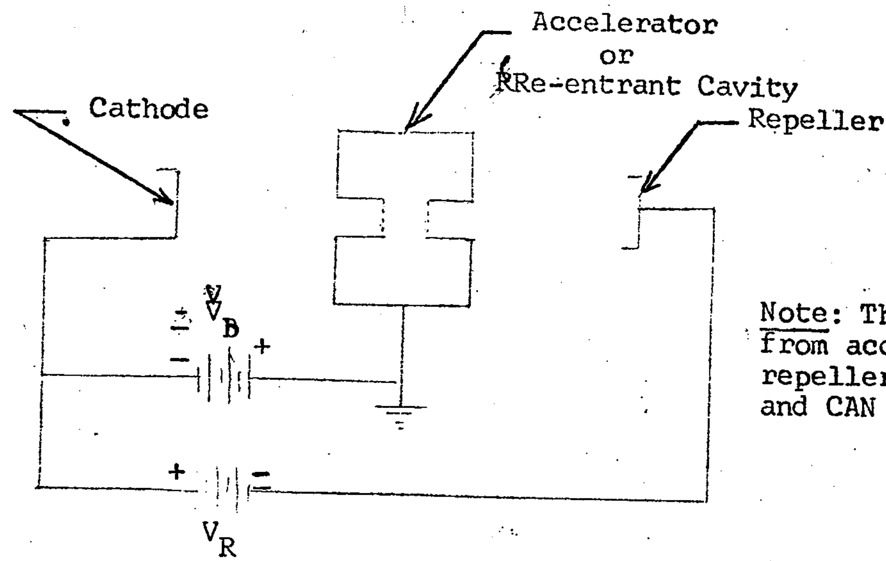
MICROWAVE EXPERIMENT
PROCEDURE

WEEK 1 - Reflex Klystron Operation

During the first week you will operate a reflex klystron which will be the source of X-band (8.2-12.4 GHz) microwaves used in the experiment. You will measure its relative output power as a function of frequency and become familiar with the range of operation possible with this 2k25 tube.

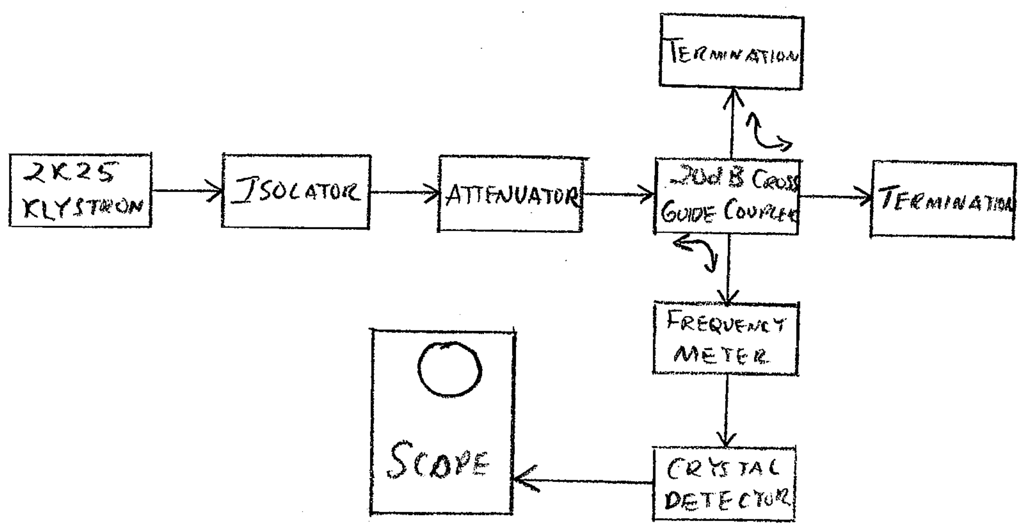
Simply described, the reflex klystron makes use of the physics of electrons under the influence of an accelerating field and a reflecting field, interacting with a resonant cavity to produce a signal of very high frequency. Referring to Fig. 1-1, where a simple klystron is shown schematically, it can be seen that electrons are drawn by the positive potential of the accelerator into the region of the reflector, where they are repelled back. During this motion the electrons pass through the center of the resonant cavity twice, where they lose energy if they are decelerated. Now consider an alternating field in the cavity such that it is directed to decelerate the electrons during their transition through the opening. Such a signal will be enhanced, i.e., will absorb energy from the electron beam and an oscillation can develop at the frequency with which the electrons pass through the cavity, providing that the cavity is tuned to this frequency. So it can be seen that the frequency of osciallation is dependent on cavity setting, while the power output is dependent on proper adjustment of V_R and V_B.



Note: The voltages from accelerator to repeller are in series and CAN BE DANGEROUS.

Fig. (1-1) Schematic of reflex Klystron

When you come to class the first week the experiment will be set up as shown below:



Review the Klystron Turn-On Procedure given on p.5 of the introductory material and go over proper operation and safety with the instructor before beginning

1. Turn-on klystron and vary the repeller voltage until power is detected by the crystal detector output shown on the oscilloscope.

2. Measure the output frequency with the frequency meter.

3. Map out the repeller voltages which produce microwave power (do not go below -50 volts!)

4. Use the sawtooth generator to continuously display the mode structure of the klystron output.

5. Measure output frequency vs. repeller voltage within one mode.

6. Determine the approximate $\cancel{\frac{\Delta\omega}{\omega_o}}$ $\omega_o/\Delta\omega$ of a mode to estimate the klystron cavity Q.

7. Vary the size of the klystron cavity and find the value of both the highest and lowest frequency of operation for this 2k25 tube.

WEEK 2 - Propagation of Microwaves in a Waveguide and Measurement of SWR

Review the Introduction to Waveguide Theory in the introductory material. We will be using the TE_{10} mode in the guide.

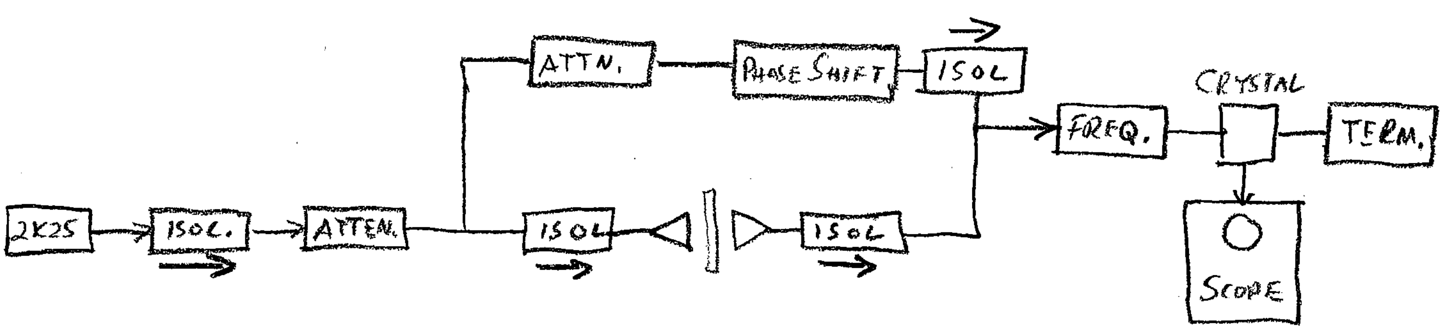
1) Using the "Slotted Guide Section," map out the standing wave pattern inside the wave guide and determine the wave length of waves propagating inside the wave guide. Use an open guide terminated with aluminum foil to generate a strong SWR pattern.

2) Vary the operating frequency over the full range of the tube and map out the dispersion characteristics ($\frac{\omega}{K}$ vs. ω) of the guide. Compare with the theoretical results.

3) Terminate the open guide with various materials (air, BN_3, polyethylene, etc.) and measure the SWR from each. Calculate the reflection coefficient and use this to compute the change in the index of refraction from guide to material. This can provide a crude measure of the index of refraction of these materials for microwaves.

WEEK 3 - Interferometer Measurements of the Index of Refraction

Set up a simple fringe shift interferometer which will allow you to insert thin samples into the measurement arm:



Measure the index of refraction for the supplied materials for two different frequencies.

AN INTRODUCTION TO MICROWAVE TECHNIQUES

APPLIED PHYSICS
LAB

CHAPTER 3

GENERAL INFORMATION

I. Introduction

The essential feature of the microwave region on the electromagnetic spectrum is, as the name implies, their short wavelengths. Typically these lengths are comparable in orders of magnitude to laboratory equipment dimensions. (In the microwave region the frequency varies between one to hundreds of GHz.) This means that neither the low frequency (KVL, KCL, lumped elements) nor the high frequency (geometric optics) approximations are normally valid substitutes for Maxwell's Equations. However, both approximations are frequently useful when appropriately modified, and help to develop physical understanding and engineering design procedures.

Because of their high frequency resultant bandwidth and quasi-optical properties, microwaves have been extremely useful in engineering practice. Uses include all classes of radar (weather, airport surveillance, civilian and military, shipborne and airborne), telecommunications, long-distance telephone and mobile television transmission, etc.

Another important application is in the area of radio astronomy.

Many materials have important properties which are observable in the microwave region. Molecular rotational energy levels are separated by energies corresponding to microwave frequencies. In convenient laboratory magnetic fields (~ 1000 gauss) the free electron gyration frequency (cyclotron frequency) is in the microwave region, as is the characteristic frequency of spin transitions.

II. Laboratory Equipment

The following list is a summary of laboratory equipment that will be used in the microwave experiments, with short descriptions on the methods of their operation. For more thorough discussion, the student is referred to the given references.

1. Wave guides: Rectangular copper (or brass) tubing used to guide microwave signals from point to point. The theory of propagation in such waveguides was introduced in EE E3401; notes summarizing the theory are provided (Chapter 2). It is important to realize that waveguides are constructed to conform to close dimensional tolerances and should not be abused.

2. Klystron: Special type of vacuum tube whose operation will be described at the first session. A Klystron generates or amplifies coherent electromagnetic energy at microwave frequencies and will be studied in the first experiment. Because of its method of operation it requires a special type of power supply.

3. Klystron Power Supply: Similar in operation to power supplies (voltage sources) the student is already acquainted with, it is, however, designed to supply all the electrode potentials necessary to operate the klystron tube. It also supplies appropriate a.c. modulation for utilizing the klystron most efficiently. This modulation is applied to the reflector of the tube and causes the potential of the reflector to vary according to the type of modulation applied. There is an output jack provided on all power supplies for observing the modulation on the scope.

4. Crystal Detector: Contained in tuned detectors and slotted line probes, it rectifies the sine wave signal, and measuring devices such as scope, VTVM, etc. which do not respond to signals at gigahertzian frequencies, show the average value (proportional to the envelope) of the rectified signal. Thus a C.W. signal would appear on the scope as a straight horizontal line somewhat displaced from the ground line (unless the scope is A.C. coupled, in which case nothing would show). Also, the A.C. voltmeters used in the laboratory would not indicate the signal (only changes in it). Thus, C.W. is not usually used with crystal detectors.

5. Attenuators: By inserting material into a waveguide which is lossy at microwave frequencies, it is possible to reduce the level of the signal transmitted through (past) the guide. Such devices are similar in purpose to potentiometers used at low frequencies, but find wide use in

high frequency work since it normally is not possible to regulate the power output of a generator without affecting most of its other charactteristics, such as frequency and stability, and since low-frequency potentiometers do not work at high frequencies.

6. Couplers: In distributing microwave signals to a number of different circuits, couplers are used. These consist of two waveguides with a common wallin which slots or holes have been machined so that a fixed amount of signal propagating in one guide is coupled into the other. These devices are designed so that they are unilateral and do not have the same effect on signals traveling in opposite directions. Their operation will be described more fully in class.

7. Frequency Meter: This device is composed of a cylindrical cavity the depth of which may be varied; it is coupled into a waveguide section through a small hole called an iris. When the frequency of the signal in the guide is at the frequency of resonance of the cavity (determined by the dimensions of the cavity), the cavity resonates and absorbs energy from the guide. Detecting the signal bypassing the iris makes it possible to determine the signal frequency in the following manner: vary the dimensions of the cavity until a sharp dip in the detected signal occurs. This indicates the cavity is at resonance and absorbing energy from the guide. The frequency of the signal can be determinémined by the set-in of the cavity (the frequency meters used in the laboratory are calibrated and the frequency is read directly in gigaHertz).

8. Slotted Section: This is a section of waveguide with a slot along the center of one of the broad walls, so that a probe may be inserted into the guide to sample the field. For efficacious measurements, it is important that the probe does not perturb the fields to any great extent; thus care should be exercised to insure that the probe is inserted no deeper than required to obtain a reasonable signal from a "crystal detector" mounted inside the probe.

The probe, which is a straight thin wire conductor "antenna", is mounted so that its position along the guide can be varied and its relative position may be read off a scale in centimeters. The crystal detector is a semiconductor diode which rectifies the current induced on the probe imbedded in the oscillating electromagnetic field; the average voltage across the crystal is then a measure of the energy density. A crystal detector can also be used in special mounts inside waveguides.

10. Termination (matched load): A microwave termination is a specially designed absorber which absorbs power incident on it with a minimum of reflection. It is normally placed at a point in the microwave system at which it is desired to terminate the waveguide without introducing reflections from the termination, so that the waveguide appears to be infinitely long, or terminated in its characteristic impedance.

III. Sources of Experimental Error

There are three common sources of error in microwave experiments. The first is due to the power and frequency versus load characteristics of a klystron oscillator. As can be shown experimentally, varying a component which affects the load impedance seen by the klystron will generally cause variations in the power and frequency delivered. This effect can usually be minimized by the insertion of a fixed or variable attenuator of at least 3 db between the klystron and the system under test.

A second common error is associated with the probe used for detection in a waveguide. Since the probe is a conductor it is evident that any insertion into the waveguide will perturb the desired fields. In order to reduce these effects, insertion of any probe should be minimized to that required for a reasonable signal to noise ratio in the detector.

A third source of error is due to damaged waveguide hardware. Care should be exercised by the student in handling all equipment since damaged components will perturb the desired fields and consequently reduce experimental accuracy.

IV. Instructions for Using Klystron and Power Supply

A. Turn-on Procedure. To protect the klystron, its supply and its user, keep the following order, except that #3 can be done at any time.

1. Before turning on power supply, make sure that the cable is securely connected between power supply and klystron, that all equipment is properly grounded, that the beam voltage is off, and that the reflector voltage knob is set several hundred volts negative (at least -350 V, or more negative).
2. Turn on line voltage. This supplies the klystron heater filament. Allow about one minute for warmup.

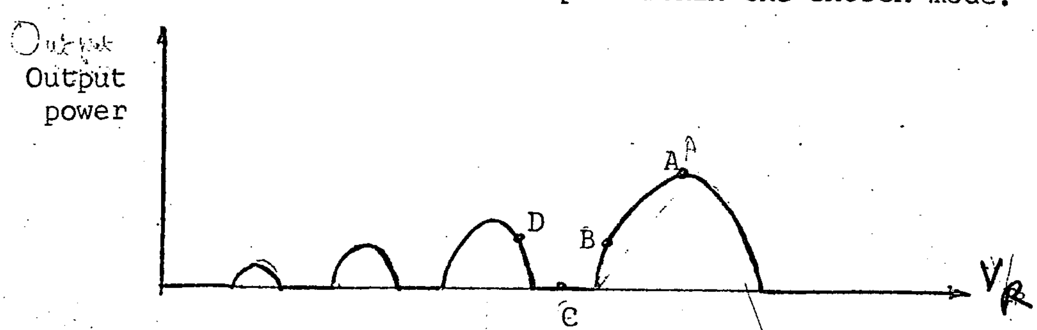
3. Set reflector modulation as desired. This may be changed at any time, as can the modulation amplitude and frequency. See B below.

4. Check that the beam voltage control is fully counterclockwise (ccw) and turn on the beam voltage.

5. Increase the beam voltage to appropriate value (usually 300 volts, but could be anywhere from 250 to 375). Check that the beam current is in proper range (25 - 30 mA for beam voltage of 300 volts). If the current is too high, decrease beam voltage and call instructor. [Note: the power supplies have built-in meters to read beam voltage and current.]

6. Increase reflector voltage (less negative) until output is obtained. <u>However, the reflector must be kept below -50 V</u> to avoid damage due to large currents (electrons must be adequately repelled). Normal operation occurs for values between -100 and -350 volts. See notes on modulation (B below) for discussion of detecting outputs.

7. You will notice that as the reflector voltage is varied the output will appear, disappear, and reappear, etc. This illustrates the various modes of operation of the klystron. See Figure G-1. Although any of these modes may be used, those for most negative V_R (-250 to -350 volts) are recommended. The reflector voltage should be set for the maximum output within the chosen mode.



Schematic of Output power vs Reflector Voltage for Klystron
Figure G-1

8. Adjust the screw on the klystron mount for maximum output. This moves a short circuit in the waveguide and hence affects the

9. impedance seen by the klystron, and will therefore change the shape and positions of the various modes in Figure G-1.

9. Repeat steps 7 and 8 until a final maximum is obtained.

B. Modulation and Detection

1. The CW (continuous-wave) modulation setting keeps the reflector voltage at a constant value, and hence allows the klystron to generate a continuous single-frequency sine wave.

2. The square wave modulation periodically switches the reflector voltage between its set value and another value which depends on the modulation amplitude. On some power supplies the frequency is 1000 Hertz (hence the name 1000 cycle modulation) while on others it may be varied between 400 and 2000 Hertz. If one value is at the center of a mode [point A in Figure G-1], while the other is at a voltage for which the klystron does not oscillate (between modes; e.g., point C), then the output will be a set of pulses of a high frequency sine wave. However, there is the danger of "double-moding" in which the two voltages are in two parts of the same mode [such as points A and B in Figure G-1] or on two modes [points A and D]. Such circumstances should be avoided not only because of reduced detected modulated output, but because the two signals are generally of different frequencies.

3. Sawtooth (on some power supplies called FM) modulation causes a continuous variation of reflector voltage. (On some power supplies this could be a sine wave modulation. Usually FM is at 60 Hertz). The same sawtooth modulation could be placed on the x-axis of an oscilloscope, so that the output as a function of V_R can be viewed directly. If the sweep amplitude is large, the output passes through the various modes, whose structure can thus be studied. If it is small (confined to one mode) it serves to modulate the klystron's frequency which is a function of V_R. This effect can be illustrated by turning a frequency meter and

watching the dip move through the mode as seen on the scope.

For square wave or sawtooth modulation, the envelope of the output would appear on the scope (use D.C. coupling to check on double-moding -- for single-mode operation, one part of the square wave should be at baseline level), while the R.M.S. value of the envelope would appear on A.C. voltmeters.

4. A thermistor responds to the heat effect of the incident wave. Hence, it will respond to any type of modulation, but it is most useful for C.W. or Pulsed applications -- the thermistor does not respond rapidly, and when connected to a power meter will yield the <u>average</u> power, calibrated in watts.

C. <u>Klystron Turnoff Procedure</u>

1. Set reflector voltage more negative than -350 volts.

2. Turn beam voltage control ccw and turn beam voltage off.

3. Turn off power supply.

VI. Miscellaneous Precautions

1. When assembling the waveguides, be careful not to scratch the flanges. Also, make sure all connections are tight.

2. When making power measurements, consult the instruction manual for your power meter. Disregard of the instructions may result in destruction of the thermistor.

3. Always make sure that directional couplers are properly oriented.

4. Note that the reflector voltage must always be kept negative. If the reflector voltage is allowed to become more positive than -50 volts, the reflector may be severely damaged.

5. The beam and reflector voltages are in series and can typically add to between 500 and 1000 volts. In the presence of dangerously high voltages, it is expected that mature electrical engineers will take necessary precautions to avoid contact with klystron electrodes and leads to power supply.

VII. References (Annotated)

1. Collin, R.L., Foundations for Microwave Engineering, McGraw Hill, 1966. Best all round reference for course. Up to date text, good format, etc. Generally contains clear discussions of difficult concepts as well as complete mathemataic

concepts as well as complete mathematics leaving out relatively few steps. Excellent on impedance and Smith Chart.

2. Ramo, S., Whinnery, J.R., and Van Duzer, T., Fields and Waves in Communication Electronics, Wiley, 1965; or Ramo and Whinnery, Fields and Waves in Modern Radio, Wiley 1944 and 1953. An intermediate-level text on electromagnetic theory, transmission lines, microwave networks,
3. and antennas. The Smith Chart is discussed in Section 1.20.

3. Reich, H.J., Ordung, P.F., Krauss, H.L., and Skalnik, J.G., Microwave Theory and Techniques, D. Van Nostrand, 1953, or Boston Technical Publishers 1965. Contains electromagnetic theory, theory of transmission lines, microwave networks, and antennas. Also discusses theory and operation of microwave generators (klystrons in Chap.10), resonators, and measuring instruments (Chap.8), and has a list of possible microwave laboratory experiments with procedures, keyed to the theory given elsewhere in the book.

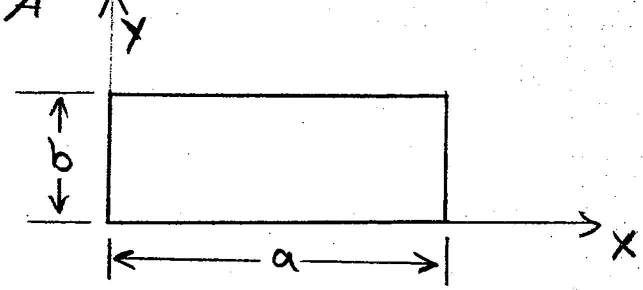
4. Schelkunoff, S.A., Electromagnetic Fields, Blaisdell, 1963, and notes distributed in E.E. E3401 and in this course. Basic theory of transmission lines and waveguides.

5. Millman, J., Vacuum Tube and Semiconductor Electronics, McGraw-Hill, 1958. The section on electron bunching (2-15) gives theory of klystron operation.

6. Ginzton, E.L., Microwave Measurements, McGraw-Hill, 1957. This is a good book on the devices studied in this course -- gives description of the design, theory, and operation of klystrons (Chap.1), crystal detectors (Chap.2), thermistors (Chap.3), etc., and discusses the methods of measurement of power (Chap.3), impedance (Chap.5), frequency (Chaps.7-8), etc. Chapter 4 has a discussion of the impedance concept for waveguides and transmission lines and on the use of the Smith Chart.

7. Montgomery, C.G., Technique of Microwave Measurement;
Montgomery, C.G., Dicke, R.H., and Purcell, E.M., Principles of Microwave Circuits;
Ragan, G.L., Microwave Transmission Circuits; M.I.T. Radiation Laboratory Series, vols. 11, 8, 9, McGraw Hill, 1947; or Boston Technical Publishers, 1964. These three books include much of what is known concerning the theory, design, operationaand application of microwave devices. The first includes microwave generators and devices to measure microwave power, frequency, and impedance. The second contains a review of electromagnetic waves and theoretical treatment of microwave circuits and elements including diaphragms, posts, cavities, dielectrics in waveguide, and microwave junctions. Chapter 3 has some discussion on waveguides as transmission lines and on the impedance concept for waveguides. The third book, mainly practical, considers waveguides, transmission lines, waveguide junctions, and filters -- the Smith Chart is discussed in Section 2.11.

MICROWAVE DATA



For TE_{no} modes in rectangular waveguides

$$\vec{E}_n(x,z,t) = E_n \sin\left(\frac{n\pi x}{a}\right) e^{i(\omega t - K_g z)}\ \hat{e}_y$$

$$\text{for} \quad \omega > \omega_c = c\left(\frac{n\pi}{a}\right)$$

$$\text{where} \quad \omega = 2\pi f \quad \text{and} \quad K_g = \frac{2\pi}{\lambda_g}$$

$$\lambda_{\text{guide}} = \frac{\lambda_{\text{free space}}}{\sqrt{1-\left(\frac{\omega_c}{\omega}\right)^2}}$$

$$V_{\text{phase in guide}} = \frac{\omega}{K_g} = \lambda_g f = \frac{c}{\sqrt{1-\left(\frac{\omega_c}{\omega}\right)^2}}$$

$$V_{\text{group in guide}} = \frac{d\omega}{dK_g} = c\sqrt{1-\left(\frac{\omega_c}{\omega}\right)^2}$$

$$\text{Attenuation in dB} = 10 \log_{10}\left(\frac{P_1}{P_2}\right)$$

$$SWR = \frac{\text{Power maximum}}{\text{Power minimum}} = \left[\frac{\sqrt{P_i}+\sqrt{P_r}}{\sqrt{P_i}-\sqrt{P_r}}\right]^2$$

[REMEMBER $V_{\text{DIODE DETECTOR}}$ IS PROPORTIONAL TO POWER $\propto |\vec{E}|^2$]

TRG Letter Designation	EIA WG Designation	JAN WG Designation	Recommended Operating Range for TE_{10} Mode Frequency (Gc.)	Wavelength (mm)	Cut-Off for TE_{10} Mode Frequency (Gc.)	Wavelength (mm)	Theoretical CW Power Breakdown Lowest to Highest Frequency (KW) *	Theoretical Attenuation Lowest to Highest Frequency (db/ft)	Material Alloy	Mating Flange Type **	Dimensions (Inches) Inside	Tol.	Nominal Wall Thickness
X	WR 90	RG 52/U	8.20–12.4	36.6–24.2	6.557	45.72	730–1100	.042–.030	Copper	UG 39/U UG 40A/U	.900 x .400	± .003	.050
K_u	WR 62	RG 91/U	12.4–18.0	24.2–16.6	9.486	31.60	440–600	.064–.047	Copper	UG 419/U UG 541/U	.622 x .311	± .0025	.040
K	WR 42	RG 53/U	18.0–26.5	16.6–11.3	14.047	21.34	160–240	.17–.11	Copper	UG 595/U UG 596/U	.420 x .170	± .0020	.040
A	WR 28	RG 96/U	26.5–40.0	11.3–7.5	21.081	14.22	95–145	0.22–0.15	Silver	UG 381/U UG 599/U UG 600/U	.280 x .140	± .0015	.040
B	WR 22	RG 97/U	33.0–50.0	9.1–6.0	26.342	11.38	62–90	0.31–0.21	Silver	UG 383/U	.224 x .112	± .0010	0.40
U	WR 19	—	40.0–60.0	7.5–5.0	31.357	9.56	47–64	0.39–0.27	Silver	—	.188 x .094	± .0010	.040
V	WR 15	RG 98/U	50.0–75.0	6.0–4.0	39.863	7.52	29–42	0.57–0.39	Silver	UG 385/U	.148 x .074	± .0010	.040
E	WR 12	RG 99/U	60.0–90.0	5.0–3.3	48.350	6.20	20–29	0.78–0.53	Silver	UG 387/U	.122 x .061	± .0005	.040
W	WR 10	—	75.0–110.0	4.0–2.7	59.010	5.08	14–20	1.02–0.71	Silver	—	.100 x .050	± .0005	.040
F	WR 8	RG 138/U	90.0–140.0	3.3–2.1	73.764	4.06	8.5–13.5	1.52–0.98	Silver	Special	.080 x .040	± .0003	.030
D	WR 7	RG 136/U	110.0–170.0	2.7–1.8	90.786	3.30	5.8–9.0	2.12–1.35	Silver	Special	.065 x .0325	± .00025	.030
G	WR 5	RG 135/U	140.0–220.0	2.1–1.4	115.71	2.59	3.7–6.1	3.05–1.93	Silver	Special	.051 x .0255	± .00025	.030
H	WR 4	RG 137/U	170.0–260.0	1.8–1.2	137.24	2.18	2.8–4.6	3.75–2.50	Silver	Special	.043 x .0215	± .00020	.030
J	WR 3	RG 139/U	220.0–325.0	1.4–0.92	173.56	1.73	2.0–3.3	5.09–3.56	Silver	Special	.034 x .017	± .00020	.030

* Based on standard atmospheric pressure and temperature

**) Cover and Choke listed for A-band and Lower Frequencies Contact for higher Frequencies

REFERENCE TABLE OF CIRCULAR WAVEGUIDE, TE_{01} MODE
(TRG Standard Copper Circular Waveguides)

TRG Letter Designation	Inside Diameter (Inches)	Recommended Operating Range Frequency (Gc)	Theoretical CW Power Breakdown, Lowest to Highest Frequency (KW)	Theoretical Attenuation Lowest to Highest Frequency (db/ft)
A	.634	26.5–40.0	770–1,110	0.050–0.018
B	.527	33.0–50.0	580–870	0.061–0.022
V	.353	50.0–75.0	270–390	0.105–0.039
E	.291	60.0–90.0	180–260	0.145–0.054
W	.250	75.0–110.0	150–200	0.145–0.062
F	.180	90.0–140.0	55–99	0.41–0.12
G	.123	140.0–220.0	31–47	0.56–0.18

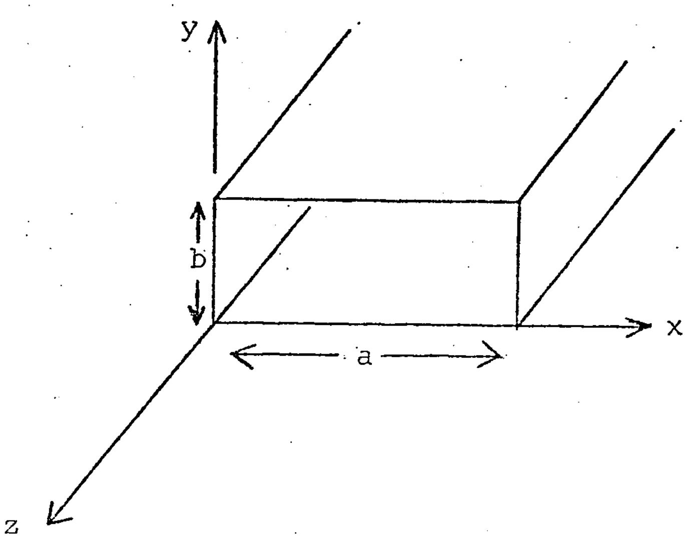
Introduction to Waveguide Theory

As with all electromagnetic waves, microwaves are propagating electromagnetic fields. Such fields, whether in free space, in various media, or in waveguides, must satisfy Maxwell's Equations. These equations have an infinite variety of possible solutions of many different types.

In this set of note, the particular waves which propagate in the waveguides used in the microwave laboratory will be derived from first principles, using the separation of variables technique to solve the partial differential form of Maxwell's Equations. Certain properties of these "transverse electric" waves will also be derived. It will then be shown that for a certain frequency range, in which only one transverse-electric wave propagates, the waveguide operates like a transmission line. Thus, in the study of reflections and impedance in waveguides, it is possible to greatly simplify what would otherwise be a fields analysis by making use of conventional transmission-line techniques and voltages.

General Equations

Let us consider a rectangular waveguide with a uniform cross-section as shown.



This type of structure is capable of supporting propagating electromagnetic fields. Maxwell's equations for source-free space ($\vec{J} = 0$) and time harmonic fields are:

<u>Ampere-Maxwell Law</u> $\oint \vec{H}\cdot\vec{d\ell} = j\omega \iint_S \vec{D}\cdot\vec{dS} \Longleftrightarrow \nabla \times \vec{H} = j\omega\varepsilon\vec{E}$

<u>Faraday-Maxwell Law</u> $\oint \vec{E}\cdot d\ell = -j\omega \iint_S \vec{B}\cdot d\vec{S} \Longleftrightarrow \nabla \times \vec{E} = -j\omega\mu\vec{H}$

(Note: the integral equations are completely equivalent to the differential equations as can be shown by applying Stokes theorem to the above integral equations.)

In rectangular coordinates, the two vector equations become the six scalar equations:

$$\frac{\partial H_z}{\partial y} - \frac{\partial H_y}{\partial z} = j\omega\varepsilon_o E_x \quad (1) \qquad \frac{\partial E_z}{\partial y} - \frac{\partial E_z}{\partial z} = -j\omega\mu_o H_x \quad (4)$$

$$\frac{\partial H_x}{\partial z} - \frac{\partial H_z}{\partial x} = j\omega\varepsilon_o E_y \quad (2) \qquad \frac{\partial E_x}{\partial z} - \frac{\partial E_z}{\partial x} = -j\omega\mu_o H_y \quad (5)$$

$$\frac{\partial H_y}{\partial x} - \frac{\partial H_x}{\partial y} = j\omega\varepsilon_o E_z \quad (3) \qquad \frac{\partial E_y}{\partial x} - \frac{\partial E_x}{\partial y} = -j\omega\mu_o H_z \quad (6)$$

where ε and μ have been replaced by their free space values ε_o and μ_o.

We will now consider the case in which the excitation is such that there is no variation in the fields as y changes, i.e., $\partial/\partial y = 0$ (we thus limit ourselves to consideration of modes with zero eigenvalues for y-variations). Equations 1-6 are then uncoupled into two sets of equations, each set involving only its own components:

$$\frac{\partial H_y}{\partial z} = -j\omega\varepsilon_o E_x \quad (1a) \qquad \frac{\partial E_y}{\partial z} = j\omega\mu_o H_x \quad (4a)$$

$$\frac{\partial H_y}{\partial x} = j\omega\varepsilon_o E_z \quad (3a) \qquad \frac{\partial E_y}{\partial x} = -j\omega\mu_o H_z \quad (6a)$$

$$\frac{\partial E_x}{\partial z} - \frac{\partial E_z}{\partial x} = -j\omega\mu_o H_y \quad (5a) \qquad \frac{\partial H_x}{\partial z} - \frac{\partial H_x}{\partial x} = -j\omega\varepsilon_o E_y \quad (2a)$$

The equations on the left are for the so-called TM modes (transverse magnetic, since the only component of magnetic intensity, H_y, is in the x-y plane which is perpendicular to the direction of propagation) and contain only the components E_x, E_z, H_y. Similarly the others are called TE mode equations (transverse electric, since the only electric intensity is in the plane perpendicular to the direction of propagation) and contain only the components H_x, H_z, E_y. The equations are said to be uncoupled since the variables in one set of equations do not depend on those in the other set. It should be noted that although setting $\frac{\partial}{\partial y} = 0$ made this decoupling most evident, it is not a necessary condition. A more complete derivation still yields orthogonal TE and TM modes.

Transverse Electric Modes

Now let us calculate the normal modes (eigenfunctions) of the TE equations (the "dominant" mode is a TE mode).

$$\frac{\partial}{\partial z}\ [\text{equation 4a}] \Rightarrow \frac{1}{j\omega\mu_o}\frac{\partial^2 E_y}{\partial z^2} = \frac{\partial H_x}{\partial z}$$

$$\frac{\partial}{\partial x}\ [\text{equation 6a}] \Rightarrow \frac{-1}{j\omega\mu_o}\frac{\partial^2 E_y}{\partial x^2} = \frac{\partial H_z}{\partial x}$$

Substituting into equation 2a yields

$$\frac{\partial^2 E_y}{\partial x^2} + \frac{\partial^2 E_y}{\partial z^2} = -\omega^2\mu_o\epsilon_o E_y \equiv -\beta^2 E_y \tag{7}$$

where $\beta^2 = \omega^2\mu_o\epsilon_o$.

Using the technique of separation of variables:

$$E_y(x,z) = X(x)Z(z)\ ^*$$

yields

$$X''Z + Z''X = -\beta^2 XZ$$

or

$$\frac{X''}{X} + \frac{Z''}{Z} = -\beta^2 \tag{8}$$

---

*[In general, $E_y(x,z,t) = X_{(x)}Z_{(z)}T_{(t)}$, but here, $T_{(t)} = e^{j\omega t}$, and of course is suppressed.]

Since X''/X is dependent only on x and Z''/Z is dependent only on z these two expressions must be constant ($-k^2$ and Γ^2 are separation constants):

$$\frac{X''}{X} = -k^2 \Rightarrow X(x) = A \sin kx + B \cos kx$$

and

$$\frac{Z''}{Z} = \Gamma^2 \Rightarrow Z(z) = Ce^{\Gamma z} + De^{-\Gamma z}$$

From equation (8) we obtain the "separation equation"

$$-k^2 + \Gamma^2 = -\beta^2$$

so that

$$\Gamma = \sqrt{k^2 - \beta^2} = \sqrt{k^2 - \omega^2 \mu_o \epsilon_o}$$

Hence, the solution for $E_y(x,z)$ is

$$E_y(x,z) = (A \sin kx + B \cos kx)(Ce^{\Gamma z} + De^{-\Gamma z}) \qquad (9)$$

Boundary Conditions

The tangential electric intensity across a boundary must be continuous. Assuming perfectly conducting walls, the component $E_y\Big|_{x=0} = E_y\Big|_{x=a} = 0$ since E_y must be $= 0$ in the perfectly conducting wall.

$$E_y(0,z) = 0 \Rightarrow B \cos(0)\ (Ce^{\Gamma z} + De^{-\Gamma z}) = 0 \text{ for all } z \Rightarrow \underline{\underline{B = 0}}$$

$$E_y(a,z) = 0 \Rightarrow A \sin ka\ (Ce^{\Gamma z} + De^{-\Gamma z}) = 0 \text{ for all } z$$

For a non-zero solution, $\sin ka = 0$ so $ka = n\pi \quad n = 0, 1, 2, \ldots$

$$\therefore \Gamma_n = \sqrt{\left(\frac{n\pi}{a}\right)^2 - \omega^2 \mu_o \epsilon_o} \qquad (10)$$

(If we had not required $\frac{\partial}{\partial y} = 0$, we would have found

$$\Gamma_n = \sqrt{\left(\frac{n\pi}{a}\right)^2 + \left(\frac{m\pi}{b}\right)^2 - \omega^2 \mu_o \epsilon_o}$$

Furthermore, we would have had to write $E = X(x)Y(y)Z(z)$ and would have found $Y(y) \sim A'\sin k'y + B'\cos k'y$ and boundary conditions implied $k'b = m\pi$. We have thus chosen $m=0$ by taking $\frac{\partial}{\partial y} = 0$.)

Let us now consider the physical interpretation of the functional dependence on the $\hat{z}$ coordinate.

<u>Case I</u>. $\Gamma_n = j\beta_n$, pure imaginary. $\omega^2\mu_o\epsilon_o > \left(\frac{n\pi}{a}\right)^2$. Then,

$$Z(z) = Ce^{j\beta_n z} + De^{-j\beta_n z}.$$

But for time harmonic fields, the time dependence is of the form $T(t) = e^{j\omega t}$. Hence

$$T(t)Z(z) = Ce^{j(\omega t+\beta_n z)} + De^{j(\omega t-\beta_n z)} \tag{11}$$

These represent wave propagation in the negative z and positive z directions respectively. This can be seen from the following arguments.

1. As ωt increases, $\beta_n z$ must decrease for the phase of the first term to remain constant. Hence, the wave must be travelling in the -z direction since we assume $\beta_n > 0$.
2. For the second term, as ωt increases, $\beta_n z$ increases to maintain constant phase; therefore the wave is travelling in the +z direction.

We will only consider the case of a wave travelling in the +z direction. It is clear that a wave could also be propagating in the -z direction and since the system is linear, we would be able to superimpose the two in order to form the complete solution. Thus,

$$\text{for } \Gamma_n = j\beta_n, \text{ e.g., } \omega^2\mu_o\epsilon_o > \left(\frac{n\pi}{a}\right)^2,$$

$$E_{y_n}(x,z,t) = E_n \sin\frac{n\pi x}{a}\, e^{j(\omega t-\beta_n z)}$$

<u>Case II</u>. Γ_n = real -- $\omega^2\mu_o\epsilon_o < \left(\frac{n\pi}{a}\right)^2$

$$\text{for } \Gamma_n \text{ real, } Z(z) = Ce^{\Gamma_n z} + De^{-\Gamma_n z}$$

If the waveguide extends indefinitely in the +z direction, we must have C = 0 since we cannot allow the solution to be infinite at $z = \infty$. Similarly, for waveguides extending to $z = -\infty$, we must have D = 0. The resulting waves are exponentially damped in z , of the form $e^{\pm\Gamma_n z}$.

Hence, the complete solution is:

for propagating waves:

$$E_{Y_n}(x,z,t) = E_n \sin\frac{n\pi x}{a}\, e^{j(\omega t-\beta_n z)} \qquad \omega^2\mu_o\epsilon_o > \left(\frac{n\pi}{a}\right)^2 \tag{12a}$$

for exponentially damped waves (z > 0):

$$E_{Y_n}(x,z,t) = E_n \sin\frac{n\pi x}{a}\, e^{-\Gamma_n z} e^{j\omega t} \qquad \omega^2\mu_o\epsilon_o < \left(\frac{n\pi}{a}\right)^2 \tag{12b}$$

The frequency defined by $\omega_c^{\,2} = \left(\frac{n\pi}{a}\right)^2/\mu_o\epsilon_o$ is called the "cutoff frequency" and is an important property of guided waves. It is the frequency below which waves will not propagate. (Can you see a physical reason for ω_c suggested by its value? Is there an upper frequency limit for guided wave propagation?)

Each n represents a possible solution or "mode of propagation." These TE modes are denoted TE_{no}.

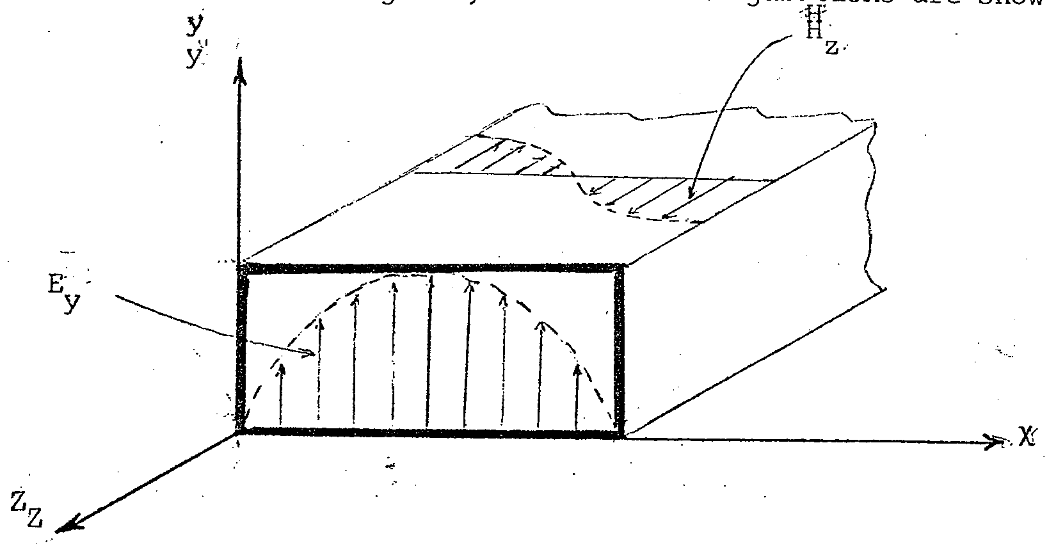
<u>Example</u> Consider TE_{10} waves (n = 1)

$$E_y = \cancel{E_1 \sin} \frac{\pi x}{a}\, e^{j(\omega t - \beta_{10}z)}$$

$$H_z = \frac{-\pi/a\, E_1}{j\omega\mu_o}\cos\frac{\pi x}{a}\, e^{j(\omega t - \beta_{10}z)} \tag{13}$$

$$H_x^x = \frac{-\beta_{10}E_1}{\omega\mu_o}\sin\frac{\pi x}{a}\, e^{j(\omega t - \beta_{10}z)}$$

Consider a section of waveguide; the field configurations are shown:

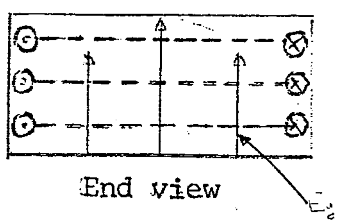


A further examination of the field equations for the TE_{10} mode (equations 13) indicates:

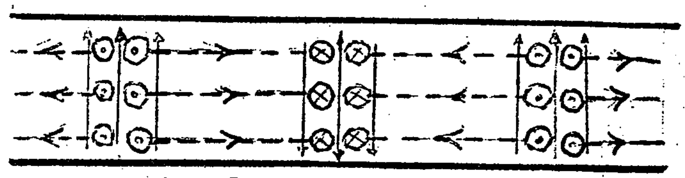
(a) Charges are zero on side walls ($E_z = E_x = 0$).

(b) Charge distribution on top and bottom corresponding to E_y ($\rho_s = \epsilon E_y$ Coulomb/m^2).

These charges causes an electric displacement current, which in turn will be surrounded by magnetic fields in the form of closed paths, so that there are components H_x and H_z.

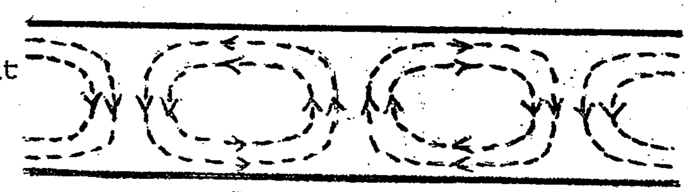


End view



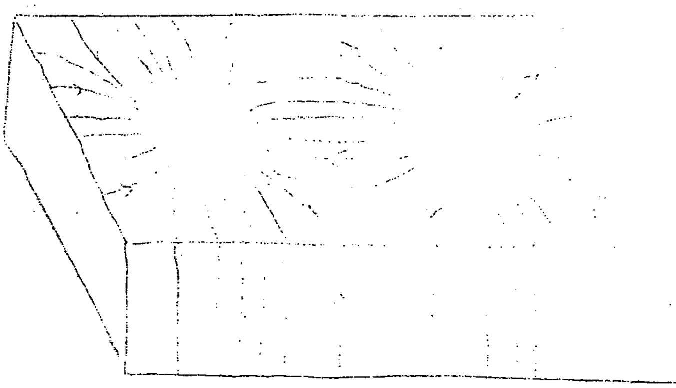
Front-side view

Dashed lines represent Magnetic Fields Surrounding Electric Displacement Current



Top view

Component H_x corresponds to a longitudinal surface current flow down the guide on the top and its opposite on the bottom, in agreement with the boundary condition $J = n \times H$. Similarly H_z corresponds to a current flow arround the periphery of the guide; these are sketch in the figure below.



Currents Flow in Walls for TE_{10} Mode.

The dependence on x of H_x is the same as that of E_y. Note that

$$E_y\Big|_{\substack{x=0\\x=a}} = 0$$

$$H_z\Big|_{x=a/2} \equiv 0$$

$$E_y \text{ is maximum at } x = a/2$$

For waves of the TE_{10} mode to propagate, we must have

$$\omega^2 > \left(\frac{\pi}{a}\right)^2 / \mu_o \epsilon_o \ .$$

Since the frequency ω is related to the free-space wavelength λ through

$$\omega \quad \mu_o \epsilon_o \equiv \beta = \frac{2\pi}{\lambda} \tag{14}$$

this equation becomes

$$\left(\frac{2\pi}{\lambda}\right)^2 > \left(\frac{\pi}{a}\right)^2 / 1$$

which reduces to

$$\lambda/2 < a \qquad .$$

which reduces to

$$\lambda/2 < a$$

For the TE_{20} (n=2) mode to propagate, we must have

$$\omega^2 > \left(\frac{2\pi}{a}\right)^2 / \mu_o \epsilon_o \quad \text{or}$$

$$\left(\frac{2\pi}{\lambda}\right)^2 > \left(\frac{2\pi}{a}\right)^2 \quad \text{or } \lambda < a.$$

It is easy to see that for a given "a" dimension, all TE_{no} modes are cutoff (do not propagate) if $\lambda > 2a$, while if $2a > \lambda > a$, then the TE_{10} mode will propagate but the TE_{no} for $n > 1$ are all cutoff. It can also be shown that the cutoff frequencies for TE_{nm} modes for $m \neq 0$ (for fields <u>not</u> satisfying $\frac{\partial}{\partial \lambda} = 0$) are greater than that for the TE_{20} mode; the same is true for transverse magnetic modes.

Thus, when the waveguide is excited with electromagnetic energy in the frequency range defined by $2a > \lambda > a$ (called <u>dominant-mode</u> operation), only one mode can propagate. This TE_{10} mode is called the "dominant" mode.

Dispersion Characteristics

In free space (and also in elementary transmission lines), the frequency $f = \frac{\omega}{2\pi}$ and wavelength λ are related through

$$\lambda \equiv \frac{2\pi}{\beta} = \frac{c}{f} \qquad (15)$$

where $c \equiv 1/\sqrt{\mu_o \epsilon_o}$ is the speed of light. Physically, this equation describes the wavelength as the ratio of the distance per unit time which a point on the wave travels to the number of such points which pass a given point per unit time.

In waveguides, a similar relation should hold:

$$\lambda_g = \frac{v_\varphi}{f} = \frac{2\pi}{\beta_g} \tag{16}$$

where the subscript g implies that the wavelength and propagation constant in the guide may be <u>different</u> from those in free space, mainly because the "phase velocity" v_φ is different. Equation (10) gives the relationship between λ_g and λ [or between v_φ and c] for TE_{no} waves for a given f:

$$\beta_g{}^2 = -\frac{n^2\pi^2}{a^2} + \beta^2$$

so

$$\lambda_g = \frac{2\pi}{\sqrt{\beta^2 - \left(\frac{n\pi}{a}\right)^2}} = \frac{\lambda}{\sqrt{1 - \left(\frac{n\pi}{\beta a}\right)^2}} \tag{17a}$$

since $\lambda = 2\pi/\beta$ (from eq. 15). Also,

$$v_\varphi = \lambda_g f = \frac{2\pi f}{\beta\sqrt{1 - \left(\frac{n\pi}{\beta a}\right)^2}} = \frac{c}{\sqrt{1 - \left(\frac{n\pi}{\beta a}\right)^2}} \tag{17b}$$

since $2\pi f/\beta = c$ (again from eq. 15). Thus, both the wavelength and phase velocity are related to their free-space values through the factor

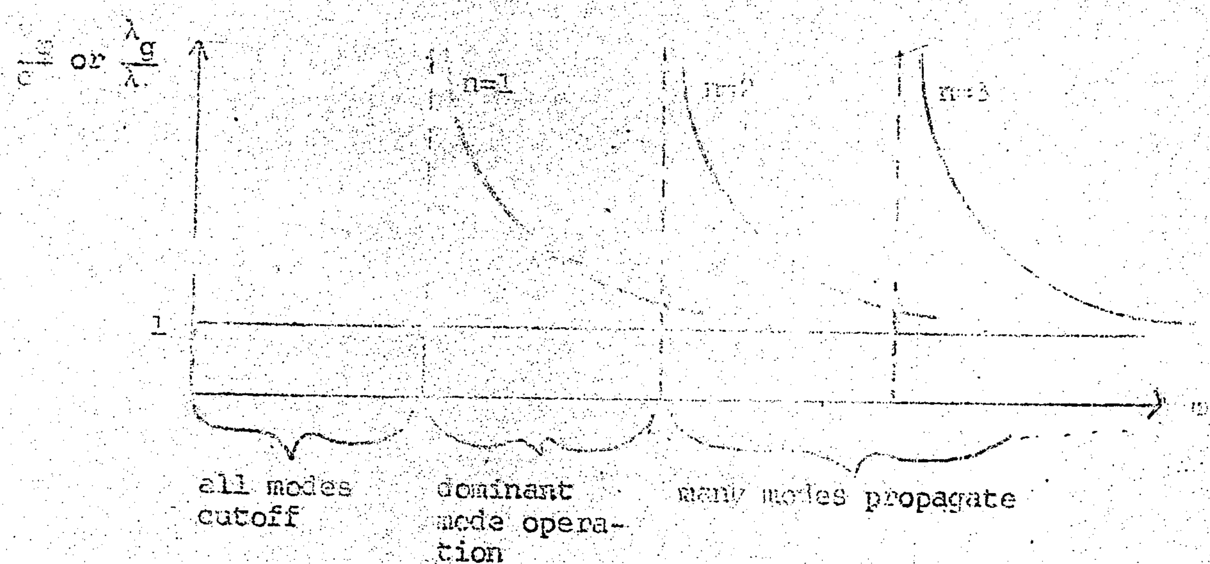
$$\sqrt{1 - \left(\frac{n\pi}{\beta a}\right)^2} = \sqrt{1 - \left(\frac{\omega_c}{\omega}\right)^2} = \sqrt{1 - \left(\frac{f_c}{f}\right)^2} = \sqrt{1 - \left(\frac{\lambda}{\lambda_c}\right)^2}$$

where $f_c \equiv \frac{c}{2a} n$ is the cutoff frequency, $\omega_c = 2\pi f_c$, and we have used $\omega = \beta c$ and $\lambda = \frac{c}{f}$. The cutoff wavelength λ_c is defined by $\lambda_c = \frac{2a}{n} = \frac{c}{f_c}$ which is the <u>free-space</u> wavelength at the frequency f_c.

We see that the velocity is always greater than c for $f > f_c$ (waveguides of the type we are discussing are called "fast wave" structures for this reason); in fact, both v_φ and λ approach ∞

as $f \to f_c$. This behavior can be explained on the basis of reflection of waves between the walls of the waveguide, as discussed in class -- no energy can propagate at a velocity greater than c. On the other hand, at high frequencies, λ_g and v_φ approach their free-space values.

For various values of n, schematic curves of v_φ or λ_g vs. ω are shown below. These are called dispersion curves.



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PLASTICS/CERAMICS FOR ELECTRONICS

Emerson & Cuming, Inc.

MICROWAVE PRODUCTS DIVISION
CANTON, MASSACHUSETTS

TECHNICAL BULLETIN 8-2-18

ECCOSORB® SF

Thin, Flexible, Weather Impervious, Resonant Absorber

ECCOSORB SF is a series of thin resonant absorbers which reflect 1% or less of normally incident energy (at least 20 dB down) at design frequencies in the range 2 to 16 GHz. These absorbers are also available on special order over the frequency range from 17 to 22 GHz. At frequencies in the range 1 to 2 GHz, reflectivity at the resonant frequency is slightly less as shown inthe tabulation. Based on a silicone rubber composition, the absorber will withstand temperatures from -65°F to +325°F (-54°C to +163°C). It is flexible. The higher frequency members of the series drape readily. It is unaffected by severe outdoor exposure; it is completely impervious to moisture.

At frequencies other than the resonant frequency, ECCOSORB SF is reasonably effective as the tabulation shows. The material also performs well at incidence angles other than normal. For example, those members of the series which exhibit normal-incidence reflectivity of -25 dB at the resonant frequency, will at that frequency maintain reflectivity of at least -20 dB out to 30° incidence and at least 16 dB out to 45°.

ECCOSORB SF is ideally suited to applications where extreme thinness is essential and where very broadband performance is not required. ECCOSORB SF finds use in lining radar nacelles (particularly where high power is present), for attachment to masts of ships, aircraft fuselages, etc., to reduce reflections and for lining magnetron housings to prevent detuning.

ECCOSORB SF is supplied in sheets 12" x 12" (30.5cm x 30.5cm). Thickness and weight are indicated in the tabulation. (The tabulation is somewhat different from that which appeared in previous technical bulletins. Thicknesses and weights are different. ECCOSORB SF in accordance with the Technical Bulletin (Revised 6-73) are available on special order.) The material is gray in color. The designation is keyed to the resonant frequency, e.g. ECCOSORB SF-5.5 resonates at 5.5 GHz, ECCOSORB SF-2.5 resonates at 2.5 GHz.

In order to function properly ECCOSORB SF should be intimately backed with a metal surface. Clean the metallic surface thoroughly. Apply a thin coat of Primer S-11 to it. Coverage is about 100 sq.ft./pound (20 m^2/kg). Allow to dry. Apply to the primed metal surface ECCOSIL® TP50 (see Technical Bulletin 13-2-7). Coverage is about 10 sq.ft./pound (2 m^2/kg). Press the ECCOSORB SF into place. Hold it in place until ECCOSIL TP50 has cured. If ECCOSORB SF is to be applied to a non-metallic surface, first apply aluminum foil and then proceed as above. Foil can be bonded using ECCOSIL TP50 but should be primed with Primer S-11. The ECCOSORB SF product line is available with an integral reflective backing on special order.

ECCOSORB SF-U for High Abrasion Resistance -- A urethane rubber version of ECCOSORB SF has been introduced. It has greater abrasion resistance with only a slight loss of flexibility. The table on the reverse side does not apply. Contact local Emerson & Cuming, Inc. office for more details. Also, ECCOSORB SF-U has greater physical strength and fuel resistance than ECCOSORB SF. It can be bonded with epoxy adhesives and has a maximum operating temperature of 225°F (107°C).

ECCOSORB SF-RB for High Incidence -- ECCOSORB SF-RB is a modified version which has been designed to have minimal reflectivity at a single design frequency and at parallel polarization in the incidence angle range from 60° to 85°. Reflection reduction from perfect reflection at the design frequency and at 85° incidence is 6 dB or more, peaking to values as high as 15 dB at 75° incidence and dropping to 6 dB at 60°. This absorber has been successfully flight tested when attached to the belly of an aircraft flying at speeds up to Mach 2. Absorbers with design frequencies from 5 to 16 GHz are available. Typically an X-band version has a thickness of 0.20" (0.51cm) and weighs 2 pounds/sq.ft. (9.8 kg/m^2). A Ku-band version is 0.125" (0.32cm) thick and weighs 1.6 pounds/sq.ft. (7.8 kg/m^2).

This information, while believed to be completely reliable, is not to be taken as warranty for which we assume legal responsibility nor as permission or recommendation to practice any patented invention without license. It is offered for consideration, investigation, and verification.

(Tabulation On Other Side)

Rev. 6/77

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FREQUENCY, GHz

Designation	Nominal Thickness Inch	Nominal Thickness cm	Nominal Weight lb/ft^2	Nominal Weight kg/m^2
SF-1.0	0.190	0.48	5.0	24.4
SF-1.5	0.150	0.38	3.9	19.1
SF-2.0	0.115	0.29	3.0	14.6
SF-2.5	0.085	0.22	2.2	10.7
SF-3.0	0.073	0.19	1.9	9.3
SF-3.5	0.064	0.16	1.7	8.3
SF-4.0	0.105	0.27	2.1	10.3
SF-4.5	0.091	0.23	1.9	9.3
SF-5.0	0.080	0.20	1.6	7.8
SF-5.5	0.092	0.23	1.7	8.3
SF-6.0	0.086	0.22	1.6	7.8
SF-6.5	0.082	0.21	1.5	7.3
SF-7.0	0.078	0.20	1.4	6.8
SF-7.5	0.074	0.19	1.3	6.3
SF-8.0	0.070	0.18	1.3	6.3
SF-8.5	0.066	0.17	1.2	5.9
SF-9.0	0.084	0.21	1.3	6.3
SF-9.5	0.078	0.20	1.2	5.9
SF-10.0	0.076	0.19	1.2	5.9
SF-10.5	0.072	0.18	1.1	5.4
SF-11.0	0.070	0.18	1.1	5.4
SF-12.0	0.066	0.17	1.0	4.9
SF-13.0	0.062	0.16	0.9	4.4
SF-14.0	0.058	0.15	0.9	4.4
SF-15.0	0.054	0.14	0.8	3.9
SF-16.0	0.050	0.13	0.8	3.9

Designation	1.0	1.5	2.0	2.5	3.0	3.5	4.0	4.5	5.0	5.5	6.0	6.5	7.0	7.5	8.0
SF-1.0	12	10	9	9	8										
SF-1.5	12	16	13	10	9	8									
SF-2.0	11	14	18	14	12	7	7								
SF-2.5	7	11	17	22	17	14	9	7							
SF-3.0	6	10	12	18	25	18	15	10	7						
SF-3.5		8	12	16	20	25	20	17	12	9					
SF-4.0			7	12	15	19	21	17	16	11	8				
SF-4.5				11	15	17	20	25	20	16	13	9			
SF-5.0					7	10	13	19	25	20	14	11	8		
SF-5.5		2	3	4	6	8	10	13	19	25	20	14	10	9	9
SF-6.0			2	3	5	6	8	10	13	19	25	20	14	10	10
SF-6.5			2	3	4	5	6	8	10	13	20	25	20	15	13
SF-7.0				2	3	4	5	6	8	10	13	20	25	20	15
SF-7.5					2	3	4	5	6	8	10	14	20	25	20
SF-8.0						2	3	5	6	7	8	10	15	20	25
SF-8.5						2	3	5	6	7	8	9	10	15	20
SF-9.0						2	3	4	5	6	7	8	9	10	15
SF-9.5						2	3	4	5	6	7	8	9	10	11
SF-10.0						2	3	4	5	5	6	7	8	9	10
SF-10.5							2	3	4	4	5	6	7	8	9
SF-11.0							2	3	3	4	4	5	5	6	8
SF-12.0															
SF-13.0															
SF-14.0															
SF-15.0															
SF-16.0															

Designation	8.5	9.0	9.5	10.0	10.5	11.0	11.5	12.0	12.5	13.0	13.5	14.0	14.5	15.0	15.5	16.0
SF-1.0																
SF-1.5																
SF-2.0																
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SF-3.0																
SF-3.5																
SF-4.0																
SF-4.5																
SF-5.0																
SF-5.5	8	8	7	8	8	8										
SF-6.0	9	8	7	6	6	7										
SF-6.5	10	9	8	7	6	6										
SF-7.0	13	10	9	8	7	6										
SF-7.5	16	12	10	9	7	7										
SF-8.0	20	16	12	10	9	8										
SF-8.5	25	20	16	12	10	9										
SF-9.0	20	25	20	16	12	10										
SF-9.5	15	20	25	20	16	12										
SF-10.0	12	15	19	25	19	16										
SF-10.5	10	11	14	19	25	20										
SF-11.0	9	10	11	14	18	23	18									
SF-12.0				10	12	16	20	25	20	16	13	9				
SF-13.0						10	12	17	20	25	20	17				
SF-14.0								9	13	17	20	25	20	17	14	10
SF-15.0										10	13	17	20	25	20	17
SF-16.0												10	15	18	20	25

NOMINAL REFLECTIVITY, METAL BACKED, AT NORMAL INCIDENCE

DB DOWN FROM 100 PERCENT REFLECTION

electric and magnetic fields are opposite that of the O-BWA. No sole plate is required in the O-BWA, of course.

The O-BWA can be converted into an O-BWO by means of a properly adjusted feedback network between the input and output terminals, as in any amplifier device. With a loop gain of unity and a phase shift of 2π n radians, proper conditions exist for sustained electrical oscillations.

9–8 APPLICATIONS OF O-TYPE DEVICES

The O-type device uses electric and magnetic fields, but these fields are generally parallel to each other. These tubes are also termed linear devices, probably because they can operate over wide dynamic signal ranges. They generally operate at lower power levels than M-type devices, but do not achieve the efficiencies of the M-type device. An example of the range of cw power level capability of the O-type device is given in Fig. 9-15. It should be noted that the reflex klystron can cover a broad range of microwave and millimeter wave frequencies. It has been commonly used as a local oscillator in radar receivers. Since local oscillators (LO's) can operate in the milliwatt (or tens of milliwatts) range, the reflex klystron should be useful as a receiver LO up to 50 GHZ. Note also that the solid-state Impatt diode is competitive with the reflex klystron. (See curves (1) and (b) of Fig. 9-15.)

Linear-beam tubes have also been developed that can produce high-peak powers. For example, with a 90 kv supply voltage, 1 MW peak power levels can be obtained from either a klystron or a twystron tube. Typical efficiencies are on the order of 30% for the linear-beam pulse-power amplifier. The usable dynamic range of these tubes can vary from 40 to 80 dB. High-power coupled-cavity travelling-wave tubes have been developed that can produce peak power outputs of up to 200 kw at 2 GHz, but at relatively modest efficiency.

The linear-beam tube tends to generate less noise than the M-type devices and can have a larger bandwidth, along with increased dynamic range. Thus if high power over a large bandwidth is required, a twystron or a TWT design would be suitable. (As we have seen, the slow-wave structure makes this possible, being nonresonant.) The linear-beam tube concept is a superior amplifier, but the crossed-field device is smaller and more efficient, if its other deficiencies can be tolerated. Since the linear-beam device operates at a lower noise level, it is useful as a small, low-power device for receiver applications. For example, low-noise TWT amplifiers are in common use for the sensitive first stages of a radar receiver. The helix-type tube would be preferable for this application, as it has the higher bandwidth.

10 Masers and Lasers

10–1 INTRODUCTION

This chapter is devoted to the description of microwave and millimeter wave oscillators and amplifiers that are based upon Maser and Laser principles. The acronym Maser stands for Microwave Amplification by Stimulated Emission of Radiation, and the acronym LASER stands for Light Amplification by Stimulated Emission of Radiation. To analyze these types of devices, it is necessary to use nonclassical, or quantum, mechanics. The operation of the O- and M-type tubes of Chapters 8 and 9 were described in terms of free electron charges and the motions of these charges. Chapters 4 and 5 dealt with solid-state electron concepts and the hole and electron motions in a semiconductor. In this chapter a new set of principles applies, the principles of quantum electronics in the various electronic materials. These principles will be described in simple, basic terms, so that the fundamental ideas can be grasped. Since some lasers can operate in the high end of the millimeter band, they must be considered in this text.

Masers operate in microwave cavities, and lasers operate in optical cavities, so that both microwave and optical cavity principles are involved. This presents no real problem, as optical waves and microwaves are both electromagnetic (EM) waves and obey the same EM laws. With this in mind, it will be no surprise to learn that masers were developed first and that lasers were derived from them. The laser has had the greater success, and thus, most of the chapter is devoted to laser principles.

10–2 QUANTUM-ELECTRONIC CONCEPTS

The spectrum of EM waves, ranging from AM broadcast waves to the cosmic waves, is given in Fig. 10-1. An EM wave can be described

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MICROWAVE PRODUCTS DIVISION
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ECCOSORB® SF

Thin, Flexible, Weather Impervious, Resonant Absorber

ECCOSORB SF is a series of thin resonant absorbers which reflect 1% or less of normally incident energy (at least 20 dB down) at design frequencies in the range 2 to 16 GHz. These absorbers are also available on special order over the frequency range from 17 to 22 GHz. At frequencies in the range 1 to 2 GHz, reflectivity at the resonant frequency is slightly less as shown inthe tabulation. Based on a silicone rubber composition, the absorber will withstand temperatures from -65°F to +325°F (-54°C to +163°C). It is flexible. The higher frequency members of the series drape readily. It is unaffected by severe outdoor exposure; it is completely impervious to moisture.

At frequencies other than the resonant frequency, ECCOSORB SF is reasonably effective as the tabulation shows. The material also performs well at incidence angles other than normal. For example, those members of the series which exhibit normal-incidence reflectivity of -25 dB at the resonant frequency, will at that frequency maintain reflectivity of at least -20 dB out to 30° incidence and at least 16 dB out to 45°.

ECCOSORB SF is ideally suited to applications where extreme thinness is essential and where very broadband performance is not required. ECCOSORB SF finds use in lining radar nacelles (particularly where high power is present), for attachment to masts of ships, aircraft fuselages, etc., to reduce reflections and for lining magnetron housings to prevent detuning.

ECCOSORB SF is supplied in sheets 12" x 12" (30.5cm x 30.5cm). Thickness and weight are indicated in the tabulatior (The tabulation is somewhat different from that which appeared in previous technical bulletins. Thicknesses and weigh are different. ECCOSORB SF in accordance with the Technical Bulletin (Revised 6-73) are available on special order. The material is gray in color. The designation is keyed to the resonant frequency, e.g. ECCOSORB SF-5.5 resonates at 5.5 GHz, ECCOSORB SF-2.5 resonates at 2.5 GHz.

In order to function properly ECCOSORB SF should be intimately backed with a metal surface. Clean the metallic surface thoroughly. Apply a thin coat of Primer S-11 to it. Coverage is about 100 sq. ft./pound (20 m^2/kg). Allow to dry. Apply to the primed metal surface ECCOSIL® TP50 (see Technical Bulletin 13-2-7). Coverage is about 10 sq. ft. pound (2 m^2/kg). Press the ECCOSORB SF into place. Hold it in place until ECCOSIL TP50 has cured. If ECCOSORI SF is to be applied to a non-metallic surface, first apply aluminum foil and then proceed as above. Foil can be bonded using ECCOSIL TP50 but should be primed with Primer S-11. The ECCOSORB SF product line is available with an integral reflective backing on special order.

ECCOSORB SF-U for High Abrasion Resistance -- A urethane rubber version of ECCOSORB SF has been introduced. It has greater abrasion resistance with only a slight loss of flexibility. The table on the reverse side does not apply. Contact local Emerson & Cuming, Inc. office for more details. Also, ECCOSORB SF-U has greater physical strength and fuel resistance than ECCOSORB SF. It can be bonded with epoxy adhesives and has a maximum operating tempera ture of 225°F (107°C).

ECCOSORB SF-RB for High Incidence -- ECCOSORB SF-RB is a modified version which has been designed to have minimal reflectivity at a single design frequency and at parallel polarization in the incidence angle range from 60° to 85°. Reflection reduction from perfect reflection at the design frequency and at 85° incidence is 6 dB or more, peaki to values as high as 15 dB at 75° incidence and dropping to 6 dB at 60°. This absorber has been successfully flight tested when attached to the belly of an aircraft flying at speeds up to Mach 2. Absorbers with design frequencies fror 5 to 16 GHz are available. Typically an X-band version has a thickness of 0.20" (0.51cm) and weighs 2 pounds/sq. ft. (9.8 kg/m^2). A Ku-band version is 0.125" (0.32cm) thick and weighs 1.6 pounds/sq. ft. (7.8 kg/m^2).

(Tabulation On Other Side)

FREQUENCY, GHz

Designation	Nominal Thickness Inch	Nominal Thickness cm	Nominal Weight lb/ft^2	Nominal Weight kg/m^2	1.0	1.5	2.0	2.5	3.0	3.5	4.0	4.5	5.0	5.5	6.0	6.5	7.0	7.5	8.0	8.5	9.0	9.5	10.0	10.5	11.0	11.5	12.0	12.5	13.0	13.5	14.0	14.5	15.0	15.5	16.0
SF-1.0	0.190	0.48	5.0	24.4	12	10	9	9	8																										
SF-1.5	0.150	0.38	3.9	19.1	12	16	13	10	9	8																									
SF-2.0	0.115	0.29	3.0	14.6	11	14	18	14	12	7	7																								
SF-2.5	0.085	0.22	2.2	10.7	7	11	17	22	17	14	9	7																							
SF-3.0	0.073	0.19	1.9	9.3	6	10	12	18	25	18	15	10	7																						
SF-3.5	0.064	0.16	1.7	8.3		8	12	16	20	25	20	17	12	9																					
SF-4.0	0.105	0.27	2.1	10.3			7	12	15	19	21	17	16	11	8																				
SF-4.5	0.091	0.23	1.9	9.3				11	15	17	20	25	20	16	13	9																			
SF-5.0	0.080	0.20	1.6	7.8					7	10	13	19	25	20	14	11	8																		
SF-5.5	0.092	0.23	1.7	8.3		2	3	4	6	8	10	13	19	25	20	14	10	9	9	3	8	7	8	8	8										
SF-6.0	0.086	0.22	1.6	7.8			2	3	5	6	8	10	13	19	25	20	14	10	10	9	8	7	6	6	7										
SF-6.5	0.082	0.21	1.5	7.3			2	3	4	5	6	8	10	13	20	25	20	15	13	10	9	8	7	6	6										
SF-7.0	0.078	0.20	1.4	6.8				2	3	4	5	6	8	10	13	20	25	20	15	13	10	9	8	7	6										
SF-7.5	0.074	0.19	1.3	6.3					2	3	4	5	6	8	10	14	20	25	20	16	12	10	9	7	7										
SF-8.0	0.070	0.18	1.3	6.3						2	3	5	6	7	8	10	15	20	25	20	16	12	10	9	8										
SF-8.5	0.066	0.17	1.2	5.9						2	3	5	6	7	8	9	10	15	20	25	20	16	12	10	9										
SF-9.0	0.084	0.21	1.3	6.3						2	3	4	5	6	7	8	9	10	15	20	25	20	16	12	10										
SF-9.5	0.078	0.20	1.2	5.9						2	3	4	5	6	7	8	9	10	11	15	20	25	20	16	12										
SF-10.0	0.076	0.19	1.2	5.9						2	3	4	5	5	6	7	8	9	10	12	15	19	25	19	16										
SF-10.5	0.072	0.18	1.1	5.4							2	3	4	4	5	6	7	8	9	10	11	14	19	25	20										
SF-11.0	0.070	0.18	1.1	5.4							2	3	3	4	4	5	5	6	8	9	10	11	14	18	23	18									
SF-12.0	0.066	0.17	1.0	4.9																			10	12	16	20	25	20	16	13	9				
SF-13.0	0.062	0.16	0.9	4.4																					10	12	17	20	25	20	17				
SF-14.0	0.058	0.15	0.9	4.4																							9	13	17	20	25	20	17	14	10
SF-15.0	0.054	0.14	0.8	3.9																									10	13	17	20	25	20	17
SF-16.0	0.050	0.13	0.8	3.9																											10	15	18	20	25

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