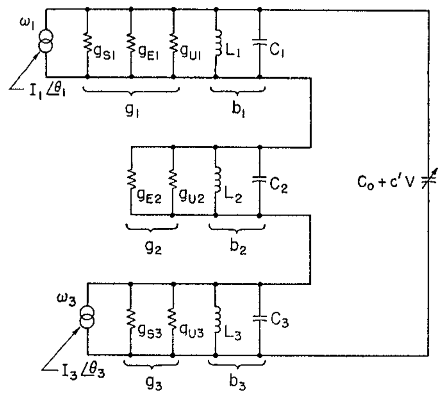

Fig. 2.

From the treatment of Heffner,[3] one can show that

$$V_{21} = -\frac{j\omega_1 c' V_3}{g_1 g_2 (1-\alpha)} I_2^*.$$

The total signal power delivered to g_{E1}, the external load in the ω_1 circuit, will be

$$S_0 = (V_{11} + V_{21})(V_{11}^* + V_{21}^*)G_{E1}.$$

Assuming $I_2 = Ie^{j(\omega_2 t + \phi)}$, and $I_1 = Ie^{j\omega_1 t}$, and assuming the phase of the pump to be zero, this can be rewritten as

$$S_0 = G_{11}S_1 + G_{21}S_2 + \frac{2g_{E1}\omega_1 c' V_3 I^2}{g_1^2 g_2 (1-\alpha)^2} \cos\left(\phi + \frac{\pi}{2}\right),$$

where

$$S_1 = \frac{|I_1|^2}{4g_{S1}} = \text{signal power incident at frequency } \omega_1$$

$$S_2 = \frac{|I_2|^2}{4g_{S2}} = \text{signal power incident at frequency } \omega_2$$

$$G_{11} = \frac{4g_{E1}g_{S1}}{g_1^2(1-\alpha)^2}$$

$$G_{21} = \frac{\omega_1}{\omega_2} \frac{4g_{S2}g_{E1}}{g_1 g_2} \frac{\alpha}{(1-\alpha)^2}.$$

Since for small signals

$$\alpha \doteq \frac{\omega_1 \omega_2 (c')^2 V_3^2}{g_1 g_2},$$

we obtain

$$S_0 = G_{11}S_1 + G_{21}S_2 + 2\sqrt{G_{11}S_1G_{21}S_2} \cos\left(\phi + \frac{\pi}{2}\right).$$

The incident signal power is given by

$$S_i = S_1 + S_2.$$

But, for equal input couplings $S_1 = S_2$, so that the gain for coherent inputs at ω_1 and ω_2 with output at ω_1 is

$$G_{DSB} = \frac{S_0}{S_i} = \frac{1}{2}\left[G_{11} + G_{21} + 2\sqrt{G_{11}G_{21}} \cos\left(\phi + \frac{\pi}{2}\right)\right].$$

A similar analysis could be carried out for a nonlinear resistance mixer; it would then be found that the factor $\cos(\phi + \pi/2)$ in the preceding should be replaced by the factor $\cos(\phi + \pi)$.

Duplexing Systems at Microwave Frequencies*

A. F. HARVEY†

Summary—**The paper reviews the various methods of duplexing at microwave frequencies. General principles, including the use of passive and solid-state devices, are first discussed. The characteristics of gaseous-discharge duplexing tubes of both self- and externally-excited types are examined and data for typical examples given. The various arrangements of discharge tube duplexers and methods of measuring their performance are described. The survey concludes with a bibliography.**

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List of Principal Symbols

B_0 = applied magnetic flux density, weber/m^2.
c = speed of light in vacuo $= (\epsilon_0\mu_0)^{-1/2} = 3\times10^8$ m/sec.
D_a = ambipolar diffusion coefficient, m^2/sec.
e = charge on electron $= 1.60207\times10^{-19}$ coulomb.
h_a = attachment probability.
m = mass of electron $= 9.1085\times10^{-31}$ kg.
N = density of electrons, m^{-3}.
N_L = Loschmidt's constant $= 2.687\times10^{25}$ m^{-3} atm^{-1}.

p = pressure, mm Hg.
r = radius, meter.
t = time, second.
v_d = drift velocity, m/sec.
Z = impedance, ohms.
α = attenuation coefficient, nepers/m.
α_r = recombination coefficient, m^3/ion-sec.
β = phase-change coefficient, rad/m.
γ = propagation coefficient $= \alpha + j\beta$.
ϵ = dielectric constant.
ϵ_0 = electric space constant $= 8.85416 \times 10^{-12}$ farad/m.
η_a = attachment cross section, meter^2.
λ_0 = wavelength in free space, meter.
λ_g = guide wavelength, meter.
μ_0 = magnetic space constant $= 4\pi \times 10^{-7}$ henry/m.
v_a = attachment frequency, c/s.
v_c = electron-ion collision frequency, c/s.
σ = conductivity, mhos/m.
τ = time constant, second.
ω = angular frequency, rad/sec.
ω_C = cyclotron angular frequency, rad/sec.
∇ = Laplacian operator.

Recovery times quoted are for 3 db loss; values for 6 db are somewhat shorter.

I. Principles of Duplexing

A. Passive Circuit

Duplex operation of a microwave system permits the passage of signals, having a common frequency, along a transmission line; it should not be confused with diplexing, which depends upon the difference in frequency of the two signals. The particular example considered is the use of a common circuit, such as an antenna, by both a transmitter and a receiver. Desirable qualities of a duplexer include low attenuation and high discrimination over a wide frequency band, freedom from manual adjustments, reliability and long life.

With CW operation, the duplexer is required to keep the transmitter disconnected from the receiver at all times and yet allow maximum coupling between the transmitter and antenna and between the antenna and receiver. With high-power pulsed operation the discrimination required to give adequate protection [111] to the receiver is about 60–80 db. Further practical requirements are that protection should be afforded from occasional random pulses received from nearby equipments and that the duplexer should recover to the receiving state in a time of 1–50 μsec according to the application.

If the first stage of the receiver is a semiconductor point-contact diode then, to avoid burn out, the leakage power of the duplexer must be maintained below about 50 mw. The thermal time constant of the diode contact is around 10^{-7} seconds and any transient effect in the duplexer must have an energy not exceeding 0.01 erg for temporary and 0.1 erg for permanent deterioration. Some recent low-noise RF amplifiers are even more sensitive [16] to leakage and, moreover, much of their advantage is lost if the duplexer itself adds appreciable noise.

A passive duplexer can, for example, be constructed from lossless four-terminal-pair junctions [114], [115] and Fig. 1 shows a hybrid T so employed. Energy from the transmitter divides between the matched termination on one arm and the antenna on the other while energy received by the antenna divides equally between the transmitter and receiver. The resulting 6-db total loss may be shown to be the minimum for such a simple circuit. The discrimination between transmitter and receiver is reduced by imperfections of the junction or mismatches in the termination and antenna. Such residual leakage can be balanced out, over a small frequency band, to a level of less than −100 db.

Passive duplexers based on circularly-polarized waves have negligible attenuation and, in addition, provide isolation of the transmitter from impedance changes of the antenna. The transmitted wave possesses one hand of polarization and the received wave, due to reflection, possesses the other hand: discrimination can thus be achieved by a suitable circuit. The arrangement shown in Fig. 2(a) contains two-orthogonal transducers from TE_{01} rectangular to TE_{11} circular mode. The circular waveguide contains a quarter-wave plate and Fig. 2(b) gives the phases of the electric field components of the transmitted energy. Fig. 2(c) shows that a reflected

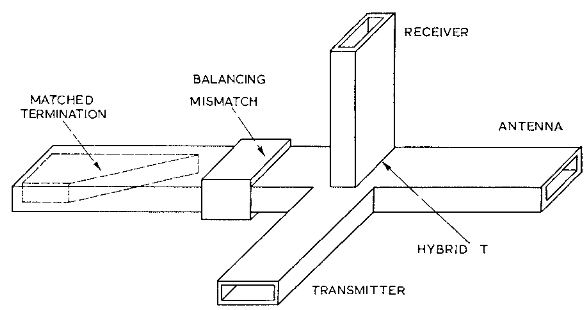


Fig. 1—Simple passive duplexer. This device gives a 3-db loss on both transmission and reception.

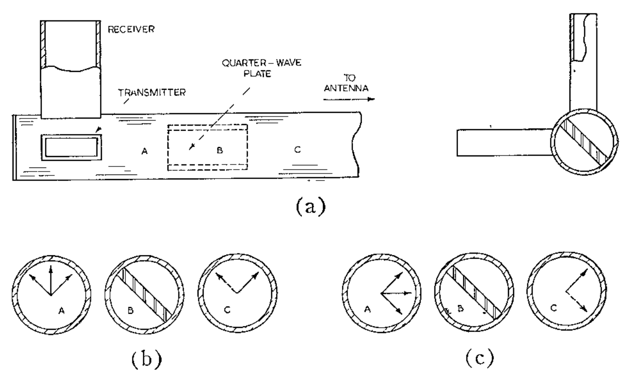


Fig. 2—Passive duplexer using circular polarization. (a) Assembly. (b) Electric vectors during transmission. (c) Electric vectors during reception.

signal is converted, after passage through the plate, to a plane wave, polarized so that it enters the receiver arm.

The conversion from plane- to circular-polarization may be carried out at the antenna [69] by the use of, for example, a polarized mirror [100]. Alternatively, the transmitter and receiver can be connected to opposite rectangular arms of a turnstile six-terminal-pair junction, the remaining pair of arms being terminated in short circuits so that they differ in length by $\frac{1}{4}\lambda_g$. At a frequency of 9 gc (gigacycles), such a duplexer has given [85] an isolation better than 40 db over a band of 100 mc. Another adjustment of the turnstile junction enables duplexing to be achieved with crossed linear polarizations of the transmitted and received signals.

The prefix for mega ($=10^6$) has been changed from m to M. This is to avoid serious confusion later on between milliwatt (mw) and megawatt (now Mw).

The restriction on the polarization is removed by the use of ferrite circulators. In the four port example shown in the inset of Fig. 3, energy from the transmitter T proceeds to the antenna A while energy from the antenna proceeds to the receiver R. A dummy load L is connected to the fourth port to absorb any leakage. The discrimination of an experimental Faraday-rotation duplexer [53] for 35 gc is shown by Fig. 3 to be better than 30 db over a 0.7 per cent band around the balanced frequency. Circulators for 9 gc, in which the ferrite is placed in the short-circuited circular guide of a turnstile junction, have shown [20] a similar discrimination with an insertion loss of 0.4 db, a VSWR of 1.25 and an overall length of 1.5 inches. Such duplexers are very satisfactory for CW systems of about 50-watts mean power.

With the duplexer of Fig. 3 and a transmitter giving 19 kw in 0.2-μsec pulses the finite frequency spectrum led to a leakage of 400 mw: such a high value could be reduced by broadband design. Duplexers for high powers must ensure that the ferrite, which is exposed to the full transmitted power, can withstand high voltages and dissipate, with the aid of air- or water-cooling, the heat generated consequent upon its attenuation.

B. Solid-State Switches

The high degree of protection required by pulsed operation may be obtained by time-division duplexing in which an attenuator or switch is energized in synchronism with the transmitter pulse. Low noise and high operating speed can be achieved by placing suitable solid-state devices in the receiver arm of a passive duplexer. To avoid a leakage transient due to the finite response time it is desirable to pretrigger the switch so that it has full insertion loss before the initiation of the transmitter pulse. Alternatively the transient may be removed by power limiters based, for example, on travelling-wave tubes [29] or ferrites [110].

The point-contact semiconductor diode is suitable [4], [33–35], [87] as a switch since the impedance, including that of its mount, may be varied from a low inductive to a high capacitive value with a bias change of, say, -0.6 to $+0.6$ volt. Improved performance is obtained [18] by connecting the diode in parallel with the shunt capacitance of a high-pass series m-derived filter. Typical parameters of such a switch are: frequency, 500–1000 mc; bandwidth, 20 mc; discrimination, >55 db; insertion loss, 2 db and operating time, $<\frac{1}{2}$ μsec.

In the duplexer [38] shown in Fig. 4, one of two ferrite circulators is reversed during the reception period: this switched circulator handles only low powers. Ferrite switches [124] operate by change of the magnetic field and some difficulty arises in making the rapidly varying field [21] penetrate the surrounding metal wall of the waveguide. At the lower frequencies sufficient room exists to place [113] the magnetizing coil around the actual ferrite. In Faraday-rotation devices a helical slot may be cut to follow the plane of polarization of the transmitted wave: this design has the disadvantage that in the closed, or reversed, position the cross-polarized components reduce the discrimination unless RF chokes are employed [93].

More usual practice takes advantage of the different penetration depths obtaining at the modulation and microwave frequencies. Films, 0.0001-inch thick, can be made on plastics such as Perspex and include electroplated silver [53] and vacuum-evaporated aluminum [11]. The switching power can be reduced by the use of toroidal-shaped [121] ferrites. Successful ferrite switches have been constructed for 9 gc [5], 55 gc [31] and 70 gc [122]. A typical example for the first frequency would have [72], [73] maximum and minimum attenuation better than 35 db and 1 db, with a switching time of $\frac{1}{4}$ μsec.

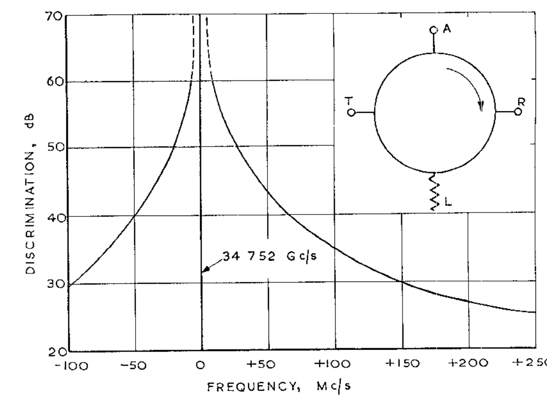


Fig. 3—Discrimination of a ferrite duplexer. The inset shows the arrangement of ports. Input power 40 kw. Ferrite-type B3.

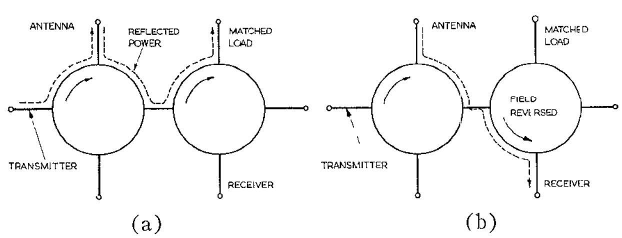


Fig. 4—Duplexer using a switched ferrite circulator. The arrangement uses one switched and one unswitched circulator. (a) Transmission. (b) Reception.

Balanced duplexers incorporate hybrid junctions which provide additional discrimination, thus giving improved protection and greater bandwidth. The power division and standing waves inherent in this duplexer result in the transmission line being exposed to twice the transmitter power and the switching devices to one half. The duplexer of Fig. 5 employs ferrite stubs whose electrical length is normally $\frac{1}{2}\lambda_g$ so that a signal from the antenna proceeds without reflection to the receiver. During the transmission period the stubs are switched to a length of $\frac{3}{4}\lambda_g$. Complete reflection occurs and the $\frac{1}{2}\lambda_g$ relative spacing of the stubs ensures that the transmitter power proceeds to the antenna.

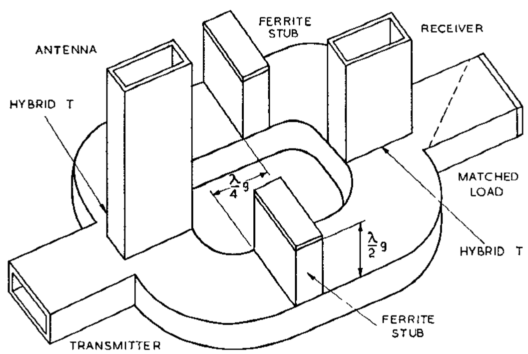


Fig. 5—Balanced duplexer with ferrite switches. The electrical length of the stubs is switched from $\frac{1}{2}\lambda_g$ to $\frac{3}{4}\lambda_g$ during transmission.

There is insufficient experience on the use of solid-state devices in duplexers for pulsed operation to form a reliable estimate of their merits. They appear to be suitable for applications where low noise, very rapid response, light weight and long life are required. With their present state of development, however, the usual duplexer for high-power pulsed operation incorporates, often exclusively, devices based on the phenomena of gaseous electronics.

II. Characteristics of Duplexing Tubes

A. *Circuit Arrangement*

The electrical discharges which provide the basis of duplexing tubes are due to ionization processes taking place at low pressure, at ordinary temperature and usually in rare gases. Their characteristics may be examined under the states of breakdown, the discharge proper and the afterglow. Breakdown of a gas takes place under the influence of an applied field which is sufficiently large to cause cumulative ionization [19], [74], [119]. The typical dc gaseous discharge consists of four main regions: cathode dark space, negative glow, Faraday dark space and positive column. The negative glow is associated [12] with much of the potential drop, high electron density and low electron collision frequency. The positive column is associated [43, 123] with low electron density and high collision frequency. In an HF discharge [2] the electrodes play only a minor role and the discharge presents a smooth appearance. The electron and current densities can reach very high levels and may exceed 10^{15} electrons/cm^3 and 15 a/cm^2 respectively.

The gaseous discharge in duplexing tubes may be externally excited by dc or HF means or internally excited by electric fields at the operating microwave frequency [15], [59]–[61]. The properties resulting from the interaction between the electrons and the microwave field are well known [41]: they endow the discharge with a conductivity given by [62], [79], [82]

$$\sigma = \frac{Ne^2\nu_c}{m(\omega^2 + \nu_c^2)} \qquad (1)$$

and a relative permittivity given by

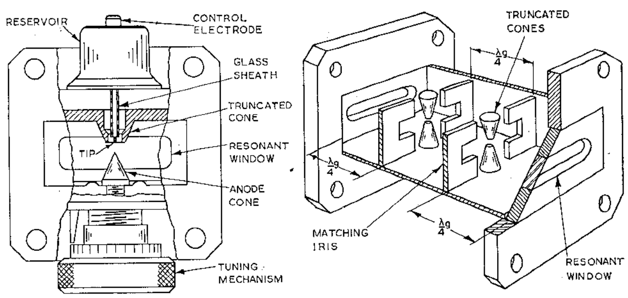
$$\epsilon = 1 - \frac{Ne^2}{m(\omega^2 + \nu_c^2)\epsilon_0}. \qquad (2)$$

Differentiation of (1) shows that the conductivity is a maximum when $\omega = v_c$, the phase angle being then 45°. The transfer of energy from the RF field to the discharge is then a maximum and corresponds to minimum values of breakdown power, arc loss and leakage through the discharge. Fluctuations in the velocity and collision-rate of the electrons result in noise which may be expressed as an equivalent temperature [17], [95].

Duplexing tubes require a microwave structure, coupled to the transmission line and designed so that optimum interaction takes place between the gaseous discharge and the RF electric field. In tubes with dc or IF excitation, efficient interaction is achieved by working in the negative glow. The arc-leakage power of RF excited tubes tends to be constant over a wide range of incident powers but, as with other parameters, it is dependent upon the nature of the discharge. In general, high electron density and other desirable properties, coupled with long working life, are provided by a basic filling of one of the heavier rare gases to a pressure of 2–20-mm Hg.

A distributed structure contains the discharge in a portion of transmission line, sealed at either end by low Q-factor windows. The design may be in the form of a coaxial line [37], [90], [91], [106], interaction taking place in a gaseous discharge established between the inner and outer conductors. At frequencies above 10 gc, the difficulty in constructing close-tolerance transducers is avoided by employing rectangular or circular [112] waveguides. The interaction distance must usually be several wavelengths and bulkiness at frequencies lower than 10 gc can be avoided by the use of slow-wave structures. One example [103] for 3 gc employs an interdigital line, the discharge taking place at the tips of the fingers. In general, distributed-structure tubes must employ either heated cathodes or very high pulsed voltages to produce a sufficiently intense electron density but, on the other hand, large microwave powers can be controlled.

Resonant structures provide a much increased interaction between the gaseous discharge and the RF field. Practical shapes tend to give nonhomogeneous discharges with rapid variation [96] of electron density so that theory cannot be accurately applied. A typical high Q-factor circuit consists, as shown in Fig. 6(a), of a single cavity, usually resonant in a modified TM_{010} mode. The interaction region is between the tips of the two cones and Q-factors are around 300. Such structures tend to give a leakage due to transmission in modes not possessing voltage maxima at the gap, but since this is about -70 db it shows up only at very high incident powers. Tuning of the cavity is achieved by adjustment of the cone gap.



(a) (b)

Fig. 6—Resonant structures using gas discharges. (a) High Q-factor tunable. (b) Low Q-factor broadband. The frequency is about 9.5 gc.

Fig. 6(b) shows an alternative low Q-factor structure which contains two cone assemblies, each being tuned by an inductive iris. The Q-factor of each element, loaded by the terminating impedance of the waveguide, is determined by the gap between the cones and the dimensions of the iris and is about 4. At either end of the structure, resonant slots, with a Q-factor half that of the cones, are included: these serve as terminations of a binomial filter and increase the bandwidth. The resonant elements are spaced $\frac{1}{4}\lambda_g$ apart, at which distance some interference due to evanescent modes is present. In such tubes, breakdown first occurs at the cones and this leads rapidly, with high incident powers, to breakdown of the gas in the vicinity of the input window. The combined attenuation of the window and gap discharges is of the order 80–100 db. Broad-band tubes are tending to supplant the high Q-factor types since, in particular, they are capable of operation at very high powers and avoid signal loss associated with incorrect tuning.

At the lower frequencies increased flexibility of design and operation is achieved by separation of the discharge tube and the circuit elements. In the tuneable single-cavity arrangement the discharge takes place in a cell in which the electrodes are brought out via disk seals. An early example [22] was used over the range 2.30–3.75 gc with different external structures. Fine tuning is achieved with moveable plungers and coupling to an external transmission line may be by loop or aperture. More recently, this plug-in feature has been applied [128] to broad-band structures so that, at each stage of the device, an independent choice can be made of the discharge characteristics.

B. Priming the Discharge

The statistical time lag occurring between the application of the exciting field and the complete buildup of the gaseous discharge is undesirable in practical tubes. The delay may be shortened by maintaining a continuous weak ionization in the breakdown region which, in the case of prepulsed externally-excited tubes, can satisfactorily be provided from a radioactive substance. About 10 microcuries of tritium, which is a source of low-energy β-particles with a half-life of 12.26 years, forms a safe and convenient priming agent.

In self-excited tubes, the time lag allows microwave energy to pass unattenuated to give a transient of duration about 5×10^{-9} seconds. The energy in this spike is minimized by correct design of the resonant structure and choice of the nature and pressure of the gas filling but, in addition, a large initial ionization of the order 10^8 electrons/cm^3 is necessary. This ionization is generally provided by the glow discharge of a keep-alive or primer. The primer must have negative polarity and may be positioned either inside and coaxial with the suitably hollowed RF electrode or at the side of the interaction space. The electron density must not be so high as to cause undue interaction loss in the low-level condition. Moreover, the discharge allows electrons with random energies to enter the RF field, thus forming a noise source internal to the receiving system. In practice currents around 100 μa are employed.

Keep-alive circuits are prone to relaxation oscillations and, under certain conditions, wander of the discharge and glow-to-arc transitions. These effects result in periods of low electron density and hence large spike energy. The tendency towards oscillation is reduced by placing one megohm of the total current-limiting resistance close to the primer terminals. A further precaution involves operating the discharge in the abnormal region [109] by covering the electrode with a glass or ceramic sleeve down to the end, which is exposed by grinding off the insulation. Such an electrode of restricted area also has the advantage of reducing wander by localizing the discharge. An arc is characterized by low potential and high current and can thus be maintained only by the stored energy in the distributed capacitance of the electrode: it extinguishes after a time of the order 10^{-4} μsec. With an electrode capacitance of 10 $\mu\mu$f and running conditions of 500 volts, 100 μa the glow discharge will not restart for 50 μsec. These glow-to-arc transitions can be avoided by ensuring that the electrode surfaces do not become contaminated.

Sputtering of the keep-alive cathode causes slow clean-up of the gas filling with consequent deterioration in the tube properties. This erosion can be reduced [65], [126] by proper design of the electrode shape and choice of the metal constituting the keep-alive tip. Kovar is in general use but rhodium plating or the provision of stainless steel tips reduces the effect. Gas clean-up also occurs under the influence of the RF discharge, especially with high powers. The working life of the tube can be prolonged by, for example, the inclusion of a gas reservoir. This gas clean-up may be studied [26] in experimental tubes by including a radioactive tracer such as krypton-85 with the usual gas filling.

The interaction loss, noise and shortened life due to the keep-alive discharge may be improved by maintaining the current small and increasing it, by means of a prepulse, just prior to the start of the transmitter pulse. In this way substantial priming can be achieved so that the spike energy is diminished while the small residual discharge protects the receiver by ensuring that occasional random pulses, although large enough to cause temporary deterioration of the diode contact, are not large enough to lead to permanent damage.

C. Quenching the Discharge

In the afterglow of the discharge the electrons, whose average energy is initially high, tend to acquire, at a rate determined by the probability [45] of collision for momentum transfer, a Maxwellian distiribution of velocity with an equilibrium temperature that of the gas. The early afterglow, which is of promising interest in the study of duplexing tubes, is thus characterized by a time-varying electron energy distribution function. Moreover, the electron density is reduced by three physical mechanisms, which may operate singly or together, according to the relation

$$dN/dt = D_a\nabla^2 N - \alpha_r N^2 - h_a \nu_c N. \tag{3}$$

Diffusion [3] involves the motion of the electrons to the discharge boundary and thus depends upon the mobility of the particles and the path length traversed. The decay is exponential with a time constant which, for a spherical container of radius r , is given by

$$\tau = r^2/\pi^2 D_a. \tag{4}$$

A typical value of D_o, reduced to a gas pressure of 1 mm Hg, is 900 for argon [8]. Recombination between an electron and a positive ion involves [83] the release of energy in another form: one example of such a process is surface recombination which leads to translational kinetic energy of a third particle. The decay leads to a linear relation between $1/N$ and t and experiments [7], [66] on helium ions at 3 gc gave values of α_r about 1.7×10^{-8} cm³/ion-sec. Attachment involves [81] the formation of a negative ion [84] on capture of an electron by a neutral gas molecule. Experimentally, three attachment parameters can be measured: the attachment cross section η_a, the attachment probability h_a and the attachment coefficient η/p. The attachment frequency can be expressed in terms of one of these parameters by

$$\nu_a = h_a\nu_c = \frac{pN_L}{760}\eta_a v_d = p\left(\frac{\eta}{p}\right)v_d. \tag{5}$$

In strongly attaching gases, the frequency of attachment is of order 10^6 to 10^8 times the pressure [1], [36], [50], [52].

Recovery of a duplexing tube at the end of the discharge period requires the electron concentration in the interaction space to fall rapidly to a negligible value. The selection of working pressure below 100 mm Hg means that loss of electrons by diffusion to the walls of the container is slow [83]. If, however, short diffusion paths and adequate clearances are provided, surface recombination enables recovery times of 30–50 μsec to be achieved with pure rare gas fillings. Solid inserts packed around the electrodes or interaction space form a suitable recovery agent, provided the material is of low permittivity and dielectric loss and is capable of withstanding high temperature. Quartz crystals, hard-glass fragments and hollow ceramic beads are easy to handle and pack while permitting the insertion of pretuning plungers. Rapid deionization is also achieved with quartz wool which, consisting of very fine fibres, has a large surface per volume ratio. This material is only convenient for simple shapes since, unless it is carefully packed, voids occur which result in increased diffusion time.

Quenching of the discharge by electron attachment is an efficient process if enhanced by the introduction, in element or compound form, of hydrogen, oxygen, nitrogen and the halogens. Experiments [125] with powers up to 200 kw in the range 3–35 gc have shown, for example, that, with an initial filling of argon at 11 mm Hg, the inclusion of hydrogen at a partial pressure of 1 mm Hg reduced the recovery time from 200 μsec to 4 μsec. Although effective as a quenching agent, hydrogen cleans up rather quickly: this disadvantage may be overcome by arranging a continuous supply from, for example, a replenisher such as titanium hydride.

Recovery times as short as one microsecond are achieved by the addition of water vapor which has a very large recombination coefficient but also some attendant disadvantages. The products of dissociation migrate in any dc field present towards the metallic electrodes with which they react chemically [57]. The dominant surface reaction results [13] in the formation of nonconducting oxides. Such a film on the cathode greatly facilitates [68] glow-to-arc transitions which tend to occur [32] in bursts of 10–20 with intervals between bursts of 5–60 minutes. In certain circumstances the chemical reactions lead to the formation of a conducting oxide which continues until short circuiting of the primer gap occurs.

These chemical reactions lead, as may be expected, to a gradual clean-up of water vapor. Long life may be attained by giving the tube a higher partial pressure of water vapor than is initially necessary but this process leads to worsening of other parameters such as arc loss and leakage energy. Water vapor replenishers maintain a sensibly-constant partial pressure and thus increase the working life. A typical replenisher consists [27] of carefully cleaned pure iron filaments on which 0.2 per cent by weight of water vapor has been absorbed. When placed in the tube envelope the partial pressure remains near 0.2 mm Hg over the temperature range $-20°$C to $+120°$C for a long period.

III. Externally-Excited Tubes

A. Phase Change

If a gaseous discharge completely fills a waveguide the propagation coefficient

$$\gamma = \alpha + j\beta \tag{6}$$

has components given by [92]

$$2\alpha^2 = \left|\frac{\omega^2}{c^2}(\epsilon - 1) - \left(\frac{2\pi}{\lambda_g}\right)^2 + j\omega\mu_0\sigma\right| + \left\{\frac{\omega^2}{c^2}(\epsilon - 1) - \left(\frac{2\pi}{\lambda_g}\right)^2\right\} \tag{7}$$

$$2\beta^2 = \left|\frac{\omega^2}{c^2}(\epsilon - 1) - \left(\frac{2\pi}{\lambda_g}\right)^2 + j\omega\mu_0\sigma\right| - \left\{\frac{\omega^2}{c^2}(\epsilon - 1) - \left(\frac{2\pi}{\lambda_g}\right)^2\right\}. \tag{8}$$

If $\nu_c \ll \omega$ so that

$$\sigma = 0 : \epsilon - 1 = -\omega_p^2/\omega^2 \tag{9}$$

and, further, if

$$\omega_p^2/c^2 < (2\pi/\lambda_g)^2 \tag{10}$$

then

$$\alpha = 0 : \beta = \left[\left(\frac{2\pi}{\lambda_g}\right)^2 - \frac{\omega_p^2}{c^2}\right]^{1/2}. \tag{11}$$

The change in phase coefficient due to the discharge is given by

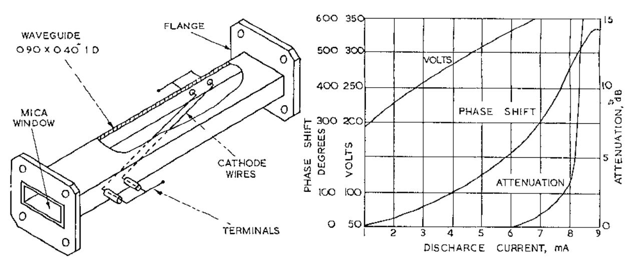
$$\Delta = \frac{2\pi}{\lambda_g} - \left[\left(\frac{2\pi}{\lambda_g}\right)^2 - \frac{\omega_p^2}{c^2}\right]^{1/2} \simeq (\omega_p^2/c^2)(\lambda_g/4\pi). \tag{12}$$

If $\Delta \ll (4\pi/\lambda_g)$ the ratio (Δ/λ_g) should be independent of frequency for constant electron density. If

$$\omega_p^2/c^2 \gg (2\pi/\lambda_g)^2 \tag{13}$$

then $\beta = 0$ and the waveguide is cut off.

These effects have been put to practical use [97] in the phase changer shown in Fig. 7(a). Interaction takes place between the RF field and the negative glow surrounding two filamentary cathodes: these are spaced $\frac{1}{4}\lambda_g$ apart axially to improve impedance matching. The device, when filled with krypton to a pressure of 1 mm Hg, gave, at 9.55 gc, the values of potential drop and phase change vs discharge current shown in Fig. 7(b). The phase change was independent of the microwave input level up to 100 mw. For the discharge length of 8.4 cm and a phase change of 180°, (12) requires an electron density about 0.45 the cutoff value. Such calculations are confirmed by the attenuation values shown in Fig. 7(b): these are quite low until the condition of waveguide cutoff is approached. This particular microwave arrangement is broad-band and achieved satisfactory performance over the range 8.9–10.2 gc: the life of the tube was 500 hours with a discharge current of 6 ma.



(a) (b)

Fig. 7—Gas-discharge phase-change tube. (a) General arrangement. (b) Properties at a frequency of 9.55 gc.

If a magnetic field is applied orthogonal to the RF electric field, the electrons acquire a helical motion with angular frequency

$$\omega_C = eB_0/m. \tag{14}$$

As the magnetic field is varied through this resonance value the breakdown voltage exhibits a sharp dip [14], [30], [70], [71], [89], [120]. As an example of use in duplexers, a switch tube based on this property has shown [116] a discrimination of 60 db with an arc loss of 0.4 db: the input power was 250 kw peak at 2.85 gc and the low-level insertion loss was 0.1 db. Another switch [117] is controlled by the turning on and off of a magnet field set at the resonance value. The breakdown power varies with gas pressure and, for example, at 2.84 gc with argon, the values with and without the applied field are respectively 0.01 kw and 1 kw for 0.3 mm. Hg and 0.03 kw and 60 kw for 0.03 mm Hg. The firing time is of the order of 0.01 μsec. The nonreciprocal properties [10], [23], [39], [42], [44], [48], [86] of magnetized discharges have, however, not found much application in passive duplexers since they appear [98] to be inferior to ferrites in respect to manufacturing cost, response time, power handling capacity, noise output and life.

B. Attenuators

Gaseous discharges can form the basis of variable attenuators but have the disadvantage of an associated noise power equivalent to a temperature of up to 10,000°K. In one distributed structure [56] for 9.5 gc, the central conductor of a coaxial line is a heated oxide cathode. The attenuation, although large, is rather dependent on cathode temperature: for this reason, cold cathode tubes are preferable. A mixture of hydrogen with the rare gas quickens the response and reduces the discharge current for a given attenuation. Another design for the same frequency consists [12] of an oxide cathode, 2 inches long and 0.02-inch diameter, mounted in a 0.9 inch by 0.4-inch waveguide. Krypton at 2-mm Hg is a suitable filling but the attenuation showed a sharp peak at a cathode temperature corresponding to a glow-to-arc transition. In a version for 50 gc, the discharge is made between the two broad walls of the waveguide, these being insulated from the rest of the structure: an attenuation of 30 db can be achieved with a discharge current of 10 ma.

An attenuator [99] employing a resonant structure has already been shown in Fig. 6(a): it has a Q-factor of 295 and is tuneable over the range 8.9–9.5 gc. The control electrode is glass sheathed so that only the 0.028-inch diameter magnesium tip is exposed. With a filling of krypton at 10 mm Hg and about 80 volts between the electrodes, the attenuations at currents of 100 μa and 200 μa were 15 db and 30 db respectively for an RF input of 50 μw. The attenuation is sensitive to the level of incident power, falling by a factor of 4 when the latter was increased from 3 mw to 100 mw. A two-gap broadband structure enables similar attenuations, with an input VSWR of 1.2, to be maintained over a frequency band of 650 mc.

C. Pulsed Switches

When the ionization in an attenuator is so intense that large reflection and loss occur, the device more closely resembles a switch: its use is normally confined to pulse operation with low duty cycles. In one novel switch [40] the positive column of a gaseous discharge formed part of the inner conductor of a coaxial line. In the absence of the discharge, the outer conductor is too small to sustain waveguide propagation and no power passes. This principle has been used [51] to provide a wide-band multiway switch of low insertion loss. The noise inherent in the required dense discharge was kept small by pulse operation, the gas and tube conditions being such that the recovery time is of the order of one millisecond; dc gaseous discharges have also formed the basis of waveguide [118] and cavity [80], [112] switches. For example, the negative-glow CV2379 tube yields [28] more than 20 db attenuation for 0.5-watt excitation power. The recovery time is 12–20 μsec and passive protection is afforded against unsynchronized pulses by virtue of its steady "simmering" electrode current of 50 μa. The life exceeds 1500 hours.

The disadvantages of internal electrodes may be avoided by employing HF excitation of the discharge. One such tube, shown in Fig. 8(a), consists [76] of a 5-mm internal-diameter precision quartz tube. An insert, ground to a dumbbell shape, is sealed off to contain air at atmospheric pressure and the intervening space filled by xenon at 60 mm Hg with a small amount of tritium. Ionization is achieved by applying a 30 mc voltage between the waveguide and an excitation electrode shown in Fig. 8(b). Two tubes may be fed from a high-output impedance push-pull oscillator whose output is about 200 watts peak. The complete switch, with each tube forming a low Q-factor resonant element, is shown in Fig. 8(c). It provides a pulsed attenuation of at least 50 db with a recovery time less than 30 μsec: the VSWR is better than 1.15 over a 10 per cent frequency band and the insertion loss is below 0.1 db. For operation at 10 gc the insert may be omitted while at 35 gc the tube forms a distributed structure along the waveguide, the excitation electrodes being placed at both ends.

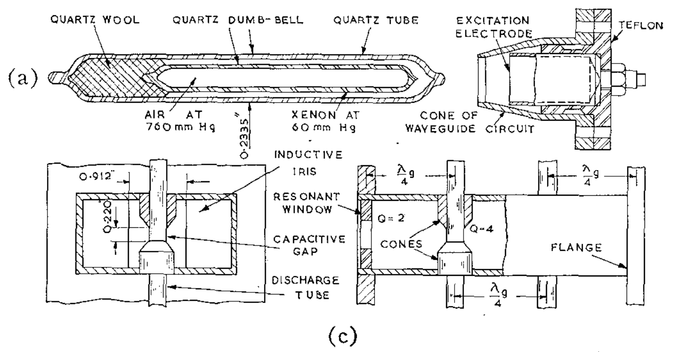


Fig. 8—Pulsed attenuator with HF excited discharge. (a) Discharge tube. (b) Excitation system. (c) Two views of the assembly for 3 gc.

IV. Self-Excited Tubes

A. Single-Function Switches

Self-excited tubes operate by differences in the levels of the incident microwave energy. They require high electric fields which are usually the result of short pulses with fast rise times and thus their design tends to be empirical. Such tubes generally employ resonant structures and may be broadly divided into single- and multiple-function classes.

A simple gas switch consists of a resonant iris in a waveguide. At low power levels, energy is allowed to pass with only small reflection but, at high power levels, the electron density in the discharge results in a large reflection of the incident power with corresponding large attenuation of the transmitted power. Such a discharge element is rarely used by itself but forms the basis of more complicated tubes. For example, in the ATR (or TB) tube shown in Fig. 9(a), the iris is terminated on one side by a short-circuited $\frac{1}{4}\lambda_g$ line. The device thus behaves, at the input window, as an open circuit in the low-level condition and a short circuit under power breakdown. The cavity dimensions and window susceptance are chosen to give the maximum impedance under low-level operation. Technical data on a few typical ATR tubes are given in Table I. The 1B36 and VX8104 tubes are of circular section and are provided with a machined bevel for contact with the main waveguide: these tubes also include a deformable diaphragm as the short circuit, thus permitting a small amount of preset tuning during manufacture. ATR tubes are usually filled with argon to about 10 mm Hg, no recovery or priming agents being found necessary.

The desirable characteristics of a pre-TR tube include large reflection and small arc loss when ionized, short recovery time, low insertion loss and ability to withstand high incident powers. Spike leakage is not important and thus keep-alives are not provided. An early example, the 1B38 for 2.8 gc, consisted [109] of two resonant discharge gaps spaced $\frac{1}{2}\lambda_g$ apart. Recent tubes [105] for 1.2 gc based on this design incorporate window of ceramic sheet or gas filled quartz containers.

Plug-in tubes have advantages of simplicity, bandwidth and long life. They are finding increasing applica-

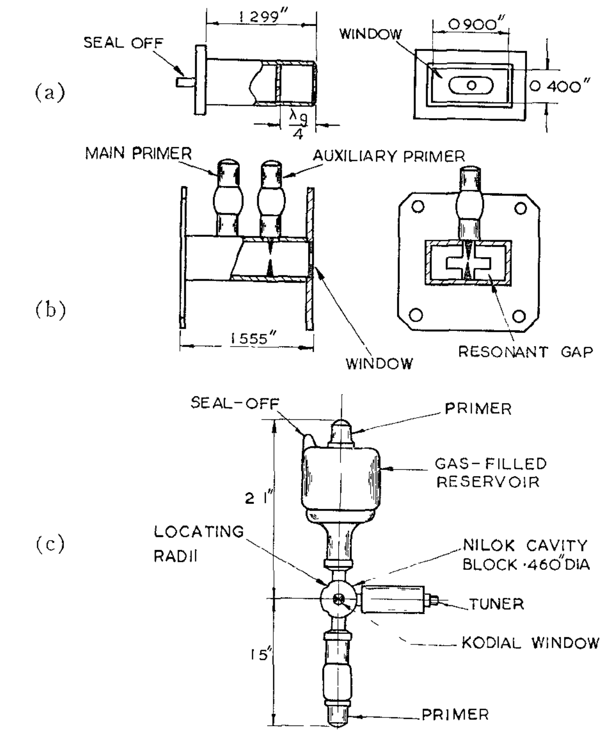


Fig. 9—Typical duplexing tubes. (a) CV2308 ATR tube for 9.5 gc. (b) CV2311 broadband TR tube for 9.5 gc. (c) CV2300 tunable TR tube for 35 gc.

TABLE I

TYPICAL ATR TUBES

Type number	1B44	CV2308	1B36	VX8104
Frequency, gc	2.6–3.1	9.0–9.6	23–25	34–36
Loaded window Q-factor	5.5	6.5	7.5	7.5
Breakdown power, kw	10	4	2	2
Arc loss, db	0.5	0.5	0.5	0.5
Maximum power, kw	—	100	—	50

TABLE II

TYPICAL PLUG-IN PRE-TR TUBES

Type number	CV2339	CV2430	CV2482	VX1023
Frequency, gc	2.5–4.0	2.5–4.0	8–12	34.36
Loaded gap Q-factor	1.2	1.0	matched	matched
Breakdown power, kw	4	1	15	8
Insertion loss, db	0.05	0.07	0.1	—
Arc loss, db	0.15	0.4	0.2	—
Recovery agent	steatite	water vapor	quartz wool	high pressure
Recovery time, μsec	30	5	5	4
Spike, ergs/pulse	1500	1000	150	2000
Flat, watts	100	100	180	—
Maximum power, mw	1.5	0.5	0.5	0.1

TABLE III

TYPICAL PASSIVE-PROTECTION TUBES

Type	VX3262	CV2359
Frequency, gc	2.5–3.0	8.9–9.6
Construction	plug-in	two-gap
Loaded gap Q-factor	5.3	4
Breakdown power, mw	70	100
Insertion loss, db	0.10	1.0
Recovery agent	water vapor	none
Recovery time, μsec	5	40
Interaction loss, db	0.05	0.1
Spike, ergs/pulse	0.15	0.1
Flat, mw	5	5
Maximum power, watts	3	10,000

tion and Table II gives technical data on typical examples. The CV2339 and CV2430 are designed so that the component parts can be assembled by high temperature brazing or cold welding, both methods allowing the tubes to be baked at 450°C during pumping to eliminate gaseous impurities. Choked mounts are required with plug-in tubes since the discharge tends to behave as the inner conductor of a coaxial line with consequent leakage of energy by radiation. The CV2482 is an argon filled quartz tube [94] ground to the correct size to match a 0.9-inch×0.4-inch waveguide over the band 8–12 gc: operation down to 3 gc is possible in an iris-matched mount.

Passive-protection tubes are associated with low leakage during the transmitting pulse and, in the receiving period, protection from unsynchronized pulses with low noise generation. The properties of typical tubes are given in Table III, the low leakage values being achieved by fitting keep-alives. In the VX3262 plug-in tube, the depreciation of receiver noise factor is maintained [28] less than 0.02 db by enclosing the keep-alive discharge as much as possible. A two-gap broadband design is used [75] for the CV2359. A weak keep-alive discharge is maintained from a side electrode and the filling is argon at 20 mm Hg. The recovery time varies from 12 μsec at 10 watts input to 100 μsec at 100 kw. At low levels the VSWR is better than 1.17 over a 5 per cent bandwidth while the life exceeds 10,000 h.

B. *Multiple-Function Switches*

1) *Tube Arrangement*: It is possible and has, in fact, long been common practice to combine, in a single unit, the functions performed by several tubes. Such a TR tube [107], [109] must possess, with high incident powers small arc loss, low spike and flat leakages and short recovery time coupled with, in the receiving condition, low insertion loss and protection against unsynchronized pulses.

At frequencies around 3–10 gc fixed tuned broadband structures are in general use. One example, the CV2311, is shown in Fig. 9(b): brief technical data for this and other TR tubes are given in Table IV. Breakdown occurs first at the cone containing the keep-alive electrode with incident powers less than one watt but until the input window breaks down, the equivalent short circuit is $\frac{1}{4}\lambda_g$ from the required position. Short recovery time in the TR tubes is ensured by partial filling with water vapor. Other broadband TR tubes include the CV2312 for 8.5–9.3 gc, the 1B63A for 8.5–9.5 gc and the three-gap 1B58A as listed.

TABLE IV
TYPICAL TR TUBES

Type	1B58A	CV2429	CV2311	1B26	CV2330
Frequency, gc	2.6–3.1	2.5–4.0	9.2–10.0	23.4–24.6	34–36
Construction	3-gap	plug-in	2-gap	tuneable	tuneable
Loaded gap Q-factor	5.5	4	—	220	150
Breakdown power, watts	10,000 (window)	0.15 (gap)	4000 (window)	—	—
Insertion loss, db	0.3	0.15	0.5	0.8	1.0
Recovery time, μsec	12	7	4	4	2
Keep-alive type	side-arm	co-axial	double	co-axial	double
Spike, ergs/pulse	0.2	0.2	0.2	0.03	0.03
Flat, mw	40	20	30	20	20
Maximum power, kw	—	0.3	250		

Tuneable-cavity tubes, such as the cell type 1B27 for 3 gc and integral-type CV221 and 1B24 for 9 gc, now find only limited application. This design does, however, lend itself to very wideband operation by means of windows of extremely low Q-factor. The troubles attendant upon a large glass area are overcome by using a window consisting of two or three parallel slots. A tube of the latter design, with a window Q-factor of 0.25, possesses [105] a total bandwidth of 40 per cent centered on 1.2 gc. The tuneable TR tube is, moreover, not so critical as regards tolerances on, for example, the position of the keep-alive electrode and hence continues to find use at the higher frequencies. Two examples have beeen given in Table IV, the CV2330 for 35 gc being illustrated in Fig. 9(c).

The CV2429, illustrated in Fig. 10, is a plug-in TR tube which allows operation, by choice of mount dimensions, at frequencies near 3 gc, in bandwidth steps of 6–10 per cent. In the CV2378 the keep-alive, and hence the water vapor replenisher, is omitted. The breakthrough is much higher, with a spike of 10 ergs/pulse and a flat of 20 watts, but the tube is suitable as the first of a pair in a TR combination. The recovery time,

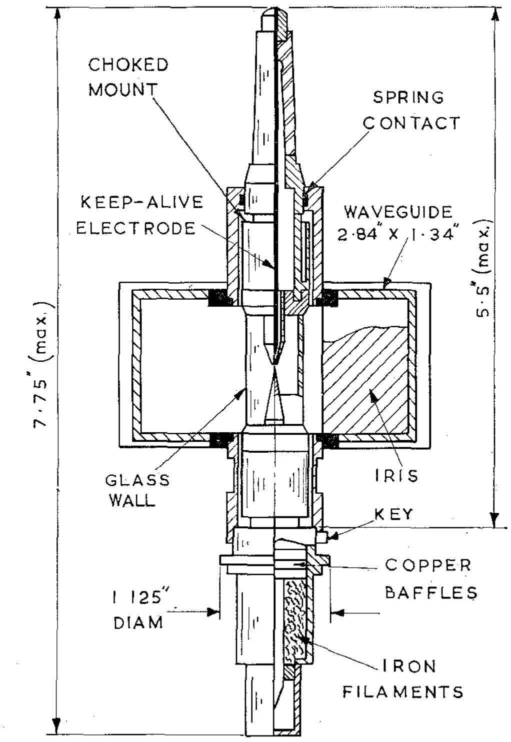
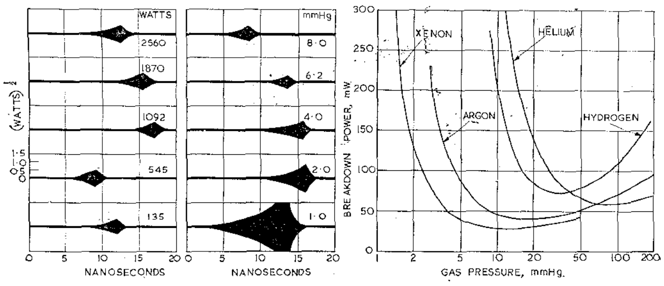


Fig. 10—Plug-in CV2429 TR tube. Frequency 3 gc. A keep-alive and water-vapor replenisher are provided.

being a function of the incident power, is often short enough for the second tube of the pair to fulfill a given requirement when a pure rare gas filling is used.

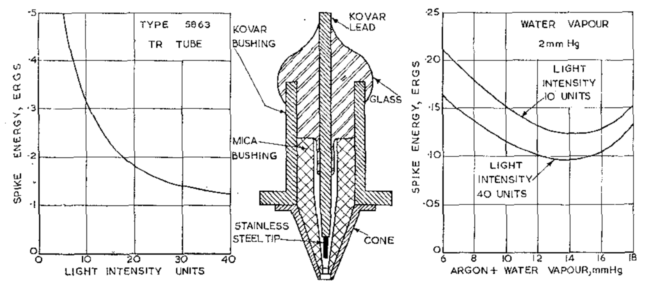
2) *Performance Control:* It will have been realized that the TR tube, containing both water vapor and a keep-alive, achieves its multiple functions only with a reduction in life. Investigations have thus been made with the view to improving this factor by an understanding of the discharge mechanisms involved. For example, direct observation of the spike by means of a high-speed oscilloscope reveals [24] the dependence of the operation of a tuneable 1B24 tube on various parameters. Fig. 11(a) shows that the spike amplitude is independent of incident power in the range 135 watts to 2.56 kw. The tube was filled with hydrogen and water vapor, both at 15 mm Hg. The spike is also constant as the keep-alive current is reduced from 100 μa to 9 μa but, at zero current, it rose to a maximum of 64 watts$^{1/2}$ before switching action was effected. The spike is sensitive to gas pressure and Fig. 11(b) shows that, with xenon, the minimum occurs at 6.2 mm Hg: with argon, it is 14.8 mm Hg and with helium, above 40 mm Hg. The breakdown power is seen from Fig. 11(c) to have a minimum at a pressure characteristic to each gas. In the case of helium the results agree well with theory [101].

Keep-alive characteristics may be studied with the aid of Fig. 12(a) which shows how the spike falls [47] with increase of initial ionization, measured in arbitrary units of light emission. Random fluctuations of electron density after about 100 h operation, due to wander of



(a) (b) (c)

Fig. 11—Breakdown performance of a TR tube. Frequency is 9.375 gc with 1B24 tube. (a) Effect of incident power. (b) Effect of gas pressure. (c) Breakdown power vs gas pressure.



(a) (b) (c)

Fig. 12—Performance of TR tube keep-alive. (a) Leakage vs electron density. (b) Improved keep-alive electrode. (c) Leakage vs total pressure.

the discharge, are avoided by fitting, as shown in Fig. 12(b), a bushing made of quartz, sapphire or mica. Fig. 12(c) shows that the spike is now less sensitive to electron density, and, in fact, such tubes can operate at 200 kw peak input for periods exceeding 1000 h. An alternative approach [13] to this problem depends upon the greatly-reduced probability of coincidence between these extinction periods and the arrival of a transmitter pulse by the use of two independent primers. Such TR tubes have been shown in Fig. 9 and experience confirms that the chance of failure, even once in, say, the 10^9 pulses involved in a typical 500-h run is substantially eliminated.

In a broadband TR tube, the arc loss and recovery time are controlled predominantly by the gas characteristics and physical conditions in the vicinity of the input window where the electron density is high. On the other hand, spike and, to a first approximation, flat leakage are determined by the characteristics near the resonant gaps. The conditions for optimizing these two groups of parameters can be independently ensured by isolation of the corresponding regions of the tube. This is confirmed [46] by experiments made on a demountable three-gap TR tube in which the window was encapsulated so that the appropriate gas filling was confined in an envelope separate and distinct from that in the main body. The incident power was at 2.8 gc in 350-kw pulses of one-microsecond duration at 1000 gc. With an argon filling, and a body pressure of 5.8 mm Hg, the spike- and flat-leakage energies remained constant at 0.3 erg/pulse and 7 mw respectively for window pressures up to at least 40 mm Hg. The recovery time, on the other hand, varied with pressure, being 160 μsec and 40 μsec at 6 mm Hg and 40 mm Hg respectively. This independence of characteristics is maintained on inclusion of quartz wool or water vapor recovery agents, making such a technique a practical, if complicated, approach to improved TR tubes performance at the lower frequencies.

V. Discharge-Tube Duplexers

A. *Branched Guide*

A conventional branched-guide discharge-tube duplexer is shown in Fig. 13, two gas switches being connected in E-plane stubs from the main transmission line. The window of the ATR tube is nearly flush with the inner wall while that of the TR tube is about $\frac{1}{4}\lambda_g$ away. On transmission, both windows break down, and the power proceeds to the antenna. In the low level condition the quiescent ATR tube presents a high admittance across the main line at a plane A such that antenna power proceeds through the TR tube and into the receiver. At powers exceeding 50 kw at 10 gc and 100 kw at 3 gc the life of the TR tube can be prolonged by the use of a pre-TR tube suitably positioned in the stub. Branched duplexers can have any combination of series or shunt stubs and may be designed [109] for two-channel working.

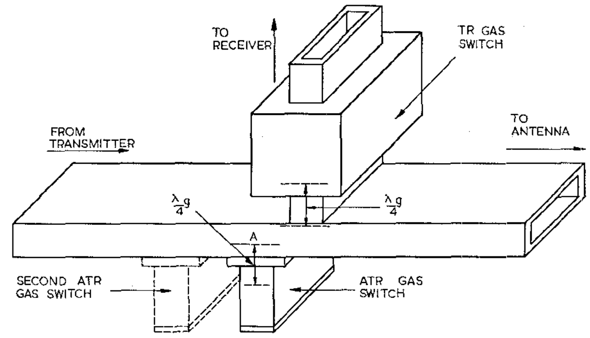


Fig. 13—Branched gaseous-discharge duplexer.

For efficient operation of, for example, a series-junction duplexer, Z_t, the impedance of the cold transmitting tube transformed along the line and combined with that of the ATR tube, should be zero. For any other value, there is a branching loss factor given by

$$L = 1 + \tfrac{1}{2}Z_t. \tag{15}$$

If, over the working frequency band, the impedance of the cold transmitting tube, or of the ATR tube during its recovery period, causes Z_t to fall outside acceptable limits then two or more ATR tubes may be employed. The exact distance between the ATR tubes depends [109] on the bandwidth required, the position of the TR tube and on whether or not the ATR tubes are stagger tuned.

The power handling capacity of the normal branched duplexer is limited by available tubes to about 300 kw at 9.5 gc and 1 mw at 3 gc. Higher capacity may be achieved in duplexers based on full transfer directional couplers. Energy flowing along the main line from the antenna is H-coupled to the auxiliary guide, and thence to the receiver, by a window aperture. Under high power conditions, the window breaks down and energy proceeds from the transmitter along the main line to the antenna. For the frequency band, 8.6–9.5 gc, the low-pressure gas filling may be [64] in the auxiliary guide or, alternatively, in the intervening space of a double walled window. The latter gave an insertion loss of 0.5 db, spike of 0.1 erg/pulse, flat of 20 mw and a recovery time of 6 μsec. In the "curtain" full-transfer duplexer [88] about 20 quartz gas tubes are spaced 1 cm apart along the aperture coupling the 2.84-inch by 1.34-inch waveguides. Over the band 2.6–3.1 gc the insertion loss is 0.3 db, recovery time 50 μsec and maximum power 23 mw peak, 16 kw mean.

The discharge tube in a duplexer protects the receiver from random external pulses only when the keep-alive is working. Under stand-by conditions of the equipment, protection can be afforded by mechanical switches or shutters. These may take the form of a spring-actuated vane which closes the waveguide aperture but which can be pulled out by a solenoid during normal operation. Short-circuiting posts have been built [105] into broadband TR tubes: while giving 40-db protection for one post and 70 db for two, the over-all cost of the tube is increased thereby.

B. Balanced Circuit

The arrangement of a balanced duplexer depends upon the type of gas switch and hybrid junction adopted. Fig. 14(a) shows two hybrid rings or tees and two broadband discharge tubes. The transmitter in arm T feeds the output arms of the hybrid in antiphase and, due to the $\frac{1}{4}\lambda_g$ separation of the gas switches, the reflections arrive back in phase so that the energy proceeds via arm A to the antenna. The leakage powers proceed via arm L to the dummy load while signals from the antenna proceed via arm R to the receiver. Alternative configurations employing 3-db directional couplers, either alone or with other hybrids, are shown in Figs. 14(b) and (c). Balanced duplexers must be followed by a passive-protection tube to safeguard the receiver against random pulses.

Good discrimination in balanced duplexers is obtained only if the phase changes through, and the spike and flat leakages of, the TR tubes are equal throughout the duration of the transmitter pulse. This condition is facilitated by the use of 3-db slot hybrids since, as shown in Fig. 15(a), both switches can then be placed [58] in one gas container. Such dual TR tubes have been made [105] for 3 gc, 5 gc and 9 gc and provide a cancellation of the spike of 4–6 db and of the flat, 6–10 db. Duplexers containing, for example, WF46 or BL-27

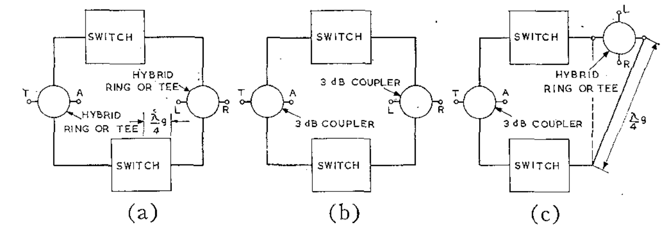


Fig. 14—Types of balanced duplexer configurations. (a) Two hybrid rings or Tees. (b) Two 3-db directional couplers. (c) One 3-db coupler plus one ring or Tee.

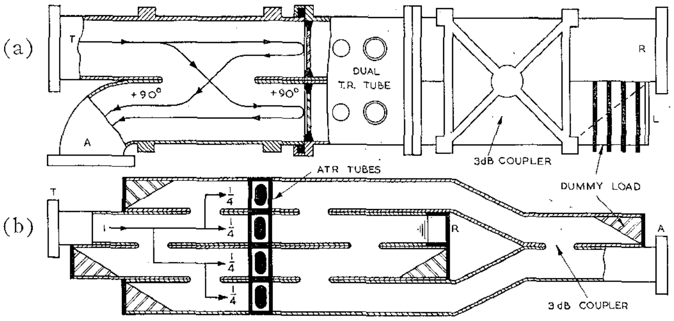


Fig. 15—Balanced duplexers with gas switches. (a) Dual TR type with 3-db couplers. (b) ATR type with power divided by four.

dual TR tubes give an isolation of 20 db, VSWR of 1.2, insertion loss of 0.8 db and an over-all spike of 0.01 erg/pulse in the range 8.5–10.0 gc.

Provided the receiver arm is fitted with a TR tube, balanced duplexers can employ pre-TR tubes with consequent increase in bandwidth and power-handling capacity. Two CV2339 tubes in a balanced circuit can handle [28] 2.5 mw peak, 3.5 kw mean at 3 gc. In a "curtain" arrangement [88] of the slot-hybrid duplexer, a number of quartz switch tubes placed in each of the coupling apertures give a good power and bandwidth performance.

Balanced duplexers may employ ATR tubes and the arrangement with two hybrid tees is rather like that of Fig. 5. For broadband performance, arrays of up to four tubes can be used [67] with advantage. The principle of power division can be extended and Fig. 15(b) shows an ATR duplexer in which the division is by a factor of 4. Such a duplexer for 2.6–3.1 gc, employing 1B44 ATR tubes with a 1B58A TR tube in the receiver arm, has given [88] satisfactory performance with powers up to 5 mw peak, 4 kw mean.

Circularly-polarized waves are employed in the balanced duplexer of Fig. 16. The input at T from the transmitter excites, via the transducer, a plane TE_{11} wave in the circular guide which is converted, by the quarter-wave plate, to circular polarization. The usual two gas switches are here combined into a single non-polarized tube which reflects the incident wave with the opposite hand of circular polarization. After retraversing the quarter-wave plate, the energy emerges from arm A and proceeds to the antenna. Leakage power from the switch proceeds to the output circuit, in this case a turnstile, and thence into the dummy load L.

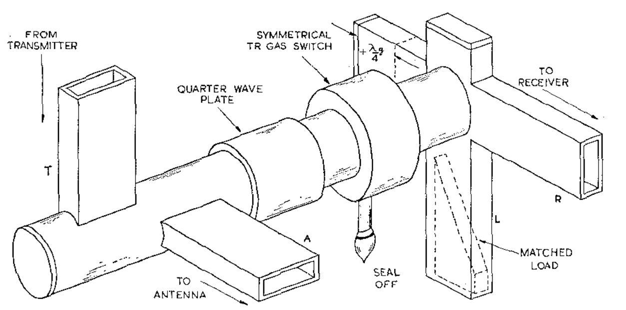


Fig. 16—Balanced duplexer using circular polarization. A quarter-wave plate is at the transmitter end and a turnstile at the receiver end.

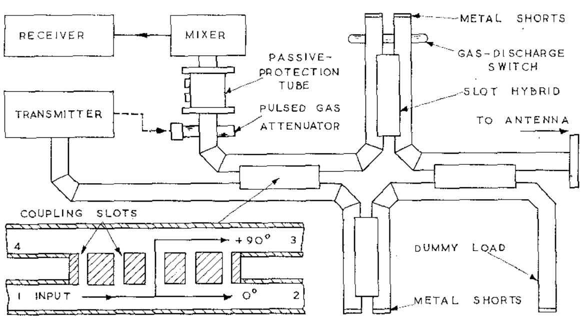


Fig. 17—Phase-change duplexer with slot hybrids. Frequency is 8.5–10.0 gc.

In the low-level condition the gas switch is transparent and signals proceed from the antenna to the receiver via arms *A* and *R*.

The polarization-twist duplexer, as its name implies, consists of a circular guide, supporting a TE_{11} mode, in which an array of, say, 20 discharge tubes provides a mechanism for rotating by 90° the plane of polarization of the transmitted wave. Suitable ports are arranged for connection to the transmitter, antenna, receiver and a dummy load. A simple krypton-filled tube, with quartz chips as the recovery agent, the CV2285, has been so employed by J. R. Thomas in unpublished work. Further protection is given by including [28], in the receiver arm, such TR tubes as the CV2378 and CV2429. The resulting duplexer is very successful as regards life and reliability in operation.

C. Individual Function

A major advance in reliability and performance has been achieved [28], [55] by evolving systems in which each duplexing function is performed by a separate tube designed specifically for that task. The duplexer can be branched or balanced and externally- and/or internally-excited tubes may be employed singly or in any combination. For example, if electrodes are placed [98] in the gas filled auxiliary guide of a full transfer directional coupler, the discharge can be made intense enough to cause cutoff. This means that the power flows uninterrupted along the main guide. Thus if an external prepulse of suitable amplitude is applied, power can be switched, during the period of transmission, from the auxiliary to the main arm and back again. Other duplexing tubes may be added to give increased protection to the receiver.

In typical arrangements [28] for a branched-guide duplexer, the single TR tube is replaced by any of the following combinations:

1) Pre-TR tube, followed by a prepulsed attenuator and a passive-protection tube.
2) Pre-TR tube, followed by an unprimed TR and a passive-protection tube.
3) Pre-TR tube, followed by an unprimed TR, prepulsed attenuator and a passive-protection tube.

A duplexer [54] for 9.5 gc based on scheme 1) employed a quartz-tube pre-TR switch [94], a prepulsed gas attenuator and a CV2359 passive-protection tube, the resultant leakage being less than 0.01 erg/pulse. An assembly for 3 gc, based on scheme 3), was made up from a CV2339, CV2378, CV2379 and VX3262 in that order: the transmitter power was 2.5 mw peak, 3.5 kw mean. At low levels, the total insertion loss was 0.5 db with a VSWR better than 1.2 over an 8 per cent bandwidth. At high levels, the spike with the CV2379 unpulsed was 0.15 erg and pulsed, 0.01 erg: the over-all recovery time was about 40 μsec.

Balanced duplexers using any of the configurations of Fig. 14 can be modified [28] by replacing each of the switches by a pre-TR/unprimed TR combination and inserting, in the receiver arm, a prepulsed attenuator followed by a passive-protection tube. The balanced duplexer of Fig. 17 employs [77] two power-sensitive phase-changers: each of these comprises a pair of 3-db directional couplers which, as the inset shows, divides incident power in arm 1 equally between arms 2 and 3 with phase changes of 0° and 90° respectively. One of the phase changers is provided with a CV2482 pre-TR tube placed $\frac{1}{4}\lambda_g$ in front of its short circuit. The electronic length of the phase changer is hence $\frac{1}{2}\lambda_g$ less for high-power signals than for low-power. Thus power proceeds from the transmitter to the antenna and from the antenna to the receiver. The VSWR in either condition is better than 1.2, and the isolation better than 25 db, over the band 8.5–10.0 gc: the power-handling capacity is about 800 kw peak, 400 watts mean. The duplexer is completed by a pulsed attenuator and a CV2359 passive-protection tube placed in the receiver arm. In a life test of such a duplexer, only six receiver diodes failed in a total period of 36,855 h.

VI. Methods of Measurement

A. Attenuation and Impedance

Measurement of the performance of both complete duplexers and the individual parts is required in the process of development and as a control during manufacture. The methods described relate specifically to gaseous-discharge duplexers but they may be applied, with modifications, to passive and solid-state designs.

Typical measurements [6], [109] include insertion loss, input VSWR and frequency bandwidth: they must usually be carried out at both high- and low-power levels. The former usually involve extensive life tests and an economy of power by a factor 2^n may be achieved [127] by using a power-multiplication circuit of n-hybrid junctions. In general, the conventional laboratory instruments must be supplemented by specialized equipment.

The arc loss of, for example, a TR tube can be measured by setting it up in a transmission line and comparing the power entering the load with that when the tube is replaced by a short circuit. In experiments at 35 gc, L. Hodgson, in unpublished work, obtained increased accuracy by a bridge method. The gas tube was placed in one arm and a calibrated attenuator in the other. A subsidiary attenuator and phase-changer were adjusted to balance the bridge with a short circuit in the tube mount. On replacing the tube, the necessary change in the calibrated attenuator to restore balance gave a measure of the arc loss. Increased accuracy, especially at high powers when the fractional arc loss becomes small, is obtained by replacing several tubes in line.

Other properties which require measurement include minimum firing power, keep-alive characteristics and leakage due to direct coupling at the fundamental and harmonic frequencies. Transmission measurements on TR tubes should be made with the keep-alive discharge on: in this way both insertion- and interaction-loss are measured. With ATR and pre-TR tubes the position of the equivalent short circuit is important: this is usually determined by bridge methods, the tube being balanced by a moveable short circuit.

B. Leakage

The spike and flat leakages of a gaseous-discharge duplexer may be separately measured by employing the cancellation principle. As shown in Fig. 18(a), the total leakage energy is first measured. Since this averages only a few microwatts, a thermistor, with suitable bridge, is usually employed. A small fraction of the RF energy is coupled out of the transmission line into the output of the duplexer, the amplitude and phase being adjusted to give cancellation of the flat as observed on a fast sweep oscilloscope: this permits the spike energy alone to be measured. In practice [6] the flat leakage is not constant during the pulse and complete cancellation is not achieved. The measured spike will, therefore, be higher than its true value. With short pulses care must be taken to equalize the path lengths traversed by the leakage and cancellation signals.

Another technique [109] employs a high-Q-factor cavity as a filter to discriminate against the energy in the relatively widely spaced sidebands of the spike. In the arrangement of Fig. 18(b) a suitably positioned rejection cavity causes the flat power to pass through the transmission cavity and be measured by a thermistor.

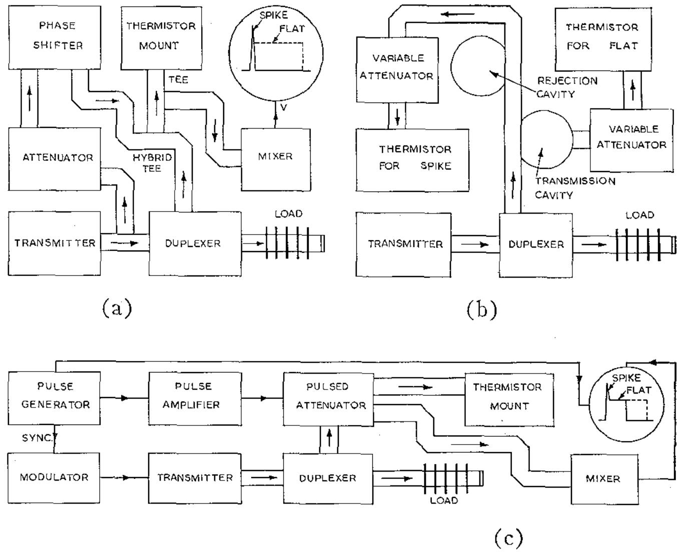


Fig. 18—Methods of measuring duplexer leakage power. (a) Cancellation of flat. (b) Separation of spike and flat by cavities. (c) Variable pulse length.

For Q-factors of, say, 1000 only a negligible portion of the spike energy passes through the cavity: most of it proceeds along the line and enters a second thermistor where it is measured. The cavities must be precisely tuned and corrections made for their small reflection and absorption losses.

Less specialized equipment is required in a further method using transmitter pulses of different durations. Measurement of the total leakage with several pulse durations enables the spike leakage to be determined by extrapolation to zero time. The requirement of a variable pulse-duration modulator is avoided by the use, as shown in Fig. 18(c), of an attenuator to which an excitation pulse can be applied at any chosen time following the spike. The rise in attenuation after pulsing must be rapid and values of 50 db/μsec have proved satisfactory. If, for example, the transmitter has a pulse duration of one microsecond, the duration of the measured flat leakage can be varied in 0.1-μsec intervals by successive reduction in the delay of the pulse generator. Extrapolation then gives, as before, the spike leakage alone.

C. Recovery Time

Measurement of the attenuation of a low-level signal as a function of time after the end of the high-power pulse gives the recovery time of a gaseous-discharge duplexer. In the arrangement shown in Fig. 19 the frequency of the signal generator is set near that of the transmitter and its output fed via a directional coupler into the duplexer. The range of times of occurrence of the probing signal should include an interval preceding the transmitter pulse to provide a reference level. The change in amplitude of the signal displayed on the oscilloscope is determined by a calibrated attenuator.

In a modification of this method, the transmitter is pulsed, for example, only once for every two signal-generator pulses: thus alternate low-level pulses come

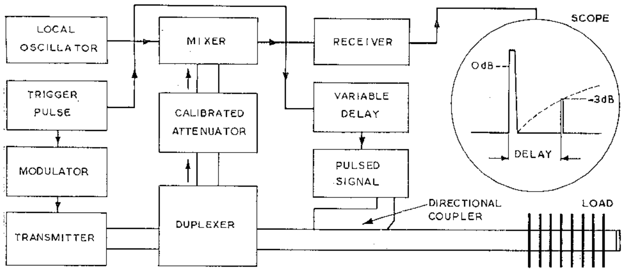


Fig. 19—Arrangement for measuring recovery time. The amplitude of a low-level signal is observed as a function of delay after the transmitter pulse.

through unattenuated. The signal on the oscilloscope shows two superimposed pulses, one unattenuated and the other affected by the recovery time of the tube. The difference in height of the pulses is a measure of the attenuation due to the recovery time. The recovery curve can be made continuously visible on the oscilloscope by arranging for the modulation pulses on the low-level generator to occur at different delay times or, better still, in rapid sequence [102]. Alternatively, by a square-wave modulating the IF amplifier at, say, 2 mc, the display can be made to show [6] a wave of this frequency varying in amplitude according to the recovery law. In all methods transmitter breakthrough must be avoided by using a resonant-cavity or waveguide-bridge filter supplemented by a blanking pulse on the receiver. These techniques are suitable for either low- or high-Q-factor TR tubes and can, with modification, be applied to gas attenuators, pre-TR and other tubes. ATR tubes, for example, can be measured by observing [6] the recovery time of the branching loss or main-line reflection by low-level probing techniques.

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Impedances of an Elliptic Waveguide (For the $_eH_1$ Mode)*

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Summary—**The power-voltage, power-current and voltage-current impedances for the elliptical waveguide for the fundamental mode ($_eH_1$ mode) are obtained by two different methods.**

The first method consists of using the exact fields inside a perfectly conducting elliptical pipe. Numerical results were obtained by numerical integration of the integrals involving Mathieu functions by the Gaussian Quadratures method by a digital computer.

In the second method approximate fields which satisfy the boundary conditions were used. By this approximate method, actual expressions for the impedances are obtained as a function of minor to major diameter ratio with no need of numerical integration.

The actual expressions for the impedance obtained by the approximate method give the impedance for elliptical waveguide within six per cent. On the basis of comparison with the exact numerical solution the expressions for the approximate impedance give the impedance of elliptical waveguide within three per cent if they are scaled by 1.03.

Introduction

CHU[1] in 1938 obtained numerical results for the exact cutoff wavelength for several modes in elliptical waveguides. He also obtained numerical results for the attenuation in elliptical waveguide.

Kihara[2] published a paper in 1947 using the variational method to determine the propagation constant of hollow pipes and cavities. Kihara was able to obtain the propagation constant of an elliptical waveguide within one per cent, in the first approximation, using trial fields.

In 1958 Harrowell[3] obtained the impedance of elliptical waveguide by using an approximate method. He showed that the magnetic field lines inside a circular waveguide were approximately ellipses. Therefore, he was able to introduce conducting ellipses without disturbing the fields. Harrowell did not mention within what degree of accuracy his impedances would compare with the exact value.

Harrowell's voltage-current impedance agrees perfectly with our exact impedance, but his impedances involving power differ from our exact values; this difference is greater for eccentricities close to unity.[4]

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