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# Slotted-Waveguide Amplifiers for Millimeter Waves

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Slotted waveguides and ribbon electron beams are studied as electron-beam interaction circuits instead of the usual circular-beam devices. The slotted waveguide and ribbon beam can be many wavelengths long, and therefore the effective beam cross-sectional area is greatly increased over that of the circular-beam helix, coupled-cavity, or klystron tubes. This increased beam area results in more beam current, more beam power, and less stringent requirements on cathode current density and focusing fields. The slotted waveguides are made by milling either grooves or threads into a slab of metal, and are therefore rugged and self-aligning. Both theory and experiments are described.

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# SLOTTED-WAVEGUIDE AMPLIFIERS FOR MILLIMETER WAVES

## INTRODUCTION

Wide-bandwidth power of a kilowatt or more near 95 GHz is needed for several applications. At 95 Ghz semiconductor devices cannot yet produce a kilowatt of power together with bandwidth. In the usual linear beam tubes such as the traveling wave, coupled-cavity, folded waveguide, and klystron tubes, the channel diameter of the electron beam and circuit is inherently limited to approximately 0.1 wavelength at the highest operating frequency. This diameter (0.013" at 95Ghz) is too small for high-power operation, beam focusing, and ease of construction. The fast-wave gyro-traveling-wave and gyro-klystron tubes utilize waveguide, and therefore have a larger interaction area and can produce higher power at 95 Ghz, but the bandwidth is narrow because they must operate very close to the waveguide cut-off frequency. They also use relativitistic electron beams and superconductor magnets. Thus, to obtain both bandwidth and power without the use of high-voltage and superconductor magnets, it becomes necessary to devise other designs that utilize circuits with larger interaction areas.

The approach used in this investigation is to increase the interaction area in one dimension by using either a slotted $TE_{10}$ waveguide or a slotted parallel-plate transmission line together with a ribbon electron beam that passes through the slots as shown in Figure 1. The slots are on both sides and are located in the center of

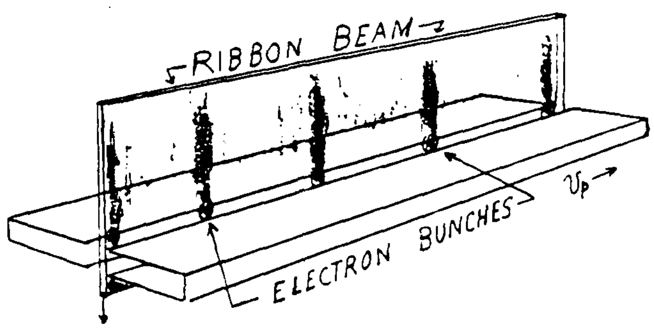


*Figure 1. Traveling Waves of Bunched Electrons entering the slotted waveguide.*

the waveguide where the RF electric fields are highest. It can be shown both experimentally and theoretically that the slots can be made approximately as wide as the waveguide height (or spacing of the parallel plate transmission line) without significant modification of the RF fields. If the electrons along the width of the

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continuous ribbon electron beam are properly phased and bunched, then as they pass through the slotted waveguide, they add energy and provide gain to the electromagnetic wave that is moving down the slotted transmission line. Although the thickness of the electron beam passing through the slots in the transmission line will still be limited to the 0.1 wavelength, the width of the ribbon electron beam can be many wavelengths long and as long as the slotted waveguide. Therefore, the interaction area and beam cross-sectional area are much larger due to the extra degree of freedom in the width of the ribbon electron beam. The traveling-wave bunching, phasing, and focusing of the ribbon electron beam that passes through the slotted waveguide (or transmission line) then becomes the major design problem.

If the ribbon beam were continous with no bunching, then the electrons would give up energy on the positive half cycle of the RF wave, but would extract energy on the negative half cycle. Thus, the net energy exchange would be zero. To provide a net increase and gain in the RF power, it is necessary to phase or bunch the electrons to enter the slotted waveguide only during the positive half cycle of the RF wave. The ribbon electron beam can be bunched by either velocity modulation or density modulation. With the velocity modulation method, used in these experiments, the beam velocity is changed successively by the incoming RF sine wave. As the ribbon beam moves down the tube, the faster electrons catch up with the slower electrons of the previous cycle. The overall result is the formation of dense electron bunches at a fixed distance down the tube similar to a klystron. In the density-modulation bunching method, the traveling wave is applied to an anode (or grid) near the cathode with the anode and cathode acting as a transmission line. A DC bias on the cathode allows electron emission and bunches of current to occur only near the crest of the positive cycle of the RF sine wave similar to a klystrode. The disadvantage of density modulation is that the large transit time between the thermionic cathode and anode causes distortion and deterioration of the bunches at gigahertz frequencies. If the electron emission source were field emission instead of thermionic emission, the transit time would be negligibly small, but field-emission cathodes are not yet practical. Three velocity-modulated design approaches, as well as experimental results, are described in this Report.

TRAVELING-WAVE KLYSTRON

Practical, low-cost millimeter-wave circuits require simple and smooth machining because of the extremely small dimensions. Slotted waveguides are made by milling square grooves into a flat slab of metal, cutting the metallic slab in half, and then placing one half above the other with the grooves in alignment as shown in Figure 2. The ribbon electron beam passes between the metallic slabs. In the

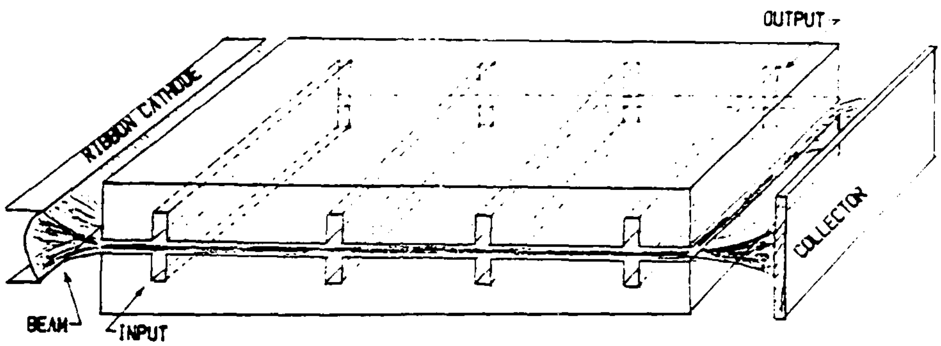


*Figure 2. Traveling-wave Klystron using slotted waveguides milled from a slab of metal and using a convergent ribbon electron beam. The Standing-Wave klystron utilizes the same structure, but with the waveguides shorted at each end.*

traveling-wave klystron, the slotted waveguides are terminated in their characteristic impedance with graphite wedges so that only traveling waves are present without reflected signals that would cause standing waves. The input signal is applied to the first slotted waveguide. This input signal velocity modulates the ribbon electron beam with a traveling-wave phase pattern. This velocity modulation becomes density modulation as the ribbon beam travels within the space between the waveguide sections. The intermediate waveguide lines provide additional modulation and serve as additional "bunchers" of the ribbon electron beam in a similar manner to a conventional 4-cavity klystron. The electron-density bunching along the ribbon forms discrete traveling electron beams bunched at the wave crest similar to a row of telephone poles moving at the phase velocity of the wave. The final output slotted waveguide acts as a klystron "catcher" or output circuit with build-up of signal along the waveguide given by eq. (11A) in Appendix A.

The traveling-wave klystron using separate electron beams has been studied by Mihran[2], Pohl[3], Vaughn[4], and others[5,6]. The objective at that time was high microwave power (10 MW) and wide bandwidth. Such tubes were large and cumbersome at microwave frequencies, but the size and shape become much more attractive at millimeter waves, especially when a continuous ribbon electron beam is used instead of separate electron guns.

An experimental tube was tested in the frequency range of 9.9 to 15 Ghz because of the availability of test equipment. Once the proper design is obtained, presumably it can be scaled to 35 or 95 Ghz. The four slotted waveguides were 6" long with an inside waveguide width of .622", a waveguide height of .100", and a slot width of .080". The waveguides were made by milling grooves with a depth of .271" and a width of .100" in a slab of metal as shown in Figure 2. The ribbon electron beam was .080" thick by 4.75" long. Rectangular focusing coils were designed and purchased. A convergent-flow electron gun was designed and made using the Herrmannsfeldt[8] computer gun program, a plot of which is shown in Figure 3. The oxide cathode was sprayed onto a nickel sleeve. Heater filaments for the oxide cathodes were designed and purchased. The rectangular box-like

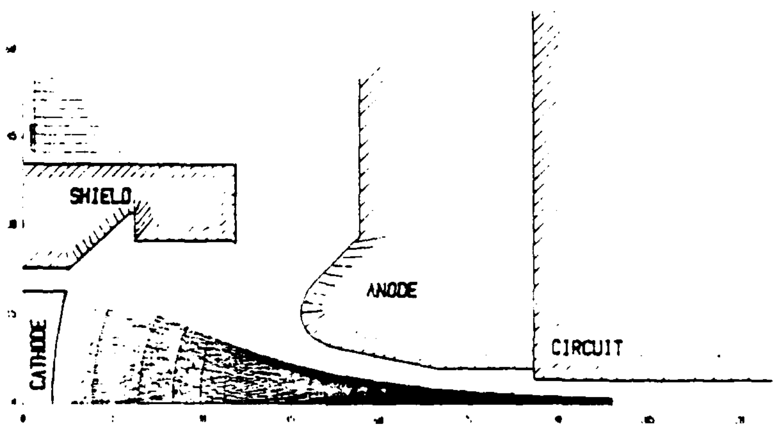


*Figure 3. Convergent Ribbon-Beam Electron Gun designed by using the EGUN computer code for a PC-AT computer as devised by Herrmannsfeldt. See reference 9.*

envelope was made of non-magnetic stainless steel. The vacuum seals were vacuum wax such that the tube could be quickly dismantled for modifications. The major problem was focusing the ribbon beam and obtaining good beam transmission through the structure. The input waveguide entered the structure at a point where a magnetic coil should have been placed. This caused a severe perturbation in the magnetic field near the electron gun. In spite of this, a gain of 19 db was obtained with 94 watts output and 1.18 watts input. The beam voltage was 10kv and the constant magnetic field was 600 gauss with a beam current of approximately .9 amperes. The output gain vs frequency is shown in Figure 4 as well as the calculated output power using eq. (11A) in Appendix A. The agreement between experiment and calculations is a lucky coincidence because the beam current measurement error was about +10%, which causes a possible error in power output of about +20%. In any case, the experiment clearly shows the expected decrease of power as the frequency is increased.

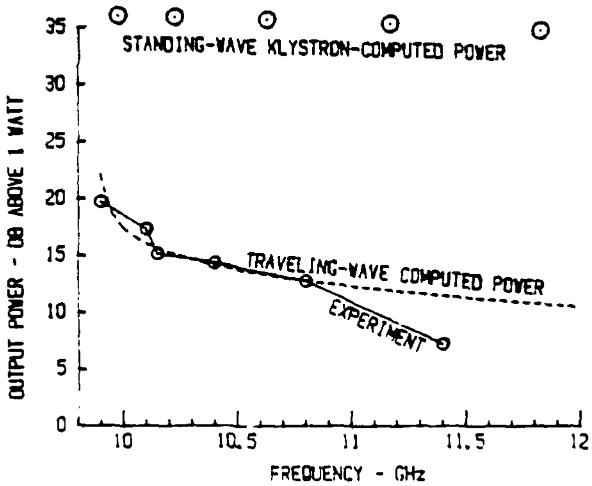


*Figure 4. Experimental and calculated output power of the Ribbon-Beam Traveling-Wave Klystron. Also calculated output of the Standing-Wave Klystron with the same beam current and same slotted waveguide.*

The above "agreement" between experiment and calculations provides guarded confidence in performance predictions based on calculations. To get an approximate idea of the performance and the design at 95 Ghz, a waveguide size of inside width .0622 by inside height of .013" was selected which has a cut-off frequency near 95 Ghz, and therefore provides a high shunt impedance and high output power at that frequency. The slot width, equal to the waveguide height, is .013". The milled grooves in the metallic slab would be .013" wide by .025" deep, a not too difficult task for a good machine shop. The length of the milled-groove waveguides was selected to be 4" long with a ribbon electron beam .013" thick by 3" wide. With a 10 to 1 beam convergence and a modest cathode loading of 2 amperes per square centimeter, the beam current would be 5 amperes with a magnetic field of approximately 1000 gauss and a beam voltage of 10 kv. If the ideal conditions of saturated beam modulation, canceled backward currents, a beam current of 5 amperes, and a beam voltage of 10 kv are assumed, then the output power and efficiency are calculated from eq. (11A) and are shown in Figure 5. It can be seen from Figure 5 that kilowatts of output power are possible from the traveling-wave klystron, but the efficiency and bandwidth are poor (the DC input beam power is 50KW). As seen in eq. (11A), the maximum output power (and efficiency) are

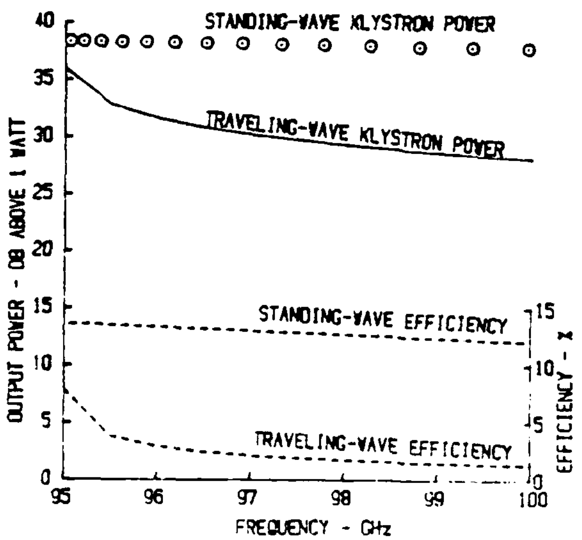


*Figure 5. Computed output power and efficiency of the Traveling-Wave and Standing-Wave Klystrons near 95 GHz.*

obtained near waveguide cut-off and when the current and current density are a maximum. Thus, the best performance depends upon high density and good focusing. It is difficult to achieve and maintain alignment of the electron gun and magnets to focus a ribbon beam that is 0.013" thick by 3" wide. There are several focusing schemes, other than a constant magnetic field, such as periodic focusing with permanent magnets and wiggler-focusing as studied by Eppley, Herrmannsfeldt, and Miller[9]. They describe a proposed 100 MW ribbon-beam klystron at 11 GHz with 1000 amperes of beam current.

The relationship for power, eq. (11A), indicates that maximum power and efficiency are obtained not only when the beam current is as high as practical, but also when the length of the slotted waveguide is long. For example, for the case of Figure 5 the computed output power at 95 GHz is 3936 watts with an efficiency of 7.9 percent. If the line length is increased from 3" to 6", while maintaining a constant beam current density, then the output power becomes 13270 watts, and the efficiency increases to 13.3 percent. There are similar gains over the bandwidth. At 95 GHz this narrow waveguide has an attenuation of about 7 db/meter, and also voltage breakdown in the guide would occur at a few kilowatts of power.

In the ribbon-beam klystron the electrons ideally should move straight forward with little or no sidewise displacement (< 0.2 wavelength) such that the traveling-wave bunching patterns on the beam are maintained without spill-over or diffusion to adjacent bunches. Calculations indicate that sidewise displacements are not large as long as the electron gun and the RF fields introduce sidewise velocities less than approximately 10 volts of energy.

## STANDING-WAVE KLYSTRON

If the waveguides of the traveling-wave klystron are terminated in a short circuit, then, only standing waves will be present, and the ribbon beam will be velocity modulated with the standing-wave pattern. This, in turn, sets up standing-wave patterns in the remaining waveguide lines. The lines are similar to the resonant cavities of a klystron except that in the usual klystron the radial cavity size is only about one-half wavelength. The standing wave patterns will be present at the frequencies where the waveguide is an integral number of half-guide wavelengths in length as given by eq. (13A) in Appendix A. Therefore, at each resonant frequency $f_N$ of the waveguide lines, the tube will amplify and produce power. The bandwidth near $f_N$ will be narrow, and the power will depend on the effective shunt impedance near $f_N$. In the conventional klystron the shunt impedance is many thousands of ohms. In this situation, however, the shunt impedance is a somewhat smaller value and is given by eq. (20A) as 14000 ohms at 10 GHz and 1100 at 95 Ghz. Thus, at discrete frequencies $f_N$ there will be good output power and efficiency. At millimeter waves where the waveguides will be many wavelengths long, the discrete frequencies of amplification may begin to merge. It the tube amplification will overlap between these discrete frequencies, then a wide bandwidth amplifier will be obtained together with high resonant impedance , which implies high gain and efficiency. The output power is given by eq. (22A). Figure 4 shows the computed output power of a standing-wave klystron using the same electron beam and slotted waveguides as the traveling-wave klystron. The output power is greater because of the higher shunt impedance. Figure 5 shows the computed output power of the standing-wave klystron at 95 Ghz using the same electron beam and slotted waveguide as the traveling-wave klystron. In the 95 GHz case, the computed power of the traveling-wave klystron is also greater because of the higher shunt impedance.

## HELICAL WAVEGUIDE

Another embodiment of this extended-dimension slotted waveguide is the helical waveguide as shown in Figure 6. The slotted waveguide is continuous and acts similar to a helix-type traveling-wave tube wherein the helix is replaced by slotted helical waveguide. The slotted waveguide is made by cutting rectangular-shaped grooves or threads into the inner and outer metallic cylinders and then aligning these inner and outer grooves or threads as shown in Figure 6. A metal lathe is ideal for cutting these threads. A hollow electron beam is used, and it passes through the waveguide slots. The hollow electron beam can be many wavelengths in circumference, but the thickness of the electron beam and the slots is still limited to 0.1 wavelength. The electron-beam velocity is synchronized to transverse the distance between turns in the same interval of time that the electromagnetic wave travels an integral number of wavelengths around a single turn. Each azimuthal segment of the electron beam is bunched in a different phase from its neighboring segment. Therefore, a focusing requirement is that the hollow electron beam move parallel to the axis of the tube without appreciable rotation or azimuthal deviation.

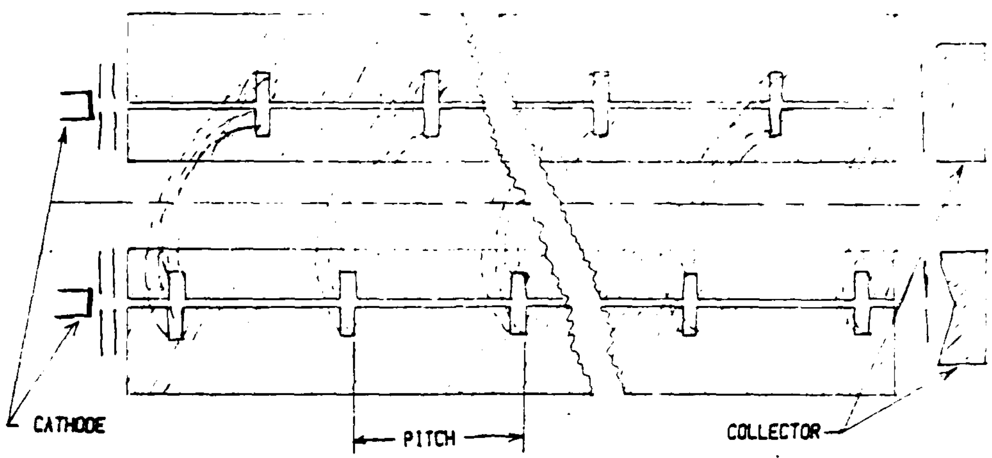


*Figure 6. Helical Waveguide. The annular electron beam moves longitudinally between the inner and outer cylinders and finally strikes the collector. Threaded grooves on the inner and outer cylinders form the helical waveguide. The spacing between the cylinders provides the slot in the slotted helical waveguide.*

Synchronization between the electron beam and the helical waveguide requires that the electron beam encounter the same RF phase as it moves from turn to turn. Straightforward analysis shows that synchronization can be obtained over most of the waveguide bandwidth. Equation (1) below is the omega-beta diagram synchronization equation. Equation (2) gives the required phase velocity for synchronization.

$$\frac{\beta_z\, p}{2\pi} = \frac{L}{\lambda_c}\sqrt{\left(\frac{f}{f_c}\right)^2 - 1} + M \qquad (1)$$

$$\frac{v_p}{c} = \frac{\frac{p}{\lambda_c}\left[\frac{f}{f_c}\right]}{\frac{L}{\lambda_c}\sqrt{\left(\frac{f}{f_c}\right)^2 - 1} \quad + \ M} \tag{2}$$

where $B_z$ is the phase shift per unit distance of the M-th space harmonic in the longitudinal direction of the beam, P is the distance between turns, or pitch, L is the mid-distance around a turn, M is the integral number of wavelengths around a turn, f is the frequency of operation, $f_c$ is the waveguide cut-off frequency, $\lambda_c$ is the waveguide cut-off wavelength, $v_p$ is the longitudinal phase velocity, and c is the velocity of light. One selects a value of M, R, and L that make $v_p/c$ in eq. (2) nearly constant over the required bandwidth. Graphic plots of equations (1) and (2) are shown in Figures 7 and 8 for the experimental tube to be described.

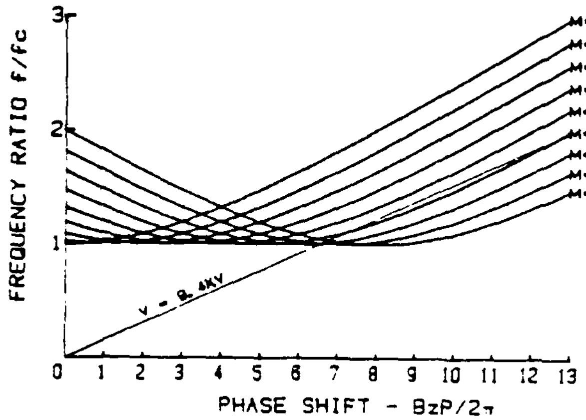


*Figure 7. The ω-β plot (frequency vs phase shift) of the helical waveguide circuit. M is the number of wavelengths around a single turn.*

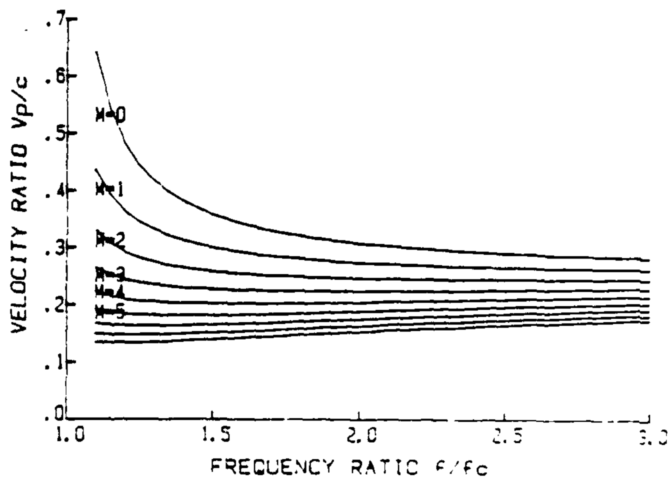


*Figure 8. This plot of phase velocity vs frequency demonstrates that the electron beam and wave remain almost synchronized over a wide frequency range.*

The small-signal analysis like that used for the traveling wave tube has been applied to this tube by Arnett[10] and by Ahn and Ganguly[11]. In this small-signal analysis the interaction impedance is reduced by the ratio b/p where b is the waveguide height and p is the pitch, or distance between turns of the helix. The wide spacing between the interaction gaps causes velocity modulation and klystron bunching to occur that is neglected in the small-signal analysis. As a result, the traveling-wave small-signal solutions give pessimistically low values of gain.

A klystron analysis was devised in which the velocity modulation of representative electrons are calculated exactly, and then the effects of all electrons are summed at the end of each turn to determine the energy given up by the electron beam to the RF field. The resulting increase in signal is added to the original signal. This analysis neglects space charge and is similar to the "large-signal analysis" used to calculate the electronic efficiency of klystrons, gyrotrons, and traveling-wave tubes. The program was written for use on a desk-top computer and takes about 2 minutes to run.

Experiment were performed in the frequency range of 10 to 18 Ghz where test equipment was available. Experimental results of gain vs frequency are shown in Figure 9. A ring-shaped oxide cathode was used that was .125 " wide by 1.625" in diameter. The cathode was immersed within the magnetic field, and the electron gun was designed to launch the electrons parallel to the axis along the lines of the magnetic field. The parameters of the experiment were:

(a) Waveguide width 0.622" by height .080" with a slot width of .050".

(b) Mid-diameter of the cathode and helix - 1.625"; M - 5; helix length - 7"

(c) Helix pitch, p - 1.417"; length around one helix turn, L-5.298"

(d) Measured cut-off frequency $f_c$-10.3 GHz; cut-off wavelength $\lambda_c$-1.15"

(e) Beam voltage 11 kv; beam current 0.1 Amp.; magnetic field 800 gauss.

There was no attenuator or isolator within the tube. The RF feed-through signal was measured, and when the electron beam was pulsed on, the resulting increase in output signal was measured as gain. The gain shown in Figure 9 was in the M-5 mode. Gain was also observed in the M-4 and M-6 modes, but, of course, at different beam voltages. Backward-wave oscillations were observed at approximately 10.4 GHz when the beam current was increased to approximately 0.13

amperes.

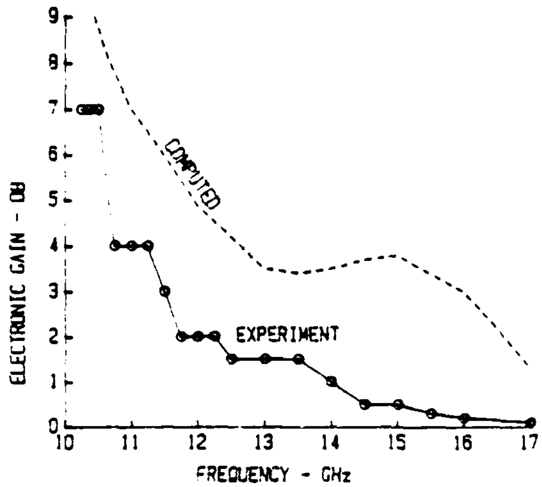


*Figure 9. Experimental and Computed Gain of the Helical Waveguide Tube. The computer calculations are from a klystron-like large-signal analysis that sums up the energy losses of individual electrons.*

The klystron analysis was used to calculate the gain of the experimental tube, and the results are plotted in Figure 9. The approximately 3 db difference can be accounted for by RF losses such as those in the input and output couplers. The small-signal traveling-wave tube analysis gives pessimistic results that indicate little or no gain (about 10 db lower gain than the klystron analysis).

The helical waveguide tube has the following disadvantages:

(a) Inherent backward-wave oscillations. The interaction impedance of the backward-wave modes is always higher than that of the operating mode.

(b) Low gain.

(c) The hollow electron beam has inherent instability. The space charge outward force is zero at the inner edge of the beam and is maximum at the outer edge.

(d) An off-axis convergent-cathode electron gun for the hollow beam is very difficult, but not impossible, to design.

## CONCLUSIONS

It has been shown that slotted waveguides can be used as circuit elements that provide more electron beam cross-sectional interaction area and therefore more beam current and more beam power than conventional traveling-wave tubes, coupled-cavity tubes, or klystrons. More specifically, calculations indicate that kilowatts of power can be obtained at 95 GHz with a modest cathode loading of 2 amps/cm$^2$, a beam convergence ratio of 10 to 1, a beam energy of 10 KV, and a magnetic field of approximately 1000 gauss. When made by milling grooves or threads into a slab of metal, the slotted waveguides are rugged and self-aligning and can dissipate lots of power due to the thick waveguide walls. A disadvantage is that the resistive losses of the waveguide are high because the waveguide height must be small to reduce the electron transit time across the gap. This can be somewhat relieved by a higher beam velocity. The traveling- or standing- wave klystron design is superior to the helical waveguide design because of the absence of both backward-wave oscillations and hollow-beam gun-design problems. All three designs have the disadvantage that the waveguide slot width and ribbon-beam thickness must be less than 0.1 wavelength at the operating frequency. This means thin ribbon beams which are not easy to align and focus.

One objective of this study was to devise a low-cost method of producing millimeter-wave power. The methods described here may reduce costs somewhat, and surely increase power, but clearly the costs of labor, parts, input and output couplers, and magnets will still be high. Some not-yet-devised lithographic method is needed to provide the necessary precision and to avoid the expense of hand-made assemblies.

## ACKNOWLEDGMENTS

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## APPENDIX A

## OUTPUT POWER ANALYSIS

Traveling-Wave Klystron:

In the traveling-wave klystron the $TE_{10}$ waveguide is terminated at both ends in its characteristic impedance. Therefore, only traveling waves are present. The objective is to calculate the output power from the slotted waveguide as a result of a known input power from an ideally bunched ribbon electron beam passing through the slotted waveguide. Such an ideal beam consists of a set of discrete beams that move at the phase velocity of the traveling wave similar to a row of moving telephone poles. These discreet beams are phased to enter the waveguide slot at the ideal time and location when the waveguide electric field E is maximum and is in a direction to slow the electrons. The energy lost by an electron is E b volts where b is the width of the waveguide. It will be assumed that:

(a) Neither the waveguide slots nor the presence of the electrons within the waveguide perturb the $TE_{10}$ fields of the waveguide.

(b) The waveguide RF fields are not large enough to turn back the electrons.

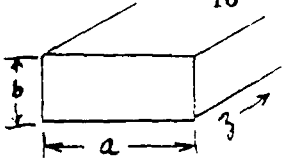
(c) The ribbon beam has a uniform thickness and linear current density of $I_L$ amperes/meter.

(d) All input current is converted to forward current with no backward current as shown in Appendix B.

(e) Space charge is neglected.

From waveguide textbooks, the power transmitted down the $TE_{10}$ waveguide, neglecting losses, is:

$$W_T = \left[\frac{ab}{4\eta}\sqrt{1-\left(\frac{f_c}{f}\right)^2}\,\right] E^2 = K\,E^2\,, \qquad (1A)$$




and the change of power with length of the waveguide is:

$$\frac{d\,W_T}{d\,z} = 2\,K\,E\,\frac{d\,E}{d\,z}\,, \quad \text{with} \quad K = \frac{ab}{4\eta}\sqrt{1-\left(\frac{f_c}{f}\right)^2}\,, \qquad (2A)$$

where $W_T$ is the average transmitted power in watts, a and b are the width and height of the waveguide, $\eta$ is 376.73, fc is the waveguide cutoff frequency, f is the operating frequency, and E is the peak electric field at the center of the waveguide in volts/meter. The attenuation per meter in the $TE_{10}$ waveguide due to resistive losses, assuming smooth walls, from textbooks[1], is:

$$\alpha = \frac{R_s}{b\,\eta\sqrt{1-\left(\frac{f_c}{f}\right)^2}} \left[ 1 + \frac{2b}{a}\left(\frac{f_c}{f}\right)^2 \right] , \tag{3A}$$

where Rs is the resistance per square of the waveguide walls ( $R_s = \sqrt{\frac{\omega\mu}{2\sigma}}$ where $\sigma$ is the conductivity of the waveguide walls in mhos/meter). The attenuation, power loss, and power transmitted are related by the equation:

$$\alpha = \frac{W_L}{2W_T} , \quad \text{or} \quad W_L = 2\,\alpha\,W_T \ . \tag{4A}$$

Thus,

$$\frac{d\,W_L}{dz} = 2\,\alpha\,W_T = 2\,\alpha\,K\,E^2 \ . \tag{5A}$$

The ideally bunched electron beam moves across the waveguide against the RF field and gives up energy E b volts to the RF field. Therefore, the RF input power $W_{in}$ from the ribbon beam is the current times the voltage:

$$\frac{W_{in}}{dz} = \frac{1}{2}\,M\,I_L\,E\,b \ , \tag{6A}$$

where $I_L$ is the current per unit length of the ribbon beam in amperes/meter, and M is a factor, to be discussed later, that accounts for transit-time corrections, imperfect bunching, and imperfect fields at the slots. The factor of 1/2 adjusts peak values to RMS values.

The change in output power plus the change in losses equals the change in input power:

$$\frac{d\,W_T}{dz} + \frac{d\,W_L}{dz} = \frac{d\,W_{in}}{dz} \ . \tag{7A}$$

When eqs (2A), (5A), and (6A) are substituted into eq. (7A), one obtains the differential equation:

$$\frac{dE}{dz} = \frac{M\, b\, I_L}{4\, K} - \alpha\, E \ , \tag{8A}$$

which when integrated becomes:

$$E = E_1\, \epsilon^{-\alpha z} + \frac{M\, b\, I_L}{4\, K\, \alpha}\left[1 - \epsilon^{-\alpha z}\right] , \tag{9A}$$

where $E_1$ is the initial electric field originating from beam noise and is considered to be negligible. When the resistive losses are assumed small, then eq. (9A) reduces to:

$$E = \frac{M\, b\, z\, I_L}{4\, K} \qquad (\text{ for } \alpha = 0 , \quad \text{and } E_1 = 0 ). \tag{10A}$$

Power output is determined from eq. (1A) with the value of E obtained from eq. (10A):

$$W_{out} = K\, E^2 = \frac{b\, \eta\, M^2\, I_L^2}{4\, a}\, \frac{1}{\sqrt{1-\left(\frac{f_c}{f}\right)^2}}\, \left[1 - \epsilon^{-\alpha\, z}\right]^2 . \tag{11A}$$

The factor M includes the usual transit-time "modulation coefficient" correction for klystrons equal to $\frac{\sin X}{X}$ where $X = \frac{\omega\, b}{2\, u}$ with u equal to the electron velocity in meters/second. Under ideal velocity-modulation conditions the ratio of the RF current to the DC beam current is 1.16 as shown in textbooks[7]. A decoupling factor should be added to allow for the imperfect fields at the slots. However, for the experiment described in this report, the factor is small. Consequently, the factor M is calculated to be:

$$M = 1.16\ \frac{\sin X}{X} , \quad \text{where } X = \frac{\omega\, b}{2\, u} , \tag{12A}$$

with u equal to the electron velocity in meters/second. This equation, together with eq. (11A), is used to calculate the output power.

The presence of the electron beam causes beam loading which can attenuate the output signal. This beam loading acts in the same way as the resistive attenuation $\alpha$ in the waveguide. It is neglected here because of the very small beam current density.

The above analysis is consistent with the analysis of Mihran[2] if it assumed that he included a b/a factor in his definition of waveguide impedance. More detailed analysis of the high-gain and high-efficiency sections of distributed amplifiers are given in references 3, 4, 5, and 6.

STANDING-WAVE KLYSTRON:

In this case the slotted $TE_{10}$ output waveguide is shorted at each end. Two traveling waves of equal amplitude moving in opposite directions exist on the waveguide. It is assumed that the incoming electrons are properly phased and bunched to enhance the standing-wave pattern along the waveguide. The structure is resonant at frequencies where the waveguide length is an integral number of half wavelengths. These resonant frequencies are:

$$f_n = f_c \sqrt{1 + \frac{a^2 n^2}{d^2}} \tag{13A}$$

where $f_n$ is the resonant frequency, n is the integral number of half wavelengths, d is the length of the slotted waveguide in meters, and a is the width of the guide. In the calculations that follow it is assumed that the frequency of operation is at one of these resonant frequencies.

The slotted waveguide shorted at each end forms a resonant cavity. By definition, the "Q" of this resonant cavity is:

$$Q = \frac{\omega_0 U}{W_{LL}}, \tag{14A}$$

where $\omega_0$ is the angular resonant frequency, U is the stored energy of the cavity, and $W_{LL}$ is the total resistive loss in the cavity walls including the end walls. The total stored energy U of the resonant cavity, on integration becomes:

$$U = \frac{\epsilon_0 a b d E^2}{8}. \tag{15A}$$

The loss $W_{LL}$ is obtained by integration of the resistive losses in the cavity walls including the end walls:

$$W_{LL} = \frac{R_s E^2}{4\,\eta^2}\,\frac{1}{(n^2a^2 + d^2)}\left\{ a\,d\left[ n^2 a^2 + d^2 \right] + 2\,b\left[ n^2 a^3 + d^3 \right] \right\}, \tag{16A}$$

and the "Q" can be obtained by substituting eqs. (15A) and (16A) into (14A):

$$Q = \frac{\pi\eta}{4\,R_s}\;\frac{2\,b\left\{ n^2 a^2 + d^2\right\}^{3/2}}{a\,d\left\{n^2 a^2 + d^2\right\} + 2\,b\left\{n^2 a^3 + d^3\right\}} \tag{17A}$$

The "Q" value given by eq. (17A) is the "unloaded Q" of a fixed length d of a waveguide having different resonant frequencies determined by n. The output power coupled to this resonant cavity plus the beam loading loss reduce the "unloaded Q" to a "loaded Q" value of about one half of the "unloaded Q". The large amount of stored energy can cause the "Q" to be very high. The equivalent shunt resistance of a parallel resonant circuit (as used in conventional klystron cavities) has no useful meaning in this situation of a cavity that is several wavelengths long.

Our objective is to determine the amount of power delivered by the bunched electron beam to the output slotted waveguide and to the coupled output load. Input power causes the electric fields within the resonant cavity to increase. These fields continue to increase until the losses plus output power equals the input power. The ideally bunched electrons move through the slots against the electric field and give up an energy of E b electron volts. The power input is the square of this voltage divided by an equivalent shunt resistance as seen by the electron beam. Thus:

$$\text{Power input} = \frac{E^2\,b^2}{2\,R}, \tag{18A}$$

where R is the equivalent shunt resistance as seen by the electron beam. The factor of 2 converts peak to RMS values. The output power plus the cavity losses (neglecting beam loading) is :

$$\text{Total output power} = 2\,W_{LL} = \text{Power input} = \frac{E^2\,b^2}{2\,R}\,. \tag{19A}$$

The factor of 2 assumes that the output load is critically coupled to the cavity, causing the load power loss to equal the cavity losses (Critical coupling is the coupling of maximum power transfer where the resistive loss is matched to the output loss). When eq. (17A) is substituted into (20A), one obtains:

$$R = \frac{E^2 b^2}{4 W_{LL}} = \frac{\eta^2 b^2}{R_s} \frac{\left(n^2 a^2 + d^2\right)}{a d \left(n^2 a^2 + d^2\right) + 2 b \left(n^2 a^3 + d^3\right)} \quad . \quad (20A)$$

The output power is:

$$W_{out} = \frac{1}{2} M^2 I_L^2 d^2 R \quad , \quad (21A)$$

where R is given by eq. (20A). The factor of 1/2 accounts for the lack of interaction at the standing-wave nodes. Combining eqs. (20A) and (21A), one obtains:

$$W_{out} = \frac{1}{2 R_s} \frac{M^2 I_L^2 d^2 \eta^2 b^2 \left(n^2 a^2 + d^2\right)}{a d \left(n^2 a^2 + d^2\right) + 2 b \left(n^2 a^3 + d^3\right)} \quad . \quad (22A)$$

Critics would like to determine the shunt resistance by means of the circuit "Q". This could be done but the stored energy, used in the deterimation of "Q", would have to be included and then extracted. It is more direct to deal only with the losses.

## APPENDIX B

## BACKWARD CURRENTS IN THE RIBBON-BEAM TRAVELING-WAVE KLYSTRON

In the traveling-wave klystron a continuous sheet electron beam has been velocity modulated by a traveling-wave electron-beam slotted-waveguide buncher similar to klystron bunching cavities. The bunched electron beam enters the output slotted waveguide, and encounters equal impedance in both directions. The current therefore splits equally and produces both a forward and a backward RF current component in the output waveguide.

Let J be the electron-beam current per unit length (linear density), and let kJ be the RF current per unit length induced in the output waveguide. At any one point along the output line the elemental current (k J dl) entering the slotted ouput waveguide divides equally in the forward and backward directions. The forward current is:

$$I_{FW} = \frac{1}{2} k J \int_0^L dl = \frac{1}{2} k J L \tag{1B}$$

where L is the length of the output line.

The backward current in the output waveguide is a result of forward signal from the input line. Therefore, there is a forward phase delay of $\beta l$ in the signal line plus an additional backward phase delay of $\beta l$ in the output line or a total of $2\beta l$ where $\beta$ is the phase shift per unit length in the waveguide. Then:

$$I_{BW} = \frac{1}{2} k J \int_0^L e^{-j2\beta l}\, dl \tag{2B}$$

$$|I_{BW}| = \frac{1}{2} k J \frac{\sin\beta L}{\beta} \tag{3B}$$

$$\frac{P_{BW}}{P_{FW}} = \frac{|I_{BW}|^2 Z_L}{|I_{FW}|^2 Z_L} = \left[\frac{\sin\beta L}{\beta L}\right]^2 \tag{4B}$$

Thus, the ratio of the backward-to-forward power decreases to zero at $\beta l = \pi$, and thereafter the rise is less than 5 percent. The value, $\beta l = \pi$ corresponds to a line length L of $\lambda/2$ . Therefore, at line lengths greater than one-half wavelength, the backward currents approach zero and can be neglected. In the next paragraph it is shown how the backward current can be redirected to become forward current.

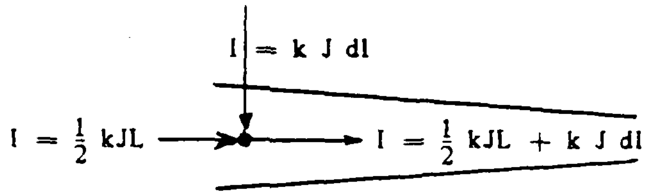



*Figure 1A. Tapered waveguide redirects backward currents.*

The electron beam acts as a constant-current source pumping current and power into the output line. If the output waveguide height is tapered with a decreasing impedance, then a greater proportion of forward current is required to maintain a continuous voltage. In Figure 1A the input and output currents at the tapered junction are shown that force the backward current to become forward current. The voltage across the guide must be continuous:

$$V = \frac{1}{2} kJL\, Z_0 = \left( \frac{1}{2} kJL + k\, J\, dl \right) Z_n \qquad (5B)$$

$$\frac{Z_n}{Z_0} = \frac{L}{L + dl} \qquad (6B)$$

where $Z_n$ is the "new" waveguide impedance and $Z_0$ is the original impedance. (This refers to the waveguide impedance $Z_0 = \eta\left[1-\left(\frac{f_c}{f}\right)^2\right]^{-1/2}$ and not the shunt interaction impedance). Thus, if the taper is according to eq. (6A), and if it is assumed there is no mismatch because of the taper, then all of the RF currents will move forward, and there will be no backward currents. This is an important result in that the output current will be doubled and the output power and efficiency will increase approximately by a factor of four. This concept was used in a multiple-beam distributed-amplifier klystron as described in reference (4).