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A Prototype Transmitter for a Microwave Refractometer

John C. Jordan

A PROTOTYPE TRANSMITTER FOR A MICROWAVE REFRACTOMETER - JORDAN

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A PROTOTYPE TRANSMITTER FOR A
MICROWAVE REFRACTOMETER

by

John C. Jordan

A Thesis
Submitted in Partial Fulfillment of the
Requirements for the Degree of
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1962

This thesis, directed and approved by the candidate's committee, has been accepted by the Graduate Committee of The University of New Mexico in partial fulfillment of the requirements for the degree of

MASTER OF SCIENCE

Stuart A. Northrop
Dean

May 15, 1962
Date

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MICROWAVE REFRACTOMETER

by

John C. Jordan

Thesis Committee

Donald C Thorn
Chairman

W W Grannemann

Ruben A. Kelly

ABSTRACT

A study of the design, development and fabrication details of a transmitter for a microwave refractometer is presented.

Atmospheric, radio frequency, refractive index is to be measured rapidly and directly up to altitudes of the order of 200,000 feet. Requirements for the transmitter with regard to its weight, volume and environment are indeed stringent since it is proposed to package the transmitter in an ogive configuration and elevate it to these altitudes utilizing a rocket propulsion system.

The basic components of the transmitter are as follows: power source, power converter, sawtooth voltage generator, klystron oscillator, resonant cavity, antenna and reflector. The self-contained power source provides primary power for the entire system. A DC-to-DC power converter is employed to furnish the required high voltages for the klystron oscillator and the sawtooth voltage generator. Generation of a carrier in the X-band is accomplished with the klystron oscillator. This carrier is frequency modulated by application of the sawtooth voltage to the repeller of the klystron. The resonant cavity produces the desired amplitude modulation of the frequency modulated signal provided it has sufficiently high Q, and the swept frequency range is chosen such that the resonant frequency of the cavity lies within this range for all anticipated values of the parameter to be measured. Information relating to the resonant frequency of the cavity is contained in the peak of the envelope of this special AM-FM signal. Suitable demodulation of this

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AM-FM signal results in an indication of the index of refraction since the resonant frequency of the cavity is inversely proportional to the index of refraction of the material (the atmosphere) contained within the cavity.

AM-FM signal results in an indication of the refractive index, since the resonant frequency of the cavity is linearly proportional to the index of refraction of the material (the atmosphere) contained within the cavity.

ACKNOWLEDGMENTS

I am indeed grateful to Dr. Donald C. Thorn for his supervision and guidance in the preparation of this thesis and to Dr. Wayne W. Grannemann and Dr. Ruben D. Kelly, who also served on the thesis committee, for their helpful suggestions.

The Signal Missile Support Agency of the White Sands Missile Range is acknowledged for supporting this study under contract DA-29-040-ORD-1238 with the Electrical Engineering Division of the University of New Mexico Engineering Experiment Station.

I also express my gratitude to Miss Karen Gordon for typing the manuscript and to Mr. Van Gilbert for preparing the illustrations.

John C. Jordan

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CHAPTER I -- INTRODUCTION

Concerted efforts to design and construct microwave devices to measure, rapidly and directly, the radio frequency refractive index of the atmosphere have been relatively recent.[1/] One of these devices developed by Crain[2/] utilizes two microwave resonant cavities, one being a reference which is sealed to prevent changes in the composition of the air inside and the other being provided with holes which permit sampling of the atmosphere. (The resonant frequency of a cavity is inversely proportional to the index of refraction of the material contained within the cavity.) The difference between the resonant frequencies of the two (resonant frequency of the reference cavity is fixed) is an indication of the index of refraction of the atmosphere contained within the sampling cavity. Another device developed by Birnbaum[3/] uses two similar cavities which are excited by a klystron oscillator that is swept linearly with time across the slightly different resonant frequencies of the two cavities. The difference between the time of passage of the oscillator between the two resonant frequencies of the cavities is an indication of the variation of the index of refraction of the air in the sampling cavity. Successful operation of both of

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[1/] Herbstreit, J. W., Radio Refractometry, Technical Note No. 66, Boulder Laboratories, National Bureau of Standards, July, 1960, for a history of radio refractometry.

[2/] Crain, C. M., Apparatus for Recording Fluctuations in the Refractive Index of the Atmosphere, Review of Scientific Instruments, Vol. 21, 1950.

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these microwave devices depends upon the utilization of a suitable resonant cavity. In addition to these microwave devices, others developed by Deam[4] and Vetter[5] also utilize the resonant cavity principle. Since suitable resonant cavities are necessary for successful operation of such microwave refractometers, it is quite logical that considerable interest has been shown in their design and construction. Recent investigations have been conducted by Thorn,[6] Thompson, et al.,[7,8] and Gilmer and Thorn.[9]

The basic requirement for these microwave refractometers, utilizing the resonant cavity principle, is to measure the atmospheric radio frequency index of refraction of the lower altitudes. It is proposed to study a device to measure this parameter up to altitudes of the order of 200,000 feet.[10] It

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[4] Deam, A. P., An Expendable Atmospheric Radio Refractometer, EERL Report 108, University of Texas, 1959.

[5] Vetter, M. J., Absolute Refractometers, National Bureau of Standards Report No. 6700, 1960.

[6] Thorn, D. C., Design of Open-Ended Resonant Cavities, University of Texas, 1958.

[7] Thompson, M. C., et al., Fabrication Techniques for Ceramic X-Band Cavity Resonators, Review of Scientific Instruments, Vol. 29, 1958.

[8] Thompson, M. C., et al., End Plate Modifications of X-Band TE_{011} Cavity Resonators, Institute of Radio Engineers Transactions, PGMTT Vol. 7, 1959.

[9] Gilmer, R. O., and D. C. Thorn, Some Design Criteria for Open-Ended Microwave Cavities, The University of New Mexico Engineering Experiment Station, Albuquerque, Technical Report EE-65, 1961.

[10] Thorn, D. C., et al., A Prototype Light-Weight Remote Microwave Refractometer, University of New Mexico Engineering Experiment Station, Albuquerque, Technical Report EE-63, 1961.

is proposed to elevate a refractometer to these altitudes utilizing a rocket propulsion system. Upon reaching the maximum altitude, the refractometer will be released and will descend under parachute retardation. The refractometer transmits an appropriately modulated signal which, upon suitable detection or demodulation, will result in a rapid and direct measurement of the index of refraction of the atmosphere. Transmission of a frequency modulated signal which has been fed through a high Q resonant cavity (whose resonant frequency depends upon the index of refraction) is the basis for the design of this refractometer transmitter. The parachute mentioned above could be metalized and serve as a reflector for the transmitter antenna.

Generation of this frequency modulated signal is accomplished by applying the sawtooth voltage to the repeller of the klystron oscillator. This frequency modulated signal is then fed through the resonant cavity which produces the desired amplitude modulation of the frequency modulated signal provided the Q of the cavity is high enough and the swept frequency range includes the resonant frequencies of the cavity that result from various indices of refraction. The information relating to the index of refraction is contained in the peak of the envelope of this AM-FM signal. The requirement for a very high Q resonant cavity is obvious since Q is directly proportional to frequency of operation (f_o) and inversely proportional to bandwidth (BW) which in this case must be quite narrow if small changes in the index of refraction are to be determined. Certainly, in order to obtain sufficient amplitude modulation of the frequency

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modulated carrier, such that peaks of the envelope may be determined, the bandwidth (3 db points) of the resonant cavity must be considerably less than the swept frequency range of the frequency modulated signal.

The volume-to-surface ratio of the resonant cavity is to some extent an indication of the Q of the cavity. The right circular cylinder configuration will, by this criterion, result in a higher Q cavity than one of rectangular cross-section. In addition to this, the hollow, open end right circular cylinder (perhaps with some modifications to prevent air stagnation) is aerodynamically suitable. Obviously, the resonant cavity is of paramount importance if successful operation of the microwave refractometer is to be realized.

Frequency of operation in the X-band has been selected for this microwave refractometer in consideration of the limited volume and weight permitted for the transmitter. Klystrons suitable for this application are available which occupy only a few cubic inches of volume and weigh only a few ounces. The resonant cavity and waveguide sections for X-band operation are also of appropriate size and weight.

The velocity of descent of the transmitter and the sweep rate of the frequency modulated signal determine the sampling rate of the index of refraction. For instance, if the transmitter is descending at a velocity of 300 feet per second and the sweep rate of the frequency modulated signal is 300 cycles per second, then the data sampling rate is one sample per foot.

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The velocity of descent of the transmitter and the sweep rate of the frequency modulated signal determine the sampling rate of the index of refraction. For instance, if the transmitter is descending at a velocity of 500 feet per second and the sweep rate of the frequency modulated signal is 500 cycles per second, then the data sampling rate is one sample per foot.

The transmitter contains the following basic components: a power source, a power converter, a klystron oscillator, a sawtooth voltage generator, a resonant cavity and an antenna and reflector. The transmitter and its various components are discussed in the subsequent chapters. The physical environment resulting from the rocket propulsion system and the extreme altitude is also described.

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CHAPTER II -- DESCRIPTION OF THE TRANSMITTER

A simplified block diagram of the transmitter is depicted in Figure 1. In consideration of the klystron filament requirements, a power source furnishing 6 volts has been selected. Available volume for the power source and the maximum permissible weight dictate utilization of cells such as the silver-zinc cell. Terminal voltage characteristics (constant voltage source with respect to load and amount of discharge) of such cells are highly desirable. The large energy-weight and energy-volume ratios that are possible with silver-zinc cells are also of great advantage.

In addition to supplying power for the filaments of the klystron, the power source must furnish power for the klystron resonator circuit, the klystron repeller circuit and the sawtooth voltage generator. Since the 6 volts from the power source are not suitable for the last three circuits, a DC-to-DC conversion is employed. Of course there is the possibility of a high voltage combination of primary cells, but such a combination would occupy more volume and weigh more than the DC-to-DC converter. This conversion is accomplished through the transformer coupled DC-to-DC power converter. A transistor switching scheme is employed in the primary of the transformer (which permits the transforming action) and diode bridge rectifiers are utilized in the transformer secondary circuits. Suitable turns ratios are, of course, a part of the transformer design. Thus, the positive DC voltage for the klystron resonator and the

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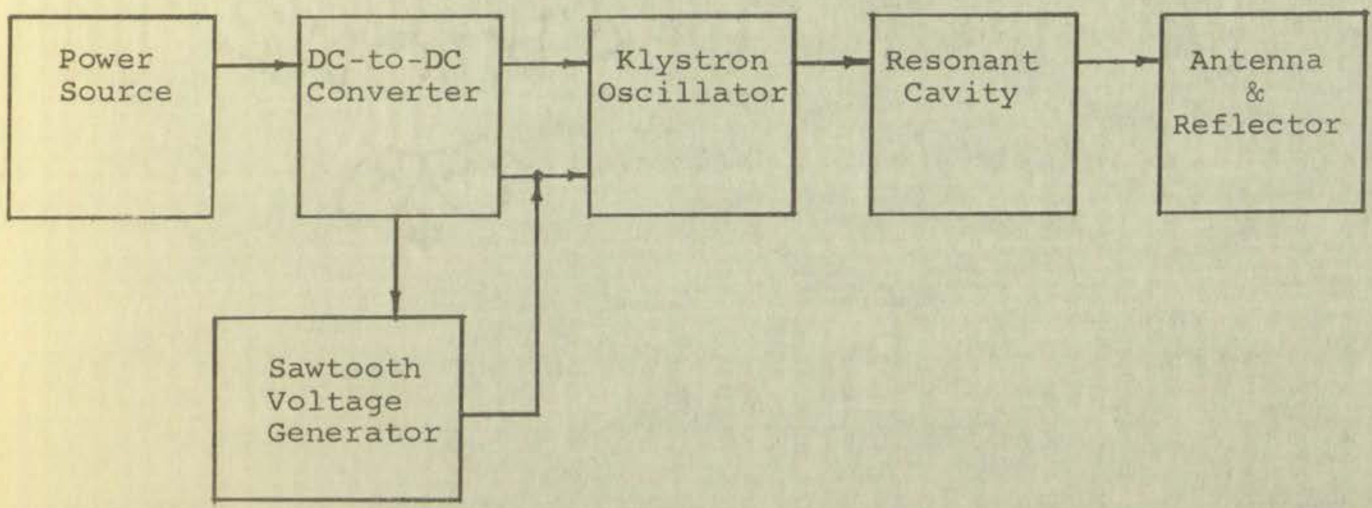

Figure 1. Simplfied Block Diagram of the Transmitter.

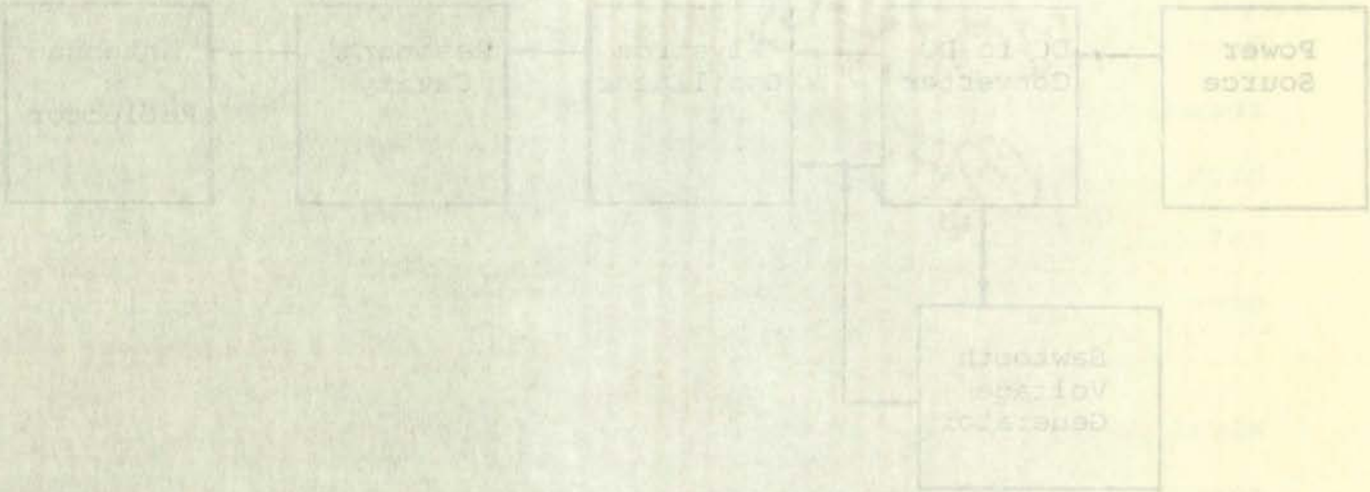

Figure 1. Simplified block diagram of the [illegible]

sawtooth voltage generator and the negative DC voltage for the klystron repeller are produced.

Application of the klystron filament voltage, the klystron repeller voltage and the klystron resonator voltage generates the carrier at frequency ω. This carrier is designated as,

$$v = V_o \sin \omega t, \tag{2.1}$$

where, V_o is the maximum voltage and ω is the radian frequency of oscillation. The sawtooth voltage, which is generated from a resistor-capacitor network utilizing a four-layer diode as a "switch," is then superimposed on the klystron repeller. Output of the klystron oscillator is now,

$$v = V_o \sin(\omega_1 + \omega_2 \frac{t}{T})t \quad \text{for, } 0 < t < T \text{ and periodic on } T \text{ thereafter,} \tag{2.2}$$

where, ω_1 is the lowest instantaneous frequency, ω_2 is the swept frequency range and T is the period of one cycle of the modulating sawtooth voltage, as shown in Figure 2a.[11]

Feeding this frequency modulated signal to the resonant cavity will result in amplitude modulation of the frequency modulated signal provided the Q of the cavity is high enough and the resonant frequency of the cavity lies within the swept frequency range. Assuming that the output signal of the klystron oscillator contained no amplitude modulation, and assuming that the resonant frequency of the cavity is close to the lowest

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[11] Dearholt, Donald W., Demodulation of a Special AM-FM Signal, The University of New Mexico Engineering Experiment Station, Albuquerque, New Mexico, Technical Report EE-34, 1960.

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Application of the Klystron filament voltage, the Klystron repeller voltage and the [illegible] voltage generates the carrier at frequency ω. This carrier is designated as

$$V = V_o \sin \omega t \quad (2\text{-}1)$$

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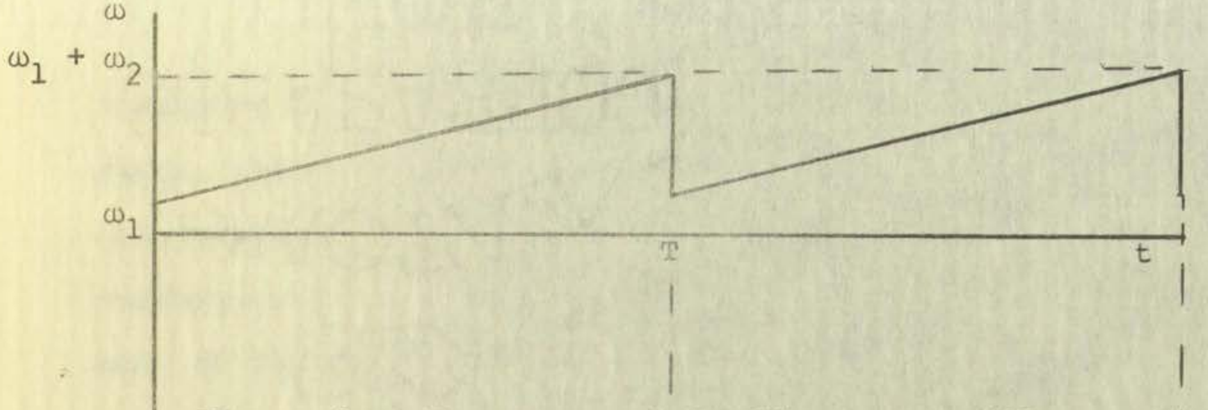

Figure 2a. Frequency of the Klystron Oscillator

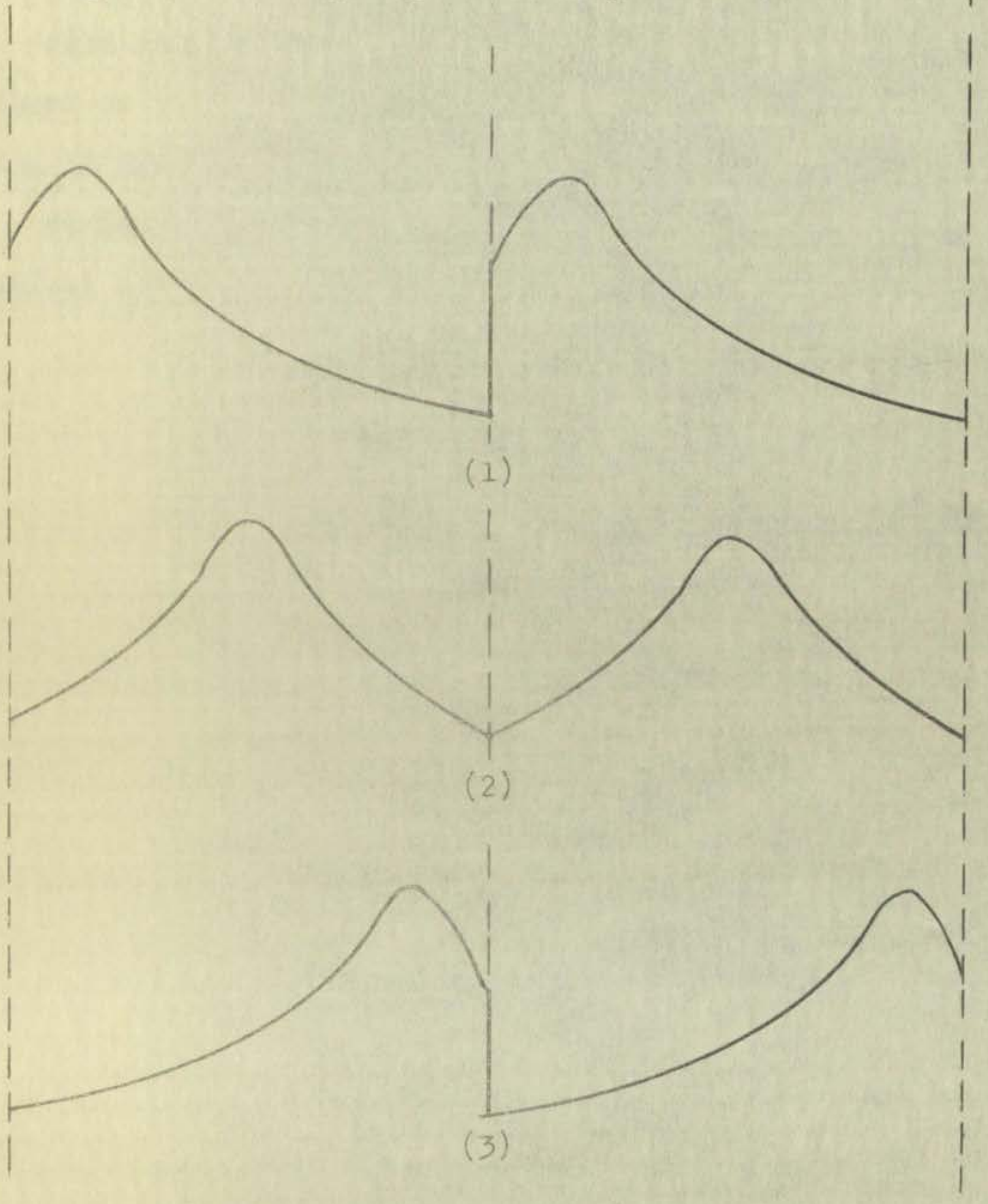

Figure 2b. Possible Outputs (envelopes) from the Resonant Cavity.

frequency of the frequency modulated signal, then in the mid-range, finally, close to the highest frequency, then output of the resonant cavity will appear as in Figure 2b as 1, 2, 3 respectively. The peaks of these envelopes represent a measurement of the indices of refraction since the resonant frequency of the cavity is inversely proportional to the index of refraction.

This special AM-FM signal is then fed to the antenna. The antenna is positioned properly with respect to the parachute which is metalized and serves as the reflector.

Suitable demodulation of the special AM-FM signal at the receiver will result in measurement of the index of refraction.

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Suitable demodulation of the special AM-FM signal at the receiver will result in measurement of the index of refraction.

CHAPTER III -- THE POWER SOURCE

One of the important requirements for the power source is that it maintain a constant voltage during the discharge cycle. The reason for this is that the selection of such a power source minimizes the need for voltage regulation in the DC-to-DC converter. Another consideration is the severe weight and volume limitations placed on the ogive package which make it necessary to choose a power source that has optimum energy to weight and energy to volume ratios. Other considerations include the requirement for the power source to operate satisfactorily at elevated temperatures in combination with low pressures.

A typical terminal voltage versus discharge cycle for the silver-zinc primary cell is shown in Figure 3.[12/] The nominal terminal voltage under load is 1.5 volts. The primary cell has a capacity of 5 ampere-hours. The load is 5 amperes. As seen from Figure 3 the terminal voltage remains approximately 1.5 volts during the discharge cycle.

The weight of such a silver-zinc cell is approximately 5 ounces and it occupies approximately 5 cubic inches. Four cells connected in series are necessary for the 6 volts utilized as the primary voltage. These four cells would weigh about 20 ounces and occupy approximately 20 cubic inches. Based on the numbers given above, it is possible to expect approximately 24 watt-hours per pound and approximately 1.5 watt-hours per cubic

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The weight of such a silver-zinc cell is approximately 5 ounces and it occupies approximately 5 cubic inches. Four cells connected in series are necessary for the 6 volts utilized as the primary voltage. These four cells would weigh about 20 ounces and occupy approximately 20 cubic inches. Based on the numbers given above, it is possible to expect approximately 24 watt-hours per pound and approximately 1.5 watt-hours per cubic

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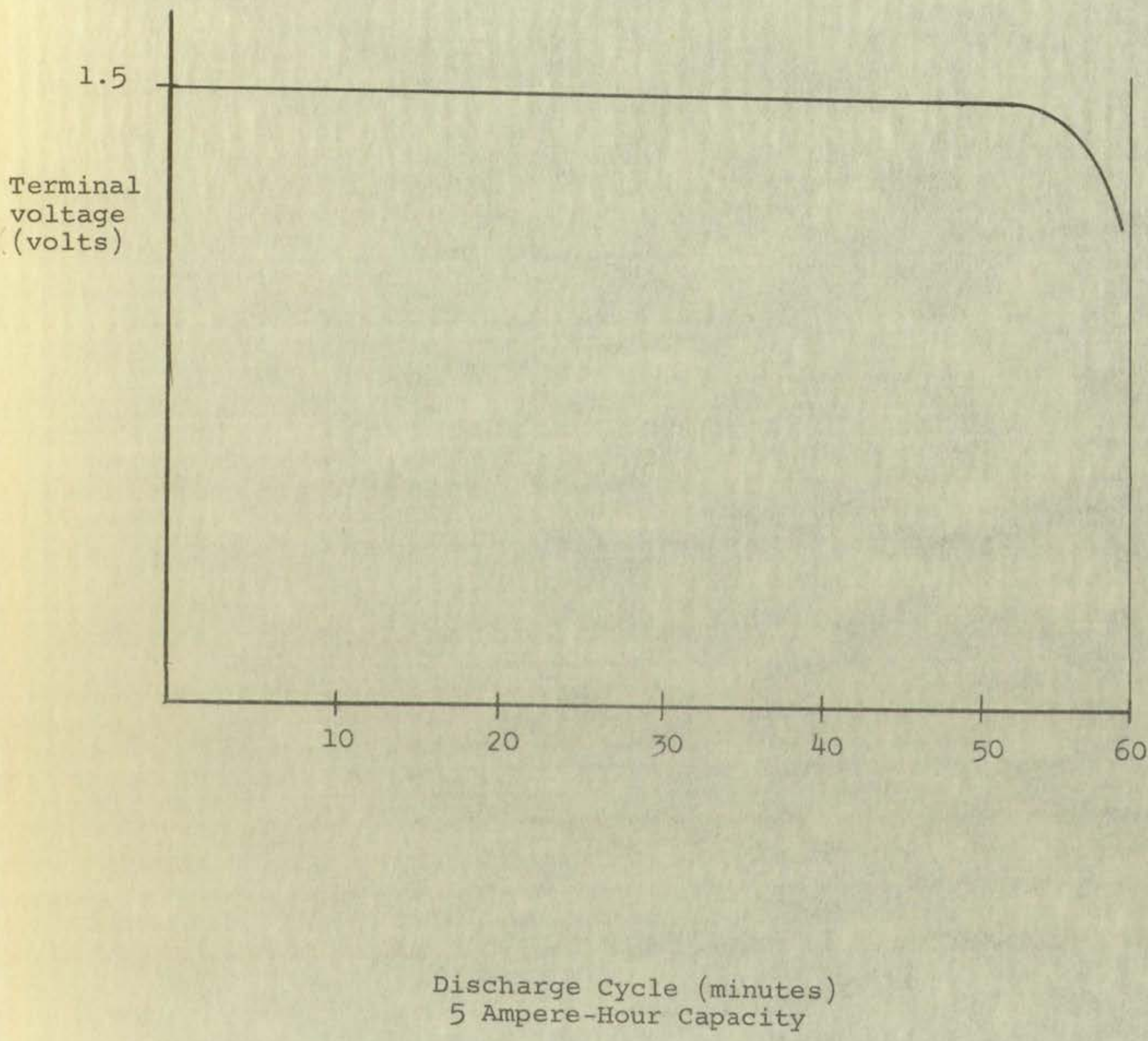

Figure 3. Typical Terminal Voltage versus Discharge Cycle for the Silver-Zinc Cell.

inch from a silver-zinc primary cell. Therefore, it appears that the silver-zinc cell has the capability of a high specific energy output and the desirable voltage regulation characteristic.

Since the power source determines the maximum power available to the microwave refractometer transmitter and an estimate can be made for the maximum weight and volume allowable for the power source, it is possible to determine the approximate maximum transmitter microwave output power in terms of the power source. Reasonable estimates for maximum permissible weight and volume are 1.5 pounds occupying 30 cubic inches. The total weight allowed for the transmitter is 7 pounds and it is to occupy a total of about 165 cubic inches. Dimensions of the ogive are given in Chapter XI, Figure 24. In view of the above considerations, a power source with a capacity of about 30 watt-hours occupying no more than 30 cubic inches could easily be assembled. It is anticipated that the transmitter will require about 1 hour to descend from the maximum altitude. The major portion of this available energy is consumed by the reflex klystron oscillator. At the present time the most promising klystron for this application appears to be the Varian type VA-242 or its equivalent. The filament power required is 6 volts at 1.2 amperes. Approximately 22.8 watts less the loss in the power converter remain for the klystron resonator circuit. All indications (Figure 18, Chapter IX) are that the DC-to-DC converter should have an efficiency of about 80%. This reduces the available klystron resonator input power from 22.8 watts to approximately

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18.2 watts. According to the manufacturer's specifications, the type VA-242 reflex klystron should generate approximately 400 milliwatts of microwave power with an input of 18 watts to the resonator circuit. It is anticipated that at least 3 db of power will be lost between the klystron oscillator and the antenna. Therefore, a maximum of perhaps 200 milliwatts of microwave power could be delivered to the antenna.

Since it is required that the power source operate satisfactorily at extremely high altitudes in combination with elevated temperatures (due to the large heat dissipation required by the transmitter), it may be necessary to package the power source in a pressurized container to prevent the electrolyte in the primary cells from boiling.

18.2 watts. According to the manufacturer's specifications, the type VA-242 reflex klystron should generate approximately 400 milliwatts of microwave power for a power input of 20 watts to the resonator circuit. It is anticipated that at least 3 db of power will be lost between the klystron output jack and the antenna. Therefore, a maximum of only about 200 milliwatts (40%) of microwave power would be radiated by the antenna.

Since it is required that the transmitter operate satisfactorily at extremely high altitudes in conjunction with elevated temperatures (due to the large heat dissipation required by the transmitter), it may be necessary to package the power source in a pressurized container to prevent the electrolyte in the primary cells from boiling.

CHAPTER IV -- DESIGN OF THE DC-TO-DC POWER CONVERTER

The power requirements for the klystron oscillator and sawtooth voltage generator necessitate a power conversion from the power source at 6 volts DC to about + 350 volts DC at approximately 40 ma and about - 160 volts DC at approximately 0 ma. Figure 4 shows the basic schematic for the transformer-coupled DC-to-DC converter.

First, consider the basic primary circuit shown in Figure 5a. Assume, upon application of the input voltage V_B, that transistor V_1 switches on and conducts until the core of transformer T_1 saturates. Then V_1 switches off and V_2 is switched on and conducts until the core of T_1 is saturated in the opposite direction. For an idealized transformer the induced voltage and resultant current and flux in the core are as depicted in Figure 5b. The induced voltage is equal to the number of turns N_p multiplied by the rate of change of flux,

$$e(t) = N_p \frac{d\varphi}{dt} \qquad (4.1)$$

As seen from Figure 5b, the total change in flux is $2\,\Phi$ and this change occurs linearly in time $\frac{T}{2}$. Thus, equation (4.1) may be written as,

$$E = N_p \frac{2\Phi}{\frac{T}{2}} \qquad (4.2)$$

Since,

$$T = \frac{1}{f} \quad \text{or} \quad \frac{T}{2} = \frac{1}{2f} \qquad (4.3)$$

CHAPTER IV. THEORY OF [illegible] DC-TO-DC [illegible]

The power requirements for the [illegible] oscillator and sawtooth voltage generator necessitate a power conversion from the power source at [illegible] volts DC to about 350 volts DC at approximately 40 ma and [illegible] volts DC at approximately [illegible] ma. Figure 4 shows the basic circuit of the transformer-coupled DC-to-DC converter.

First, consider the basic primary circuit shown in Figure 5a. Assume, upon application of the supply voltage E, that transistor V_1 switches on and conducts until the core of transformer T_1 saturates. Then V_1 switches off and V_2 is switched on and conducts until the core of T_1 is saturated in the opposite direction. For an idealized transformer the induced voltage and resultant current and flux in the core are as depicted in Figure 5b. The induced voltage is equal to the number of turns N_p multiplied by the rate of change of flux.

$$e'(t) = N_p \frac{d\phi}{dt} \qquad (4.1)$$

As seen from Figure 5b the total change in flux is 2Φ and this change occurs linearly in time $\frac{T}{2}$. Thus, equation (4.1) may be written as,

[illegible]

Since,

[illegible]

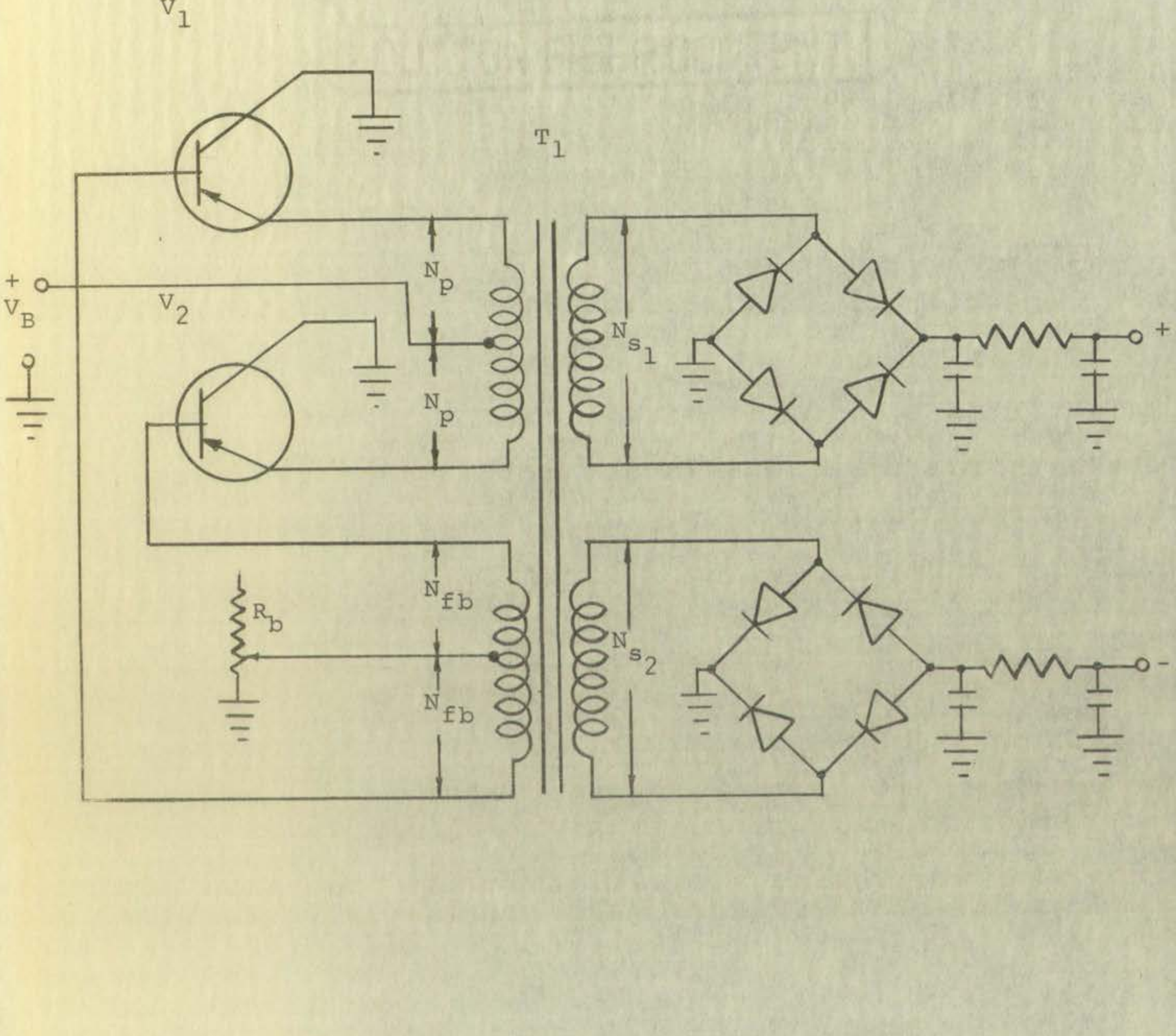

Figure 4. Basic Schematic for the Transformer-Coupled DC-to-DC Converter.

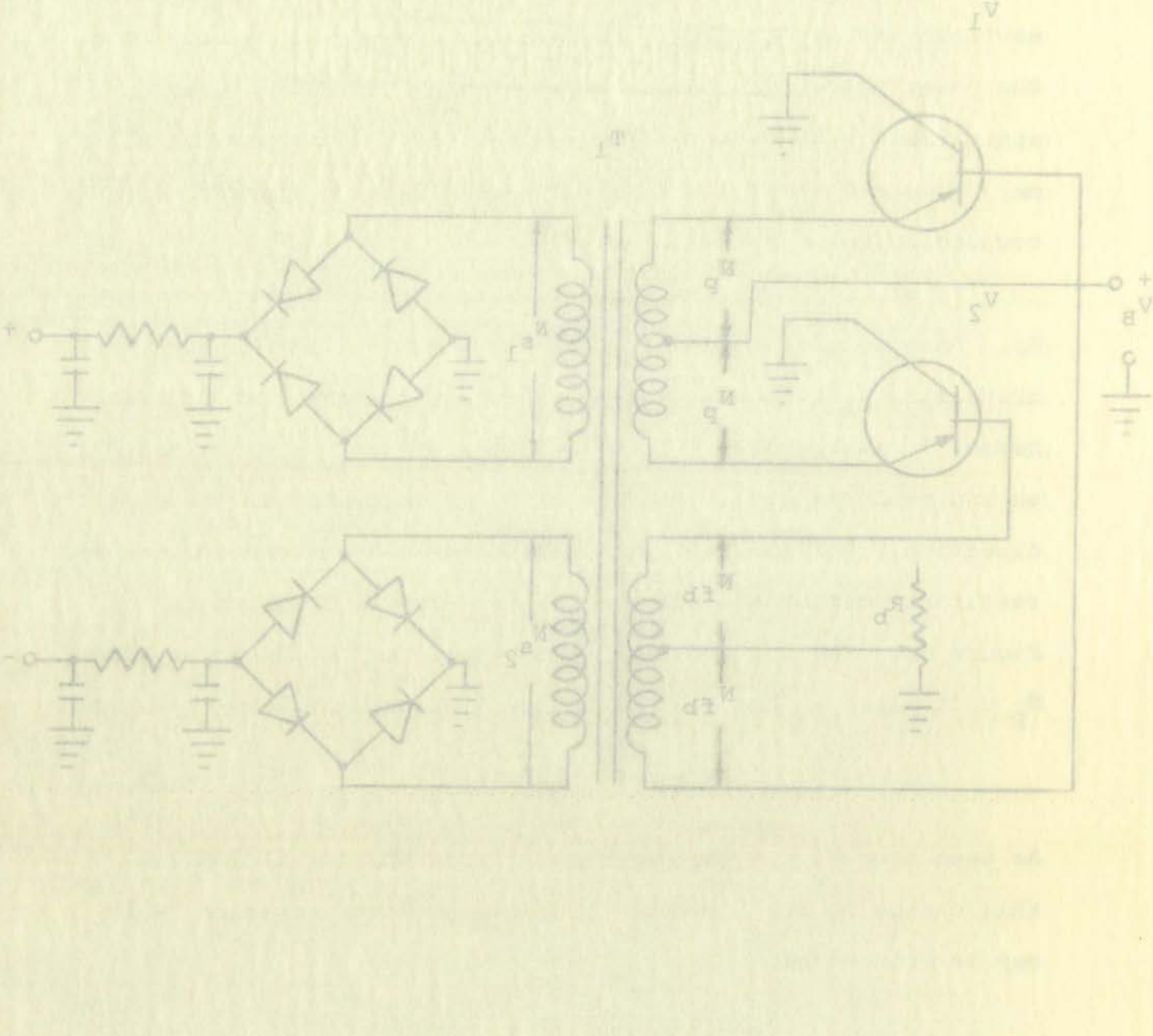

Figure 4. Basic Schematic for the Transformer-Coupled DC-to-DC Converter.

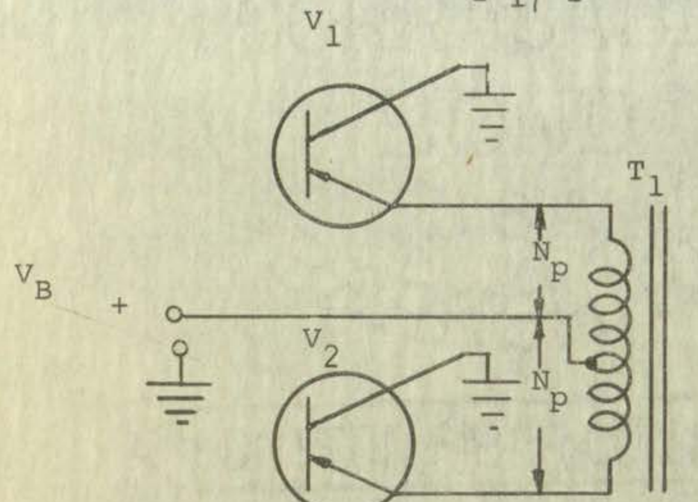

Figure 5a. Basic Primary Circuit.

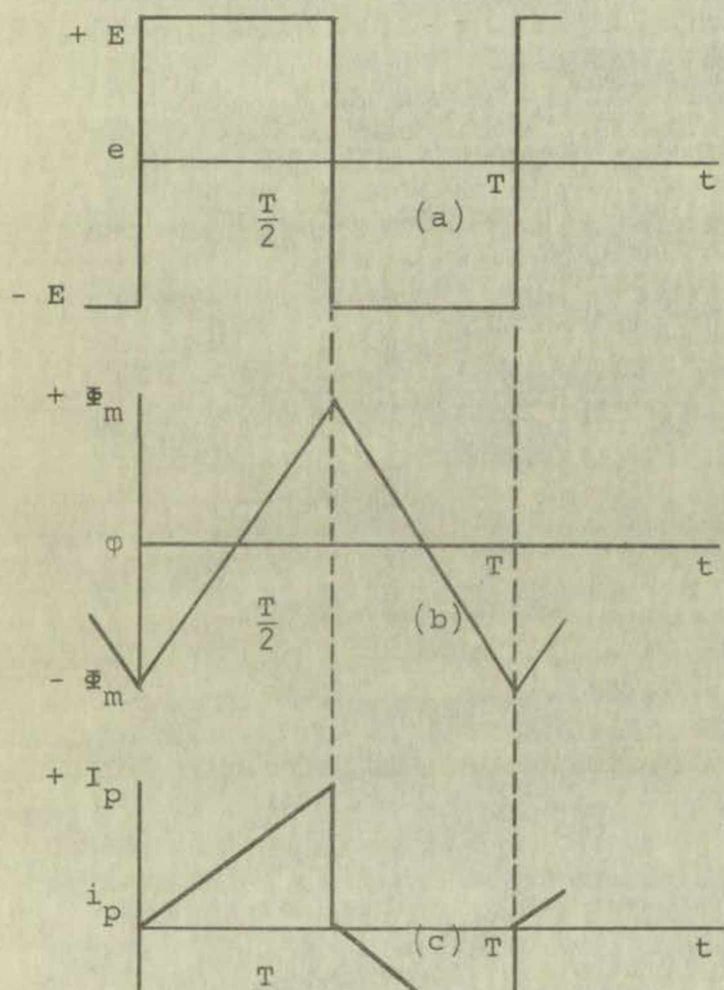

Figure 5b. Induced Voltage e and Resulting Current i_p and Flux φ Waveforms for the Primary Circuits.

equation (4.2) may be written as,

$$E = N_p 4\Phi f \tag{4.4}$$

Also, the total flux may be written as,

$$\Phi = B_m A \tag{4.5}$$

where B_m is the flux density and A is the cross-sectional area of the core. Therefore, equation (4.4) may be written as,

$$E = 4N_p B_m Af \tag{4.6}$$

where E is the induced voltage in volts, B_m is the flux density in webers per square meter, A is the cross-sectional area in square meters, and f is the frequency of operation in cycles per second. Equation (4.6) is rearranged to

$$N_p = \frac{E}{4B_m Af} \tag{4.7}$$

since it is desired to calculate the number of turns N_p required for the primary.

The inductance L_p of the primary N_p may be determined from the following considerations: a torroidial core of cross-section A and mean length m is chosen. The flux density within the volume enclosed by the winding is,[13]

$$B_m = \frac{\mu N_p I_p}{m} \tag{4.8}$$

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[13] Sears, Francis W., Electricity and Magnetism, Page 274, Addison-Wesley Press, Inc., 1951.

equation (4.2) may be rewritten as

$$E = \omega\Phi \qquad (4.4)$$

Also, the total flux may be written as

$$\Phi = B_m A \qquad (4.5)$$

where B_m is the flux density and A is the cross-sectional area of the core. Therefore, equation (4.4) may be written as

$$E = 4N\,f\,AB_m \qquad (4.6)$$

where E is the induced voltage in volts, B_m is the flux density in webers per square meter, A is the cross-sectional area in square meters, and f is the frequency of operation in cycles per second. Equation (4.6) is rearranged as

$$N_p = \frac{E}{4fAB_m} \qquad (4.7)$$

since it is desired to calculate the number of turns N_p needed for the primary.

The inductance L_p of the primary may be determined from the following considerations. A toroidal core of cross-section A and mean length m is chosen. The flux density within the volume enclosed by the winding is [13]

$$B_m = \frac{\mu N_p I}{m} \qquad (4.8)$$

[13] Sears, Francis W., Electricity and Magnetism, Page 276, Addison-Wesley Press, Inc., 1951.

where B_m is the flux density in webers per square meter, μ is the permeability of the core in henrys per meter, N_p is the number of turns on the core, I_p is the current in the winding and m is the mean length of the core. The total flux in the core is,

$$\Phi = B_m A = \frac{\mu A N_p I_p}{m} \tag{4.9}$$

where Φ is the flux in webers, A is the cross-sectional area of the core in square meters and the other terms are the same as in equation (4.8). Assuming all of the flux links with each turn, then,

$$L_p = N_p \frac{\Phi}{I_p} = \frac{\mu A N_p^{\ 2}}{m} \quad \text{henrys} \tag{4.10}$$

where the units are given as in equation (4.9). The current in the primary may be obtained from,

$$e(t) = L_p \frac{di_p}{dt} \tag{4.11}$$

where e(t) is the induced voltage. From Figure 5b it may be seen that a total change of current $2I_p$ occurs in time $\frac{T}{2}$, thus,

$$\frac{di_p}{dt} = \frac{2I_p}{\frac{T}{2}} = \frac{4I_p}{T} = 4I_p f \quad , \tag{4.12}$$

and,

$$I_p = \frac{E}{4fL_p} \quad , \tag{4.13}$$

where I_p is the maximum current in the primary in amperes, E is the induced voltage in volts, f is the frequency of operation in

where B_m is the flux density in [illegible], μ is the permeability of the core [illegible], N is the number of turns on the core [illegible] area in [illegible] and m is the mean length of the core [illegible] total flux [illegible] the core is.

$$[illegible] \qquad (4.9)$$

where Φ is the flux in webers, A [illegible] of the core in square meters and the other terms [illegible] the same as in equation (4.8). Assuming all of the flux links with each turn, then,

$$L_p = [illegible] \qquad (4.10)$$

where the units are given as in equation (4.9). The [illegible] in the primary may be obtained from

$$e(t) = [illegible] \frac{di}{dt} \qquad (4.11)$$

where e(t) is the induced voltage. From Figure [illegible] may be seen that a total change of current [illegible]

$$\frac{di}{dt} = [illegible] \qquad (4.12)$$

and,

$$[illegible] \qquad (4.13)$$

where I_p is the maximum current in the primary [illegible] the induced voltage [illegible] is the frequency of operation [illegible]

cycles per second and L_p is the inductance of the primary in henrys.

To effect the switching in the transistors V_1 and V_2 , a feedback winding is added to the primary circuit and the transistor base circuit is completed, as in Figure 6a. A description of the switching action is as follows: As before, assume that when the voltage V_B is first applied, the leakage current through V_1 is slightly larger than the leakage current through V_2 . This larger current in the primary from terminal 2 to terminal 1 (Figure 6a) induces a voltage in the feedback winding N_{fb} such that terminal 4 is positive with respect to terminal 6. This forward bias on V_1 causes the current through V_1 to increase and the reverse bias on V_2 causes the current through V_2 to decrease. This polarity amounts to positive feedback which results in fast turn on for V_1. Current through the primary from terminals 2 to 1 increases linearly until core saturation is reached. When the core saturates, the induced voltage in the feedback winding drops to zero and V_1 is switched off. The rapidly collapsing field in the core induces a voltage in the feedback winding in the opposite direction (terminal 6 positive with respect to terminal 4) which further applies reverse bias to V_1 and forward bias to V_2. Current now flows in the primary from terminals 2 to 3, increasing linearly, until the core is now saturated in the negative direction. Oscillation conditions now exist and switching is repeated with the core being saturated first at $+ \Phi_m$ and then at $- \Phi_m$. The current also varies between $+ I_p$ and $-I_p$. Switching action occurs very

cycles per second and L_p is the inductance [illegible] in henrys.

To effect the switching of the transistors V_1 and V_2 a feedback winding is added to the [illegible] to the transistor base which [illegible] [illegible] description of the switching [illegible] [illegible]. As [illegible] assume that when the voltage V is first applied, the leakage current through V_1 is slightly [illegible] than the leakage current through V_2. This larger current in the primary from terminal 2 to terminal 1 (Figure 6) induces a voltage in the feedback winding N_{fb} such that terminal [illegible] is positive with respect to terminal 6. This forward bias on [illegible] causes the current through V_1 to increase and the reverse bias on V_2 causes the current through V_2 to decrease. This process is known as positive feedback which results in fast turn-on of [illegible]. Current through the primary from terminals [illegible] to 1 increases linearly until core saturation is reached. When the core saturates, the induced voltage in the feedback winding drops to zero and [illegible] is shut off. The rapidly collapsing field in the core induces a voltage in the feedback winding in the opposite direction (terminal [illegible] positive with respect to terminal 4). [illegible] bias to V_1 and forward bias to V_2. Current now flows in the primary from terminals 2 to 3 increasing linearly until the core is now saturated in the negative direction. [illegible] conditions now exist and switching is [illegible] with the core being saturated first [illegible] also varies between [illegible] and [illegible]. [illegible] action occurs [illegible]

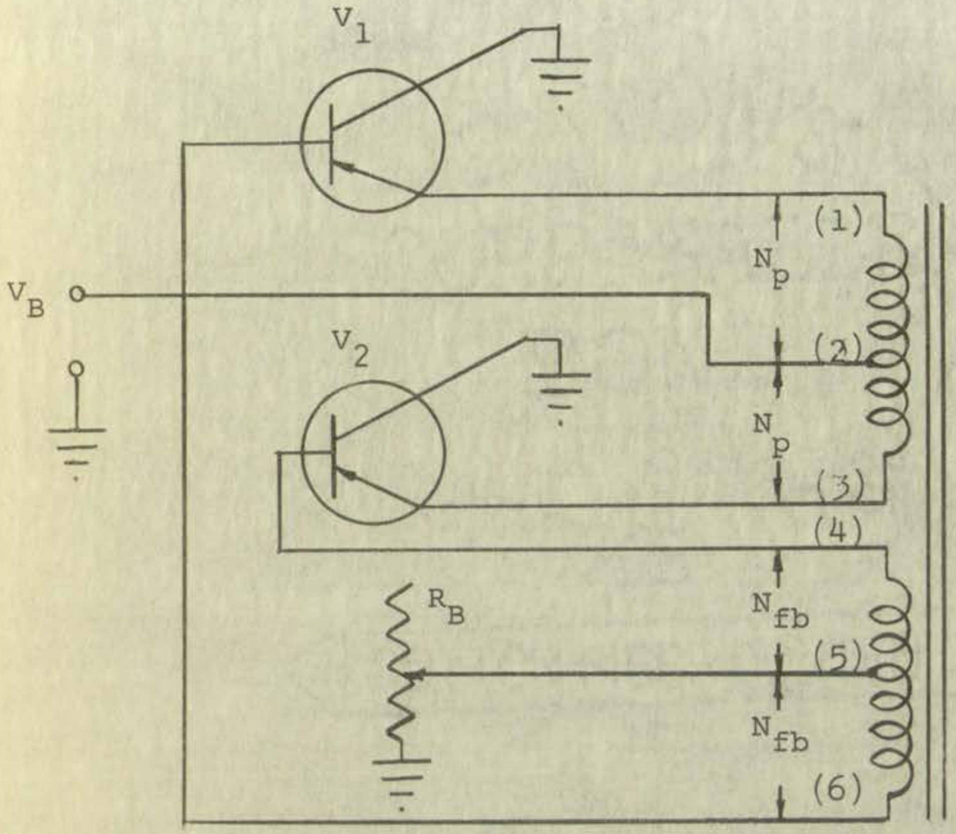

Figure 6a. Basic Primary Circuit plus the Feedback Winding.

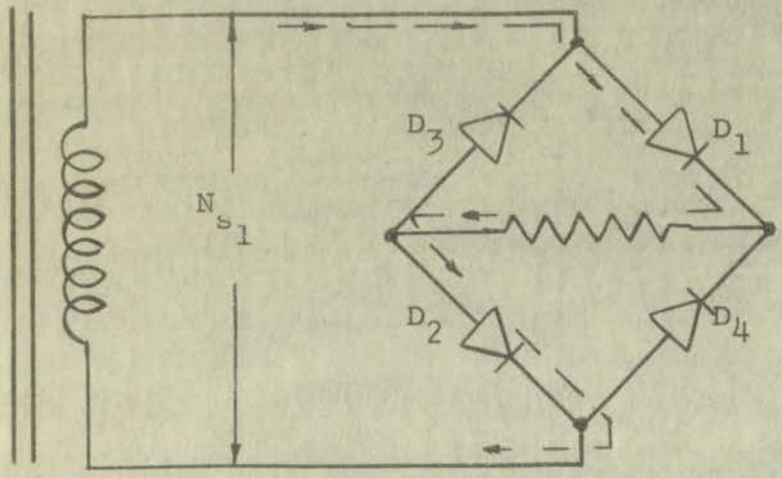

Figure 6b. Transformer Secondary with Bridge Rectifier.

rapidly and the induced primary voltage approaches a square wave depending mostly upon the hysteresis curve of the core. An ideal core would have a hysteresis curve shown in Figure 7a while Figure 7b shows a hysteresis loop of a nickel-iron core.

Now, to determine the number of turns, N_{fb} , it is necessary to know the maximum base-to-emitter voltage, $V_{be_{max}}$ which permits the peak primary current I_p to flow. This information depends upon the transistors V_1 and V_2 that are selected and may be obtained from the transistor characteristic curves. The number of turns, N_{fb} , is therefore,

$$N_{fb} = N_p\left(\frac{V_{be_{max}}}{E}\right) \tag{4.14}$$

where E is the power source voltage.

The value of the external base resistor R_b is,

$$R_b = \frac{V_{be_{max}} - V_{be}}{I_b} \tag{4.15}$$

One of the converter secondary circuits is depicted in Figure 5b. The proper turns ratio N_{s_1} is calculated from,

$$N_{s_1} = N_p\frac{E_{out}}{E_{in}} \times 1.05 \tag{4.16}$$

where N_p is the number of primary turns, E_{out} is the required DC voltage, E_{in} is the voltage of the power source and 1.05 is a factor to compensate for the I^2R losses. A full wave bridge rectifier follows the secondary winding. The bridge rectifier is chosen because, among other things, the peak inverse voltage across a nonconducting diode rectifier is E_{out} rather than $2E_{out}$ which would exist in the standard full wave rectifier. During

rapidly and the induced current becomes approximately a square wave depending mostly upon the rectangular figure of the core. An ideal core would have a hysteresis loop as shown in Figure 7a, while Figure 7b shows a hysteresis loop of a practical core.

Now, to determine the number of turns, N_{fb}, it will be necessary to know the maximum base-to-emitter voltage, V_{bem}, which permits the peak primary current I_p to flow. This information depends upon the transistors T_1 and T_2 that are selected and may be obtained from the transistor characteristic curves. The number of turns, N_{fb}, is therefore

$$N_{fb} = N_p \frac{V_{bem}}{E} \qquad (4.14)$$

where E is the power supply voltage.

The value of the external base resistor, R_b, is

$$R_b = \frac{V_{bem}}{[illegible]} \qquad (4.15)$$

One of the converter secondary circuits is depicted in Figure 5b. The proper turns ratio, N, is calculated from

$$N_s = N_p \frac{E_{out}}{E_{in}} \times 1.05 \qquad (4.16)$$

where N_p is the number of primary turns, E_{out} is the required DC voltage, E_{in} is the voltage of the power source and 1.05 is a factor to compensate for the IR losses. A full wave bridge rectifier follows the secondary winding. The bridge rectifier is chosen because, among other things, the peak inverse voltage across a nonconducting diode rectifier is E_{out} rather than $2E_{out}$ which would exist in the standard full wave rectifier. During

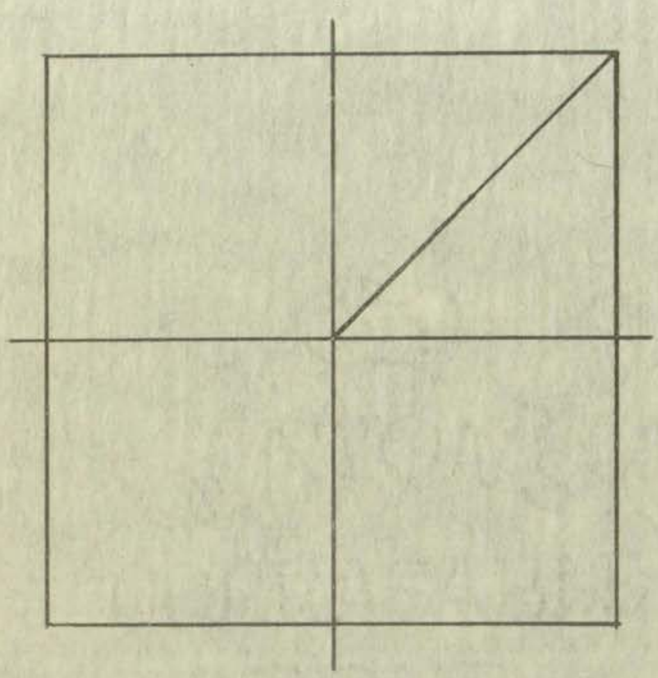

Figure 7a. Hysteresis Curve of an Ideal Core

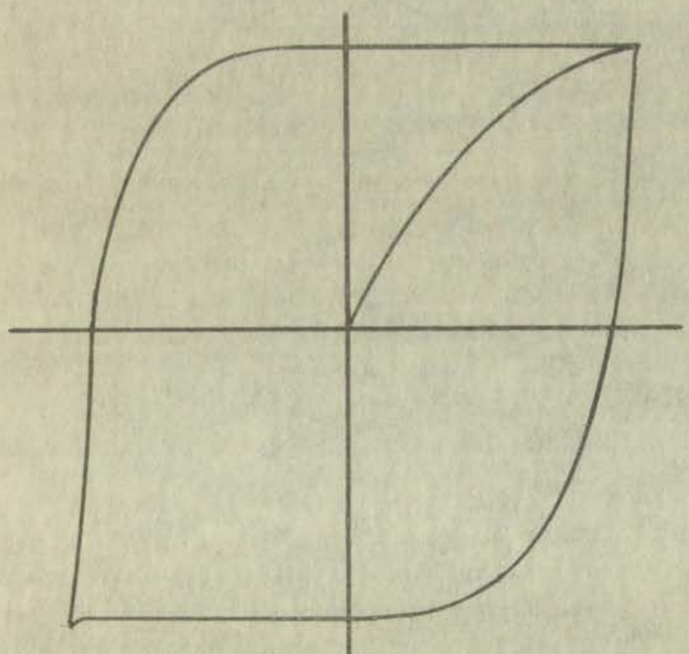

Figure 7b. Hysteresis Curve of a Nickel-Iron Core

each half cycle the circuit provides a path for current flow in alternative directions through the supply, but in the same direction through the load R_L . If the voltage across the conducting diode is considered to be zero, it is obvious, from Figure 6b, that the inverse voltage across a nonconducting diode is equal to the load voltage.

For sufficient filtering of the rectifier output voltage, the value of the capacitor C should be large enough to produce a time constant (in combination with the load R_L) that is at least one-half of the period of the power converter switching frequency, therefore,

$$C > \frac{1}{4\pi R_L f} \qquad (4.17)$$

where C is the capacitance in farads, R_L is the load in ohms, and f is the switching frequency in cycles per second of the power converter.

each half cycle the capacitor provides a path for current flow in alternative directions through the supply but in the same direction through the load R_L. If the voltage across the conducting diode is considered to be zero, it is obvious from Figure 6b, that the inverse voltage across a nonconducting diode is equal to the load voltage.

For sufficient filtering of this rectified output voltage, the value of the capacitor C should be large enough to produce a time constant (in combination with the load R_L) that is at least one-half of the period of the power converter switching frequency, therefore:

$$\frac{1}{4R_Lf} \qquad (4.17)$$

where C is the capacitance in farads, R_L is the load in ohms, and f is the switching frequency in cycles per second of the power converter.

CHAPTER V -- THE SAWTOOTH VOLTAGE GENERATOR

A compact, light-weight sawtooth voltage generator is desirable. Also, it should consume a minimum amount of power. In view of the above criteria, a sawtooth voltage generator designed around a "four-layer diode" has advantages.

The "four-layer diode" serves as a semiconductor switch. It is a two-terminal device which operates electrically in either of two states: an open or high-resistance state (greater than 10 megohms) and a closed or low-resistance state (less than 10 ohms). An applied voltage of sufficient magnitude results in switching action. The "switch" is then opened by either reverse bias or a decrease in current to a value insufficient to maintain conduction. Very rapid switching action is possible; approximately 50 nanoseconds (50×10^{-9} seconds) to close and 500 nanoseconds to open.

The diode symbol utilizes the symbol four and the slant line of the four gives the forward current direction of the device when the "switch" is closed.

Figure 8 shows a schematic for the sawtooth voltage generator. When the voltage V is applied, the capacitor C charges through R_1 until the voltage across the diode (in the high-resistant state) is sufficient to cause the diode to switch to the low-resistance state.

The LaPlace transformed differential equation which describes the voltage across the capacitor C follows. Its

CHAPTER V -- THE AVALANCHE VOLTAGE GENERATOR

A compact, light weight avalanche voltage generator is desirable. Also, it should consume a minimum amount of power. In view of the above criteria, a avalanche voltage generator designed around a "four-layer diode" has advantages.

The "four-layer diode" or p-n-p-n junction switch is a two-terminal device which switches discontinuously to either of two states: an open or high-resistance state (greater than 10 megohms) and a closed or low-resistance state (less than 10 ohms). An applied voltage of sufficient magnitude results in switching action. The "switch" is then opened by either reverse bias or a decrease in current to a value insufficient to maintain conduction. Very rapid switching action is possible; approximately 50 nanoseconds (50×10^{-9} seconds) to close and 500 nanoseconds to open.

The diode symbol utilizes the typical form and the slant line of the four gives the forward current direction of the device when the "switch" is closed.

Figure 8 shows a schematic for the avalanche voltage generator. When the voltage V is applied the capacitor C charges through R_1 until the voltage across the diode (in the high-resistant state) is sufficient to cause the diode to switch to the low-resistance state.

The LaPlace transform of the differential equation which describes the voltage across the capacitor is as follows:

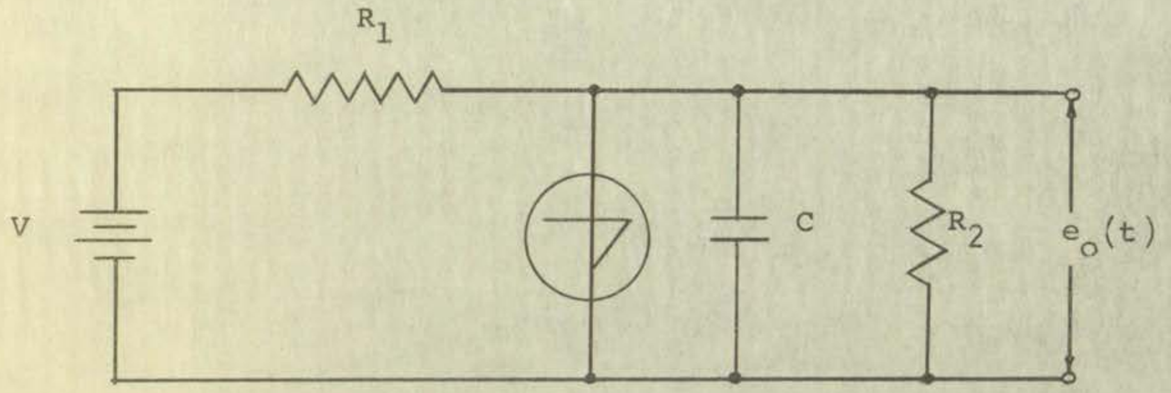

Figure 8. Schematic Diagram of the Sawtooth Voltage Generator.

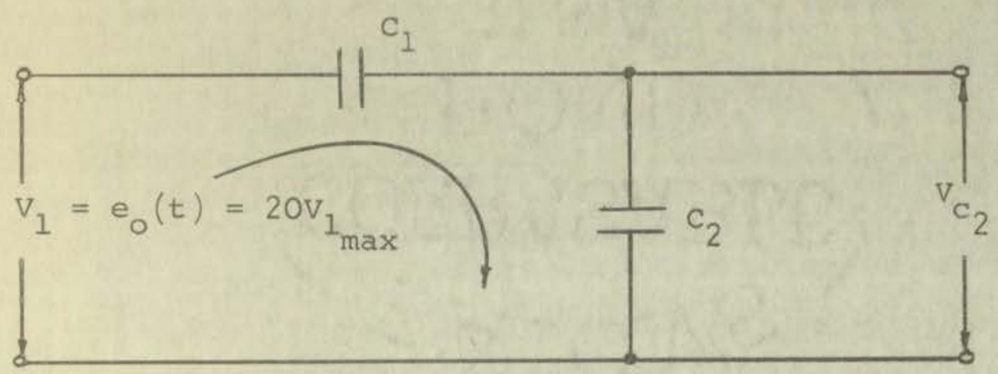

Figure 9. Capacitive Voltage Divider for the Sawtooth Voltage Generator.

solution is rearranged to determine the period t of the sawtooth voltage. The approximate average power consumed by the sawtooth voltage generator is also calculated.

Assuming the diode is "open," (high-resistance state) the following equation describes the voltage $e_o(t)$ in the LaPlace transformation $E_o(s)$:

$$E_o(s) = V(s) \left[\frac{\dfrac{R_2/Cs}{R_2 + \frac{1}{Cs}}}{R_1 + \dfrac{R_2/Cs}{R_2 + \frac{1}{Cs}}} \right] \tag{5.1}$$

Thus,
$$E_o(s) = \frac{V(s)R_2}{R_1R_2Cs + R_1 + R_2} \; . \tag{5.2}$$

Since,
$$V(s) = \frac{V}{s} \, , \tag{5.3}$$

$$E_o(s) = \frac{VR_2}{s(R_1R_2Cs + R_1 + R_2)} \tag{5.4}$$

and,
$$E_o(s) = \frac{VR_2}{(R_1 + R_2)s} - \frac{\dfrac{R_2V}{R_1 + R_2}}{s + \dfrac{R_1 + R_2}{R_1R_2C}} \tag{5.5}$$

Therefore:
$$e_o(t) = \frac{VR_2}{R_1 + R_2} \left[1 - \exp(- \frac{R_1 + R_2}{R_1R_2C})t \right] \tag{5.6}$$

Then, rearrange equation (5.6) and solve for t,

$$\exp\left[-\frac{R_1 + R_2}{R_1 R_2 C} t\right] = 1 - \frac{e_o(t)\left[R_1 + R_2\right]}{VR_2}, \quad (5.7)$$

and,

$$-\frac{R_1 + R_2}{R_1 R_2 C} t = \ln\left[1 - \frac{e_o(t)\;\; R_1 + R_2}{VR_2}\right], \quad (5.8)$$

Thus,

$$t = -\frac{R_1 R_2 C}{R_1 + R_2} \ln\left[1 - \frac{e_o(t)\left[R_1 + R_2\right]}{VR_2}\right], \quad (5.9)$$

which gives the period t of the sawtooth voltage generator.

Choosing R_1 quite large, perhaps greater than 1 megohm, will result in very low power consumption for the sawtooth generator. The average power consumed by the sawtooth voltage generator is approximately,

$$P = \frac{V^2}{R_1}, \quad (5.10)$$

Thus,

$$P = \frac{300^2}{1 \times 10^6}, \quad (5.11)$$

and,

$$P = 90 \text{ milliwatts} \quad (5.12)$$

The modulation sensitivity of the klystron oscillator is of the order of 1.5 mc/volt for repeller voltage "electronic tuning." Since the sawtooth voltage generator produces a maximum output voltage of approximately 20 volts, and it is desired to frequency

modulate the oscillator a maximum of perhaps 10 mc, it is necessary to include an appropriate voltage divider in the output circuit of the sawtooth generator. A capacitive divider is considered. (See Figure 9.) The equation relating voltage V_1 and current I is

$$V_1 = (Z_1 + Z_2)I \tag{5.13}$$

$$I = \frac{V_1}{Z_1 + Z_2} \tag{5.14}$$

$$V_{C_2} = Z_2 I \tag{5.15}$$

$$V_{C_2} = \frac{Z_2 V_1}{Z_1 + Z_2} \tag{5.16}$$

$$V_{C_2}(s) = \frac{\frac{1}{sC_2}V_1(s)}{\frac{1}{sC_1} + \frac{1}{sC_1}} \tag{5.17}$$

$$V_{C_2}(s) = \frac{V_1(s)}{sC_2} \cdot \frac{sC_1 sC_2}{sC_1 + sC_2} \tag{5.18}$$

$$V_{C_2}(s) = \frac{V_1(s)C_1}{C_1 + C_2} \tag{5.19}$$

Thus it is possible to set V_{C_2} to the desired voltage by assigning a value to C_2 and calculating C_1 . If C_1 and C_2 are small, the impedance will be high, which prevents loading of the sawtooth generator.

modulate the oscillator [illegible] it is necessary to include [illegible] in the output circuit of the sawtooth generator. A [illegible] is considered. (See Figure [illegible]) [illegible] current I is

[illegible] (5-15)

[illegible] (5-16)

[illegible] (5-17)

[illegible] (5-18)

[illegible] (5-19)

[illegible] (5-20)

Thus it is possible to [illegible] by assigning a value to C_1 and calculating [illegible] If C_1 [illegible] C_2 are small, the impedance will be high [illegible] of the sawtooth generator.

CHAPTER VI -- THE KLYSTRON OSCILLATOR

The reflex klystron is chosen to generate the microwave carrier because it is compact, rugged, and relatively easy to tune both mechanically and electronically. This results from the fact that the reflex klystron uses one simple resonant circuit which is contained within the vacuum envelope.

In this application of the reflex klystron as an oscillator the anode or resonator voltage is held constant and the variation of the frequency of oscillation is achieved by modulating the repeller. The change in output frequency with applied repeller voltage is referred to as "electronic tuning" and results in the generation of the frequency modulated carrier. The repeller voltage is negative with respect to the cathode and draws little if any current. This characteristic is of great advantage since very little, if any, power is required in the modulation circuit.

The transit-times for the electrons emitted by the cathode of the reflex klystron depend upon both the repeller-cathode and anode-cathode voltages. Oscillation of the klystron occurs for some combinations of these voltages and does not occur for others. When the anode-cathode voltage is held constant and the repeller-cathode voltage is varied, typical reflex klystron output power and frequency of oscillation vary as shown in Figure 10. As stated, the frequency of oscillation is quite sensitive to the applied repeller voltage. Frequency modulation sensitivity of about 1.5 mc/volt is typical. Changes in the repeller voltage, which produce frequency modulation of the reflex klystron oscillator, are achieved by superimposing the

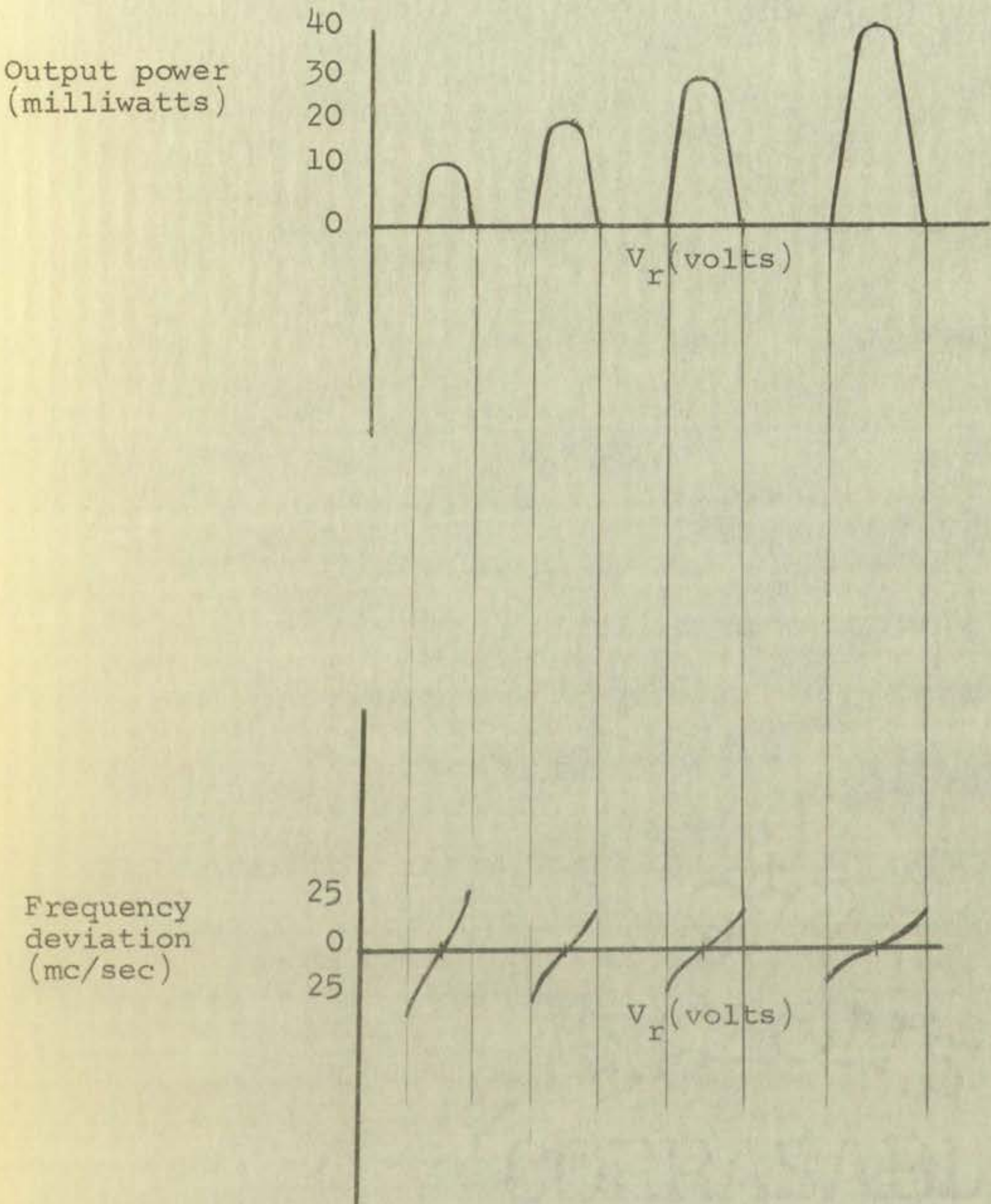

Figure 10. Reflex Klystron Output Power and Frequency Deviation as Functions of Repeller Voltage. Anode Voltage is Constant.

voltage from the sawtooth generator upon the repeller voltage. The reflex klystron oscillator frequency variation, output power and repeller voltage, all as functions of time, are shown in Figures 11a, b, and c respectively.

It is desirable to produce this frequency modulated signal from the reflex klystron with as little amplitude modulation as possible. Information about the index of refraction is contained in the amplitude modulation of the frequency modulated signal; therefore, amplitude modulation other than that produced by the resonant cavity is detrimental. As can be seen in Figures 10 and 11, the output power from the klystron varies with time. This results in amplitude modulation of the frequency modulated signal. Fortunately, the klystron has to be frequency modulated only a few megacycles in its application in the microwave refractometer and this frequency variation can be centered around the peak of the mode of oscillation. The result, of course, is a minimum of this undesirable amplitude modulation produced by the reflex klystron.

One of the major factors of concern in utilizing the reflex klystron in the microwave refractometer transmitter is the relatively large amount of power that must be dissipated as heat. To illustrate, consider the 2K25 reflex klystron. With a resonator potential of 300 VDC, the resonator current is approximately 20 milliamperes (repeller voltage is - 160 VDC). The microwave output power with these voltages is approximately 20 milliwatts. The filament power is approximately 2.5 watts. Input power to the resonator circuit is 6 watts which essentially

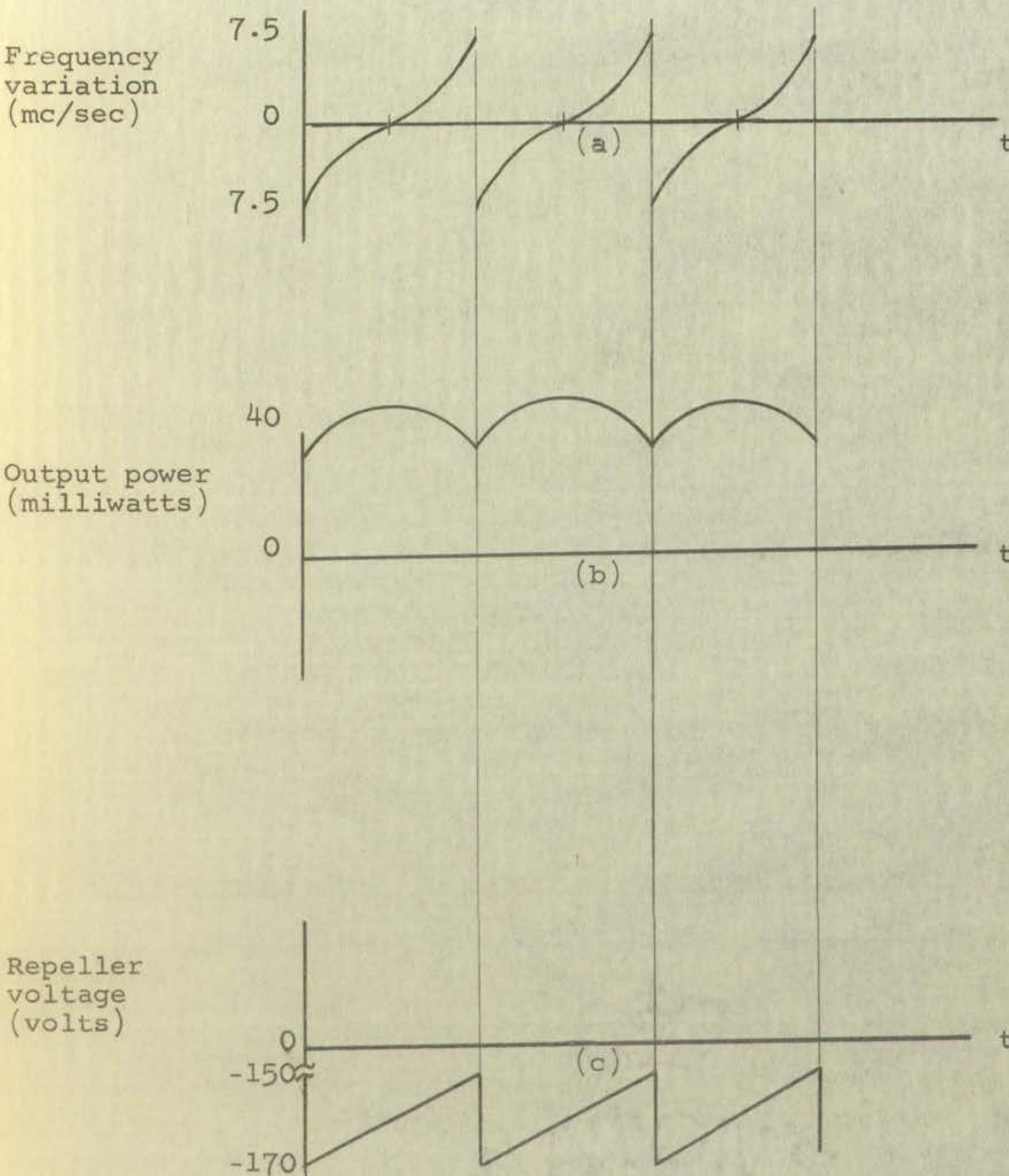

Figure 11 . Klystron Oscillator Frequency Variation,
Output Power and Repeller Voltage as Functions of Time.

must be dissipated along with the filament power as heat. At an altitude of 200,000 feet, where very little atmosphere is present to provide convection cooling, a prodigious problem obviously exists. This problem has not as yet been solved, but it is believed that there are several possible solutions, including heat sinks, cooling fins, reduced filament voltage, etc., or a combination of these.

CHAPTER VII -- THE RESONANT CAVITY

The resonant frequency f_o of a microwave cavity is determined by the dimensions of the cavity, the mode of operation and the composition of the material within the cavity. The resonant frequency f_o can be determined by,

$$f_o = \frac{k}{n} \tag{7.1}$$

where k is a constant determined by the mode of operation and the cavity dimensions and n is the index of refraction.

The empirical formula for computing the index of refraction of the atmsophere from meteorological data is,[14,15]

$$n = 1 + 77.6\left(\frac{P}{T} + \frac{4801e}{T^2}\right)10^{-6} \tag{7.2}$$

where T is the absolute temperature in degrees Kelvin, P is the pressure in millibars and e is the partial pressure of water in millibars. The index of refraction of free space (vacuum) is unity. The range of n for the atmosphere of the earth is considered to be,

$$1.00000 \leq n \leq 1.000500 \tag{7.3}$$

where the upper limit is closer to 1.000300 for practical situations. To facilitate the use of this somewhat awkward range

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[14] Smith, E.K., and S. Weintraub, "Constants in the Equation for Atmospheric Refractive Index at Radio Frequencies," Proceedings of the IRE, Vol. 41, No. 8, p. 1035 (August, 1953).

[15] Cramond, W.R., and D.C. Thorn, An Analog Computer for Calculating Atmospheric Density and Refractive Index, University of New Mexico Engineering Experiment Station, Albuquerque, New Mexico, Technical Report EE-58, 1961, for further discussion of this empirical formula.

CHAPTER VII -- THE RESONANT CAVITY

The resonant frequency f_o of a microwave cavity is determined by the dimensions of the cavity, the mode of operation and the composition of the material within the cavity. The resonant frequency f_o can be determined by,

$$f_o = \frac{k}{n} \qquad (7.1)$$

where k is a constant determined by the mode of operation and the cavity dimensions and n is the index of refraction.

The empirical formula for computing the index of refraction of the atmosphere from meteorological data is,[14,15]

$$n = 1 + 77.6\left(\frac{P}{T} + \frac{4810e}{T^2}\right)10^{-6} \qquad (7.2)$$

where T is the absolute temperature in degrees Kelvin, P is the pressure in millibars and e is the partial pressure of water in millibars. The index of refraction of free space (vacuum) is unity. The range of n for the atmosphere of the earth is considered to be,

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of values define refractivity, N, by

$$N = (n - 1) \times 10^6 \tag{7.4}$$

This definition results in the following range of N,

$$000.0 \leq N \leq 500.0 \tag{7.5}$$

for the index of refraction of the earth's atmosphere ranging from 1.000000 to 1.000500. If the operating frequency f_o is set at 10 gigacycles (10×10^9 cps), for n = 1 or N = 0,

then, $$f_o = 10 \text{ gc} = \frac{k}{\frac{N}{10^6} + 1} \tag{7.6}$$

and, $$10 \text{ gc} = \frac{k}{\frac{0}{10^6} + 1} \tag{7.7}$$

so, $$k = 10 \text{ gc} \tag{7.8}$$

Thus, $$f_o = \frac{10 \text{ gc}}{\frac{N}{10^6} + 1} \tag{7.9}$$

and, for N = 500, $$f_o = \frac{10 \text{ gc}}{\frac{500}{10^6} + 1} \tag{7.10}$$

or, $$f_o \cong 9.995 \text{ gc} \tag{7.11}$$

This, of course, is about 5 megacycles per 500 N-units or about 10 kilocycles per N-unit.

Consider the bandwidth (3 db points) of the resonant cavity as,

$$BW = \frac{f_o}{Q} \tag{7.12}$$

where f_o is the cavity resonant frequency and Q is the quality factor or the energy stored in the cavity divided by the energy loss in the cavity per cycle. If the resonant frequency f_o is 10 gc and it is desired to determine small changes in f_o corresponding to a few N-units then obviously a very high Q is required. Amplitude modulation of the FM carrier is effected by the resonant cavity. To obtain sufficient amplitude modulation such that peaks of the envelope may be determined, the bandwidth of the resonant cavity must certainly be considerably less than the swept frequency range of the frequency modulated signal. The minimum swept frequency range for 500 N-units, for base $f_o = 10^{10}$ cp, is 5 megacycles. A swept frequency range of 7 megacycles is chosen and the 500 N-units in the range $100 < N < 600$ are utilized. If a Q of 10,000 is obtained,

$$BW = \frac{f_o}{Q} \tag{7.12}$$

and

$$BW = \frac{10 \times 10^9}{10 \times 10^3} \tag{7.13}$$

Then,

$$BW = 1 \times 10^6 \text{ cps} \tag{7.14}$$

which is considerably less than the swept frequency range of 5 megacycles necessary for the 500 N-units.

The volume-to-surface ratio of the resonant cavity is to some extent an indication of the Q of the cavity.[16] The right circular cylinder (perhaps with modifications to prevent air

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[16] Thorn, Donald C., Design of an Open-Ended Resonant Cavity, The University of Texas, Austin, Texas, 1958.

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$$BW = \frac{f_o}{Q} \qquad (4.12)$$

and

$$BW = \frac{10 \times 10^9}{10 \times 10^3} \qquad (4.13)$$

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stagnation) is aerodynamically suitable. The basic configurations for the resonant cavity under consideration are shown in Figures 12 and 13.

It is possible to use either a transmission or an absorption cavity. If a transmission type cavity is used the output of the klystron oscillator (the FM signal) is fed through the resonant cavity resulting in an output from the cavity only at the resonant frequency of the cavity. Transmission of the FM signal other than at this resonant frequency is essentially cut off. With the absorption cavity the output of the klystron oscillator is fed past the resonant cavity in such a manner that the energy is absorbed only at the cavity resonant frequency and transmission of all other portions of the FM spectrum results.

In order to provide the more suitable input voltage for the discriminator in the receiver, an absorption type cavity seems desirable. This would mean that the discriminator input is relatively constant except for the portion of the frequency modulated signal that appears within the bandwidth of the resonant cavity. Of course, energy absorption takes place in this portion of the frequency spectrum. Consequently, the discriminator has an input voltage over most of the swept frequency range and is not required to lock in on narrow pulses which would be the case if a transmission type cavity is used.

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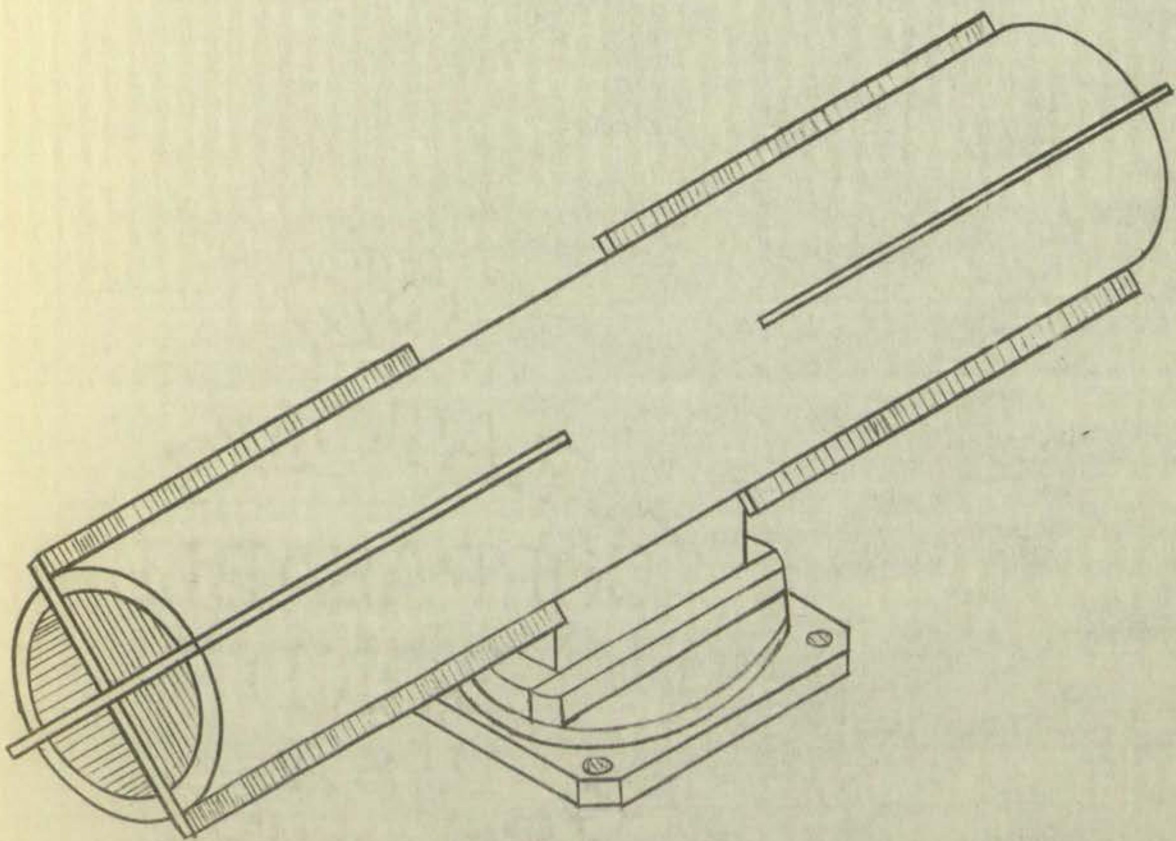

Figure 12. Resonant Cavity Configuration.

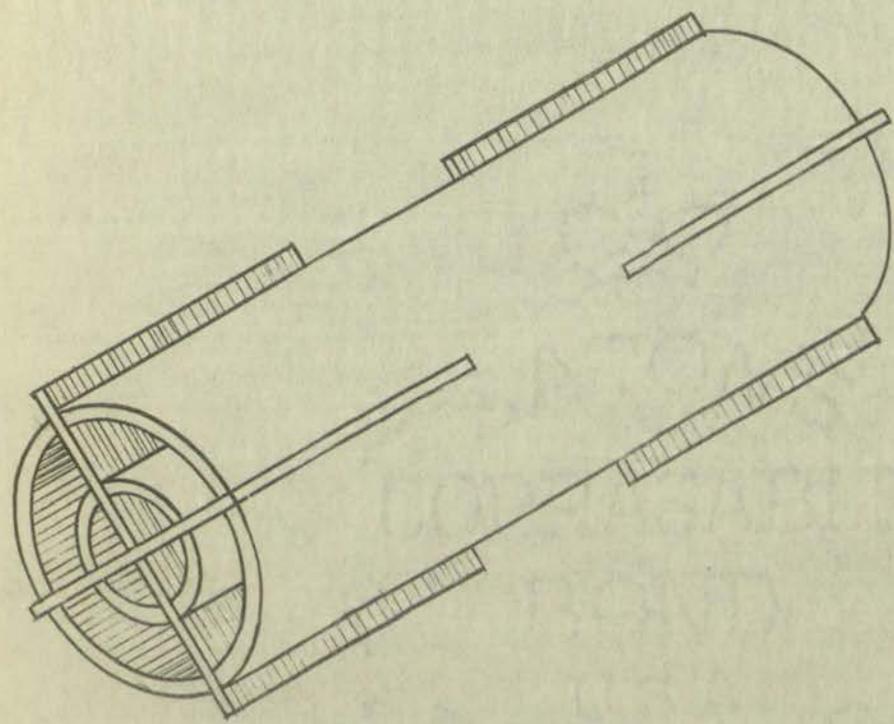

Figure 13. Another Resonant Cavity Configuration.

CHAPTER VIII -- ANTENNA AND REFLECTOR

"The gain of an antenna expresses its effectiveness in concentrating power in a given direction. It is the ratio of the peak intensity of the pattern to the intensity of an isotropic radiator emitting the same amount of power."[17/] Since the microwave refractometer transmitter has relatively low power output (perhaps a maximum of 200 milliwatts operating CW), the first inclination may be to consider a very high gain antenna in combination with a reflector that contributes additional directional gain in an effort to produce maximum radiated power. However, a more thorough study of the microwave refractometer transmitter indicates that the maximum gain that can be obtained with the antenna and reflector is not necessarily the most desirable result. The reason for this is that the transmitter is designed to be suspended beneath a parachute, which is metalized and serves as the reflector. The descent of the system other than exactly vertical descent could result in lack of reception of the transmitted signal if the beam width of the radiation pattern is quite narrow due to a highly directive antenna and reflector system. To illustrate, consider a parabolic parachute that is 20 feet in diameter. The gain associated with radiation of 3 centimeters wavelength is approximately 53 db and the beam width at the 3 db points of the radiation pattern is approximately 0.3°.[18/]

[17/] Montgomery, Carol G., Techniques of Microwave Measurements, Radiation Laboratory Series, Vol. II, Page 899, McGraw-Hill Book Co., Inc., New York, 1947.

[18/] Reference Data for Radio Engineers, International Telephone and Telegraph Co., Fourth Edition, Third Printing, Stafford Press, New York, 1957.

CHAPTER [illegible] ANTENNA AND REFLECTOR

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[17] Montgomery, Carol G., Technique of Microwave Measurements, Radiation Laboratory Series, [illegible] Hill Book Co., Inc., New York, [illegible]

[18] Reference Data for Radio Engineers, International Telephone and Telegraph Corp., [illegible] Edition, Third Printing, Stafford Press, New York, [illegible]

This highly directional gain would be extremely useful. However, to maintain the reflector and antenna in the proper organization, with respect to the receiver at the ground station such that this beam width includes the receiving antenna, appears to be quite a formidable problem. Figure 14 gives the approximate gain in db versus the beam width in degrees of a parabolic reflector. It can be seen from this curve that, even if it is necessary to create a beam width of as much as 20 or 30 degrees at the 3 db points, it is still possible to expect a few db of gain. The curve is based upon placement of the antenna at the focal point of the parabola. If the antenna is located away from the focal point, a few db of gain should still result.

Now consider the antenna or "feed system" for the reflector. Various directional antennas including the Yagi antenna, the horn and the helical antenna are possibilities. In any case, the antenna that "illuminates" the reflector should direct its energy such that it contributes to the directivity of the reflector. The directional characteristics of the antenna should be such that a minimum amount of energy is wasted and at the same time the desired distribution of field intensity is created across the reflector. A loosely wound (relatively large pitch) helical antenna of appropriate dimensions, which is fitted with a reflecting screen (perpendicular to the axis of the helix) at the input end, produces a radiation pattern that is highly directive along the axis of the helix and is also circularly polarized. This circular polarization is advantageous since the wave front directed by the reflector has polarization in the opposite direction,

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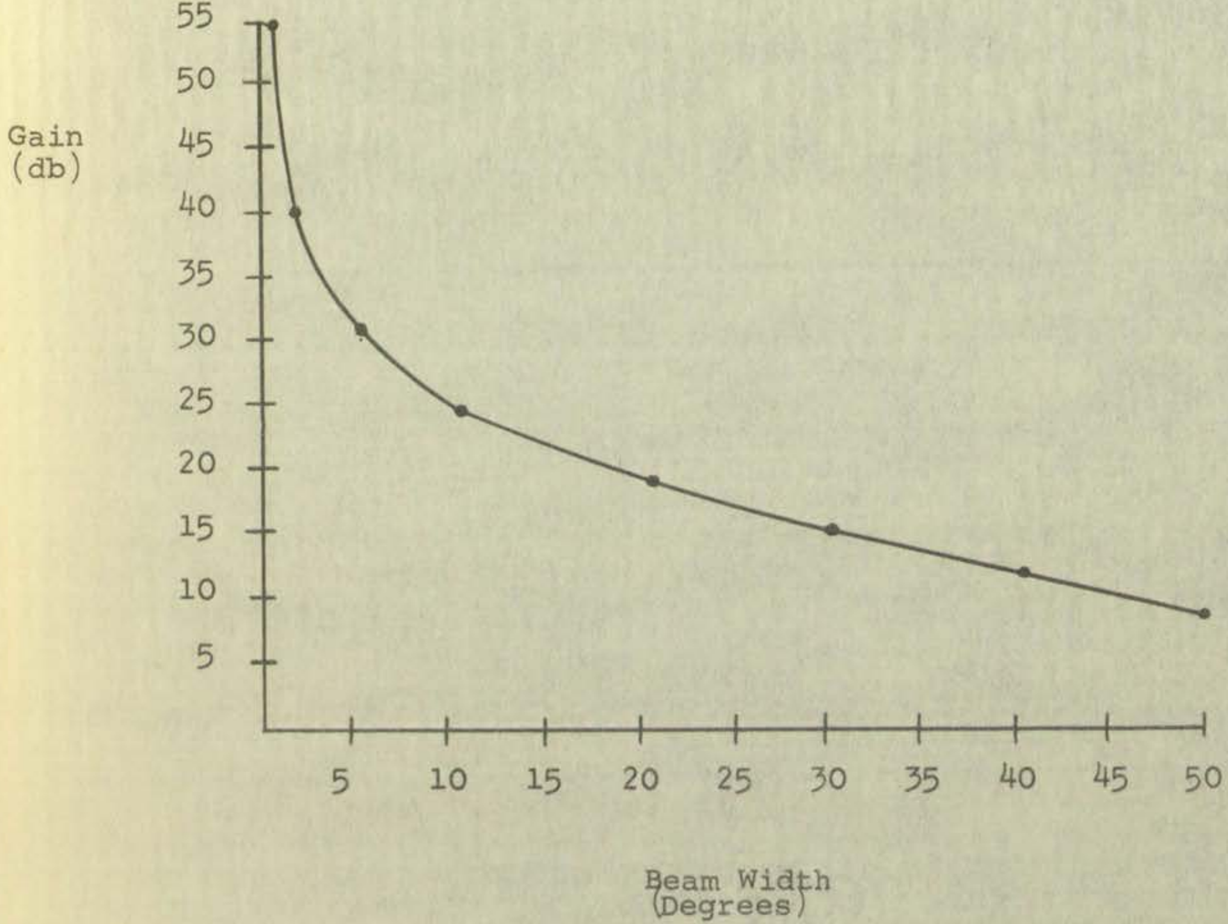

Figure 14. Approximate Gain versus Beam Width of a Parabolic Reflector.

thereby minimizing the interference caused by the helix being positioned in the path of transmission. The directive gain of the helical antenna is appreciable; a six-turn helix with a diameter of 0.30λ and a pitch of 0.3λ develops a power gain of approximately 16.5 db when utilized with a reflecting screen that is positioned perpendicularly to the helix axis at the input end.[19/] If this configuration of helical antenna is used as the feed system for the reflector additional gain is realized which results from the directivity of the reflector. The gain of the helical antenna can be made as large as possible resulting in maximum energy transfer to the reflector which in turn is utilized to shape the radiation pattern such that the required beam width exists. Unless a system can later be evolved for orientation of the package, it appears that a helix alone will be adequate; however, packaging is easier with a reflector.

[19/] Terman, Frederick E., Electronics and Radio Engineering, Fourth Edition, Page 909, McGraw-Hill Book Co., Inc., New York, 1955.

CHAPTER IX -- EXPERIMENTAL MODELS

The purpose of the experimental model is to provide a suitable system which will permit a study of the theory of operation of the microwave transmitter. The main point of interest is the generation of the desired AM-FM signal. This model will also allow an investigation of the individual operation of the DC-to-DC converter and sawtooth voltage generator as well as the simultaneous operation of these components in combination with the reflex klystron oscillator.

The experimental model transmitter consists of a 6 volts automobile battery for the power source, a DC-to-DC converter which supplies the high voltages for the sawtooth voltage generator and the reflex klystron oscillator, a crystal detector to view and measure the output of the oscillator, a 6 db attenuator to prevent "pulling" of the oscillator by the tunable resonant cavity, an additional 6 db attenuator followed by an additional crystal detector to view and measure the output of the resonant cavity, and finally, a waveguide termination. Figure 15 shows a simplified block diagram of the experimental model transmitter. The klystron was a 2K25, manufactured by RCA, and the cavity was one manufactured by Polytechnic Research and Development, having an experimental Q of approximately 9000.

First, calculations for the DC-to-DC converter follow. Also, selection of the various elements of the converter is detailed. The schematic of the experimental model transmitter is given in Figure 16. The various elements of the transmitter are identified and labeled.

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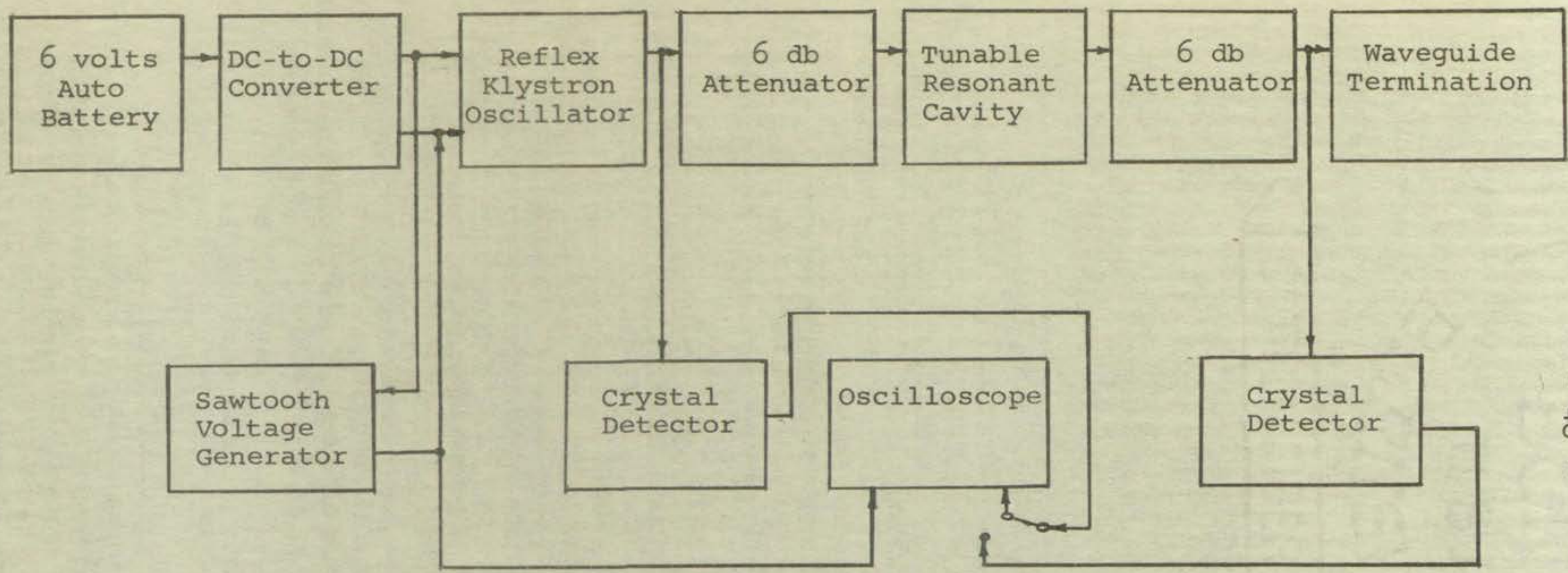

Figure 15. Simplified Block Diagram of the
Experimental Model Transmitter.

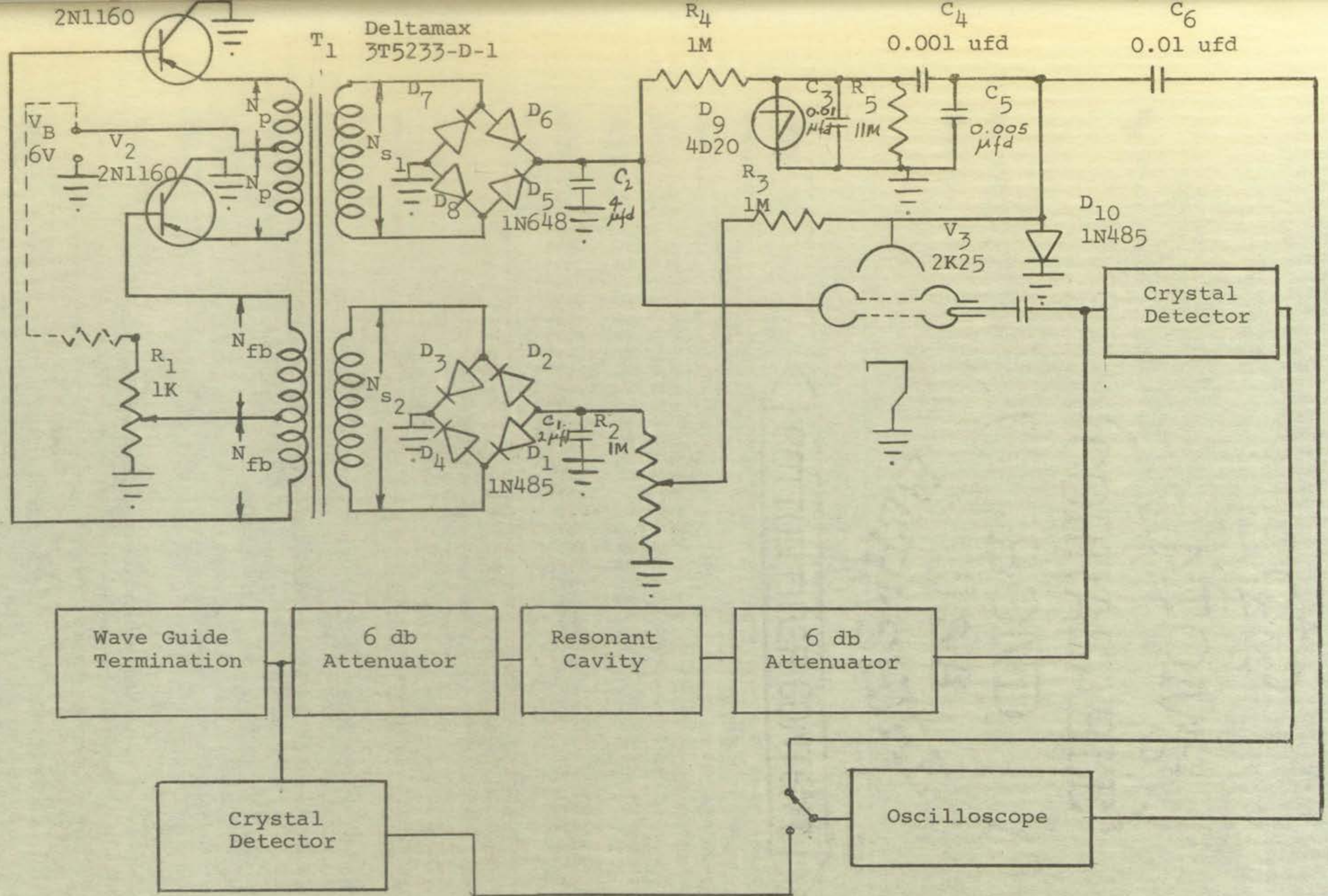

Figure 16. Schematic of the Experimental Model Transmitter.

Equation (4.7) is utilized to determine the number of turns N_p of the primary circuit,

$$N_p = \frac{E}{4B_m Af} \qquad (4.7)$$

In order to calculate N_p, the transformer core must be selected. The net core area may be roughly approximated from [20/]

$$A_C = \frac{\sqrt{W_{out}}}{5.58}\sqrt{\frac{60}{f}}\ \text{in}^2 , \qquad (9.1)$$

where A_C is the core area in square inches, W_{out} is the power output of the converter in watts, and f is the frequency of operation in cycles per second. A suitable compromise between core loss and physical size of the filter components, with respect to frequency of operation is 1250 cycles per second. The power supply output required by the klystron oscillator and saw-tooth voltage generator is approximately 6 watts. Thus,

$$A_C = \frac{\sqrt{6}}{5.58}\sqrt{\frac{60}{1250}} , \qquad (9.2)$$

and

$$A_C = 0.096 \quad \text{in}^2 . \qquad (9.3)$$

The Deltamax core #3T5233-D-1 has a core area of 0.094 square inches, which is suitable. Since A, f and B_m (given as 14 K Gauss for the Deltamax core by the manufacturer, Arnold Engineering Co.) are now determined,

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[20/] Reference Data for Radio Engineers, International Telephone and Telegraph Corporation, Fourth Edition, Third Printing, Page 274, March, 1957.

Equation (8.7) is utilized to determine the number of turns N_p of the primary winding.

$$N_p = \text{[illegible]}$$

In order to calculate N_p, the size of the transformer core must be selected. The net core area may be roughly approximated from:[20]

$$A_c = \text{[illegible]}$$

where A_c is the core area in square inches, P_{out} is the power output of the converter in watts, and f is the frequency of operation in cycles per second. A suitable compromise between core loss and physical size of the core leads to the choice, with respect to frequency of operation, of 4250 cycles per second. The power supply output required by the horizontal deflection and saw-tooth voltage generator is approximately 5 watts. Thus,

$$A_c = \text{[illegible]}$$

and

$$A_c = 0.0\text{[illegible]} \text{ in.}^2$$

The Deltamax core #5T5275-D1 has a core area of 0.0[illegible] square inches, which is suitable. Since N_p and B_m (given in a data sheet for the Deltamax core by the manufacturer, Arnold Engineering Co.) are now determined.

[20] Reference Data for Radio Engineers, International Telephone and Telegraph Corporation, Fourth Edition, Second Printing, Page 274, March, 1957.

$$N_p = \frac{6 \times 10^8}{4(14 \times 10^3)(6.45 \times 0.094)(1250)} \tag{9.4}$$

which includes the appropriate conversion factors resulting in consistent units,

$$N_p = 14.2 \quad \text{turns} \tag{9.5}$$

The inductance L_p of the primary is , from equation (4.10),

$$L_p = \frac{4\pi N_p^{\ 2}}{10^9} \times \frac{\mu A}{m} \quad , \tag{4.10}$$

$$L_p = \frac{4\pi(14)^2}{10^9} \times \frac{(12000)(0.094)(6.45)}{(0.25)(2.54)} \tag{9.6}$$

which includes the permeability, μ, as 12,000 and the mean length, m, of the core as 0.25 inches (furnished by the core manufacturer),

Therefore, $$L_p = 27.9 \text{ millihenrys.} \tag{9.7}$$

The primary exciting current I_p is, from equation (4.11),

$$I_p = \frac{E}{4fL_p} \quad . \tag{4.13}$$

When the transformer is loaded the secondary current is added, taking into consideration the expected efficiency of the transformer. Thus,

$$I_p = \frac{E}{4fL_p} + \frac{I_{AV}}{2} \quad , \tag{9.8}$$

and, $$I_p = \frac{6}{4(1250)(0.0279)} + \frac{(0.02)(350)}{(2)(0.8)} \quad . \tag{9.9}$$

Therefore, $$I_p = 0.77 \text{ amps.} \tag{9.10}$$

For the Delco 2N1160 transistor, $V_{be_{max}}$ which permits 0.77 amps in the collector circuit is 1.0 volts. Thus, the feedback winding N_{fb} is from equation (4.14),

$$N_{fb} = N_p\left(\frac{V_{be_{max}}}{E}\right) \quad , \tag{4.12}$$

and

$$N_{fb} = 14\left(\frac{1}{6}\right) \quad , \tag{9.11}$$

Therefore

$$N_{fb} = 2.33 \text{ turns} \quad . \tag{9.12}$$

The value of the external base resistance R_b is from equation (4.15),

$$R_b = \frac{V_{be_{max}} - V_{be}}{I_b} \quad , \tag{4.13}$$

and,

$$R_b = \frac{1 - 0.5}{20 \times 10^{-3}} \quad , \tag{9.13}$$

Therefore,

$$R_b = 25 \text{ ohms} \quad . \tag{9.14}$$

The converter secondary N_s , for the B + supply is from equation (4.16) ,

$$N_{s_1} = N_p \frac{E_{out}}{E_{in}} \times 1.05 \quad , \tag{4.14}$$

Thus,

$$N_{s_1} = 14\left(\frac{350}{6}\right) \times 1.05 \quad , \tag{9.15}$$

and,

$$N_{s_1} = 860 \text{ turns} \quad . \tag{9.16}$$

Using equation (4.16) again for N_{s_2} ,

$$N_{s_2} = 14\left(\frac{160}{6}\right) \times 1.05 \quad , \tag{9.17}$$

$$N_{s_2} = 392 \text{ turns} \quad . \tag{9.18}$$

The wire sizes are now determined based on the conservative rating of 700 circular mils per ampere. The results are

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as follows: #21 wire for N_p and #34 wire for N_{fb} and both secondary windings N_{s_1} and N_{s_2}.

The diode rectifiers for the B + supply are 1N684's rated at 500 PIV and 400 ma. The diodes for the B - supply are 1N485's rated at 185 PIV and 50 ma.

Equation (4.17) gives the minimum value for the capacitor input filter for the bridge rectifier. Therefore,

$$C > \frac{1}{4\pi R_L f} \tag{4.17}$$

and for the B + supply,

$$C > \frac{1}{4\pi(20 \times 10^3)(1.25 \times 10^3)} \tag{9.19}$$

$$C > 3200 \ \mu\mu\text{fd} \tag{9.20}$$

For the B - supply,

$$C > \frac{1}{4\pi(1 \times 10^6)(1.25 \times 10^3)} \tag{9.21}$$

and,

$$C > 53 \ \mu\mu\text{fd}. \tag{9.22}$$

Equation (5.9) gives the period of the sawtooth voltage generator,

$$t = -\frac{R_1 R_2 C}{R_1 + R_2} \ln\left[1 - \frac{e_o(t)(R_1 + R_2)}{VR_2}\right] \tag{5.9}$$

Set: $R_1 = 1 \times 10^6$ ohms, $R_2 = 11 \times 10^6$ ohms, and $C = 0.01 \times 10^{-6}$ farads, then,

$$t = -\frac{(1 \times 10^6)(11 \times 10^6)(0.01 \times 10^{-6})}{11 \times 10^6 + 1 \times 10^6} \ln\left[1 - \frac{(20)(11 \times 10^6 + 1 \times 10^6)}{(300)(11 \times 10^6)}\right] \tag{9.23}$$

and,

$$t = 0.76 \text{ milliseconds.} \tag{9.24}$$

The frequency of the sawtooth voltage generator will, therefore, be approximately 1300 cps.

The magnitude of the modulating sawtooth voltage follows from equation (5.19),

$$V_{C_2} = \frac{V_1 C_1}{C_1 + C_2} \tag{5.19}$$

Set: $V_{C_2} = 5$ volts and $C_2 = 0.005 \times 10^{-6}$ farads, then,

$$C_1 = \frac{(5)(0.005 \times 10^{-6})}{20 - 5} \quad , \tag{9.25}$$

and,

$$C_1 \cong 0.0016 \ \mu\text{fd.} \tag{9.26}$$

Now, consider the experimental model whose block diagram is given in Figure 15. The schematic is shown in Figure 16, where the various elements are identified and labeled.

First the output voltages of the B + and B - supplies are investigated in terms of the resulting ripples with reference to the load on each supply and the capacitor input filters on the bridge rectifiers. The output of the B + supply is 300 VDC into a 20 milliampere load. The output of the B - supply is - 160 VDC into a 0.16 milliampere load. Figure 17 shows the magnitudes of the peak-to-peak ripples that are superimposed on the B + and B - supplies as the capacitive input filters are increased. As can be seen from the curves, a capacitor input filter of 4 microfarads is probably sufficient for the B + supply and one of

and,

$$t \cong 0.75 \text{ milliseconds} \qquad (5.24)$$

The frequency of the sawtooth voltage generator will therefore be approximately 1500 cps.

The magnitude of the modulating sawtooth voltage follows from equation (5.19),

$$V_{C_2} = \frac{V C_1}{C_1 + C_2} \qquad (5.19)$$

Set: V_{C_2} = 5 volts and C_2 = 0.005 x 10^{-6} farads, then,

$$C_1 = \frac{(5)(0.005 \times 10^{-6})}{20 - 5} \qquad (5.25)$$

and,

$$C_1 \cong 0.0016 \; \mu fd. \qquad (5.26)$$

Now, consider the experimental model whose block diagram is given in Figure 15. The schematic is shown in Figure 16 where the various elements are identified and labeled.

First the output voltages of the B + and B - supplies are investigated in terms of the resulting ripples with reference to the load on each supply and the capacitor input filters on the bridge rectifiers. The output of the B + supply is 300 VDC into a 20 milliampere load. The output of the B - supply is -150 VDC into a 0.16 milliampere load. Figure 17 shows the magnitudes of the peak-to-peak ripples that are superimposed on the B + and B - supplies as the capacitive input filters are increased. As can be seen from the curves, a capacitor input filter of 4 micro-farads is probably sufficient for the B + supply and one uf

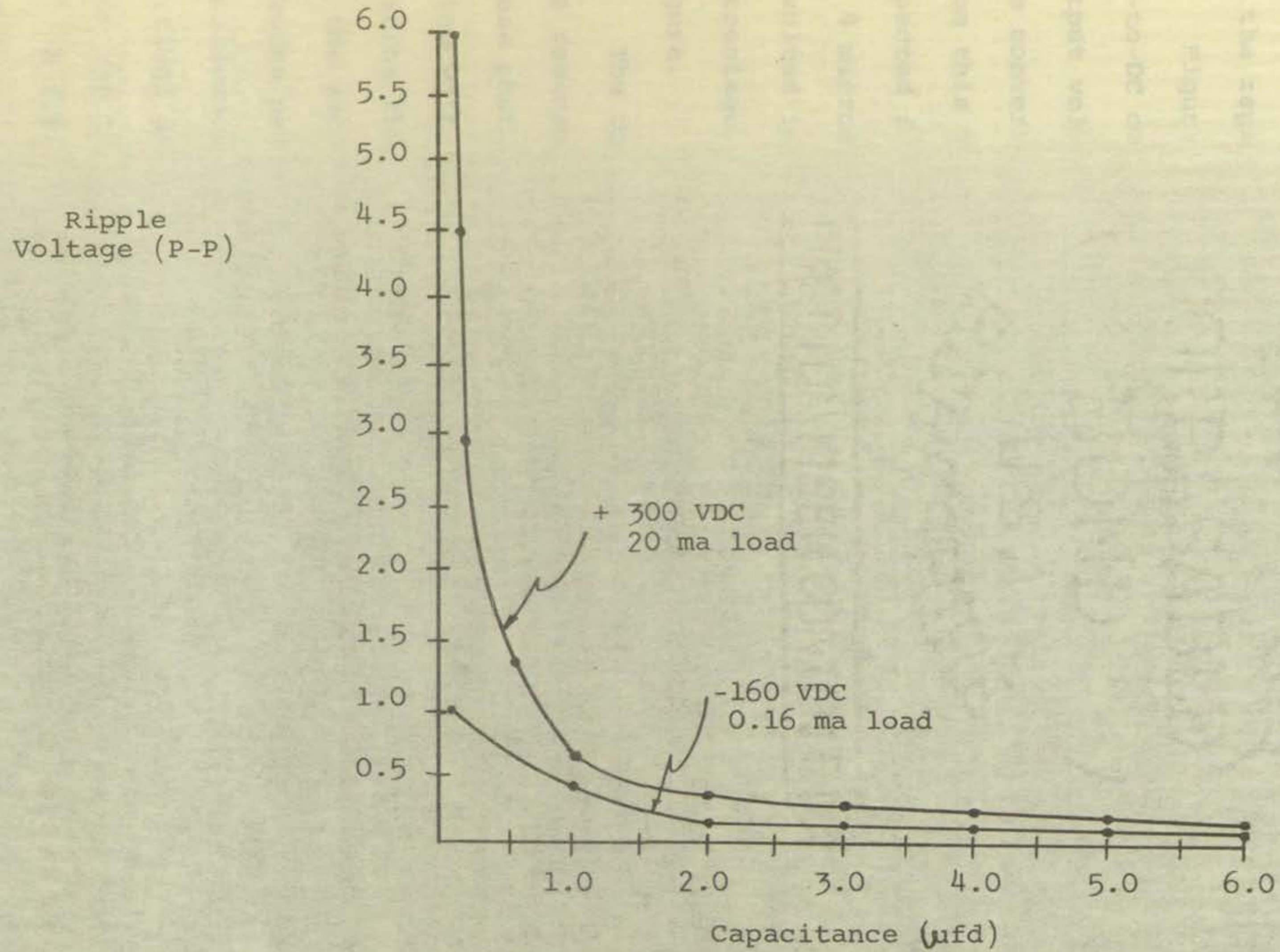


Figure 17. Ripple Voltage (P-P) versus Filter Capacitor for B + and B - Supplies with Constant Loads.

2 microfarads or more is probably sufficient for the B - supply. It is desirable to bring the ripple on the B - supply to a minimum because the output frequency of the reflex klystron oscillator is quite sensitive to changes in the voltage applied to the repeller.

Figure 18 furnishes additional information concerning the DC-to-DC converter. The converter input current, B + supply output voltage, B + supply ripple voltage, and the efficiency of the converter are plotted vs resistive load. It can be seen from this figure that an efficiency of approximately 80% can be expected from the DC-to-DC converter. A capacitive input filter of 4 microfarads was connected to the B + supply rectifiers which resulted in an output ripple voltage that is a quite small percentage of the B + DC voltage as can also be seen from this figure.

The detected outputs of the reflex klystron oscillator and the resonant cavity were photographed from an oscilloscope display. These photographs are shown in Figure 19. The outputs are shown along with the sawtooth voltage generator output which is modulating the reflex klystron oscillator. The vertical sensitivity of the oscilloscope display which shows the sawtooth voltage is 5 volts per centimeter in all cases. The sweep speed of the oscilloscope is 0.2 millisecond per centimeter in all cases. The vertical sensitivities of the displays showing the oscillator mode and the resonant cavity outputs are 0.05 volts per centimeter.

A definition for amplitude modulation of the frequency modulated microwave carrier and the block diagram of the experi-

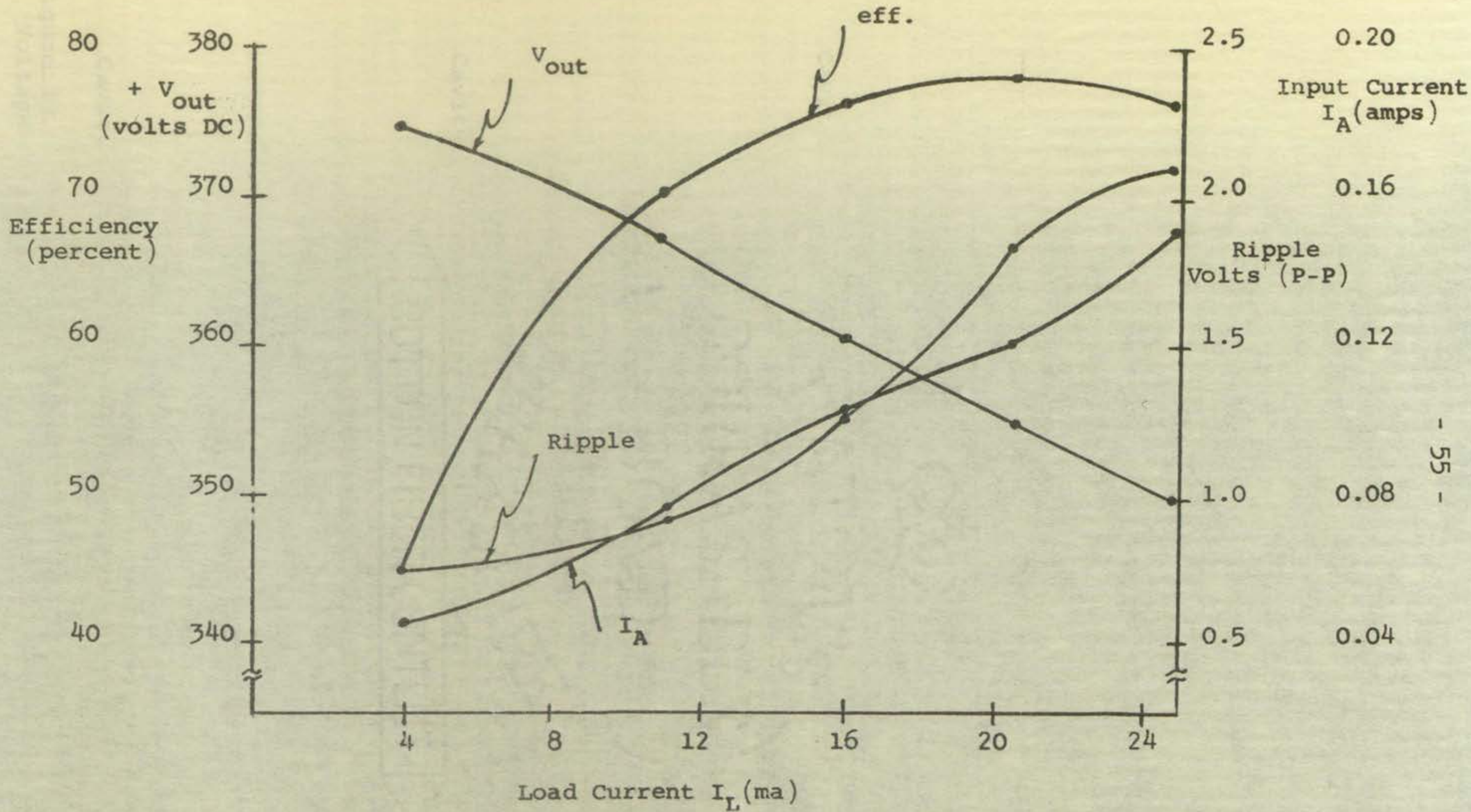
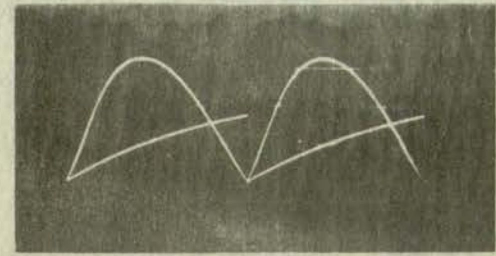
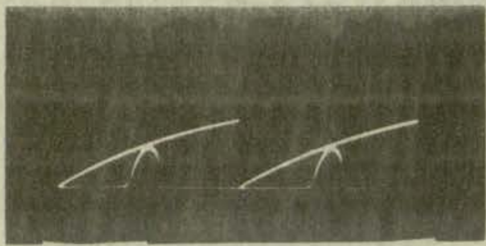


Figure 18. DC-to-DC Converter Input Current, Positive Output Voltage, Output Ripple and Efficiency versus Load Current.

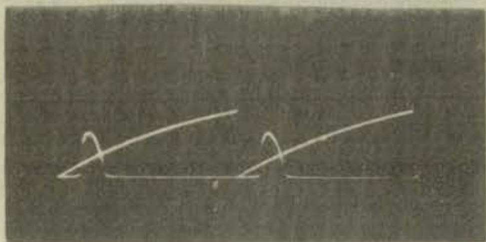


Klystron Oscillator Mode
with Sawtooth Modulating Voltage.



Sawtooth Modulating Voltage
with Output of Transmission Cavity.
Cavity Resonant Frequency at Mid-Range of FM Signal.

Sawtooth Modulating Voltage
with Output of Transmission Cavity.
Cavity Resonant Frequency at Low End of FM Signal.



Sawtooth Modulating Voltage
with Output of Transmission Cavity.
Cavity Resonant Frequency at High End of FM Signal.

Figure 19. Photographs of the Klystron Oscillator, Sawtooth Voltage Generator and Resonant Cavity Output Waveforms.

With [illegible]
Cavity Resonant [illegible]

[illegible]
With [illegible]
Cavity Resonant [illegible]

[illegible]
With [illegible] Cavity
Cavity Resonant [illegible]

Figure 19. Photographs [illegible]
Voltage Generation [illegible]

mental model using the resonant cavity as a transmission cavity is shown in Figure 20. The cavity is then utilized as an absorption cavity and again a definition for amplitude modulation and also the experimental model are shown in Figure 21. The results of these experiments are shown in Figure 22. While conducting these experiments, the repetition rate of the sawtooth voltage generator was varied from 10 cycles per second to 10,000 cycles per second. No noticeable variation in amplitude modulation resulted from the changes in the repetition rate of the sawtooth voltage generator. In other words, amplitude modulation of some given value remains essentially constant as the repetition rate is varied from 10 cycles per second to 10,000 cycles per second.

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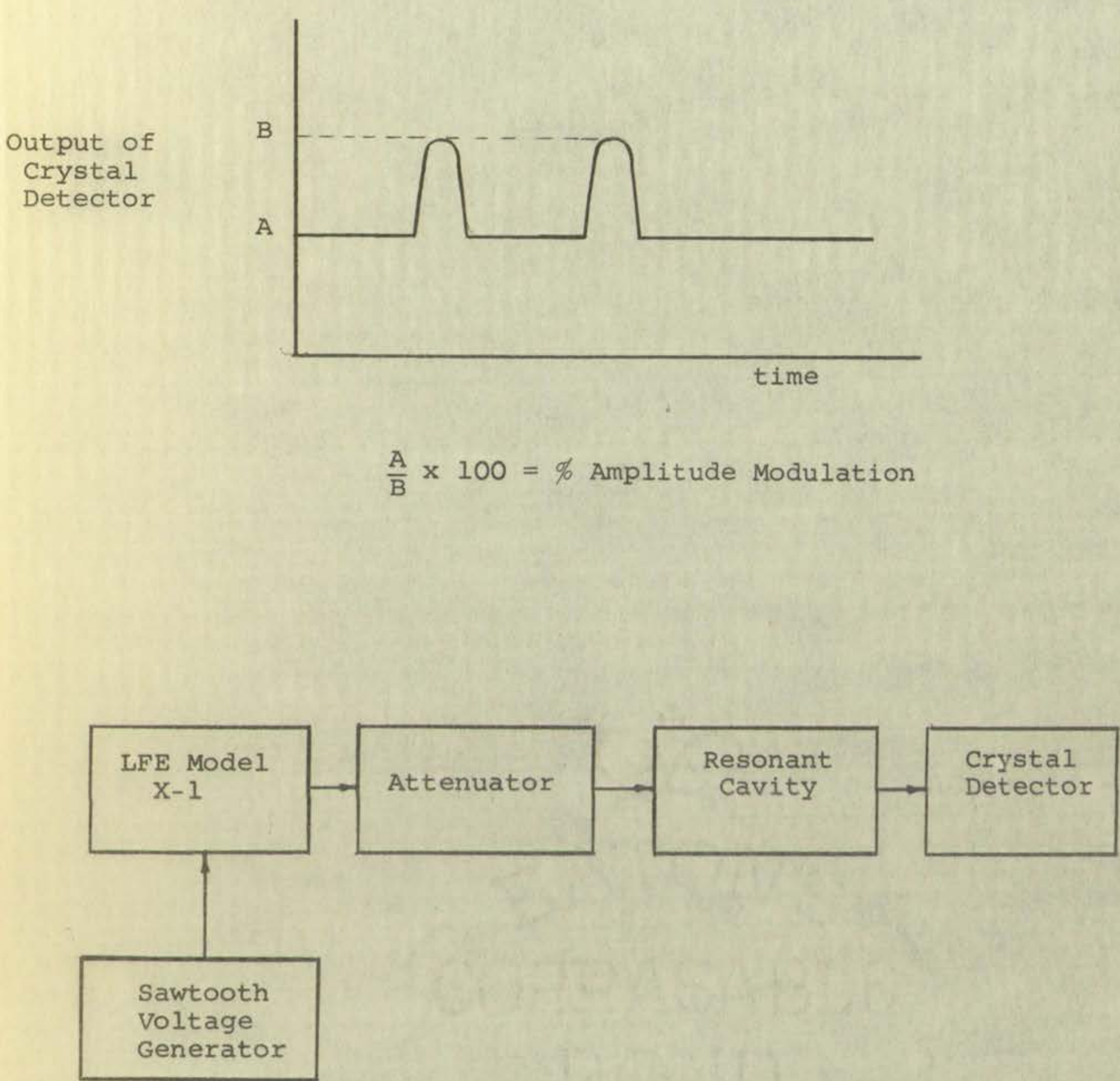
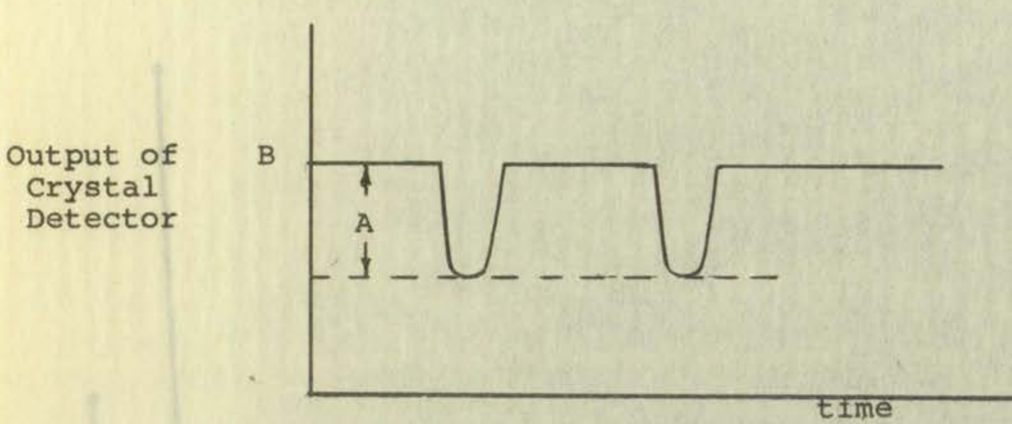


Figure 20. Definition of Amplitude Modulation and Block Diagram of Experimental Model using Transmission Cavity PR&D Model 585B.



$\frac{A}{B} \times 100 = \%$ Amplitude Modulation

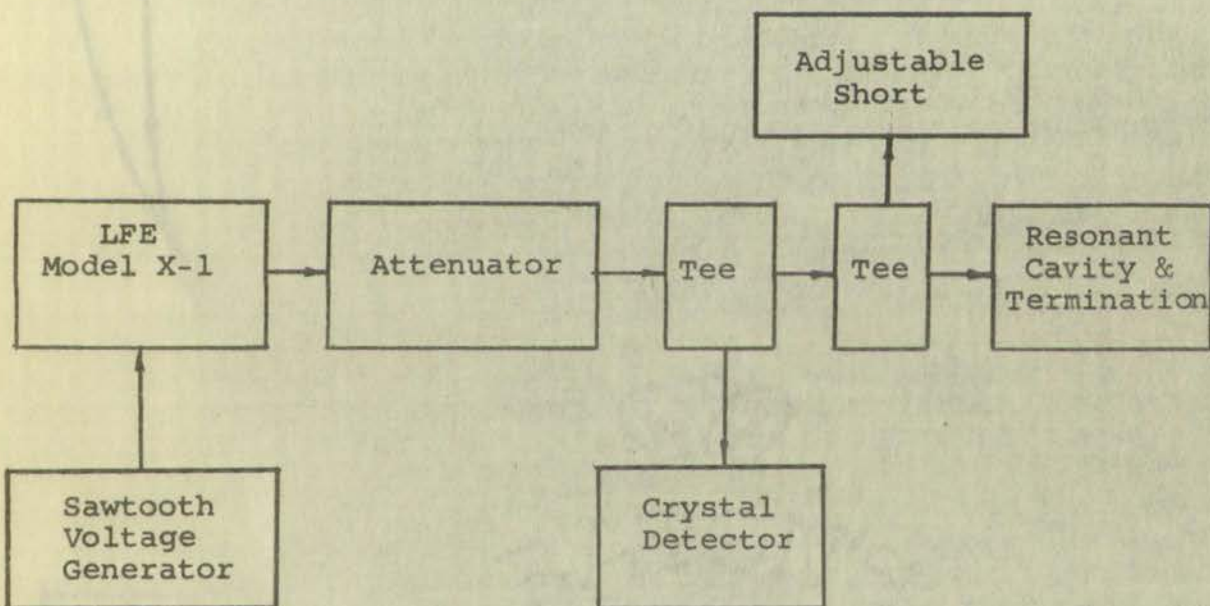


Figure 21. Definition of Amplitude Modulation and Block Diagram of Experimental Model using PR&D Cavity Model 585B as an Absorption Cavity.

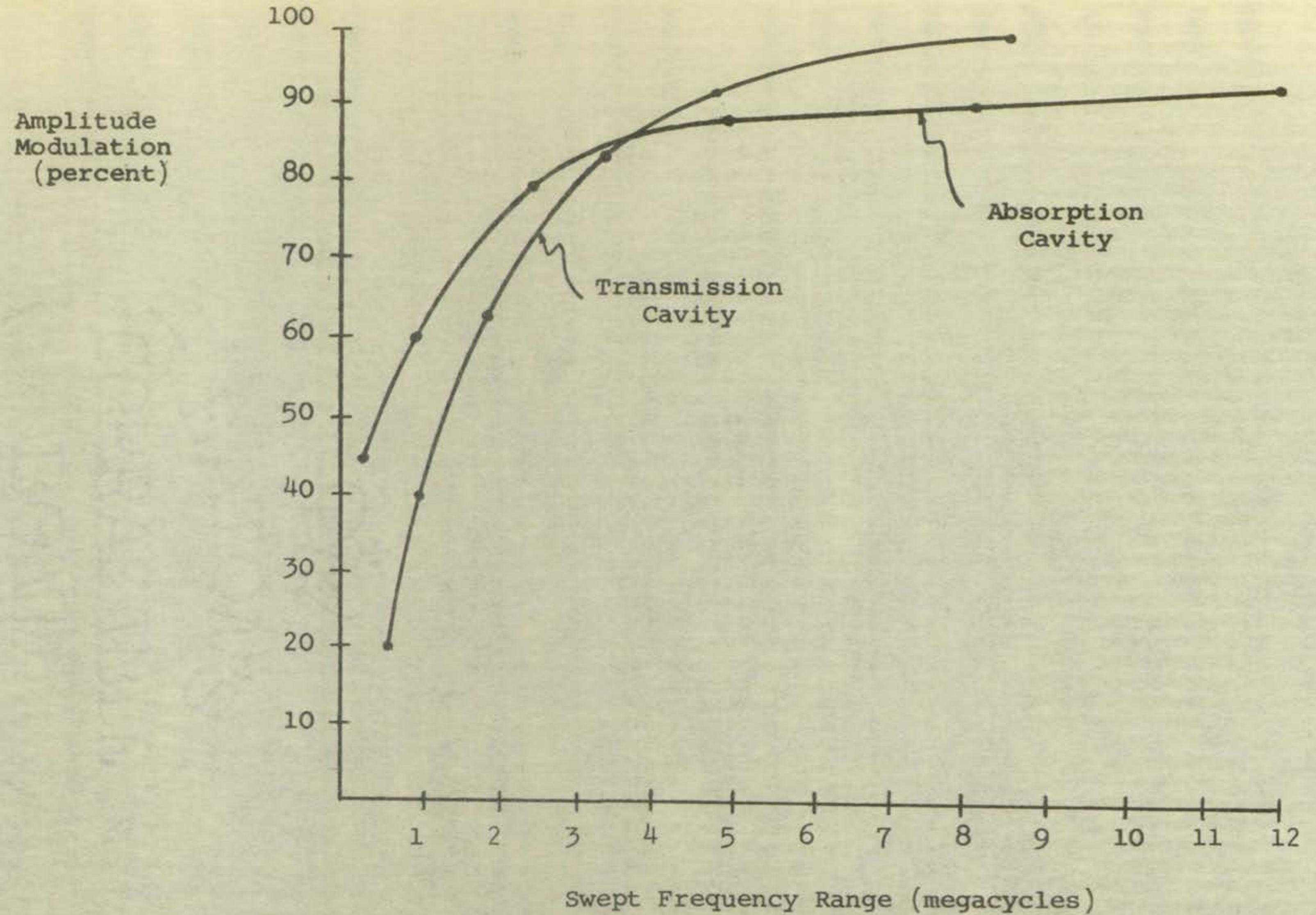


Figure 22. Amplitude Modulation by Resonant Cavities versus Swept Frequency Range of the Klystron Oscillator.

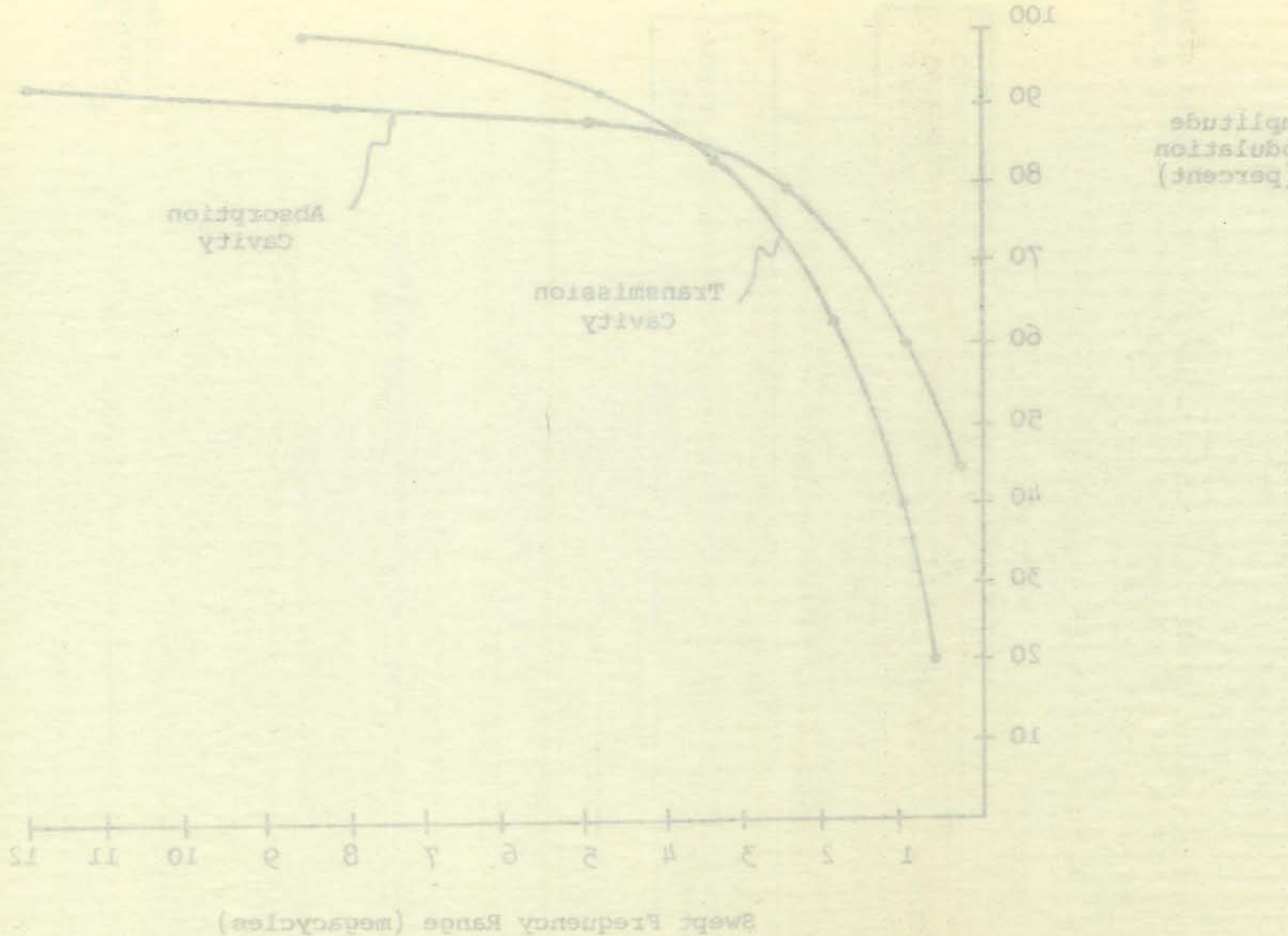


Figure 22. Amplitude Modulation by Resonant Cavities versus Swept Frequency Range of the Klystron Oscillator.

CHAPTER X -- THE PHYSICAL ENVIRONMENT

As stated previously, it is proposed to elevate the microwave refractometer to altitudes of the order of 200,000 feet above sea level utilizing a rocket propulsion system. The microwave refractometer must survive and operate satisfactorily in the physical environment resulting from such an effort. This physical environment includes mechanically and acoustically induced vibrations, mechanically induced shock, temperature variations, altitude (pressure) variations, etc., which are detailed in the following paragraphs. The effects of this physical environment upon the microwave refractometer are also considered.

First, consideration of the sequence of operation for the rocket propulsion system is instructive in determining the possible sources of this physical environment. At time $t = 0$ the rocket is ignited and a mechanical shock pulse of approximately 80 g's in magnitude and duration of approximately 10 milliseconds is created. If this shock pulse is considered to be a half sine wave, the fundamental frequency of the pulse is 50 cps. As the missile leaves the launcher barrel an additional shock pulse of 2 g's of short duration appears. At rocket "burn out" ($t \cong 27.5$ seconds) a 9 g's shock pulse is generated and free flight of the missile follows for the rest of the approximately 127 seconds at which time the nose cone containing the microwave refractometer transmitter is blown free. During the

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nose-cone separation a charge (ignited by a fuse which in turn was ignited by the rocket engine) builds up pressure until the pins, previously securing the nose cone to the missile, shear. The shock pulse which occurs at this time is not definitely known but a pulse of approximately 20 g's in magnitude appears to be a reasonable estimate. During the flight of the missile, mechanical vibrations in the frequency spectrum of 20 to 2000 cps at 10 g's are present. The missile has a resonant frequency between 800 and 900 cps. Consequently, the most severe vibrations transmitted to the refractometer will probably occur in this range. It is certainly significant that the microwave refractometer transmitter must only survive these shocks and vibrations which occur during its elevation to the maximum altitude--with no consideration given to its functional capabilities during this period--since data is not recorded until descent under parachute retardation begins. The effects of shock and vibration during descent of the transmitter are certainly at a minimum, if they exist at all, since the source for this environment has now been eliminated except for the possibilities such as the excitation of a mechanical resonance in the open end resonant cavity due to the flow of air through the cavity as the atmosphere is sampled.

The ambient temperature variation from the surface to 200,000 feet could conceivably be as much as + 40°C at the surface to - 80°C at 200,000 feet. This change of 120°C takes place in a period of approximately 2 minutes when the vehicle is launched. Therefore, it may be required to consider this thermal shock in the transmitter design or to package the transmitter

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The ambient temperature variation from the surface to 200,000 feet could conceivably be as much as + 40°C at the surface to - 80°C at 200,000 feet. This change of 120°C takes place in a period of approximately 2 minutes when the vehicle is launched. Therefore, it may be required to consider this thermal shock in the transmitter design or to package the transmitter

such that the thermal shock is no longer detrimental. "Potting" the transmitter in some material such as polystyrene will alleviate, at least to some extent, this problem. When considering the temperature environment a problem appears that is much more significant than the thermal shock, namely, the lack of convection cooling of the transmitter at the extremely high altitudes. As described in Chapter VI, the reflex klystron generates a relatively large quantity of heat. At the higher altitudes, where very little atmosphere exists to provide convection cooling, this problem of heat dissipation becomes most significant. Damage to the microwave refractometer transmitter of a catastrophic nature could obviously result unless this area of the physical environment is thoroughly investigated and proper precautions are taken to eliminate or at least sufficiently mitigate its effects.

The extremely low pressures at these high altitudes in combination with the heat associated with the system produce another situation with respect to the transmitter which merits consideration. The electrolyte in the power source, at low pressure and elevated temperature, will likely tend to boil. Packaging the power source in a pressurized container could perhaps prevent this situation from occurring.

At the present time an entirely sufficient description of the physical environment and its effects is not available. These preceding paragraphs are therefore necessarily presented in a heuristic manner.

CHAPTER XI -- THE TRANSMITTER PACKAGE

The considerations relative to the transmitter package include its weight and volume and its ability to survive the physical environment described in the preceding chapter.

A sketch of the basic configuration for the microwave refractometer transmitter is given in Figure 23. This sketch illustrates a possible placement of various components of the system which would allow the transmitter to be placed in the ogive. A cross-section of the ogive (with dimensions) is presented in Figure 24. The transmitter along with the parachute is sketched in Figure 25 in an effort to delineate the conceptual configuration of the complete system.

A total weight of 7 pounds is permitted for the microwave refractometer transmitter. This maximum weight is established by the Federal Aviation Agency. It specifies that no more than 7 pounds may be released in the atmosphere under parachute retardation in an experiment such as this microwave refractometer transmitter. The weights of some of the components of the transmitter have been established: the power source shall have a maximum weight of approximately 1.5 pounds; the DC-to-DC power converter shall have a maximum weight of approximately 1.0 pound; the klystron shall have a maximum weight of approximately 0.5 pound; the waveguide sections shall have a maximum weight of 1.0 pound; and the structural members shall have a maximum weight of approximately 1.0 pound. Approximately 2.0 pounds remain for the resonant cavity.

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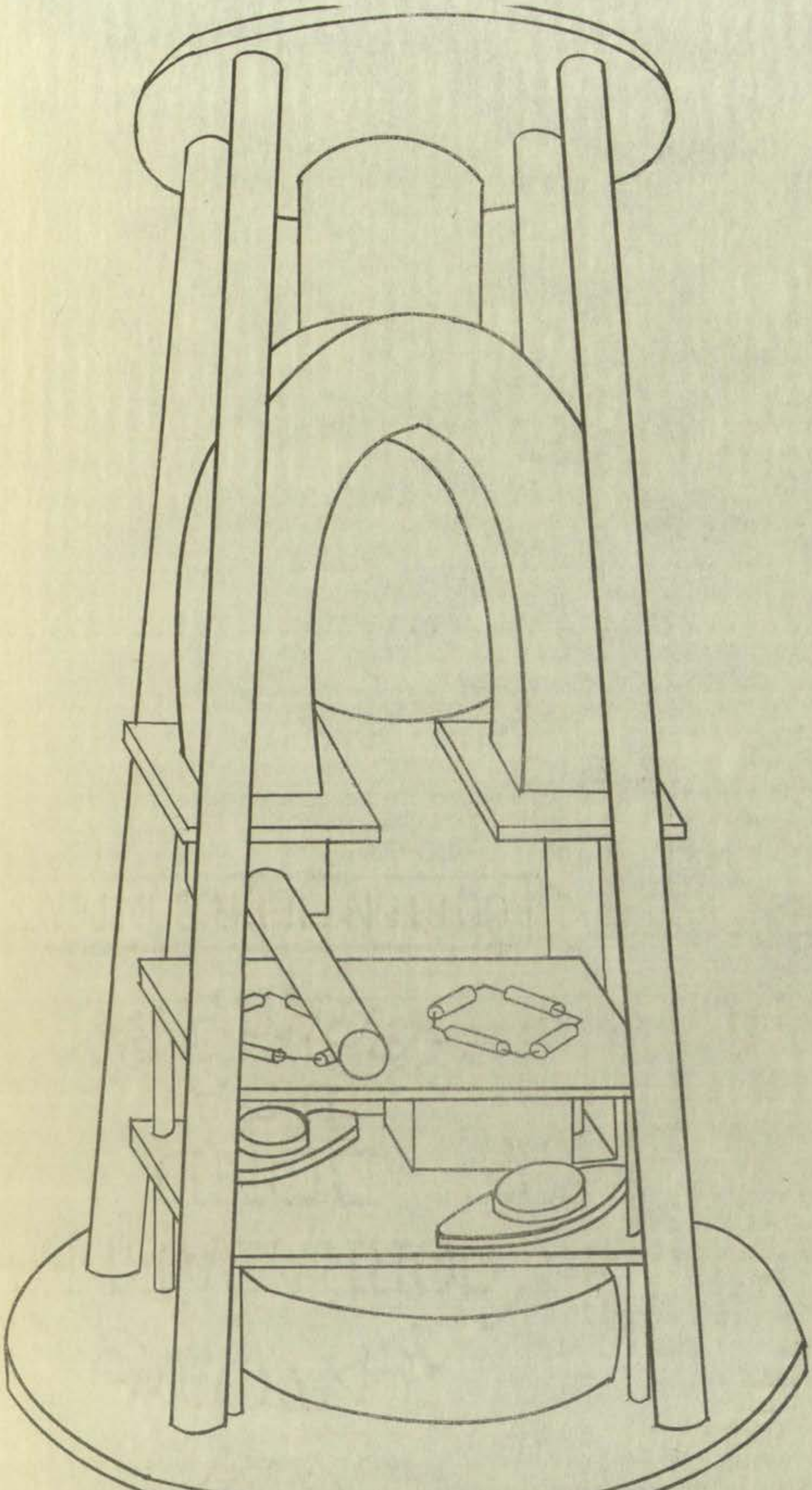

Figure 23. Basic Configuration of the Microwave Refractometer Transmitter.

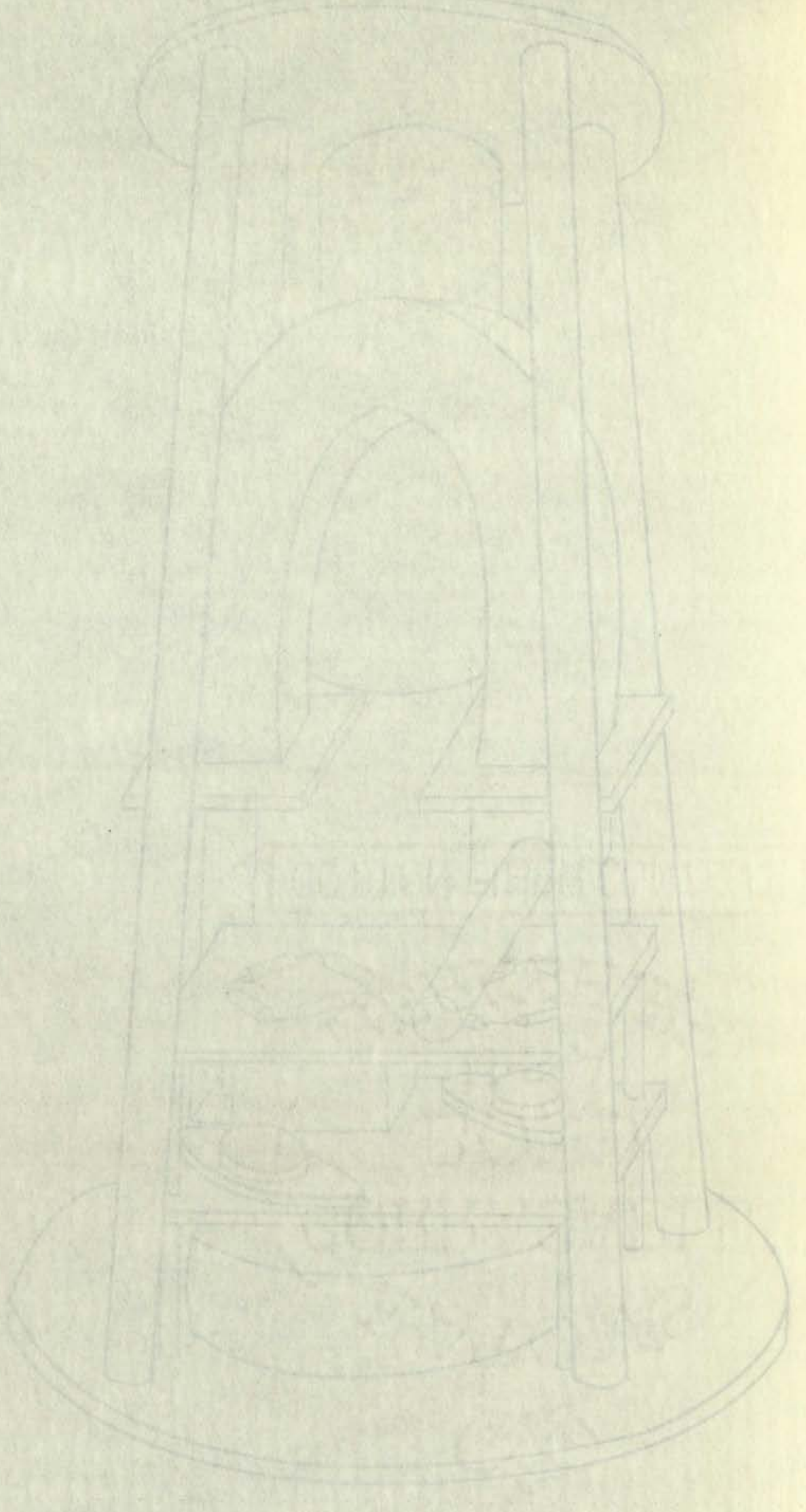

Figure 23. [illegible]
[illegible]

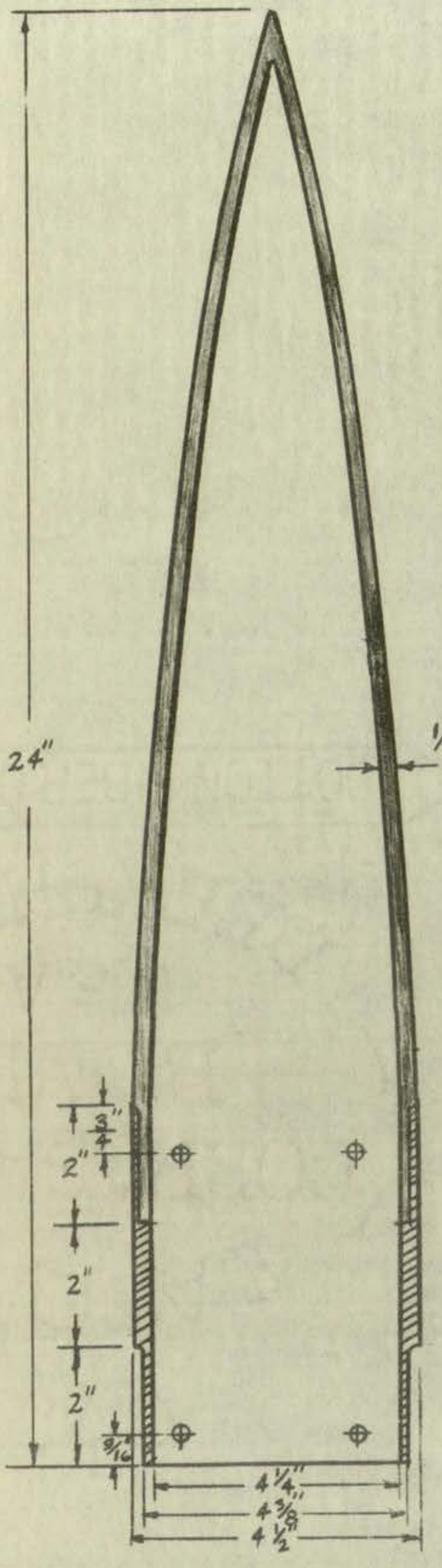

Figure 24. Cross-Section of the Ogive.

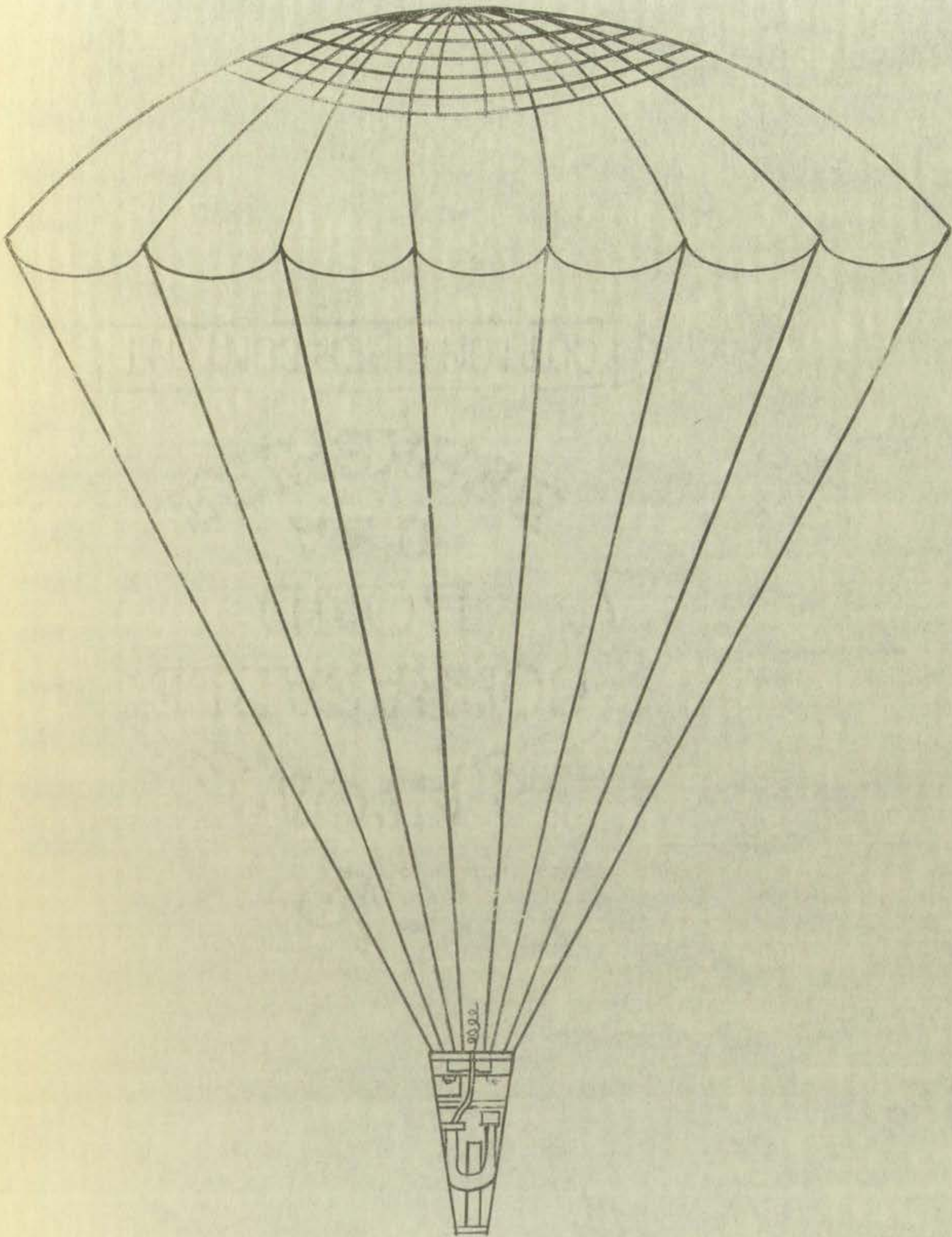

Figure 25. Sketch of the Complete Microwave Refractometer Transmitter.

The ogive, Figure 24, contains the microwave refractometer transmitter during its transition to the high altitudes. The available volume of the ogive which will be occupied by the transmitter is approximately 165 cubic inches. The volume for the configuration of the transmitter given in Figure 23 is within this maximum of 165 cubic inches.

As stated in the preceding chapter, one of the major points of interest associated with the physical environment which the transmitter must survive is the large heat dissipation. The high g mechanical shock which occurs at launch and the mechanical vibration that follows are also apparently extremely detrimental. The extent of the pernicious effects of these environments is not presently well defined. Simulation of these environments in the laboratory, at least to some extent, is possible. Certainly, much useful information could be obtained from a consistent, well organized, environmental test program.

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CHAPTER XII -- SUMMARY AND CONCLUSIONS

An effort has been made to investigate one possibility of generating a suitable radio frequency signal which contains information describing the refractive index of the earth's atmosphere. The effort has been centered around a transmitter operating in the microwave portion of the radio frequency spectrum. The X-band region of the microwave spectrum was selected because of the severe limitations of weight and volume placed upon the transmitter. A sufficiently detailed description of the transmitter is presented such that a microwave refractometer of this general configuration is shown to be feasible.

The design criteria for some of the various components of the transmitter have been established, while others still merit extensive investigation. The power source, the DC-to-DC power converter and the sawtooth voltage generator are sufficiently detailed such that these components may be effectively fitted into the microwave refractometer transmitter. The reflex klystron is apparently the logical choice to generate the microwave carrier. However, pulsed operation of the klystron rather than CW operation may very well be advantageous in consideration of the possible peak power output. As stated, perhaps 200 milliwatts of average microwave power output could be realized. If the resonator circuit of the reflex klystron is gated, peak power output of perhaps 50 watts could be achieved and still maintain a sufficient sampling rate of the index of refraction. Of course, the transient problems associated with gating the resonator circuit would have to be resolved and consideration would also

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have to be given to the amount of power required to gate this circuit.

Another scheme that might be employed in connection with the reflex klystron oscillator is to utilize the ripple voltage that appears in the B - supply to frequency modulate the oscillator. If a sufficient load is placed upon the B - supply and a small capacitive input filter is used, the ripple voltage would be large enough in magnitude to sweep the oscillator over the required range. However, this technique results in undesirable power consumption. Other methods could possibly be devised that would produce large ripple voltage and still not consume appreciable power. For instance, a nonsymmetrical winding of the two halves of the B - secondary might generate the necessary magnitude of ripple.

If the sawtooth voltage generator, as presently described, is utilized to modulate the reflex klystron oscillator, some useful information could possibly be obtained by introducing a temperature sensitive resistive element (perhaps a thermistor) in the circuit which generates the time constant. This resistive element could be placed in the vicinity of the klystron in an effort to determine the extremes of temperature of this component at the high altitudes. Changes in temperature of this component would then result in changes in the sweep rate of the system. As stated previously, the temperature of the klystron (and of course other portions of the system) is apparently one of the most critical environmental parameters associated with the microwave refractometer transmitter in the proposed application.

The need for extensive investigation, in an effort to define adequately the physical environment and its effects upon the transmitter, is evident. A certain amount of information concerning this environment could possibly be ascertained in an environmental test laboratory. However, the final analysis will depend upon field tests of the microwave refractometer transmitter. The initial field tests could begin with a study of the environmental effects on the power source, the DC-to-DC converter, the sawtooth voltage generator and the reflex klystron oscillator. These parts of the system could be effectively utilized in the study of the environment and its effects. Then, once these conditions are sufficiently established and circumvention of these problems is accomplished, the effects of the physical environment on the complete microwave refractometer transmitter could possibly be resolved.

Another area of further study which is necessary to optimize the microwave refractometer transmitter is involved with the amplitude modulation of the frequency modulated carrier. If an absorption resonant cavity is utilized the amount of amplitude modulation that is achieved depends upon the swept frequency range of the klystron oscillator, the Q of the resonant cavity and the amount of coupling of the resonant cavity to the microwave system. For a given swept frequency range it is possible to realize optimum amplitude modulation and optimum resolution of the resonant frequency of the cavity if the proper compromise between the Q of the cavity and the coupling of the cavity is reached. To illustrate, consider a system which contains a very

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high Q resonant cavity that is loosely coupled to the system. This situation will result in a low amplitude modulation index but will produce the capability of high resolution with respect to very small changes in the resonant frequency of the cavity. High resolution is of course desirable but sufficient modulation index is also necessary in order to insure a suitable signal-to-noise ratio. If the coupling of the cavity to the system is increased, the Q of the cavity will decrease. Therefore, a compromise must be reached between the coupling of the cavity and the Q of the cavity. The need for an investigation to define this compromise is evident. A successful investigation could result in the optimum amplitude modulation index that can be achieved in combination with the optimum resolution of the resonant frequency of the cavity.

APPENDIX A

It seems appropriate to outline a microwave refractometer receiver that could be used with the transmitter since successful operation of the transmitter certainly depends upon suitable reception and demodulation of the transmitted AM-FM signal. The block diagram of one possible receiver is given in Figure 26. The AM-FM microwave signal reaching the receiving antenna will most likely be quite small in magnitude. The traveling-wave tube amplifier is selected as the pre-amplifier because considerable gain (presumably 25 or 30 db) may be expected with a noise figure of perhaps 10 db. A cascaded traveling-wave tube amplifier of perhaps 2 or 3 stages may be necessary. The received signal level will depend on the orientation of the package. The AM-FM microwave signal is then introduced into the crystal mixer along with another microwave signal from the local oscillator, which must be very stable. The local oscillator frequency is adjusted to produce a 43 megacycles output from the mixer. (A center frequency of 43 megacycles was chosen only because a broad-band 43 megacycles amplifier was available for experimental investigations.) This AM-FM signal output from the crystal mixer is then fed to a splitter to separate the AM information from the FM signal.

First, consider the AM channel. This channel contains information pertaining to the resonant frequency of the cavity in the peak of the signal. The signal is first amplified, then detected, and fed to the pulse shaper. The pulse shaper operates on this signal and its output is a very narrow pulse which defines the peak of the AM signal. Greater resolution results from

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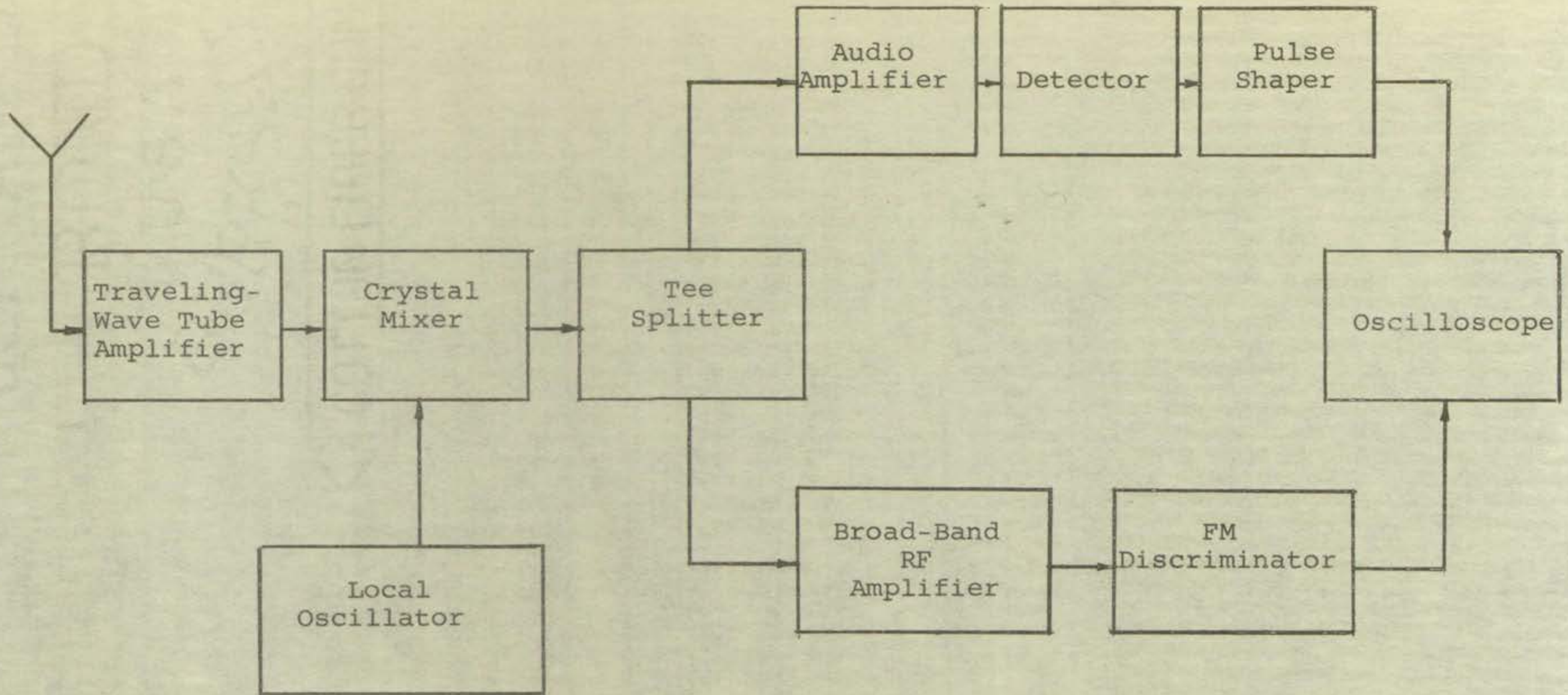

Figure 26. Block Diagram of a Receiver for the Microwave Refractometer Transmitter.

generation of this very narrow pulse. The output of the pulse shaper is then fed to the vertical deflection input of an oscilloscope. The FM signal is amplified and fed to a FM discriminator whose output is a voltage proportional to frequency of the FM signal. In this manner the transmitter modulating sawtooth voltage is reconstituted. The sawtooth voltage output from the discriminator is fed to the horizontal deflection input of the oscilloscope. If a dual-trace oscilloscope is used, the discriminator output could be displayed on one trace while the pulse shaper output is displayed on the other trace. The result would be a display as given in Figure 27. The vertical sensitivities of the traces for the pulse shaper output and the discriminator output is 2 volts/centimeter. The sweep speed of the oscilloscope is 0.2 milliseconds/centimeter. (These photographs were taken of waveforms generated by an experimental model of the receiver whose input was generated by the experimental model transmitter.) The position of the pulse from the pulse shaper, with respect to the sawtooth voltage output from the discriminator, is a determination of the resonant frequency of the cavity with respect to the swept frequency of the klystron oscillator. The index of refraction is therefore determined since the resonant frequency of the cavity is inversely proportional to the index of refraction.

Of additional interest, with respect to the refractometer receiver, is the bandwidth required in the pre-amplifier which feeds the FM discriminator. The input signal to the receiver is mixed with the local oscillator. (The local oscillator is

generation of this very narrow pulse. The output of the pulse shaper is then fed to the vertical deflection input of an oscilloscope. The FM signal is amplified and fed to a FM discriminator whose output is a voltage proportional to frequency of the FM signal. In this manner the transmitter modulating sawtooth voltage is reconstituted. The sawtooth voltage output from the discriminator is fed to the horizontal deflection input of the oscilloscope. If a dual-trace oscilloscope is used, the discriminator output could be displayed on one trace while the pulse shaper output is displayed on the other trace. The result would be a display as given in Figure 27. The vertical sensitivities of the traces for the pulse shaper output and the discriminator output is 2 volts/centimeter. The sweep speed of the oscilloscope is 0.2 milliseconds/centimeter. (These photographs were taken of waveforms generated by an experimental model of the receiver whose input was generated by the experimental model transmitter.) The position of the pulse from the pulse shaper with respect to the sawtooth voltage output from the discriminator is a determination of the resonant frequency of the cavity with respect to the swept frequency of the klystron oscillator. The index of refraction is therefore determined since the resonant frequency of the cavity is inversely proportional to the index of refraction.

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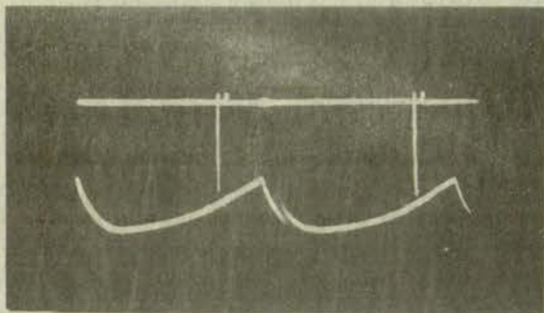


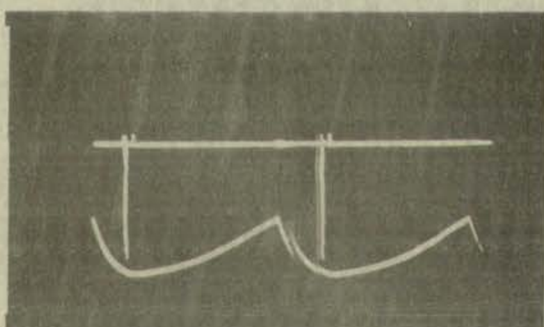
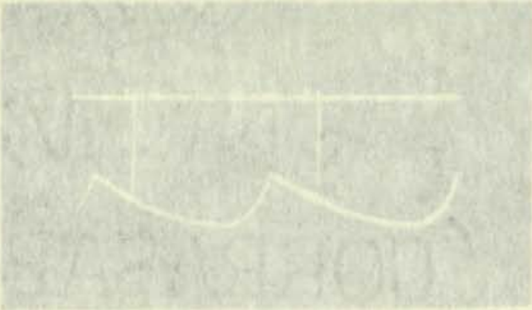


Figure 27. Photographs of the Receiver's Pulse Shaper and Discriminator Output Waveforms.



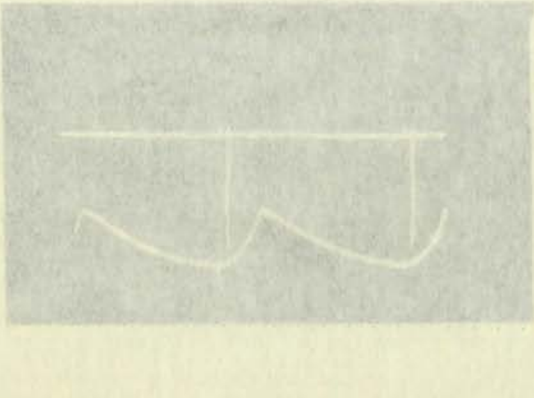


Figure 27. Photographs of the Receiver's Pulse Shaper
and Discriminated Output Waveforms.

adjusted to produce a difference frequency from the mixer of 43 megacycles.) This 43 megacycles signal is fed to a splitter where the FM portion of the signal is separated from the AM portion. The FM portion of the signal is amplified and fed to a discriminator. This amplifier is the circuit now being considered. For a given swept frequency range of the klystron oscillator in the transmitter, it is necessary to determine what bandwidth is required in this amplifier (in terms of the swept frequency range) which will accommodate the signal and allow the discriminator to reconstitute the modulating sawtooth voltage which was applied to the klystron repeller in the transmitter. According to Post[21/] the bandwidth of the pre-amplifier should be approximately $2\Delta f$ where Δf is the maximum frequency deviation of the frequency modulated wave. This means, of course, that the bandwidth of the pre-amplifier should be approximately 14 megacycles if the frequency deviation of the FM wave is 7 megacycles. The 7 megacycles deviation is equivalent to 700 N-units change in the index of refraction. The 700 N-units change is larger than the expected change in the index of refraction of the atmosphere, which is for practical purposes approximately 300 N-units. Obviously, the swept frequency range could be reduced. However, the larger swept frequency range has some advantage in terms of the amplitude modulation which is produced by the resonant cavity.

Some consideration has been given to other designs for the microwave refractometer receiver which could eliminate the

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[21/] Post, D. F., Spectrum Analysis of Waveform Transmitted from Airborne Refractometer, Private Communication to D. C. Thorn, September, 1961. (A memorandum report.)

adjusted to produce a different frequency from the... megacycles.) This ... where the FM portion of the signal is separated from the AM portion. The FM portion of the signal is amplified and applied to a discriminator. This output is the originally applied modulation... ed. For a given swept frequency range of the Klystron oscillator in the transmitter, it is necessary to determine what bandwidth is required in this amplifier (i.e. before the discriminator range) which will accommodate the signal and allow the discriminator to reconstitute the modulating sawtooth voltage which was applied to the Klystron repeller in the transmitter. According to Post[24] the bandwidth of the pre-amplifier should be approximately $2\Delta f$ where Δf is the maximum frequency deviation of the frequency modulated wave. This means, of course, that the bandwidth of the pre-amplifier should be approximately 14 megacycles if the frequency deviation of the FM were 7 megacycles. The 7 megacycles deviation is equivalent to 700 N-units change in the index of refraction. The 700 N-units change is larger than the expected change in the index of refraction of the atmosphere which is for practical purposes approximately 300 N-units. Obviously, the swept frequency range could be reduced. However, the larger swept frequency range has some advantage in terms of the amplitude modulation which is produced by the reference cavity.

Some consideration has been given to other schemes for the microwave refractometer problem, which could eliminate a...

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[24] Post, E. F., *Radiometeorological Measurements Made with an Airborne Refractometer*, ... Electronic..., September, 1954. (Unpublished Report.)

requirement of reconstituting the modulating sawtooth voltage. Since the transmitter has weight and volume limitations and the receiver does not, the receiver could certainly be expanded if this would allow more flexibility in the design of the transmitter. It should be understood that this receiver design is more sophisticated than would be required for a similar refractometer which was not subjected to such stringent environmental conditions. The receiver is intended to correct for as many potential transmitter instabilities as possible.

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