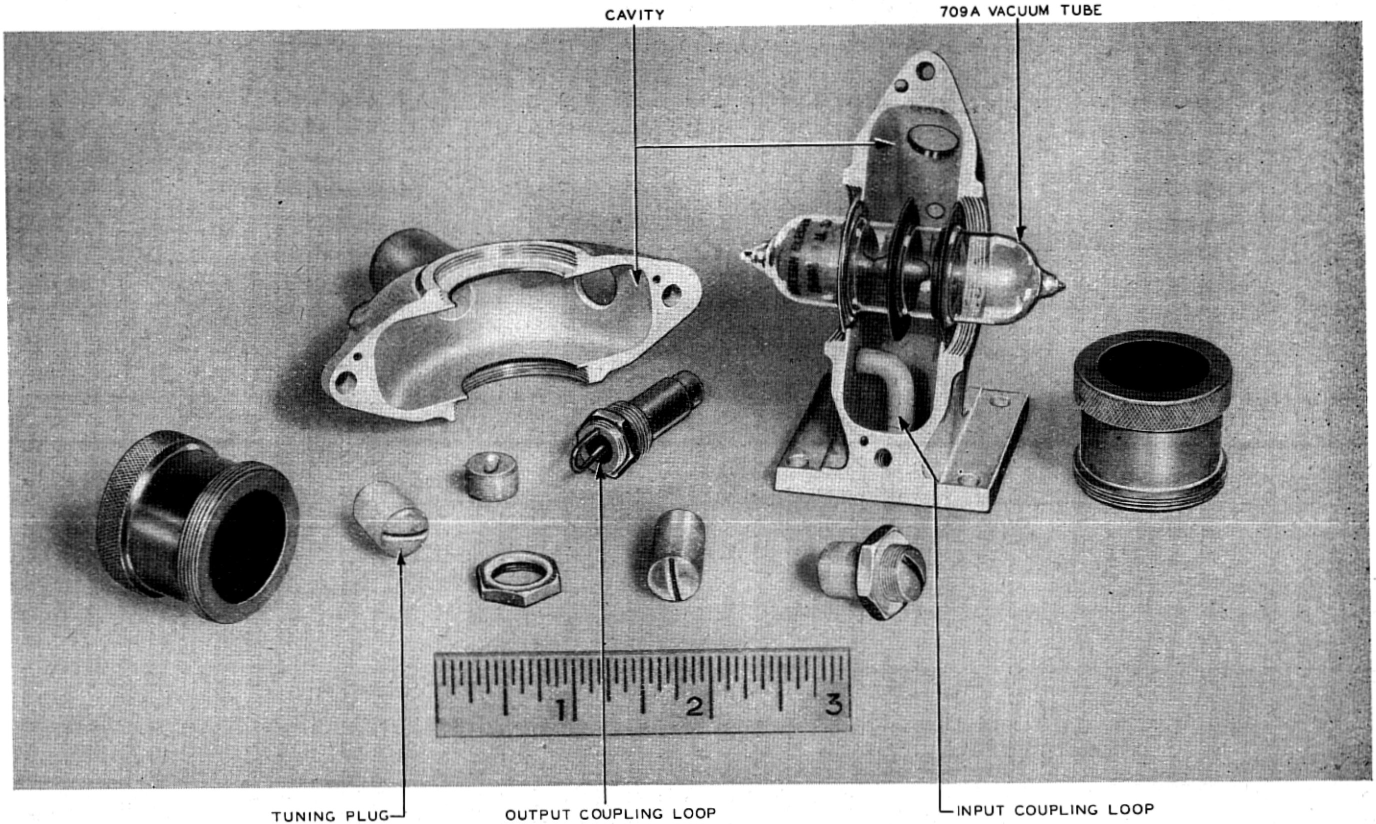

Fig. 1—The 709A vacuum tube and its associated cavity

The radar on the U.S.S. Boise in the battle off Savo Island on October 11–12, 1942 employed a 702A vacuum tube.

Three developments soon led to the need for much improved TR boxes. One of these was the rapid progress which was being made in increasing the peak output power from the magnetron. The second was improvements in the silicon point contact rectifier, the so-called crystal detector, which increased its reliability and convenience and at the same time reduced its conversion loss and noise figure as compared with the vacuum tube converter. The third was the development of still higher frequency systems to

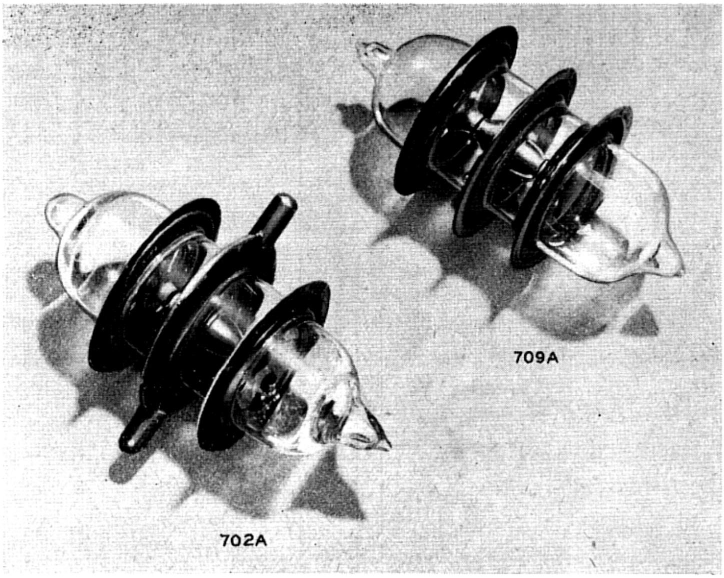

Fig. 2—The 702A and 709A vacuum tubes

achieve either greater antenna directivity or smaller size. Since satisfactory vacuum tube converters were not available for these frequencies, the silicon rectifier had to be used. Unfortunately the silicon rectifier, as then available, was subject to permanent damage if subjected to but very small amounts of power as compared with the magnetron power levels.

An active program of work was initiated at the Bell Telephone Laboratories to obtain designs of TR boxes offering adequate protection for contact rectifiers at any power levels then available or contemplated. Three tubes were developed, the 721A, 724B and 1B23 vacuum tubes shown in Fig. 3. These tubes are used at frequencies in the vicinity of 3000 megacycles, 10,000 megacycles, and 1000 megacycles respectively. They are all of the

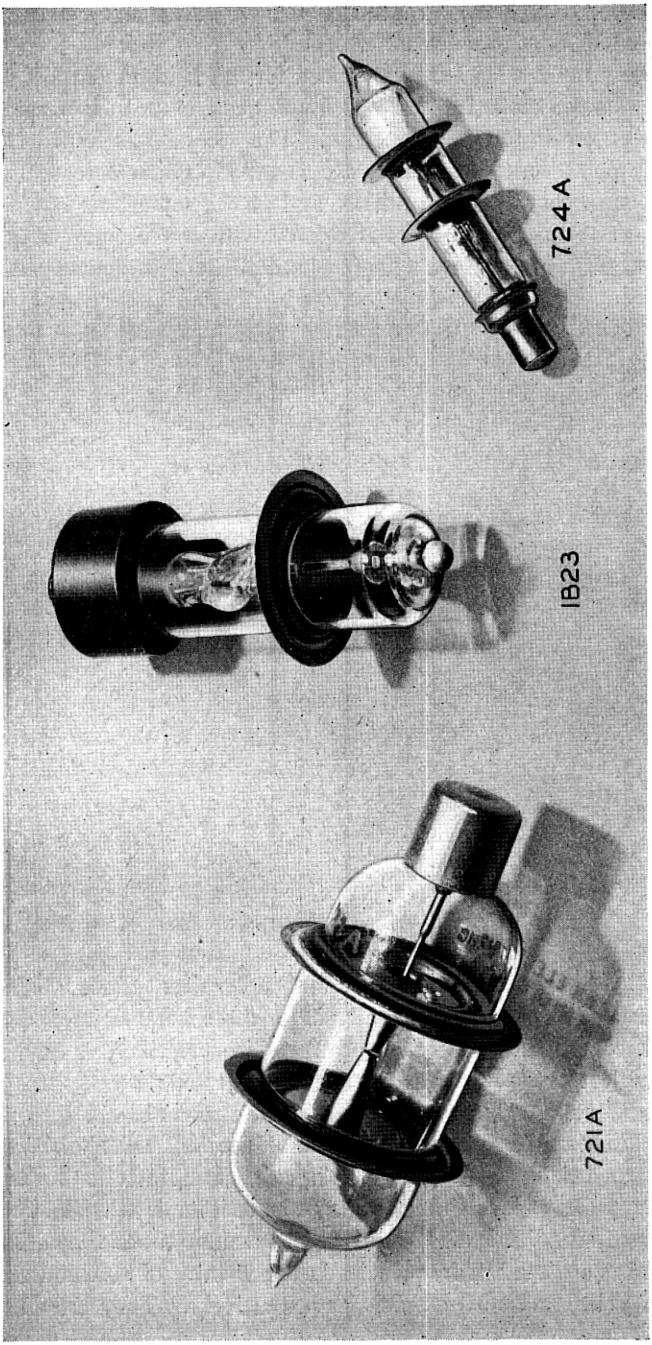

Fig. 3—The 721A, 724B and 1B23 vacuum tube

separate cavity type in which contact through the vacuum envelope is made by means of thin copper discs. More recently other designs of tubes have appeared in which the entire cavity is evacuated. Over 400,000 tubes of the three types discussed in this paper were manufactured in 1944 alone and substantially all of the American-made radars which saw active service employed one or more of these tubes.

The TR tubes used in American radars are, of course, no more essential than are the magnetrons, the beating oscillators, and the many other special parts which go to make up the modern radar. Nevertheless it is interesting to note that the 721A tube was an essential part of the radar equipment on almost every major ship in the United States Fleet, that the 724B tube was an essential part of the bombing equipment on nearly every bomber used against Japan, including the planes which carried the atomic bombs, and that the capture of Okinawa, to name a single case, would have been much more expensive in men's lives without equipment depending upon the 1B23 tube.

Method of Operation

The 709A tube as shown in Fig. 1 was operated in what has come to be known as a shunt branching circuit. Its operation can be explained in terms of Fig. 4. During transmission, energy flows from the transmitter along the coaxial line toward the antenna. Some of this energy enters the branch leading to the receiver where it encounters the TR box. This consists of a resonant cavity with a pair of spark gap electrodes arranged so that the maximum resonant voltage is built up across the gap. Since the voltage across the gap is then limited by the discharge voltage and the voltage applied to the receiver is still further reduced by an equivalent step-down ratio of the output coupling in the resonant cavity, the receiver input power is held to a small value. The power dissipated in the gas discharge, and therefore abstracted from the transmitted signal is kept small by the impedance mismatch. The discharge itself takes the form of a small pale blue glow between the electrodes. The effect of the discharge is to place a low impedance (predominantly resistive) across the maximum impedance point of the cavity. This results in the appearance of a still lower apparent impedance across the input to the cavity. If the length L_1 is an odd number of quarter wavelengths, the apparent impedance of the receiver branch at the branching point becomes very high in comparison with the impedance of the antenna and is therefore unable to abstract much power from the line.

At the end of the transmitting period, the conductance of the gas discharge falls rapidly to a very low value since the small received voltages will be insufficient to maintain the discharge. Signals arriving at the antenna can then be transmitted through the TR box to the receiver. However, in the circuit shown in Fig. 4, the receiver is still bridged by the transmitter. It is a fortunate fact that the internal impedance of many magnetrons (the most common type of transmitting tube) becomes very low when they are in the inoperative condition so that the tube is nearly the equivalent of a short circuit. By adjusting the length L_2, until this equivalent short circuit position is an odd number of quarter wavelengths from the junction point "B", the shunting impedance at "B" can be made very high so that only a small part of the received energy is lost. In the event that this change of impedance of the transmitter is not sufficient, a second TR switch commonly known as an ATR, may be introduced to perform this function as will be described later. During the receiving period, some loss will occur in the TR box resonant cavity as a result of the inherent resistive and dielectric losses. An additional loss will occur immediately after the

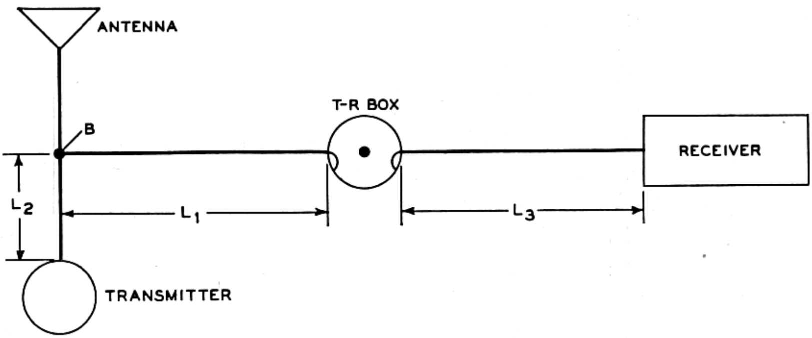

Fig. 4—The elements of a shunt branching circuit

transmitting period because of the loss producing particles (free electrons) which remain for a time in the discharge gap. The combined losses must be kept small so as not to impair the performance of the system.

Most modern radars employ series branching circuits instead of the shunt branching circuit just described. A coaxial line example of such a system (from the SCR-545) employing the 721A tube is shown in Fig. 5. As shown in Fig. 6 the cavity is coupled to the coaxial line by means of a window which can be slid along on a slot in the outer conductor of the coaxial line leading from the transmitter to the antenna. Fig. 7 is an exploded view of the cavity. Such a cavity is in effect in series with the line as the currents existing in the outer conductor of the coaxial line are interrupted by the window. During the transmitting period the low impedance at this window limits the voltage across it to a small value and prevents serious loss of transmitter power.

Reception in the series branching circuit of Fig. 5 is achieved by adjusting

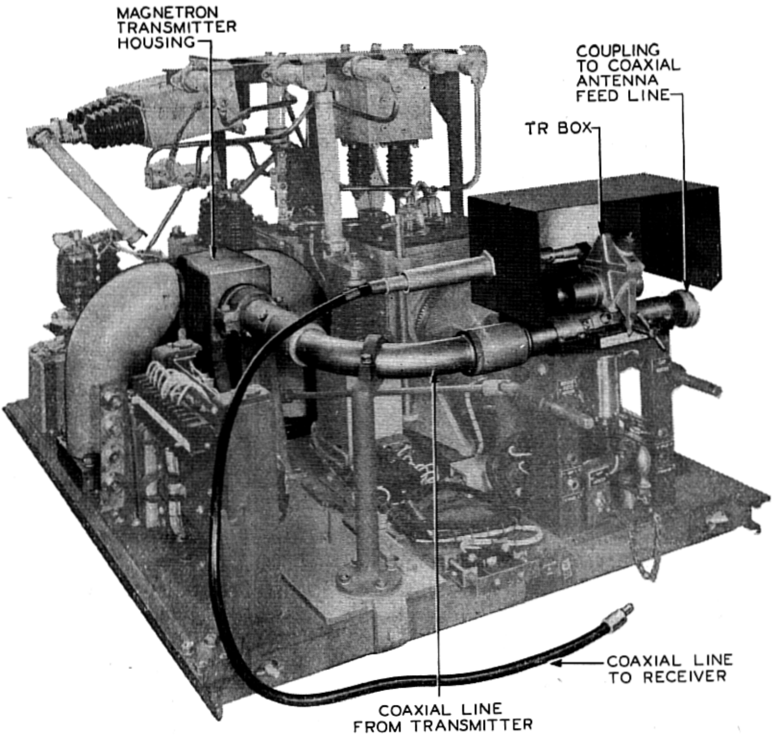

Fig. 5—The series branching circuit employing the 721A tube used in the SCR545

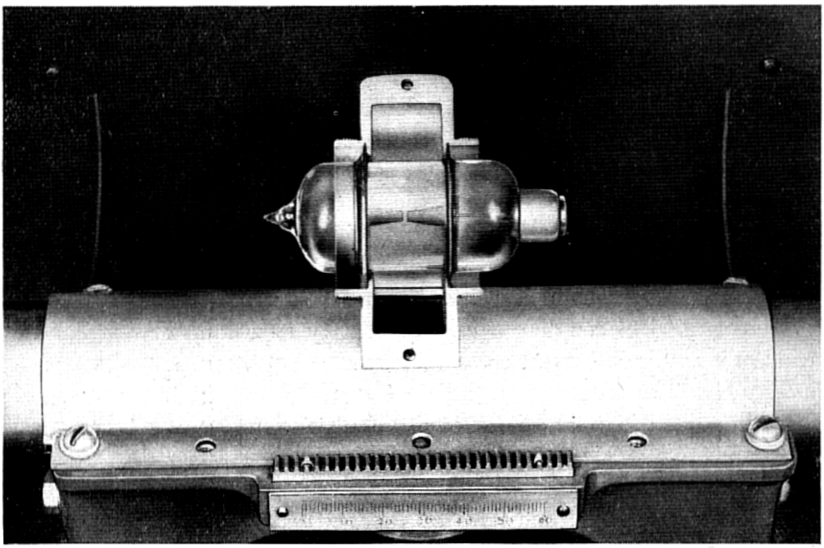

Fig. 6—Closeup view of the cavity shown in Fig. 5, partly disassembled to show coupling window

the position of the TR cavity along the slotted section of the line until the window is located the correct distance from the transmitting tube. This now corresponds to an even number of quarter wavelengths between the equivalent short-circuit plane and the junction, so that the maximum current is caused to enter the cavity. The output to the receiver is obtained from a small coupling loop in the TR cavity.

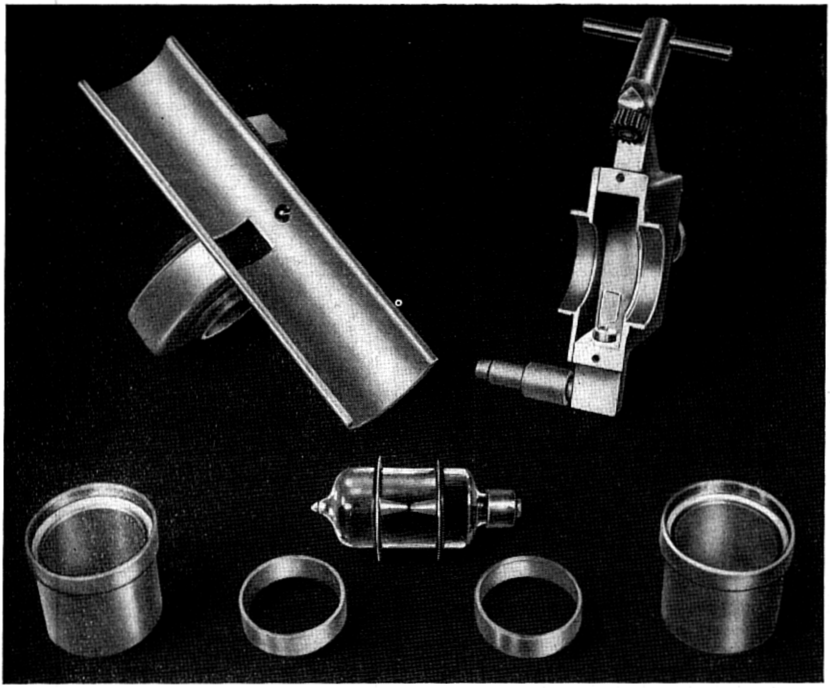

Fig. 7—Exploded view of the cavity of Fig. 6

A similar coaxial line series branching circuit using the 1B23 is shown in Fig. 8. The method of inserting the tube in the cavity is shown in Fig. 9 while Fig. 10 is an exploded view showing the details.

This method of coupling a resonant cavity to a transmission line by a window is not limited to the coaxial line case. A wave guide system is shown in Fig. 11. The distance between the TR cavity and the transmitting tube is again adjusted by sliding the cavity and its window along over a section of the rectangular wave guide containing a slot.

The ATR. In all of the systems so far described use is made of the impedance mismatch conditions at the magnetron or other transmitting tube terminals to prevent serious loss during the receiving period. If the magnetron "cold" impedance does not differ greatly from the surge impedance of the transmission line used, it may not be possible to avoid loss of reflected signal into the transmitter line. Also in some cases, an unreasonable amount of adjustment must be provided in the position of the TR cavity with respect to the transmitter to make up for large variations which may be encountered in transmitter "cold" impedance. Both difficulties

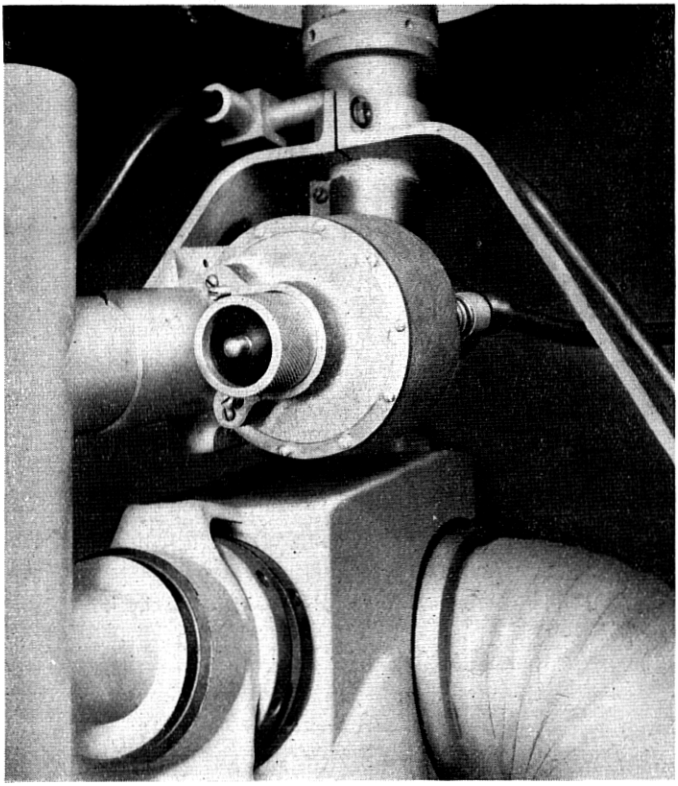

Fig. 8—Series branching circuit using the 1B23 vacuum tube

may be avoided by the use of a second gas discharge tube located between the transmitter and the TR and at an odd number of quarter wavelengths from the TR junction. This second tube is referred to as the anti-T-R tube (usually abbreviated to ATR), or sometimes as the RT tube. The use of an ATR tube was not found to be necessary in most of the systems which employed the 721A tube.

With the advent of still higher frequency systems, for which the 724B tube was designed, the "cold" impedance difficulties just mentioned made

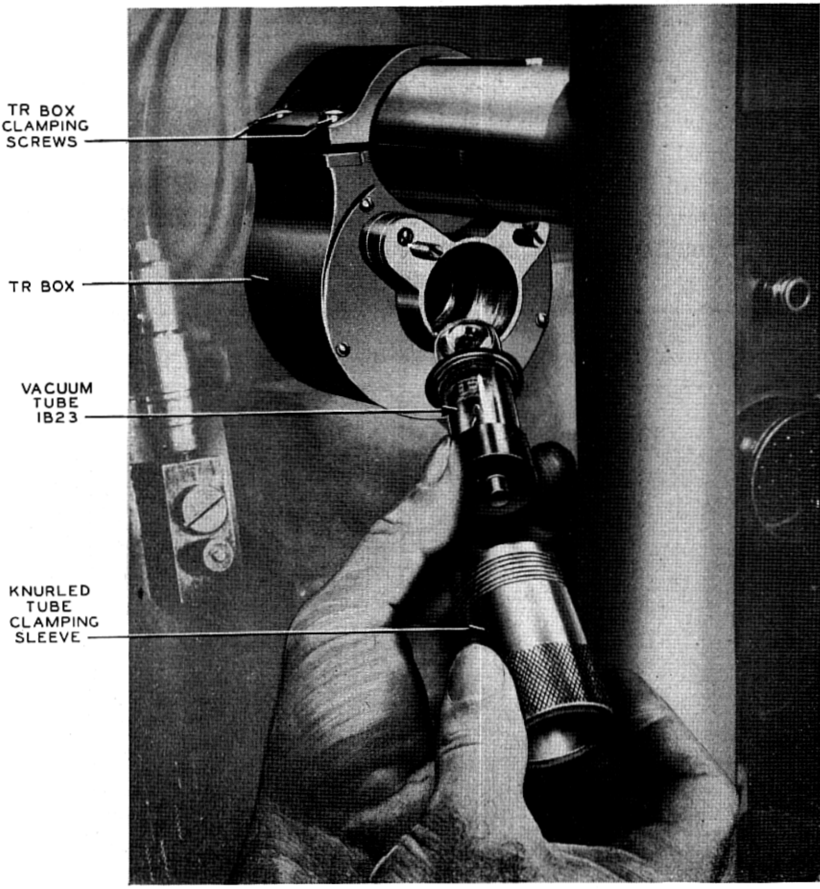

Fig. 9—Method of inserting the 1B23 into the circuit

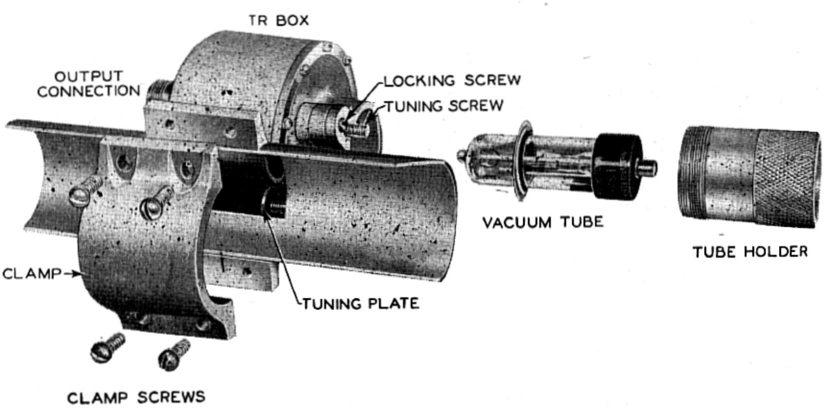

Fig. 10—Exploded view of the 1B23 cavity

it seem advisable to employ an ATR tube. The general arrangements of the circuit elements in one of these systems is shown in Fig. 12. The main wave guide section is shown removed from the rest of the equipment

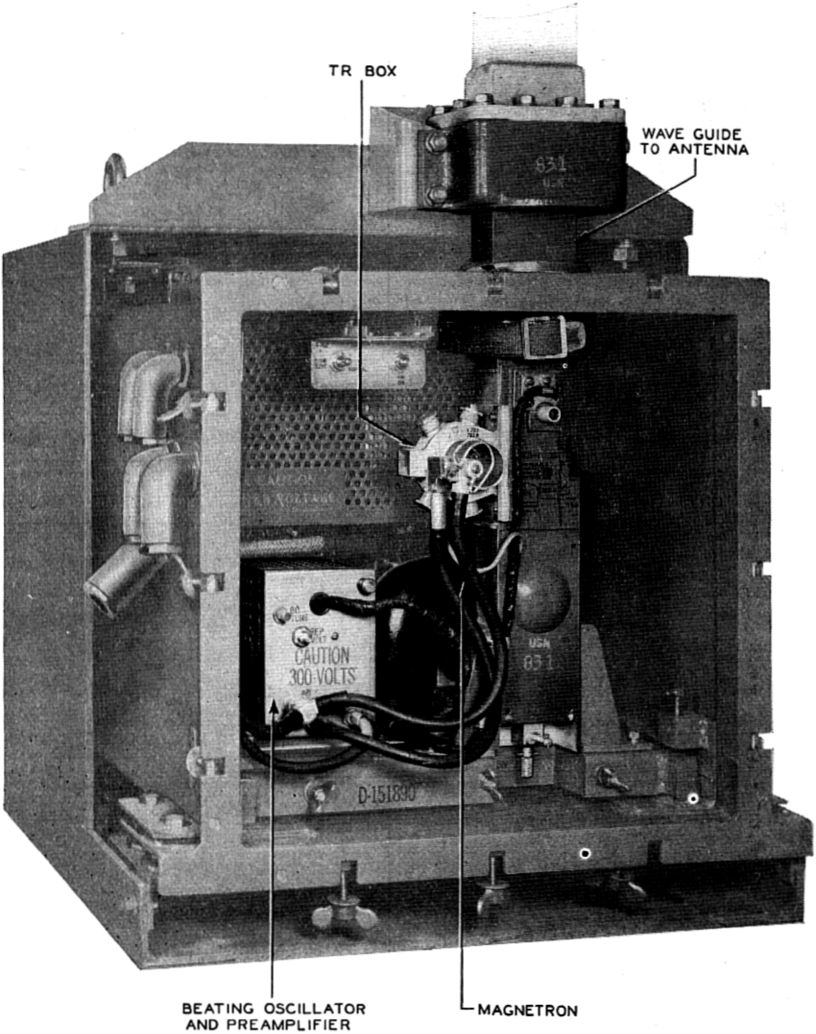

Fig. 11—The wave guide system of the SL radar employing the 721A tube

in Fig. 13 while Fig. 14 is an exploded view revealing the details of the cavity construction. The two cavities are, of course, coupled to the wave guide by means of windows. The wave guide branch leading to the receiver is in

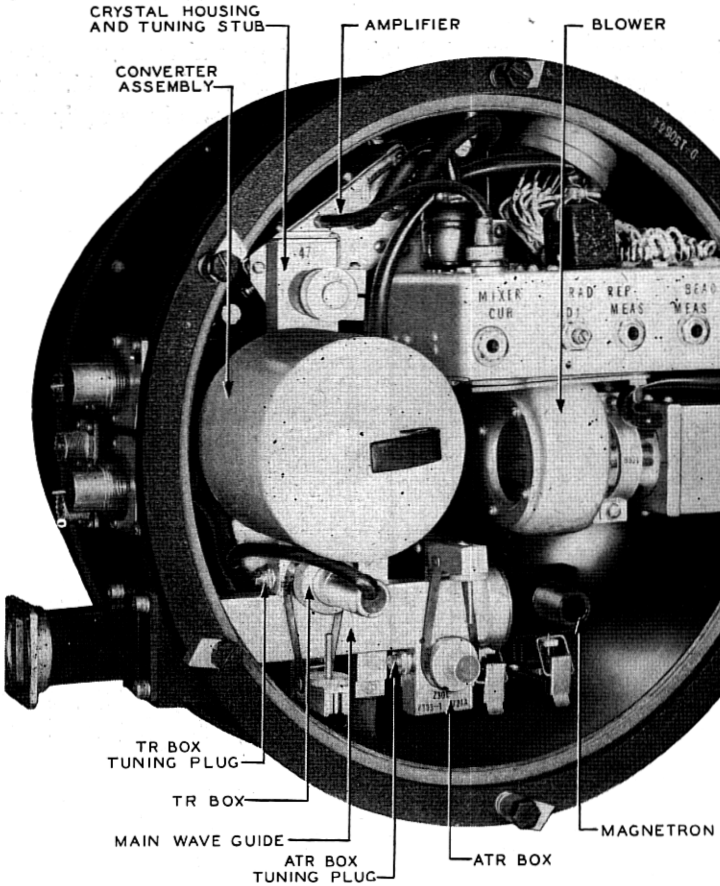

Fig. 12—A general view of a wave guide system employing a 724B tube in the TR box and a second 724B tube in the ATR

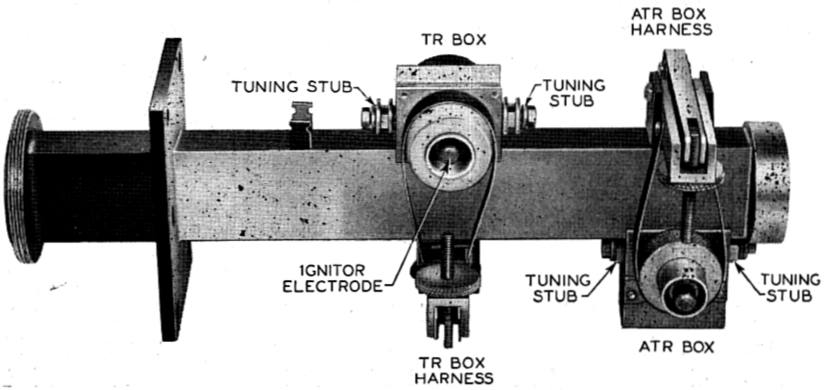

Fig. 13—The main wave guide of Fig. 12 removed from the rest of the equipment

turn coupled to the TR cavity by another window. The input and output windows of the TR cavity are adjusted in size to provide an impedance match to the line during the receiving period. The window to the ATR is, however, adjusted so that it presents a high impedance, that is much greater than the surge impedance, during the receiving period. This high impedance is effectively in series with the magnetron impedance. The resulting high impedance is located at an odd number of quarter wavelengths from the TR and so presents a very low impedance in series with the receiving branch.

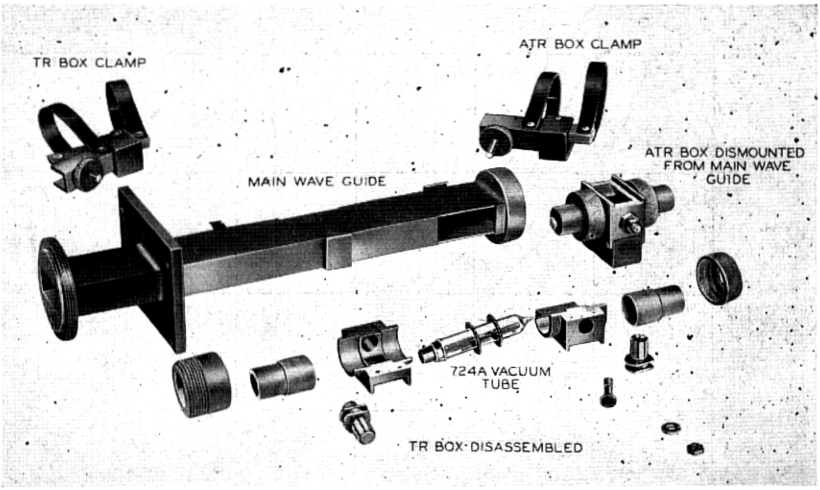

Fig. 14—An exploded view of Fig. 13

Both the TR box and the ATR box must be tuned to resonance at the magnetron frequency. Broad-band ATR boxes using very low Q circuits have been designed which require no adjustment over a 5% band. Such boxes, which obviously are very advantageous in tunable systems, do not use the copper-disc-seal tubes which form the principal subject matter of the present paper, and will not be discussed further.

TR Box Performance

The performance of a TR box can be described in terms of four parameters which are related to the four duplexing requirements mentioned earlier. These parameters are respectively: (1) the high level loss, which is the transmitting power loss in the TR tube; (2) the leakage power, which is the amount of transmitter power which reaches the receiver; (3) the recovery time, which measures the rate at which the TR box recovers its low level behavior after the termination of the transmitting period; and (4) the low-level loss, which describes the loss of the received signal including (a) the loss in the TR box itself and (b) the loss in the transmitting branch. These parameters are interrelated and conflicting. For example, the interdependence of the leakage power and the low-level loss may be computed on the basis of a somewhat idealized TR box as is done in Appendix A and the results presented in the form of the curves of Fig. 15. It is customary to design the cavities for matched input conditions ($\sigma = 1$), for obvious reasons, and for a low-level loss of one to two db. The relationship between the transmitting power dissipated in the TR tube and the low-level loss is shown

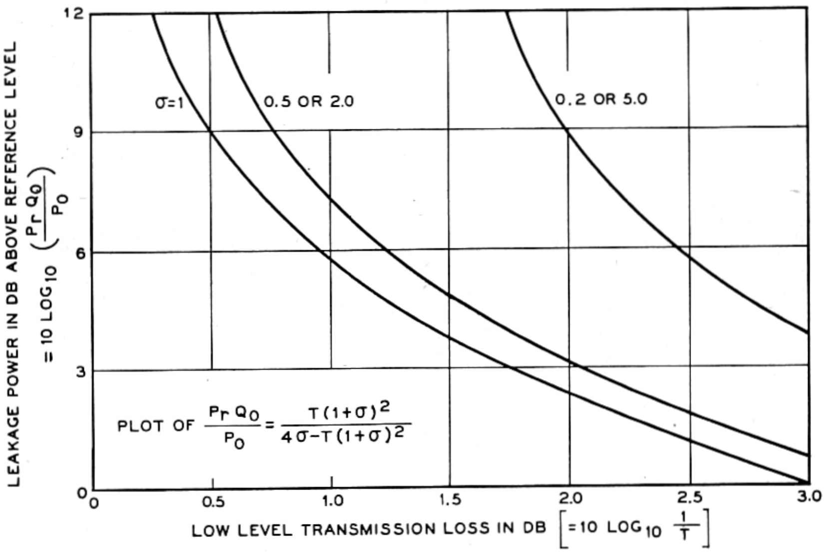

Fig. 15—The variation in leakage power with the low level loss adjustment

in Fig. 16. This curve may also be used to determine the effect of the low-level loss adjustment of a TR cavity on the recovery time characteristic since recovery time depends upon the gas discharge power rather than upon the transmitter power per se. In spite of this interdependence, it will be convenient to consider the different operating parameters separately in the sections to follow. The receiver protection aspect will be considered first.

Receiver Protection. The receiver protection problem is complicated by the fact that power reaches the receiver through the TR box by three different mechanisms. As shown in Fig. 17, the observed leakage power pattern is composed of three parts known respectively as the spike, the normal flat leakage and the direct coupling. The spike is the result of the transient condition existing at the beginning of each pulse. Normal leakage power can be thought of as due to the finite voltage drop across the gas discharge while the direct coupling is that component of the leakage power which would be present if the voltage drop across the discharge were zero.

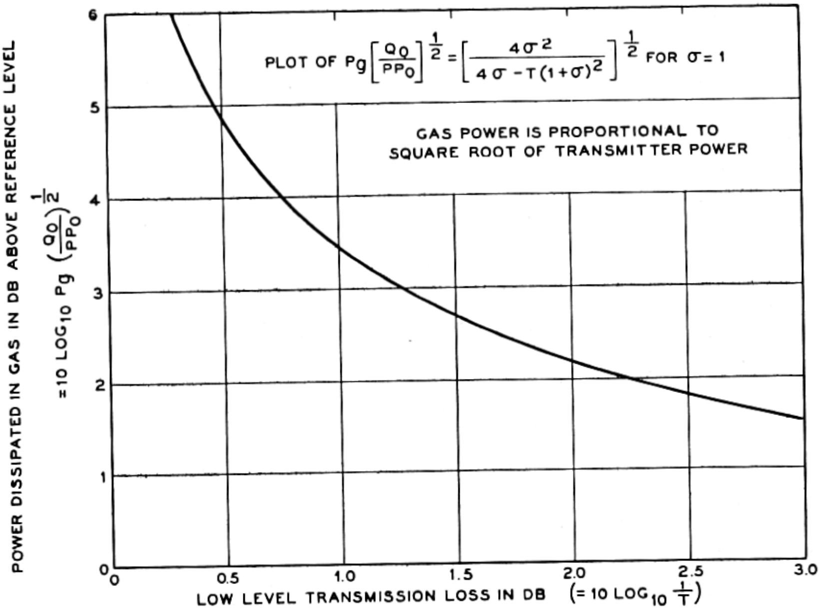

Fig. 16—The variation in gas discharge power with the low level loss adjustment

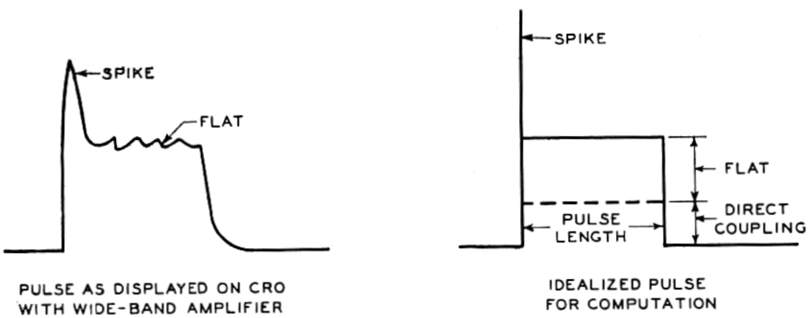

Fig. 17—The shape of the leakage power pulse

The Spike. Because of the rapid rate of rise and fall of the spike, the observable deflection on an oscilloscope is dependent upon the bandwidth of the video amplifier and upon the energy in the spike. Although an observation of the true shape of the spike has not yet been made, the duration is estimated to be of the order of 10^{-9} seconds, a time interval that is probably short compared with the thermal time constant of the contact on a converter crystal. However, it is possible to measure the energy content of the spike, and such measurements indicate that this energy is fairly independent of the length of the pulse and of the transmitted power level, although it is definitely dependent upon the steepness of the wave front of the transmitted pulse. The spike clearly represents energy transfer through the TR box during the period required to establish the discharge conditions which exist during the flat. The energy contained in the spike varies between a few hundredths of an erg to perhaps one erg per pulse, depending upon a variety of factors. By way of comparison, the conventional crystal rectifiers are proof tested in manufacture with a single spike of 0.3 erg to 5.0 ergs, depending upon the crystal type. It is generally believed that the spike is more damaging than the flat in most radar systems.

The energy in the spike is found to depend upon the repetition rate of the transmitting pulses, presumably because of residual ionization in the gas discharge gap. At low repetition rates (that is less than roughly 1,000 pulses per second), the spike energy may be materially decreased by a d-c glow discharge near the radio frequency gap. This discharge provides a continuous supply of ions and free electrons and so aids in establishing the desired condition in the r-f discharge path. A discharge is supplied in all the standard TR tubes. An auxiliary electrode called the "igniter" or "keep-alive" is used as the cathode, with the back or inside portion of one of the high frequency electrodes acting as the anode. A small amount of radioactive material is placed in the tube to insure that the igniter discharge starts on the application of the igniter voltage. Fig. 18 is a plot of the way in which the spike energy varies with the repetition rate both with and without an igniter discharge. Igniter oscillations sometimes occur as a result of the negative resistance characteristics of the igniter discharge. This causes a cyclic variation in the number of free electrons and ions with a resulting fluctuation in the spike energy. Inadequate protection may result from such oscillations. It is customary to mount a current limiting resistance very close to the igniter cap to minimize the effects of these undesirable oscillations. When such oscillations still occur they are usually evidence of an insufficiently high igniter voltage or of tube failure. The margin of safety in the igniter operation may be increased by increasing the discharge current but at the expense of greatly reduced tube life.

When a radar system is first turned on, the first pulse occurs without the benefit of residual ions in the discharge, and for the first few pulses the spike energy may easily reach dangerously high values. While the magnitude of this "turn on" effect is greatly reduced by the presence of the igniter discharge, it is customary to provide a "crystal gate" in the form of a shutter which isolates the crystal from the TR box until after stable transmitting conditions have been reached and until the TR tube discharge has been established. The need for this additional turn-on protection is somewhat greater with the 724B than it is with the 721A tube. Another important function of the "crystal gate" is to prevent the crystal in an idle radar from being damaged by energy from other radars operating nearby.

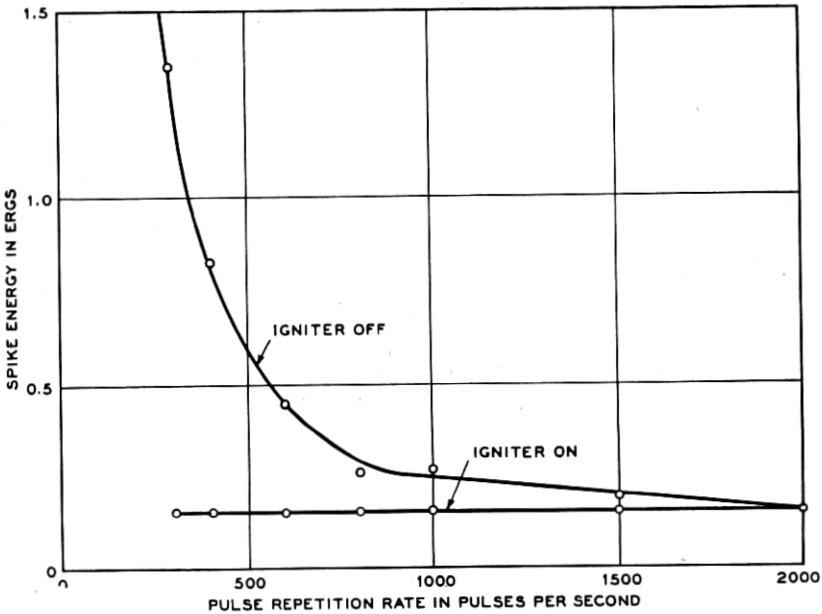

Fig. 18—The dependence of spike on the repetition rate for the 724B tube in a cavity adjusted for a 1.5 db low level loss and for a transmitter power level of 8 kw peak

The energy in the spike is a function of the effective size of the input and output coupling windows of the TR box. A convenient method of presenting this effect is to plot the spike energy as a function of the low-level transmission loss of the cavity which also depends upon the window sizes. Fig. 19 is such a plot for the 724B tube*. Comparing these experimental data with the computed flat power curve of Fig. 15, one notes that the spike energy varies at a more rapid rate than does the flat power. In both cases, the leakage decreases as the low-level loss increases and crystal protection can be purchased at the expense of receiver sensitivity.

* Based on data taken at the M.I.T. Radiation Laboratories by F. L. McMillan. Jr. and J. B. Wiesner.

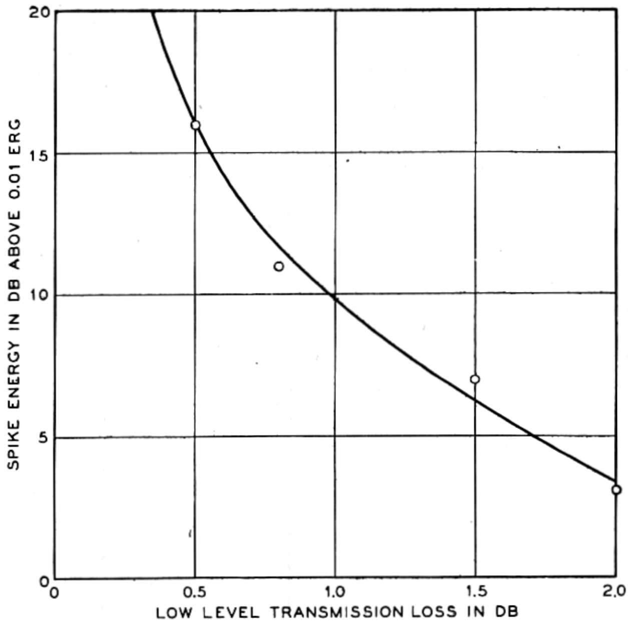

Fig. 19—The variation in spike energy with the low level loss adjustment ($\sigma = 1$) for the 724B. This experimental curve for the spike energy should be compared with the idealized flat power curves of Fig. 15

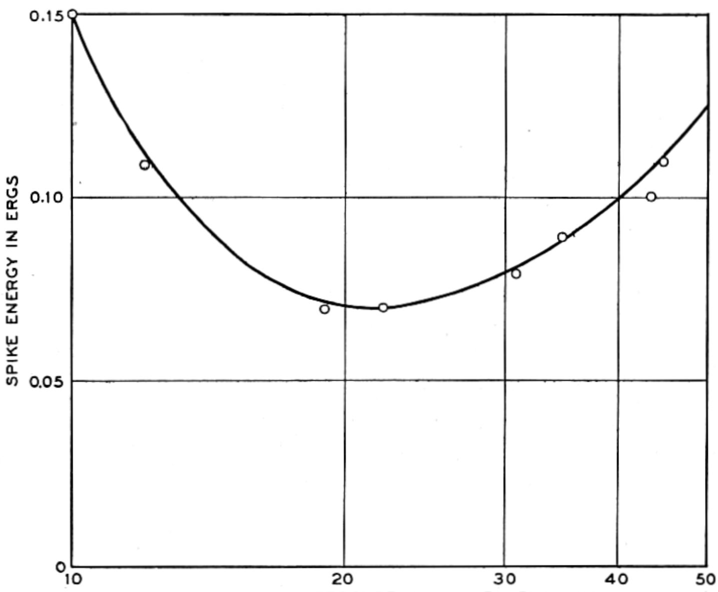

Fig. 20—The effect of gas pressure on the spike energy for the 724B

The way in which the spike energy varies with pressure of the gas in the TR tube is illustrated in Fig. 20. These data were obtained on the 724B tube structure. Other factors, yet to be discussed, prevent the use of the exact optimum pressure as determined on the basis of the spike energy only.

The Flat. The more or less flat portion of the leakage power is in reality the result of two different mechanisms of energy transfer, one of which is reasonably independent of the transmitter power level. It is this portion only with which we will now be concerned. This flat power is critically dependent on the chemical constitution and pressure of the gas within the TR tube. It can be thought of as being the power transmitted by the TR

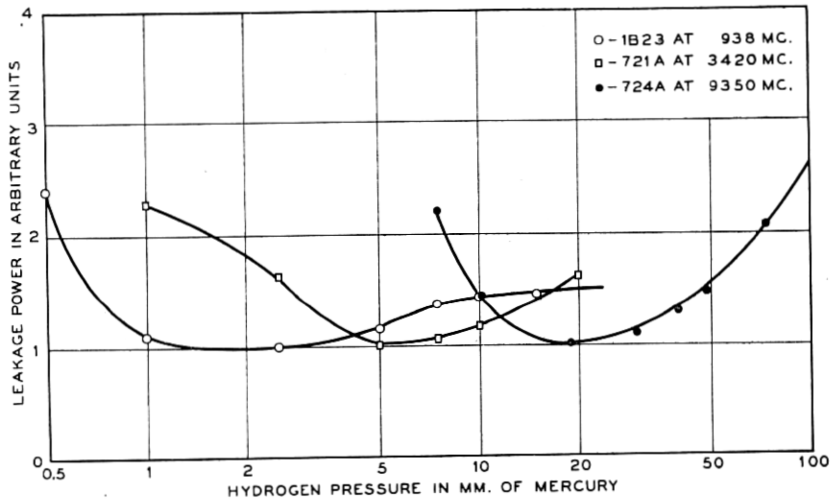

Fig. 21—Experimental curves showing the relationship between flat leakage power and gas pressure, taken with a c-w oscillator

box by virtue of the fact that the voltage drop across the gas discharge is not zero. The constancy of the flat power in spite of variations in the transmitter power level is presumably related to the similar phenomenon of a nearly constant voltage drop across a d-c gas discharge independent of the discharge current. Because of this constancy, the gas discharge parameter P_0, shown in Fig. 15, can be assumed to be a constant more-or-less independent of the transmitter power level. Reasonable values of P_0 for cavity design purposes are 20 volt-amperes for the 721A tube and 10 volt-amperes for the 724B tube. Corresponding values of the Q_0 parameters needed in interpreting Fig. 15 are 2500 for the 721A tube and 1500 for the 724B tube. Using these values the flat leakage power for a TR box using a 721A tube and having a low-level loss of 1 db would be 30 milliwatts. The corresponding flat leakage power for the 724B tube in a 1.5 db box would be 16 milliwatts. Actual measured values are usually somewhat less than these figures. As most crystals will withstand flat powers very much greater than this amount, the flat power is normally of much less importance than the spike in contributing to converter crystal failure.

Since the flat portion of the leakage power represents quasi-steady-state conditions, it is possible to simulate it for purposes of study by the use of a C.W. oscillator. Fig. 21 contains three experimental curves taken at

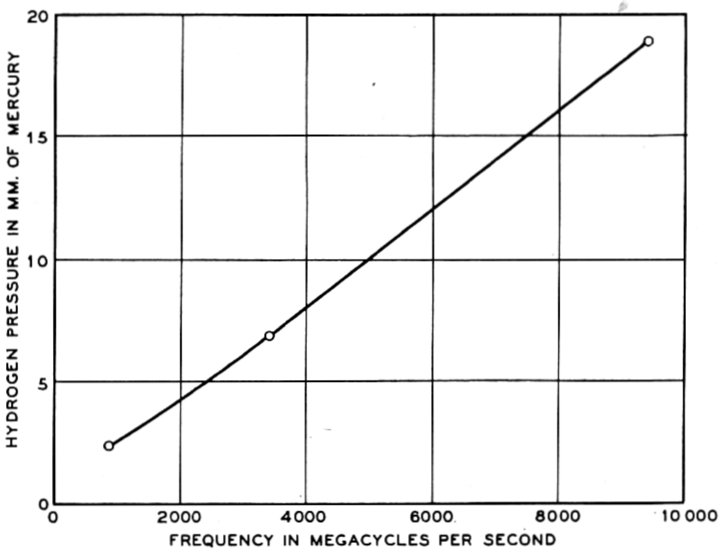

Fig. 22—The pressure for minimum leakage power as a function of frequency

three different frequencies showing the relationship between the flat leakage power and gas pressure. These curves were all taken with tubes filled with hydrogen only. Fig. 22 shows that the pressure for minimum flat leakage is proportional to the frequency. This simple law probably does not apply at frequencies much less than 1000 mc.

Water vapor is used in commercial TR tubes to improve the recovery time, as will be discussed later. The variation in flat leakage power with partial water vapor pressure as measured on a 721A type of tube containing both hydrogen and water is shown in Fig. 23. These data were taken in a radar system.

In this connection, it is of interest to note that the characteristics of the gas discharge in the TR box must of necessity be quite different from those that obtain at lower frequencies. Simple calculations indicate that the mean free path of an electron is in general of the same order as the distance between the electrodes but that very few electrons are able to reach the electrodes because of the very rapid reversals in the r-f field. Electrons therefore oscillate rapidly to and fro, losing energy to the neutral gas molecules and to positive ions through occasional collisions The positive ions do not contribute in any substantial way to the discharge current because of their large mass and correspondingly low velocity. The r-f voltage drop across the discharge is maintained at a relatively low value by the formation of more ions and free electrons by collisions between electrons and neutral

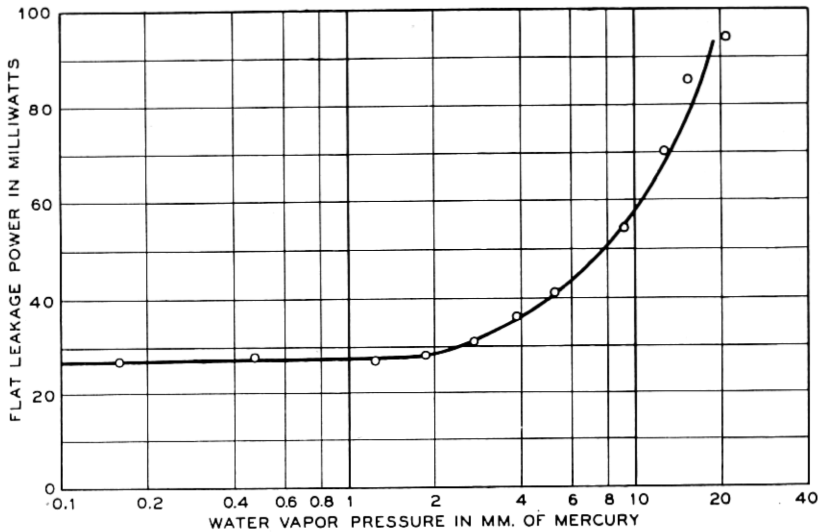

Fig. 23—The effect on leakage power of the addition of water vapor to 20 mm of hydrogen in the 721A type tube

molecules as soon as this voltage rises above some critical value. Measurements indicate that the voltage drop across the r-f gap is of the order of 80 to 100 volts for a typical TR tube. The variation in voltage drop with gap length may be inferred from the flat power measurements recorded in Fig. 24.

Direct Coupling. At very high transmitter power levels a third component of leakage power is observed which is directly proportional to the transmitter power. This component is usually called "direct coupling". It is due to the transmission of power through the cavity in modes which do not have voltage maxima at the gas discharge gap. It can therefore be observed even when the gap in the tube is short circuited. In fact measurements made under such short-circuited gap conditions yield results comparable to the values observed for actual tubes. The direct coupling component of the total flat power and the gas discharge limited component are found to be additive. Direct coupling power is logically measured in terms of db below the transmitter power level and for the usual TR box is of the order of 60 to 70 db. An abnormal form of direct coupling which may reach dangerous values can occur under certain improper operating conditions when the magnetron produces an appreciable amount of power at other than the normal operating frequency. Some of these spurious frequencies may be in the vicinity of the resonant frequencies for these "direct coupling"

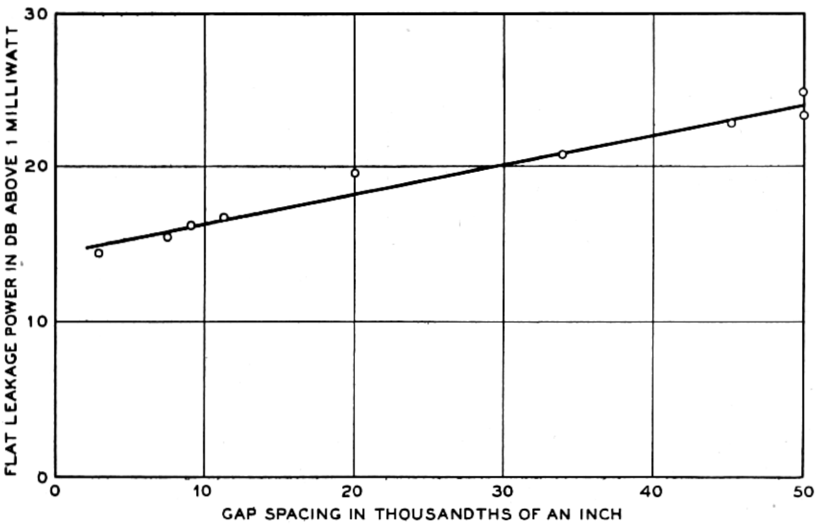

Fig. 24—The variation in flat leakage power with gap length, measured at 3000 megacycles

modes in the TR box and so be transmitted without much attenuation. Normally, direct coupling is of interest only in very high power systems.

Receiver Self-Protection. The fact was mentioned earlier that a receiver can provide itself with a certain amount of self-protection as a result of its change of impedance with level. This effect is still of use even in systems employing TR boxes. Unfortunately the apparent source impedance at the TR box output terminals is different for the different components of the leakage power so that the self-protection feature cannot be utilized for all components simultaneously. The matter is further complicated by the fact that the converter crystals themselves vary greatly in their impedance and in their variation of impedance with power level. The best designed converters as far as crystal protection is concerned are usually those which provide a certain amount of self-protection against the spike. It has been estimated that this self-protection seldom exceeds 2 db in practice.

Leakage Power Measurement. The c-w method of measuring the flat power has already been mentioned. Spike energy and direct coupling must, of course, be measured under normal high level operating conditions. Relative measurements of the spike can be made with an oscilloscope, acting ballistically, and the factors which affect the spikes can be studied in this manner. A correlation between the relative spike energies and the degree of crystal protection can be obtained by trial and from this correlation the operating conditions for adequate protection can be determined. Most of the early studies were made in this way. It is possible to deduce absolute values for spike energy, flat power, and direct coupling from measurements made when all three are present because of the different ways in which these parameters vary with the recurrence rate, pulse length and transmitter power. The method of doing this is outlined in appendix D.

A more precise method of measuring the spike energy involves the cancellation of the flat power by a signal of adjustable phase and amplitude obtained from the high-level transmission line. The average spike power is then measured directly and energy per spike computed. Most of the spike data quoted earlier were obtained in this fashion.

High-Level Loss. The power dissipated in the TR box as a result of the gas discharge is not ordinarily a large enough fraction of the total transmitter power to be of any concern. Using the P_0 values previously quoted, it is possible to compute the gas discharge power by the use of Fig. 16. At a line power of 100 kw and a low-level loss of 1 db the gas power in the 721A tube is 63 watts. The corresponding figure for a 1.5 db box using the 724B tube is 47 watts. For these cases the high-level loss is therefore less than 0.005 db. Low as this fraction is in db it still may be high enough to affect the life of the TR tube, as discussed in a later section. No trouble of this sort is ordinarily encountered with the 724B or 721A tubes. The chief cause of failure of the 1B23 is from loss of Q and this in turn is caused by the sputtering action of the high-frequency discharge.

Recovery Time. As mentioned earlier a TR box must recover its low-level properties at the end of the transmitting pulse in a very short period of time. The actual "recovery time" is in fact several orders of magnitude smaller than the deionization times of the usual gas discharge so that a quite different mechanism must be involved. While an exact theory of the recovery is beyond the scope of the present paper, a qualitative picture of the recovery process may be of interest.

During the transmitting period the free electrons provide almost all of the discharge current, and are replenished by electron-molecule collisions. At the end of the transmitting period these electrons may migrate from the discharge region, they may recombine with the positive ions, or they may be captured by molecules to form negative ions. Negative ion formation by attachment effectively removes an electron from the discharge because of the great increase in mass. It is an experimental fact that those gases which readily form such ions (of which water vapor is the most common) are the gases which exhibit good recovery in a TR box. This process is

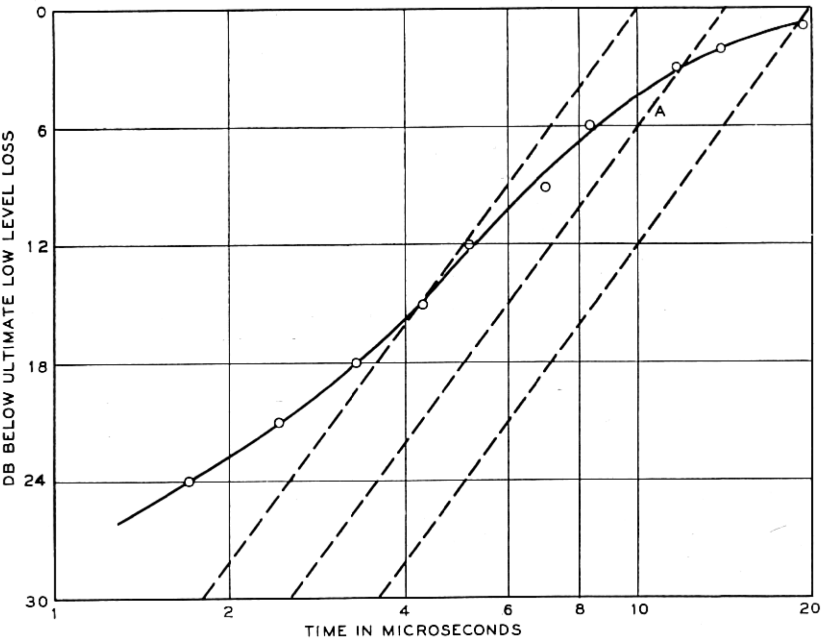

Fig. 25—A typical recovery time characteristic for the 721A tube in a TR cavity adjusted to 1.5 db low level loss with a transmitter power level of 100 kw peak

not deionization in the ordinary sense and it can take place at a surprisingly rapid rate.

Of course, immediately upon the termination of the transmitting pulse, the cloud of free electrons will cause an extremely high loss to any reflected signal but the loss will rapidly decrease to some limiting value set by the fixed losses in the TR cavity itself.

A typical recovery curve for the 721A tube is shown in Fig. 25. This curve has a particularly fortunate shape in that the variation in loss with distance, or more correctly with time, is at approximately the same rate as the variation in the reflected signal level with distance for a target of fixed size. The importance of this can be understood by considering the way in which the reflected signal intensity varies as an object of fixed size approaches a radar set from a great distance. Such an object as seen by a given radar set may be represented on Fig. 25 by a straight line having a slope of 12 db per factor of two in distance. Several such lines are shown. Considering line A it will be observed that this target can first be seen at a distance corresponding to 12 microseconds. Since this target line always remains below the TR recovery line for times shorter than that at the intersection point, an object once seen will remain in view continuously as it approaches

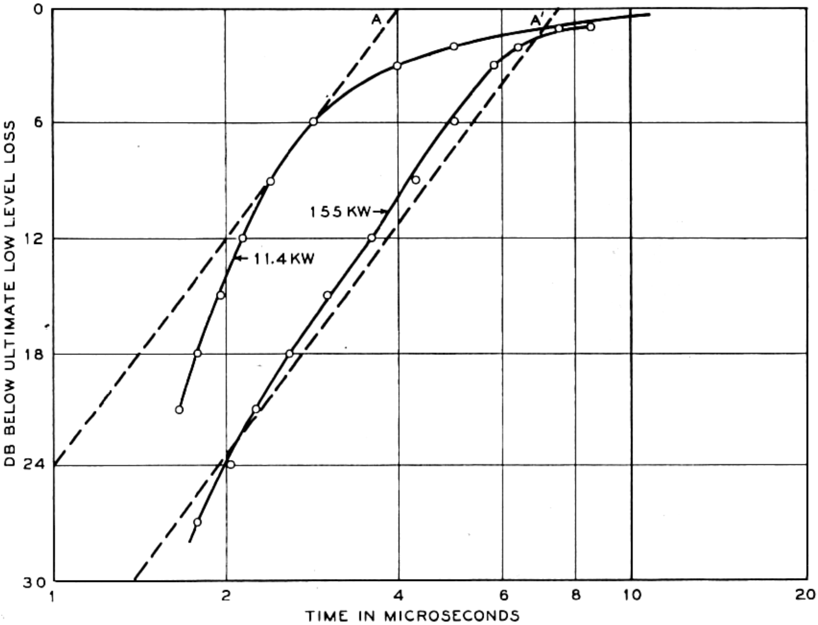

Fig. 26—The recovery time characteristic at two different transmitter power levels for the 724B tube in a 1.5 db TR box

the transmitter even in spite of the poor recovery characteristics of the TR box. A certain amount of sensitivity-time-control action is actually provided. The recovery time characteristic frequently does not necessarily set a lower limit on the effective range of a radar system although it always sets a limit on the smallest effective target size which can be observed.

The recovery time characteristic is critically related to the transmitter output power level as shown by the data for a 724B tube shown in Fig. 26. When the recovery time curves for two different power levels are compared, the target line which is just detectable at a power level of 11.4 kw is shown as the line marked A Contrasting with the 721A behavior this target would be visible only at one point and would then be lost from sight as it approached the transmitter. This same target is represented on the plot as line A′ for a transmitter output power of 155 kw, that is being displaced vertically by approximately 12 db to take account of the difference in transmitter power. At this higher power level the target would be visible at a much greater distance (corresponding to 7 microseconds elapsed time) and would remain in view until the target distance corresponded to 2.1 microseconds time. In this case an increase in power by 12 db resulted in an increase in range by a factor of 2.8. While this would indicate that an increase in range can be obtained by increasing the transmitter power, it should not be inferred that an increase in the near-range sensitivity will always result from an increase in power. At any specified range there appears to be a unique value of transmitter power output beyond which the loss in TR box recovery more than offsets any increase in range due to higher output powers. While accurate figures are not available for the 721A tube, there is some evidence that an output power of 100 kw is already too large for ranges corresponding to elapsed times of 10 microseconds or less. Under these conditions improved operation results from a decrease in the transmitter power level. Such an effect has never been observed by the writers with the 724B tube, probably because the transmitter powers available in its operating frequency range have usually been somewhat less than that available with the 721A tube.

It should be noted, at this point, that the recovery time does not depend upon the transmitter power only, but rather upon the gas discharge power which is a function of both the transmitter power and the low-level loss adjustment of the TR box as shown by Fig. 16. A very great improvement in near range sensitivity can usually be obtained by increasing the transmitter power level and at the same time increasing the low-level loss adjustment of the cavity to limit the gas power to a value for which the recovery time is satisfactory. This of course increases the ultimate low-level loss and so adversely affects the long-range sensitivity.

The dependence of the recovery time on the ambient temperature for the 721A tube is shown in Fig. 27. The 724B tube is much less temperature dependent. This variation in recovery time with temperature is caused by the reduction in water vapor pressure through condensation, as shown by the identity of the recovery curve for a standard 721A tube at −186° C. with a special tube filled with hydrogen only.

With continued life the water vapor content of the tube decreases with the corresponding change in the recovery time characteristic. Fig. 28 shows the effect with the 721A tube. The dependence of the recovery time on the water vapor content in the 724B tube is shown in Fig. 29. Comparing this curve with Fig. 27, it will be observed that the loss of water vapor has much less effect on the recovery characteristics of the 724B tube than on the

721A tube. It should be noted, however, that the 724B tube frequently reaches the end of its useful life as a result of its failure to provide adequate receiver protection before serious loss of recovery occurs.

The ATR, if one is used, can also contribute to poor recovery as may be seen by referring to Fig. 30. These data are not necessarily representative since it is possible to adjust the length of line between the magnetron

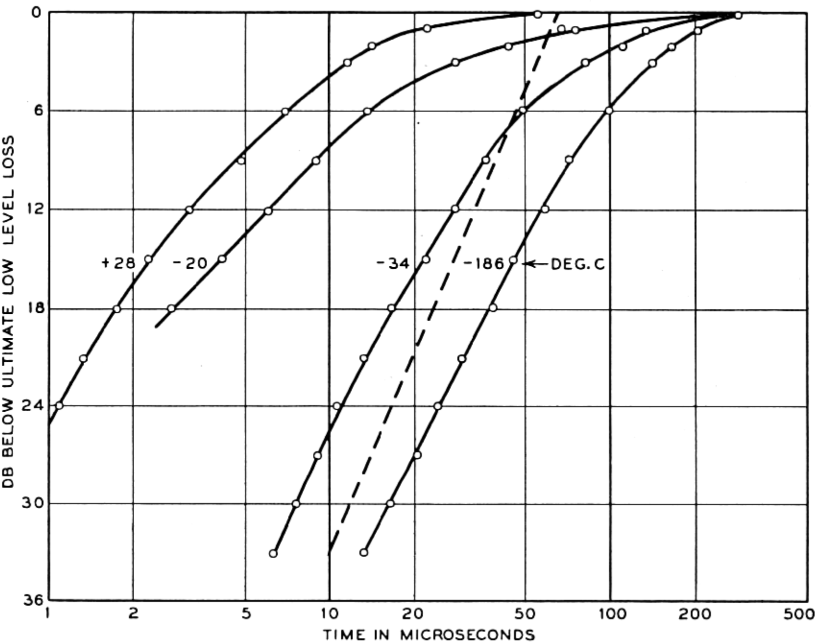

Fig. 27—The dependence of the recovery time on the ambient temperature for the 721A tube

and the ATR junction so as to minimize the effect. Nevertheless the effect is important and should not be overlooked.

Low-Level Loss. An analysis of the low-level loss must take into account two components of loss, the first resulting from power loss in the TR cavity itself and the second resulting from the fact that some power will always be absorbed by the transmitting branch of the system.

The relationships existing between the low-level loss adjustment of a TR box and its other performance characteristics have already been discussed. One aspect of the problem, not previously considered, has to do with the dependence of the performance on the Q of the cavity. This is clearly shown in Fig. 15 which has already been referred to in a different connection. From this aspect, at least, the higher the Q of the tube and its associated cavity the

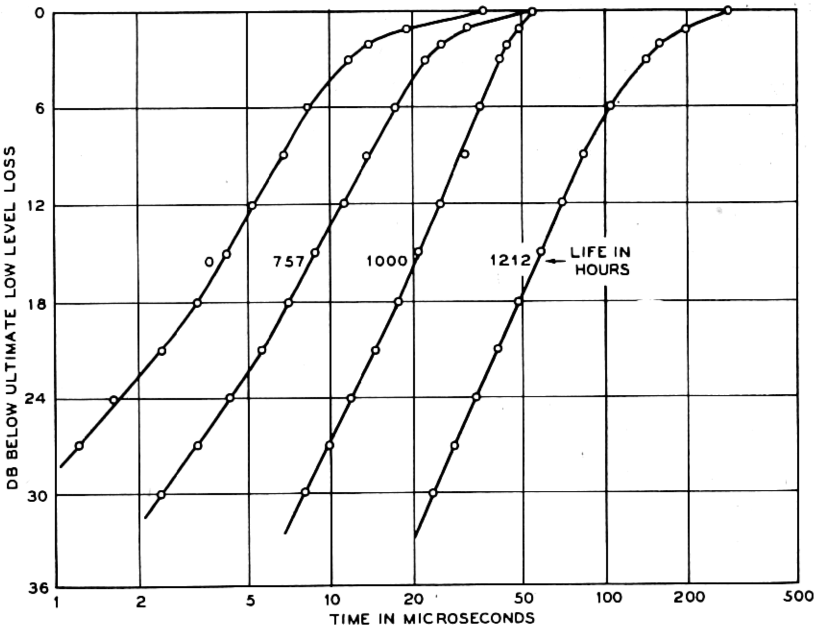

Fig. 28—The variation in recovery time with life for a typical 721A tube

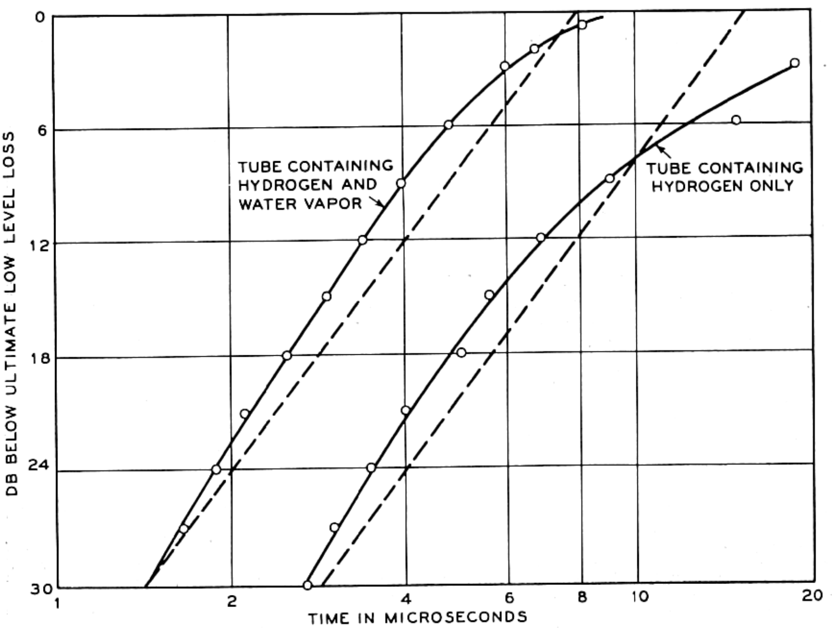

Fig. 29—The variation in recovery time with gas content for the 724B tube

better the over-all performance of the system.* Any basic improvement in Q can be reflected either in a lowering of the leakage power or in a reduction of the low-level loss, as may be desired.

Variations in the Q between tubes used in a given cavity of fixed design, are, however, seen as variations in the low-level loss only and have no

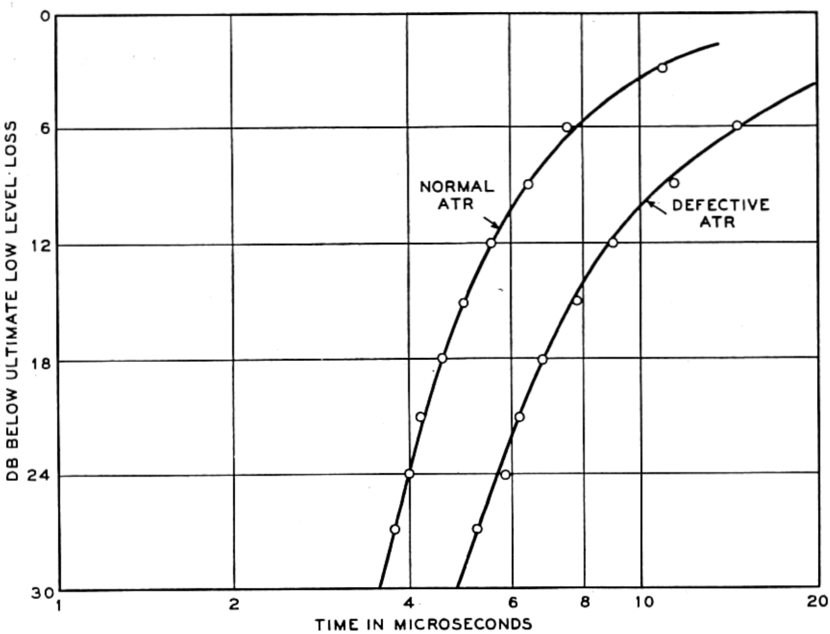

Fig. 30—The ATR recovery effect

noticeable effect on the leakage power. This may be understood by reference to the equations derived in appendix A. The performance of a somewhat idealized TR box may be expressed in terms of three design parameters δ_0, δ_1 and δ_2 which relate respectively to the properties of the cavity, its input coupling, and its output coupling; and in terms of a gas discharge parameter P_0. In terms of these new parameters the in-tune low-level tansmission of a TR box is given as a power ratio by

$$T = \frac{4\delta_1\delta_2}{[\delta_0 + \delta_1 + \delta_2]^2}. \tag{12}†$$

* As explained later in this section, band width limitations set an upper limit to the permissible Q.

† Numbered equations in the text correspond with the numbers used in the appendices.

The input standing wave ratio on the input line is given by

$$\sigma = \frac{\delta_1}{\delta_0 + \delta_2}. \tag{13}$$

The leakage power is similarly given by

$$P_r = P_0\,\delta_2 \tag{14}$$

while the gas-discharge power in terms of these parameters and the power in the transmitting line (P) is given by

$$P_g \doteq (P\,P_0\,\delta_1)^{1/2} \tag{19}$$

The parameters δ_1 and δ_2 are properties of the input and output coupling as they are geometrically related to the cavity and are substantially independent of the Q of the cavity. The parameter δ_0 is, however, the reciprocal of the intrinsic or unloaded cavity Q. Equation 12 is seen to depend upon δ_0 but equations (14) and (19) do not. The effect of variations in Q is thus demonstrated.

The over-all performance is also affected by the relative values of δ_1 and δ_2. In view of the dependence of P_r on δ_2 directly and on δ_1 indirectly through the fact that P_0 is not entirely independent of P_g, it is advantageous to adjust the values of δ_1 and δ_2 so that the input standing wave ratio (σ of equation 13) is unity. Such a condition is also very desirable for system reasons as well. When this condition is met, equation (12) reduces to

$$T = \frac{\delta_2}{\delta_1}.$$

The curve marked $\sigma = 1$ of Fig. 15 and the curve of Fig. 16 were plotted on this basis. It should be noted that matched input requires that the input window be larger than the output window.

TR boxes are unfortunately not always operated in the in-tune condition, and they must also pass a band of frequencies as fixed by the narrowness of the transmitter pulse. For these reasons the Q must not be set at too high a value. The additional low-level loss which results from off-tune operation may be computed from equation 28 of appendix A.

Incidentally, it is an experimental fact that the leakage power and the gas-discharge power are not materially altered by small departures from the in-tune adjustment, presumably because of the very low effective Q of the gas discharge.

The ATR Low-Level Loss Component. The component of low-level loss which results from losses of power to the transmitting branch depends very greatly upon the "cold impedance" of the magnetron or other transmitting tube and upon the properties of an ATR box if one is used. As shown in appendix C the loss chargeable to the ATR and the associated transmitting arm can be expressed as a factor F given by

$$F = \frac{4}{(2 + G)^2 + B^2} \tag{33}$$

where G and B are respectively the conductance and susceptance of this branch in units of the surge admittance of the transmission line. Since

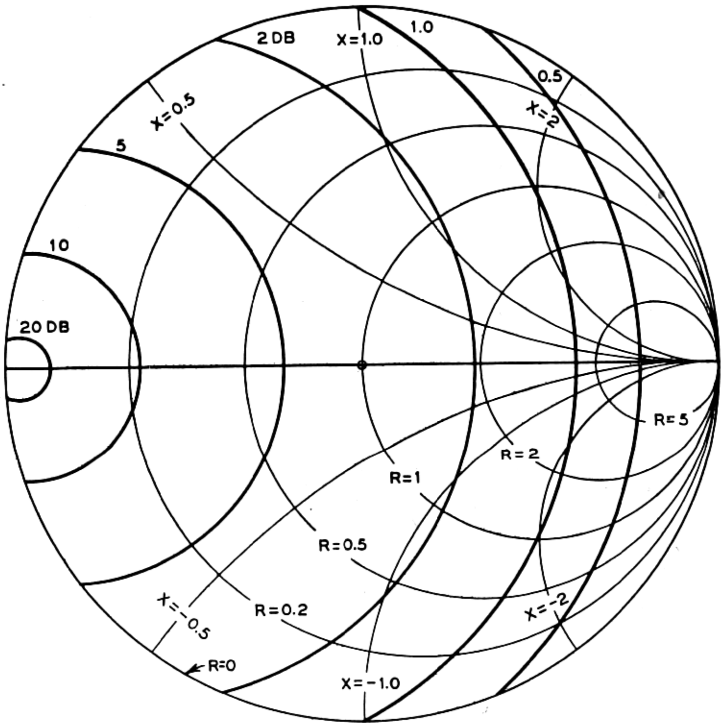

Fig. 31—Curves for constant ATR loss in db as a function of the impedance of the transmitting branch

curves for constant values of F appear on the reflection coefficient plane (Smith transmission line chart) as circles this presentation is very convenient. Fig. 31 is such a plot (impedance circles rather than admittance circles are shown).

If now an ATR is introduced having a resistive component of impedance the range of values of G and B is restricted so that a minimum value of F exists for any random value of the magnetron impedance. With variations in the magnetron cold impedance or in the effective length of line between the magnetron and the ATR junction the value of F will vary between this minimum value and some maximum value which may approach unity. For example if the ATR is adjusted to have the same gas-discharge power as that in a TR adjusted for a transmission of T, its low-level in-tune input impedance will be

$$Z = \frac{1}{1 - T}. \tag{38}$$

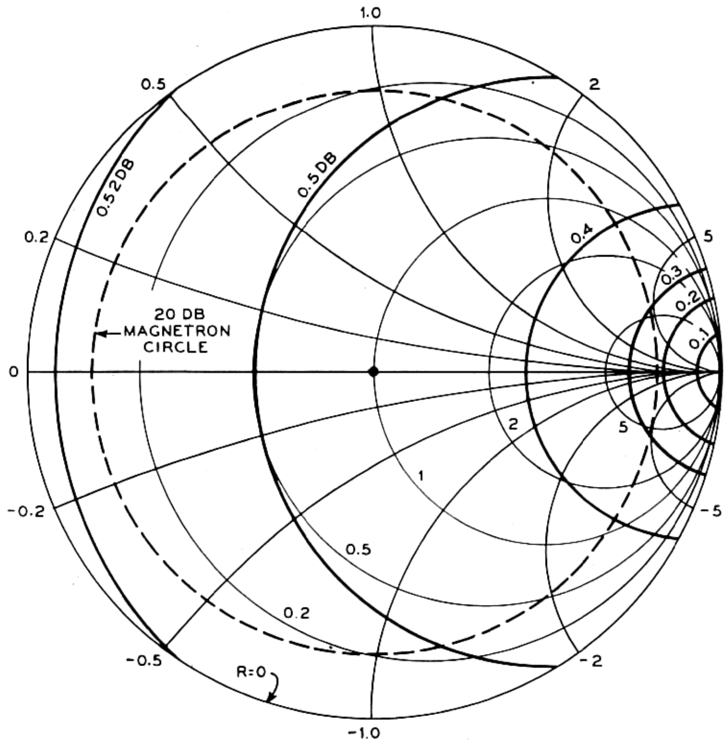

Fig. 32—The ATR low level loss as a function of the magnetron impedance with an ATR adjusted to an impedance of $8 + j0$

Actually since the gas-discharge power is usually not the limiting factor in the design it is possible to adjust the ATR to have an input impedance of 8 or 9 (in terms of the line characteristic impedance), corresponding to a TR low-level transmission of the order of 1/2 db ($T = 0.89$) and yielding an ATR loss of approximately 1/2 db.

Since the exact value of the impedance of the magnetron branch is not necessarily known it is convenient to show the dependence of the loss factor F for any given ATR on the magnetron impedance by a plot somewhat similar to Fig. 31 but transformed to the magnetron side of the ATR junction. This may be done by subtracting the ATR impedance from the values read off of Fig. 31 corresponding to desired values of F and replotting these on the reflection coefficient plane. As an example, the in-tune value of Z for one typical 724B ATR cavity is $8 + j0$. Points lying on the $R = 8$ circle on Fig. 31 will then lie on the magnetron $R = 0$ circle, the region inside being distorted and expanded to fill the entire positive R region on the reflection plane. The results are shown in Fig. 32. From this plot it is evident that the maximum possible low-level loss chargeable to the transmitting branch would be slightly more than 0.52 db and that this would occur only for a restricted range in the value of magnetron impedance. As a matter of practical interest the "cold impedance" of the usual magnetron is such as to give at most a 20-db standing wave. This restricts the possible range in impedance values to the area on Fig. 32 within the dotted circle, thus limiting the maximum loss to slightly less than 0.52 db, and imposing a minimum loss limit of 0.22 db.

This type of analysis may be extended to consider the ATR loss during the recovery period if desired although the problem becomes rather complicated as a result of the simultaneous variation in input impedance of both the TR and the ATR.

TR Box Design Considerations

The desired electrical properties for a TR box can of course be achieved in a variety of different physical structures. A construction technique which separates the gas-discharge tube from the rest of the TR box cavity offers many advantages. In the first place the cost of the entire device is kept low by reason of the fact that it is not necessary to transmit the tuning motion through the vacuum-tight tube enclosure. The replacement cost is also greatly reduced since the more complicated part of the TR box is a permanent part of the equipment. Then, the same tube structure can be used for a variety of different types of equipment operating in different wavelength bands and requiring different amounts of receiver protection by the use of different size cavities and different size coupling windows. This greatly simplifies the problem of maintaining replacement stocks. An additional factor, which was of importance during the early days of the war, is that the design of such a tube can be frozen at an early stage, before all the possible circuit aspects of the TR problem have been solved since changes in the external parts of the TR box can be made independent of the design of the replaceable tube element. The widespread use of the 721A and 724B vacuum tube is, in a sense, proof of the essential soundness of the arguments for the external cavity type of construction.

The chief difficulty to be overcome in the design of a separate cavity type of TR box has to do with the need for a low-loss contact between the internal portions of the tube and the external cavity. A copper-disc sealing technique, developed at the Bell Laboratories in connection with the construction of water cooled tubes* and later superseded by the now conventional Housekeeper seal, had previously been applied at ultra-high frequencies in the design of oscillators and amplifiers. This technique makes possible very satisfactory high-frequency connections by simply clamping the external portion of the disc between machined surfaces. The flexibility of the copper discs is sufficient to compensate for minor machining errors and for differential thermal expansions while the relative softness of the copper insures a continuous contact around the entire periphery. The goodness of contact provided by these contacts is evidenced by the fact that Q's of 4000 and greater are obtained at 3000 megacycles with discs of the 721A type. This technique was therefore adopted for the 721A tube and the 724B tube and for one electrode of the 1B23 tube. The second high-frequency electrode of the 1B23 was made in the form of a rod terminating in a ball for convenience in replacing tubes since the accompanying loss of Q can be tolerated in the frequency range where this tube is used.

In an external cavity type of TR box the over-all goodness of the design is largely determined by the design of the gas-discharge tube. It is the tube designer's responsibility to determine the optimum shape and size for the copper discs and for the glass tube envelope and to determine the optimum gas composition and pressure, with due consideration being given to such matters as mechanical ruggedness, manufactureability and freedom from undesirable ambient temperature, pressure and humidity effects.

With the copper-disc type of tube the system designer has at his disposal the ability to vary the design of the external cavity, and to arrive at any specific compromise between the various conflicting performance criteria which he feels to be the best for his particular application. For example, in systems employing vacuum tube converters it is customary to adjust the TR box for a low-level loss of 1 db or somewhat less since receiver protection is of minor interest while in systems employing crystal converters it is customary to fix the low-level loss at 1.5 db or sometimes as high as 2.0 db. Certain cavities, notably the one shown in Fig. 5, have to be designed to have an extended tuning range, in this case achieved by a piston tuner with, however, some loss in Q, while other cavities, the one shown in Fig. 11 being typical, do not require this same tuning range and a different tuning mechanism (in this case, tuning plugs) can be employed.

An extreme example, illustrating the advantages to the system designer

* W. Wilson, "A New Type of High-Power Tube," *B. S. T. J.*, vol. 1, p. 4, July 1922.

of the external cavity type of tube, is that of certain radar systems which were required to be capable of receiving signals on occasion at a frequency differing from their normal tuning. This was done by a solenoid-operated plunger which could be preset to alter the tuning of the cavity by the desired amount whenever the solenoid was energized.

The Tube Design

The 702A and 709A vacuum tubes, as previously mentioned, were put into service with little or no consideration of their real suitability. With these stop-gap designs in production the basic design problem was given serious consideration, with separate studies being made of the mechanical design considerations as they relate to the size and shape of the discs and glass of the tube, and of the gas filling.

The exact shape of the disc is determined first by the total tuning range which is to be required of the tube, and second by the necessity for maintaining the Q of the structure as high as possible. It has been shown that in a spherical resonator with coaxial cones the maximum Q occurs when the cone half-angle is nine degrees. The copper-disc tube can only roughly approximate the ideal spherical resonator; nevertheless it appears desirable to use cones of this angle. The disc spacings and diameters are so chosen that the tube resonates at the shortest wavelength at which it is to be used in a "square" cavity; i.e., one in which the inside diameter approximately equals the height. Such a cavity is about the closest practical approach to a sphere. The glass diameter is made as large as mechanical considerations permit so it is as far as possible removed from the region of high electric field intensity.

The experimental results of Fig. 24, previously noted, indicate that the leakage power of a TR box decreases as the gap spacing decreases; thus one is tempted to make the gap extremely small. Too small a gap is very troublesome, however, since such a gap has an unreasonably rapid variation of resonant frequency with gap separation, making the tuning extremely subject to change as a result of dimensional variations due to processing or to temperature changes. Accordingly one chooses a compromise gap separation. The electrode radius at the gap must be large enough to permit the radio frequency glow discharge to dissipate the required power without excessive spreading, and must be determined by experiment.

Rather than attempting to hold all of the mechanical variations in the tube (including glass thickness) to the necessary tolerances to insure the desired uniformity in tuning, the tubes are pretuned before exhaust by deforming the copper discs. The tubes are placed in a special cavity and the disc inside the envelope distorted by a tool until resonance is obtained at a specified frequency. It is quite easy to tune tubes in this way so that they are uniform to within $\pm$ 0.25%.

Unless the tube is properly designed, changes in ambient temperature may seriously affect its resonant frequency. The part of the disc which is inside the glass envelope may be considered as a diaphragm supported around its periphery by the glass which has a temperature coefficient of expansion negligibly small compared to that of the copper. An increase in temperature, which causes the copper to expand, will force the cone tip to move toward or away from the gap, depending on the initial slope of the nearly flat portion of the disc. The temperature coefficient of frequency may be either positive or negative, and will have extreme variations in magnitude from tube to tube if consideration is not given in disc design to avoid such difficulties.

A cavity made wholly of copper will have a fractional change in wavelength with temperature the same as the fractional change in length of copper (approximately fourteen parts in a million per degree centigrade). As the temperature increases, the frequency decreases. At a frequency of 1000 mc, the approximate temperature coefficient of frequency is $-.014$ megacycles per degree centigrade; at 3000 mc it is $-.042$ mc/°C; and at 10,000 mc it is $-.14$ mc/°C. Magnetrons normally have temperature coefficients of about these magnitudes. The ideal TR tube would have the same coefficient as the magnetron; practically speaking, any coefficient between zero and twice the value for copper is satisfactory.

It is practical to make a copper disc structure which has the required temperature coefficient. Fig. 33, which is a cross-section of a 721A tube, shows how temperature compensation within the tube is effected. The disc is slanted away from the center portion of the tube, so that as temperature rises the cone is carried away from the gap. At the same time the cone itself expands; the net effect is to increase the gap between the two cone tips. The angle of the slanted part of the disc must be such that the gap increases with temperature at the same rate that it would in an all-copper cavity. If this condition is fulfilled, the net result of the expansion of all the tube parts, and of the cavity itself, will be the same as if it were all made of copper. This result is achieved by an experimental series of successive approximations. A number of models are built until the angles are found which give the desired temperature coefficient of frequency.

The gas content of the tube was the subject of considerable study. As stated in the section on Recovery Time, gases which readily form negative ions are invariably the most satisfactory from that viewpoint. Gases of low ionization voltage, such as the rare gases, give excellent protection but usually have extremely poor recovery. The choice of a TR gas must of necessity be a compromise between the two requirements. Some otherwise satisfactory gases are not useful because of other characteristics. HCl is an excellent TR gas, but is very corrosive. Freon, a common refrigerant, is excellent but is unstable. In general, no gas which contains a solid elementary constituent is a satisfactory TR gas. The most satisfactory gas found was water vapor. It is cheap, stable, and easy to handle. Water vapor alone is not safe to use at a low temperature, so a small amount of hydrogen is added to ensure adequate protection when the water vapor is frozen out.

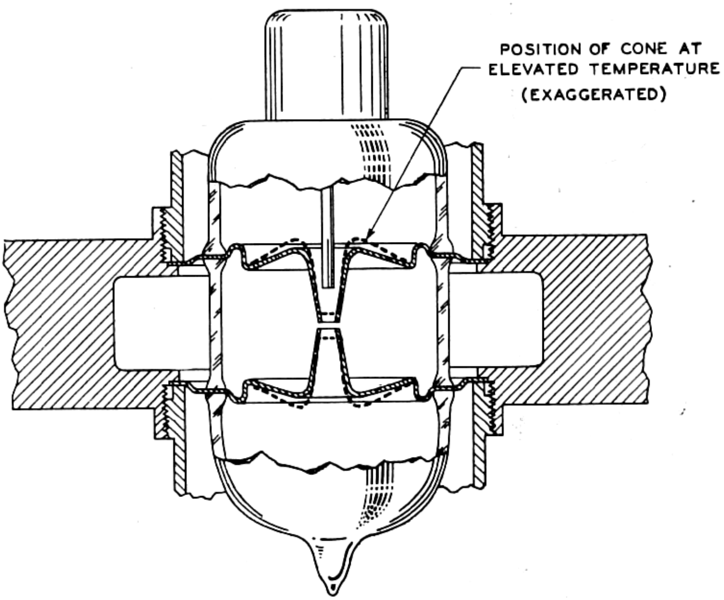

Fig. 33—Cross section of a 721A TR Tube, showing the special shape of the temperature compensated discs

The life of a TR box is in general determined by the gas volume. The radio-frequency discharge consumes no gas except at extremely high power levels; the igniter discharge accounts for the greater part of the loss of the gas initially placed in the tube. Reduction of the water vapor to hydrogen by formation of copper oxide on copper parts of the tube seems to be the principal process which goes on. This change results in no change in total pressure until the water vapor is exhausted; thereafter sputtering becomes more important and accounts for a fairly rapid hydrogen clean-up. The life of a TR tube is determined by the igniter current, which is maintained at a value as small as possible consistent with adequate spike protection, and by the volume of the tube.

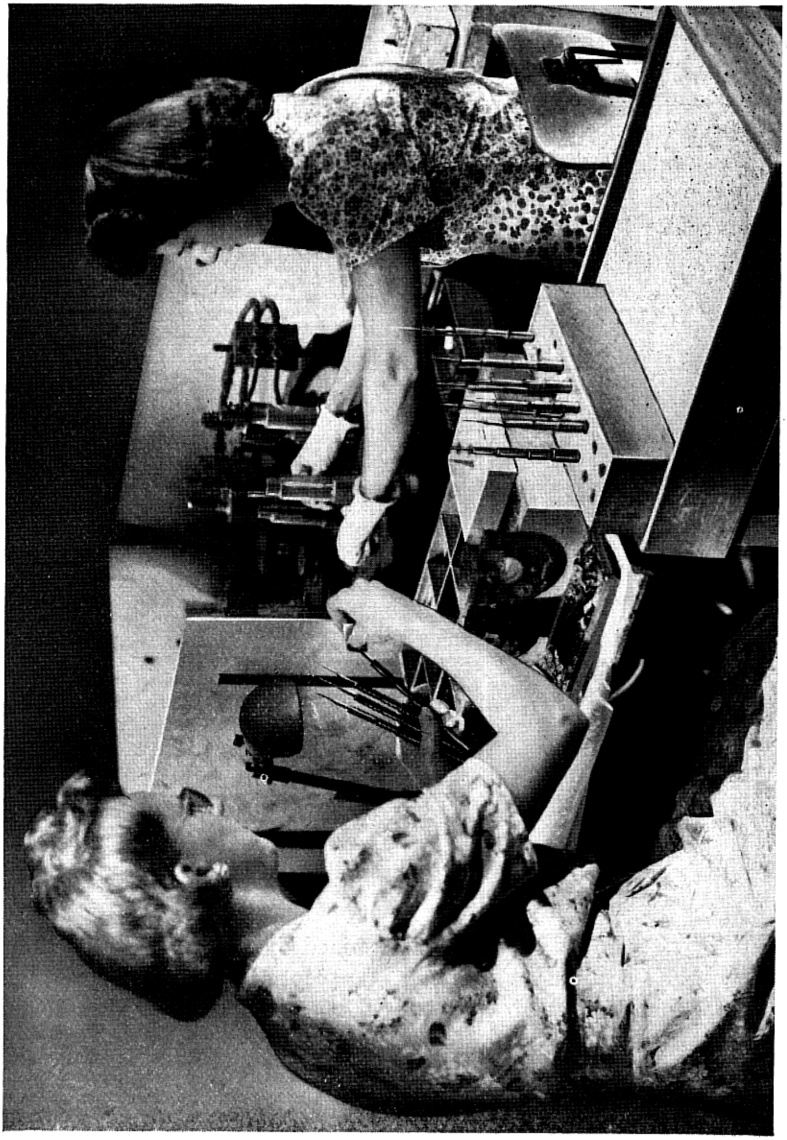

Fig. 34—Setup for making 724B copper-to-glass disc seal

The lack of water vapor in a tube which has been operated for some hundreds of hours may manifest itself by a failure in either protection or recovery time. The operating frequency determines which failure becomes important first; at long wavelength it is likely to be recovery, while at shorter wavelength the spike protection is likely to fail first.

The life of a TR tube operated without igniter is very much longer. This may be understood from the picture given above under "recovery," of the state of affairs existing in the radio-frequency discharge. Electrons do not completely traverse the gap, but oscillate about some mean position, while the positive ions hardly move at all. Thus there is little more interaction between the metal electrodes and the gas molecules with the R.F. discharge on than with it off. A few 721A's have been operated without igniter for as long as 5000 hours with no measurable change in either protection or recovery. This experiment was done at a transmitter power level of 250 kw. peak power. The best life that can be expected with the igniter operating at 100 microamperes is 500 hours, at which time the recovery time is badly deteriorated. In order to maximize the life of the tube, the initial gas filling consists of a minimum amount of hydrogen and as much water vapor as may be introduced without causing excessive leakage power (see Figs. 20 and 23).

Manufacturing and Testing

Some interesting problems occur in the manufacture of copper-disc seal TR tubes which are quite different from those encountered in the construction of more conventional tubes. The copper-disc seals are usually made by high-frequency induction heating. Close control of the spacing between discs must be maintained during the bulb-making operation in which the discs are fused to the glass parts of the tube. One way of accomplishing this is shown in Fig. 34, which depicts a machine setup for making the 724B TR tube. The parts are held by lavite forms which support and locate them during the bulb-making process. The seal is made possible by a correct choice of copper thickness. The copper disc is stressed due to forces set up by the different expansion coefficients of the glass and the copper, and if too thick will pull the seal apart. If too thin, the copper itself will tear. Nevertheless, a properly designed copper-disc seal is very strong; the copper-disc seal TR tubes will pass the JAN1-a* mechanical and thermal shock tests for glass tubes without any difficulty.

The electrical pretuning operation, referred to earlier, comes right in the middle of the manufacturing process. Before the igniter is sealed in, the bulb is placed in a special pretuning cavity. The setup includes an oscillator

* Joint Army-Navy Specification for electron tubes.

of appropriate frequency range, a wavemeter, and some device for indicating resonance. The part inside the glass of one of the copper discs is bent, by gentle tapping, until the resonant frequency of the bulb in the special cavity is within the required tolerance (which may be as small as 0.25%) of the pretuning frequency.

No heat treatment of any kind is used in the pumping of TR tubes. It is obviously unnecessary to subject the tubes to the usual baking, the principal purpose of which is to remove water vapor film from the tube parts. On

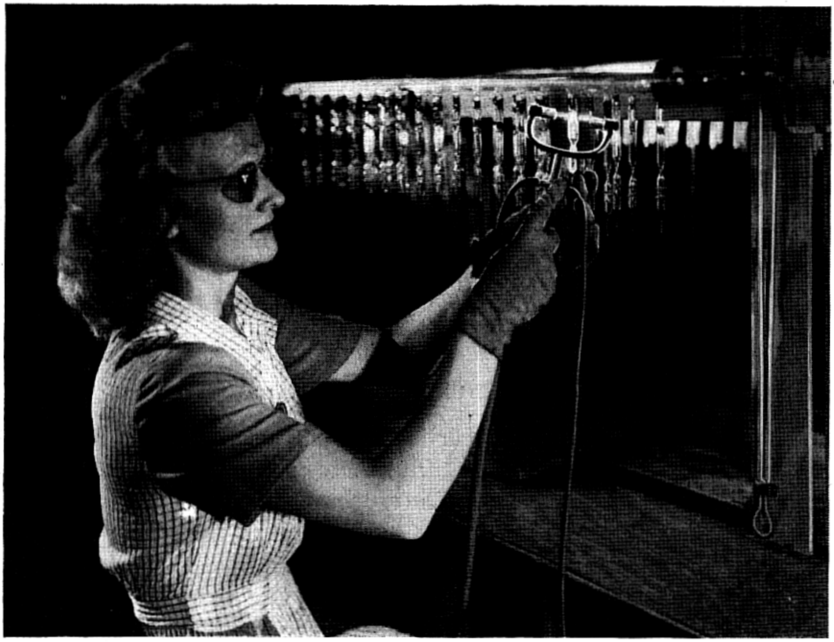

Fig. 35—Pump Station for the 724B Vacuum Tube

the contrary, it is difficult to control the water vapor pressure in tubes which have been baked as the parts absorb a surprising quantity of water. The tubes are filled to fairly high pressure, so a diffusion pump is not necessary. Fig. 35 shows a pump station used in production of the 724B.

The test procedure for TR tubes must verify that each individual tube will fulfill its fourfold function of protecting the crystal, of recovering rapidly, and of introducing neither excessive high-level loss nor excessive low-level loss. The tuning of each tube must be verified, and it must pass mechanical and dimensional tests. Fortunately a tube which is otherwise sound will never introduce excessive high-level loss, so no specific test is required.

The protection test may be made either with a c-w oscillator of sufficiently high level to ionize the TR tube gap, or with a magnetron in the equivalent of a radar microwave head. The high-level test bench used for the 724B is shown in Fig. 36. This bench uses a radar microwave head fitted to special plumbing. In either case, the leakage power is measured at a specified R.F. level. For production testing, an actual measurement of recovery time is not used. Instead a test which measures the quantity of

Fig. 36—High Level Test Bench for the 724B Vacuum Tube

water vapor in the tube in a relative way has been developed. This test involves touching some part of the tube envelope with a piece of carbon dioxide ice, so that most of the water vapor is frozen out forming a small spot of ice on the inside of the tube. Only the hydrogen remains, and the resulting change in either the leakage power or the igniter arc drop is indicative of the quantity of hydrogen and of water vapor in the tube. Careful correlation must be made between this simple dry-ice test and absolute recovery time measurements; experience has shown the dry-ice test to be reliable and in the hands of a skilled operator very informative.

Low-level loss and tuning are checked at such low level that the gas does not ionize in a cavity of restricted tuning range. Every tube must resonate within the range of the tuning adjustment, and the transmission loss through the test cavity must not be excessive.

Two additional tests are made at the time the d-c igniter characteristics are checked. One, igniter interaction, is important in the 721A, the 10 cm. TR tube. This tube has rather large openings in its cone tips, so that if the igniter electrode is sealed in too close to the cone tip, the glow discharge which surrounds it may extend out into the gap. Such a defective tube will show igniter interaction; the low-level loss through it will be more when the igniter arc is on than when it is off. Normal tubes do not show this effect.

The 724B 3 cm. TR tube has such a tiny opening in its cone tip that igniter interaction does not occur. The tube is more subject to igniter oscillations, perhaps because it is filled to a higher gas pressure. The consequences of these oscillations was explained in the section on The Spike. Igniter oscillations are usually due to an improper gas filling, and are detected by means of a cathode ray oscilloscope.

The above tests are made on each tube as it comes off the production line. Some additional tests are made on selected samples to insure that the quality is being maintained. Selected tubes are subjected to mechanical shock tests and to temperature variation tests to verify both their resistance to thermal changes and that their temperature coefficient of frequency is not excessive. Absolute recovery time, Q, and leakage power tests are made on these tubes, and some are set aside for life testing. By all these tests the important electrical properties of the tube are under constant scrutiny and the danger of shipping defective tubes is minimized. The importance of adequate testing can hardly be over-emphasized, as a defective TR tube may render a whole radar system inoperative.

Acknowledgements

Because of the very close liaison maintained during the war period between various industrial and governmental laboratories, the developments described in this paper were carried on with the constant advice and criticism of many individuals. It is not therefore possible to assign credit to specific individuals for any particular aspect of the work with any certainty. The authors would be remiss, however, were they not to call attention to the many contributions made at the M.I.T. Radiation Laboratory particularly by members of the group under the direction of Dr. J. R. Zacharias and later Dr. A. G. Hill. Colonel J. W. McRae assisted in the early formulation of the TR problem and Captain A. Eugene Anderson did much of the original development work on the 724B tube while at the Bell Laboratories and was later involved in the formulation of test methods and test limits in connection with his Signal Corps work. Mr. C. F. Crandell of the Southwestern

Bell Telephone Company, while at the Bell Laboratories, was responsible for the construction of test equipment and for most of the recovery time measurements reported in this paper. At various periods during the development work Messrs. A. B. Crawford, V. C. Rideout, G. M. Eberhardt and J. P. Schafer were closely associated with the measurement of the system performance of TR tubes. Mr. R. M. Purinton of the Bureau of Ships deserves much credit for his encouragement and assistance in the standardization program which led to the adoption of the 721A and 724B tube designs by all manufacturers. The Thermionics Branch of the Evans Signal Laboratory provided the bulk of the electrical standardization, calibration and engineering service associated with these tubes and assisted in the development of improved test methods. The magnificent production job done by the Western Electric Company and by other manufacturers, particularly by the Sylvania Electric Products Inc. in making these tubes available to the armed services also deserves mention. Perhaps the final mention should go to the many circuit design engineers both within the Bell Laboratories and elsewhere who handled the many difficult problems relating to the design and use of TR cavities in actual radar systems.

APPENDIX A

Analysis of the Idealized TR Box

Schelkunoff has shown* that the impedance of a resonant cavity can be represented in terms of its resonant frequencies as

$$Z = \sum_a \frac{Z_a}{j\left(\frac{\omega}{\omega_a} - \frac{\omega_a}{\omega}\right) + \frac{1}{Q_a}} \tag{1}$$

or in the vicinity of any single resonance as

$$Z = Z_1 + \frac{Z_n}{j\left(\frac{\omega}{\omega_n} - \frac{\omega_n}{\omega}\right) + \frac{1}{Q_n}}. \tag{2}$$

Under most conditions the Z_1 term is negligibly small. We are therefore justified in thinking of the generalized resonant cavity used as a TR switch as a shunt resonant circuit to which are coupled input and output circuits. For the moment we will consider (1) that these external circuits are resistive only, (2) that the Z_1 in equation (2) is zero; and (3) we will restrict the analysis to the in tune condition.

When the cavity is excited by energy supplied from the input circuit there exists in the cavity a certain amount of reactive power which will be

* S. A. Schelkunoff, "Representation of Impedance Functions in Terms of Resonant Frequencies," *Proc. I.R.E.*, vol. 32, pp. 83–90, February (1944).

designated by the symbol P_0. Of this power, a certain fraction δ_0 is dissipated as losses in the cavity itself where

$$\delta_0 = \frac{1}{Q_0}. \tag{3}$$

The symbol Q_0 with the subscript is further defined as the intrinsic Q, that is, the Q without external loading, to differentiate it from the more general Q_L which is the measured Q when the cavity is loaded down by external coupling. It should be noted that this definition of δ differs from the logarithmic decrement by a factor π.

When coupled to the external circuits the loaded δ is increased. On the assumption that the loading effects of the input and output irises are independent we can write

$$\delta_L = \delta_0 + \delta_1 + \delta_2 \tag{4}$$

where δ_L is the loaded δ, and δ_1 and δ_2 are respectively the input and output loadings. Physically the assumption underlying this expression is that the distribution of electromagnetic fields within the cavity is not seriously altered by the input and output coupling devices. This assumption should certainly be valid as long as the absolute values of the δ's are very small compared to unity. Since the δ's usually encountered are of the order of 10^{-3} or less, the assumption seems to be justified.

Equation (4) may be written

$$\frac{1}{Q_L} = \frac{1}{Q_0} + \delta_1 + \delta_2 \tag{5}$$

The values of δ_1 and δ_2 evidently depend upon the ratio of the apparent series resistance which the external coupling introduces into the resonant cavity to the effective reactance of the cavity, that is,

$$\delta_1 = \frac{k_1^2 R_1}{X} \tag{6}$$

where k_1 is the transformation ratio of the input coupling device, R_1 is the resistance of the input circuit and X is the cavity reactance. Similarly

$$\delta_2 = \frac{k_2^2 R_2}{X}. \tag{7}$$

The values of the δ's may be equally well considered as the ratios of the coupled conductance to the shunt susceptance of the cavity considered as a shunt resonant circuit so that equations (6) and (7) become

$$\delta_1 = \frac{G_1}{k_1^2 B} \tag{8}$$

and

$$\delta_2 = \frac{G_2}{k_2^2 B} \tag{9}$$

when the R's and X are replaced by their reciprocals and transformed from a shunt to a series circuit.

The equivalent circuit is shown in Fig. 37, where for convenience everything is referred to the cavity and the sources for receiving and transmitting are represented by constant current generators, I and I_m respectively.

The Low-Level Transmission. We are now in a position to express the low-level transmission of the cavity. For this purpose we will assume that

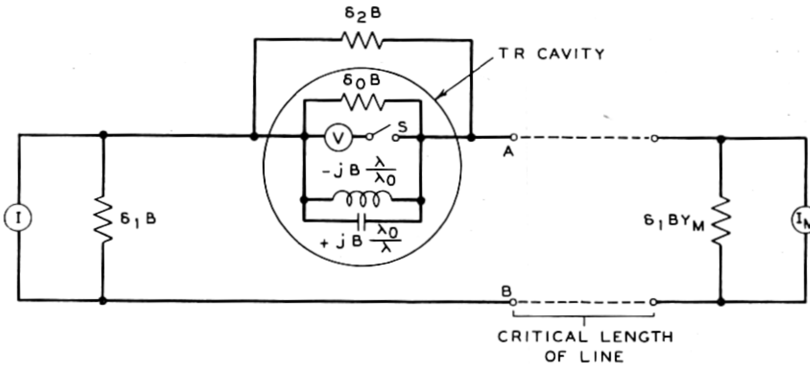

Fig. 37—Equivalent circuit of a system referred to the TR cavity

the admittance of the transmitting branch at plane AB is infinite. The available power is given by

$$P_{\text{avail}} = \frac{I^2}{4\delta_1 B}. \tag{10}$$

while the power actually going into the load is given by

$$P_{\text{out}} = \frac{I^2 \delta_2 B}{(\delta_0 + \delta_1 + \delta_2)^2 B^2} \tag{11}$$

the power transmission ratio defined as T is given by

$$T = \frac{4\delta_1 \delta_2}{(\delta_0 + \delta_1 + \delta_2)^2}. \tag{12}$$

One additional expression is desired. This is the ratio of cavity input resistance to the resistance of the input circuit. This is evidently the reciprocal of the conductance ratio and is given by

$$\sigma = \frac{\delta_1}{\delta_0 + \delta_2}. \tag{13}$$

The symbol σ is used to call attention to the fact that the in tune impedance ratio is numerically equal to the voltage standing wave ratio on the input line.

The low-level behavior of the cavity is thus defined by three equations.

$$\delta_L = \delta_0 + \delta_1 + \delta_2 \tag{4}$$

$$T = \frac{4\delta_1 \delta_2}{(\delta_0 + \delta_1 + \delta_2)^2} \tag{12}$$

$$\sigma = \frac{\delta_1}{\delta_0 + \delta_2}. \tag{13}$$

High-Level Operation. The high-level performance of the cavity containing a gas discharge can be expressed directly in terms of our original definitions. Fig. 37 still applies, the transmitter admittance changing to its operating value which is assumed to be $\delta_1 B$ at the plane AB. When the gas discharge becomes conducting, the switch S is closed, the value of the reactive power in the cavity (P_0) is set by the character of the discharge and the leakage power is given by

$$P_r = P_0\delta_2. \tag{14}$$

A constant value of P_0 is equivalent to a constant value of V in the figure. The power dissipated in the cavity walls, the gas discharge and in the output circuit must evidently be given by

$$P_1 = \frac{I_m V}{2} = (PP_0 \delta_1)^{1/2} \tag{15}$$

if

$$V << I_m/\delta_1 B.$$

Of this power an amount called the excitation power

$$P_e = P_0\delta_0 \tag{16}$$

is lost in the cavity walls. The net loss of power in the gas discharge is given by

$$P_g = P_1 - P_r - P_e \tag{17}$$

or

$$P_g = (P\, P_0\delta_1)^{1/2} - P_0(\delta_0 + \delta_2). \tag{18}$$

Since the last term is usually very small compared to the first term, we may write

$$P_g \doteq (P\, P_0\delta_1)^{1/2}. \tag{19}$$

This equation was used for plotting Fig. 16, where δ_1 is replaced by its equivalent in terms of σ, Q_0 and T.

The Derived g Parameters. For some purposes it is convenient to eliminate δ_0 from the expressions for T and σ. This may be done by defining

$$g_1 = \frac{\delta_1}{\delta_0} \tag{20}$$

and

$$g_2 = \frac{\delta_2}{\delta_0}. \tag{21}$$

Introducing these new parameters the equations become

$$\frac{Q_0}{Q_L} = 1 + g_1 + g_2 \tag{22}$$

$$T = \frac{4g_1 g_2}{(1 + g_1 + g_2)^2} \tag{23}$$

$$\sigma = \frac{g_1}{1 + g_2} \tag{24}$$

$$P_r = P_e g_2 \tag{25}$$

$$P_g = (PP_e g_1)^{1/2} \tag{26}$$

The g parameters are particularly useful in defining the behavior of a tube and cavity combination when δ_0 is a fixed quantity while the effects of changes of δ_0 are more clearly shown when the δ parameters are used. The g parameters may be determined experimentally, using equations (21) and (22) without knowing the value of δ_0, that is of Q_0. On the other hand the g's are altered if a tube is replaced by one giving a different Q value while the δ's are intrinsic properties of the coupling mechanisms and remain fixed as long as the cavity and the tube tune at the same frequency and have the same effective reactance.

Tabulation of Related Equations. In the interest of completeness a number of the more important combinations of the basic equations are listed in Table 1. Some of these are of interest for measurement purposes while others apply particularly to actual system conditions.

Off-Resonance Analysis. The analysis can be extended to predict the transmission when the cavity is detuned from resonance by introducing the necessary susceptance term in equation (11) above and solving for T. This gives for the absolute value (neglecting phase)

$$T = \frac{4\delta_1 \delta_2}{[\delta_0 + \delta_1 + \delta_2]^2 + \left[\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right]^2} \tag{27}$$

TABLE 1.—*Relations between Cavity Parameters*

	Quantity	Symbol	General Expressions: In terms of δ's & Q_0	General Expressions: In terms of σ, T, & Q_0	Special Cases: TR case $\sigma = 1$	Special Cases: ATR case $T = 0$
Low level parameters	Input standing wave ratio	σ	$\dfrac{\delta_1}{\delta_0 + \delta_2}$	σ	1	$\dfrac{Q_0}{Q_L} - 1$
	Low Level Transmission	T	$\dfrac{4\delta_1 \delta_2}{(\delta_0 + \delta_1 + \delta_2)^2}$	$\dfrac{4\sigma}{(1+\sigma)^2}\left[1 - (1+\sigma)\dfrac{Q_L}{Q_0}\right]$	$1 - 2\dfrac{Q_L}{Q_0}$	0
	Q ratio	$\dfrac{Q_0}{Q_L}$	$(\delta_0 + \delta_1 + \delta_2)Q_0$	$\dfrac{4\sigma(1+\sigma)}{4\sigma - (1+\sigma)^2 T}$	$\dfrac{2}{1-T}$	$1 + \sigma$
	Input δ	δ_1	δ_1	$\dfrac{4\sigma^2}{[4\sigma - (1+\sigma)^2 T]Q_0}$	$\dfrac{1}{(1-T)Q_0}$	$\dfrac{\sigma}{Q_0}$
	Output δ	δ_2	δ_2	$\dfrac{(1+\sigma)^2 T}{[4\sigma - (1+\sigma)^2 T]Q_0}$	$\dfrac{T}{(1-T)Q_0}$	0
High level parameters	Leakage power	P_r	$P_0 \delta_2$	$\dfrac{P_0(1+\sigma)^2 T}{Q_0[4\sigma - (1+\sigma)^2 T]}$	$\dfrac{P_0 T}{Q_0(1-T)}$	0
	Gas discharge power	P_g	$[PP_0\delta_1]^{\frac{1}{2}} - P_0(\delta_0 + \delta_2)$	$\dfrac{2\sigma}{1+\sigma}\left[\dfrac{PP_r}{T}\right]^{\frac{1}{2}} - \dfrac{4P_r\sigma}{T(1+\sigma)^2}$	$\left[\dfrac{PP_0}{(1-T)Q_0}\right]^{\frac{1}{2}} - \dfrac{P_0}{(1-T)Q_0}$	$\left[\dfrac{PP_0\sigma}{Q_0}\right]^{\frac{1}{2}} - \dfrac{P_0}{Q_0}$

where ω_0 and ω are respectively the resonant angular frequency and the operating angular frequency.

This may be rewritten as

$$T = \frac{T_0}{1 + Q_L^2\left[\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right]^2} \tag{28}$$

where T_0 is the in tune transmission and Q_L is the loaded Q, if one assumes that the δ's and Q_L remain unchanged for small departures from the resonant wavelength.

The input impedance of the cavity in terms of the input line impedance is then

$$Z = \frac{\delta_1}{j\left[\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right] + \delta_0 + \delta_2}. \tag{29}$$

The effect of other resonant modes which have been neglected in this analysis may be included by the addition of a term (σ_1) in equation (29) giving

$$Z = \sigma_1 + \frac{\delta_1}{j\left[\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right] + \delta_0 + \delta_2}. \tag{30}$$

Equation (30) and equation (2) are identical except for terminology.

APPENDIX B

Experimental Determination of "g" Parameters of Windows

The derived g-parameters which express the electrical size of a window between a resonant cavity and a surge impedance line were defined in Equations 20 and 21 of Appendix A. Numerical values of these parameters may be of some interest, together with their relation to physical dimensions of the windows. The 721A test cavity was used for an experimental determination of the relation between window width and "g." This cavity is $2\frac{1}{16}$ inches inside diameter, and is coupled by means of windows to two $\frac{9}{16}$ diameter coaxial lines. The width of the windows may be adjusted by rotating the coaxial lines so as partially to close the openings. The insertion loss through the cavity was measured at 3100 mc. by means of a superheterodyne receiver which included a calibrated attenuator in its intermediate frequency section. The windows were carefully maintained geometrically equal. In this case,

$$g = \frac{T^{1/2}}{2(1 - T^{1/2})}$$

which follows immediately from equation 23 of Appendix A on the assumption that $g_1 = g_2$. Fig. 38 shows the results; g proves to be proportional to the fifth power of the window width, over a very large range of values of g. A knowledge of this relationship permits one, with the aid of equations 23, 24, 25, 26, and 27 of Appendix A to calculate the window size

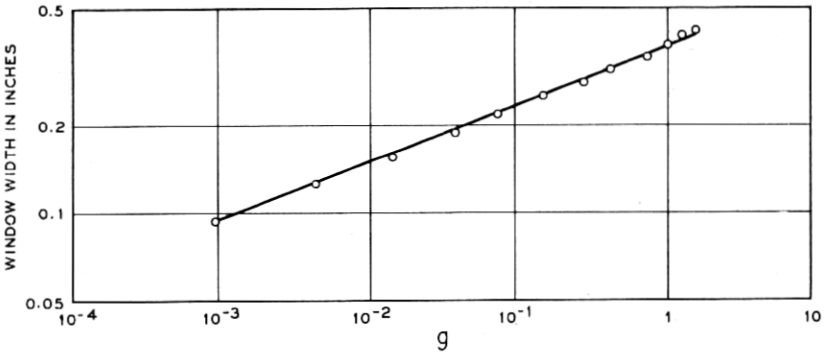

Fig. 38—The relationship between window conductance (g) and window width for the 721A test cavity

necessary to give any desired conditions of match, insertion loss, and leakage power.

APPENDIX C

The ATR Box

The value of the input impedance of the ATR is given by equation 29 with δ_2 equal to zero so that

$$Z = \frac{\delta_1}{j\left[\dfrac{\omega}{\omega_0} - \dfrac{\omega_0}{\omega}\right] + \delta_0} \tag{31}$$

which reduces to

$$Z = \frac{\delta_1}{\delta_0} \tag{32}$$

for the in tune case. This impedance is in series with the magnetron branch and hence restricts the possible range in values for the impedance at plane AB. Defining as F the fraction of the available power which is not absorbed by the ATR, then from Fig. 39 with the admittance of the receiver branch assumed to be $\delta_1 B$,

$$F = \frac{4}{(2 + G)^2 + B^2}, \tag{33}$$

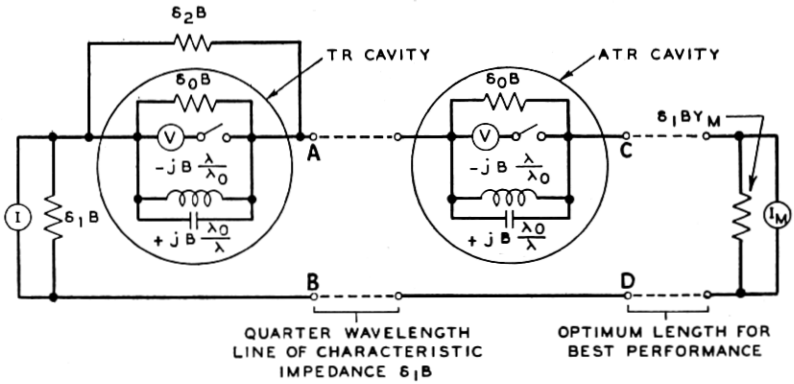

Fig. 39—Equivalent circuit of a system including an ATR

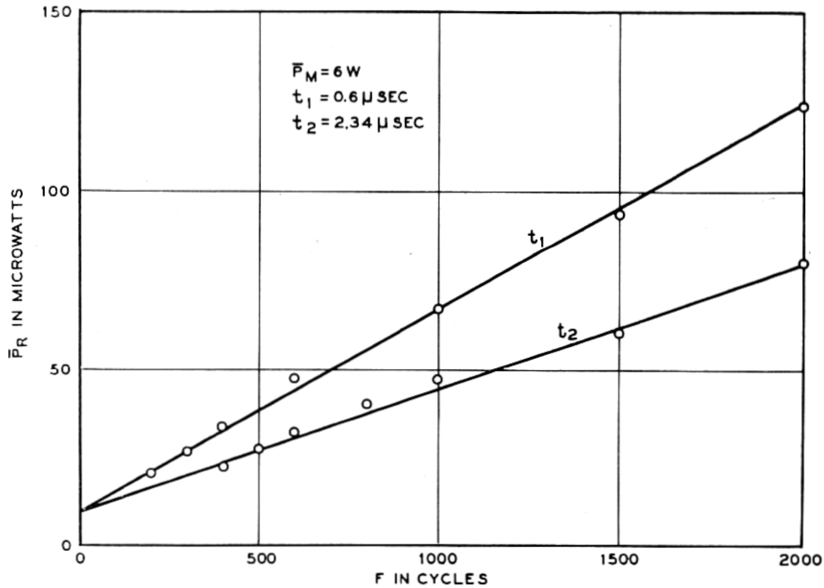

Fig. 40—Average leakage power as a function of repetition rate for two different values of pulse duration for 724B tube

where

$$G - jB = \frac{1}{(Z_m + Z)} \tag{34}$$

and Z_m is the impedance of the transmitter referred to the cavity and measured at the plane CD. The worst condition will occur when $Z_m = 0$. Under these conditions but assuming that the ATR is in tune

$$F = \frac{4}{(2 + G)^2}, \tag{35}$$

But now G is the reciprocal of the Z of equation (32) so that

$$F = \frac{4\delta_1^2}{[2\delta_1 + \delta_0]^2}. \tag{36}$$

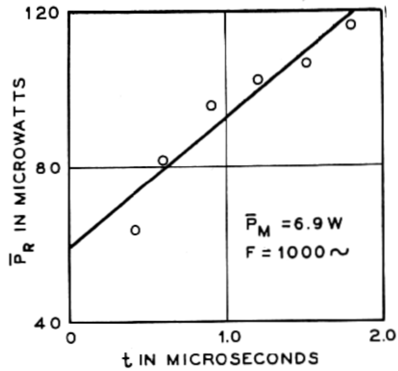

Fig. 41—Average leakage power as a function of pulse duration for the 724B tube

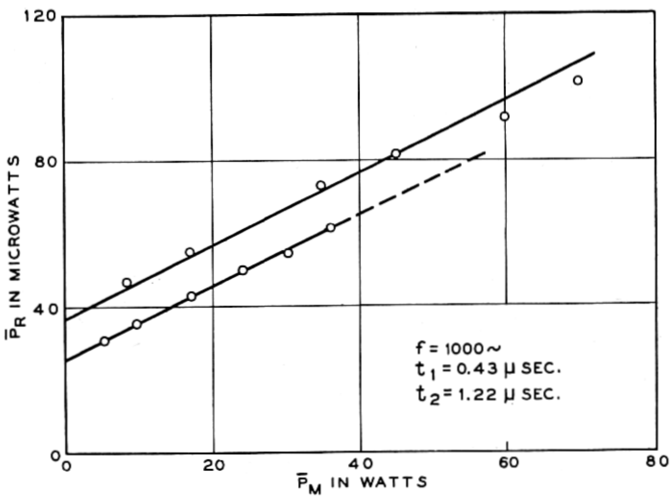

Fig. 42—Average leakage power as a function of average magnetron power for two different values of pulse duration for the 724B tube

This equation is analogous to equation (12) of Appendix A. At high levels the gas discharge power will be given by equation (19) as for the TR

$$P_g = [P\, P_0\, \delta_1]^{1/2} \tag{19}$$

If the ATR and the TR are designed to have the same value of P_g then the values of δ_1 and δ_0 must be the same so that a relationship will exist between F and T given by

$$F = \frac{4}{(3 - T)^2} \tag{37}$$

The input impedance Z to an ATR adjusted to the same gas discharge power of a TR with a transmission of T is given by

$$Z = \frac{1}{1 - T}. \tag{38}$$

APPENDIX D

The Analysis of Leakage Power Data

The section on receiver protection described the three components of leakage power which were referred to as spike, flat, and direct coupling. One may write down at once the following simple expression for leakage power:

$$\overline{P}_R = E_s f + P_F f t + \overline{P}_M T_D \tag{39}$$

where $\overline{P}_R$ is average leakage power
E_s is energy in a single spike
f is pulse repetition frequency
P_F is flat power
t is pulse duration
$\overline{P}_M$ is average magnetron power (averaged over the recurrence period) and
T_D is direct coupling insertion loss.

Experimental curves verifying the linear relationships indicated by this simple equation are shown in Figs. 40, 41, and 42. It is a straightforward operation to deduce numerical values for the three TR box leakage parameters from the slopes and intercepts of these curves.

Equation (39) was written on the assumption that gas-limited flat power and direct coupling power add linearly. If instead we assume that a phase angle θ exists between the two currents, we find:

$$\overline{P}_R = E_s f + P_F f t + T_D \overline{P}_M + 2\sqrt{\overline{P}_F f t T_D \overline{P}_M} \cos \theta \tag{40}$$

This of course is identical with equation (39) except for the cos θ term. If cos θ is not zero, we no longer expect a linear variation of $\overline{P}_R$ with f, t, or $\overline{P}_M$; the experimental curves demonstrate quite clearly that cos θ must vanish, hence θ must equal 90°.