KLYSTRON TUBES

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The quality of the materials used in the manufacture of this book is governed by continued postwar shortages.

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PREFACE

Early in 1942, a small pamphlet known as the Klystron Technical Manual was prepared to fulfill the need for a unified source of basic information about klystron tubes. Essentially the same material was included in a book with the same title which was distributed by the Sperry Gyroscope Company, Inc., in 1944.

The present book is a complete revision of the earlier publications. Considerable new material has been added and a more complete theoretical analysis of the operation of klystron tubes has been included. The additional theoretical material necessitates an increase in the mathematical content of the book, but the derivations have been simplified in order to increase the usefulness of the book. The introductory chapters assume that the reader is not acquainted with the principles of klystron tubes. Practical interpretations of the equations in the theoretical chapters are given so that an understanding of these principles may be obtained without following the mathematics in detail. However, an understanding of electronic phenomena and the fundamentals of radio is a prerequisite for the effective use of this book.

Although the book is primarily a theoretical text, rather than a handbook, several chapters on the operation of klystron tubes, power supply considerations, and microwave techniques have been included for the reader who is interested in the application of these tubes. Design information has not been included as a part of the text, but some information of this nature is given in the design charts that have been included in Appendix B to supplement the simplified illustrations in the text.

The preparation of this material for publication was made possible by the cooperation of the tube-development group of the Sperry Gyroscope Company at the Garden City Laboratories. Every member of this group and many engineers in other groups have made contributions or have helped to clarify certain points during informal discussions of klystron characteristics. The author is particularly indebted to Coleman Dodd, who read the manuscript and made innumerable suggestions for its improvement. The assistance of Robert Wathen is also gratefully appreciated. Chapters 6 and 7 on reflex klystron theory are based on the work of Dr. Edward L. Ginzton and Dr. Donald R. Hamilton. T. Moreno prepared the material for the chapter on Microwave Measurements.

Many of the ideas presented in the book are based on the early training of the author under the guidance of Dr. William W. Hansen. The ana-

lytical work of Dr. Eugene Feenberg and suggestions by Edward Barlow are also reflected in many parts of the book. Appreciation is extended to Dr. L. J. Black, Dr. P. L. Morton, Dr. J. R. Pierce, Dr. J. R. Ragazinni, and other writers whose articles on various phases of klystron theory have aided the author. Acknowledgment for several of the photographs used as illustrations is included in the text. All other photographs were furnished by the Sperry Gyroscope Company, Inc.

A. E. HARRISON.

PRINCETON, N. J.,
May, 1947.

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CONTENTS

CHAPTER 1

KLYSTRON CONSTRUCTION

INTRODUCTION

1.1. Definitions.—Electrical energy can be converted into radio-frequency energy in several ways. Conventional vacuum tubes control the electron emission in order to vary the current that flows to the anode or plate electrode of the tube. This process is known as "current density modulation." An entirely new control process known as "velocity modulation" has been developed recently and has resulted in a phenomenal extension of the radio-frequency spectrum to the superhigh-frequency region. Velocity modulation does not depend upon varying the emission from a cathode, but varies the velocity of the electrons in a beam of constant density. This new method of controlling an electron beam is the basic principle of the klystron tubes to be described in this book.

Klystrons are a family of microwave vacuum tubes that depend upon the conversion of a velocity-modulated beam into a varying current by the process of electron bunching. A klystron amplifier furnishes the simplest example of the bunching process. In an amplifier, a small input voltage alternately decreases and increases the velocity of the electrons. During the transit time of the electrons from the input to the output of the amplifier, the electron beam is converted from a constant density beam into a beam with an a-c component which delivers energy to the output circuit. Klystron tubes are not limited to a single application but are quite versatile and may be used as oscillators, amplifiers, frequency multipliers, and detectors or mixers. These tubes have helped to extend the useful range of radio frequencies into the microwave region above 1,000 megacycles and permit most of the functions of conventional triodes and pentodes to be achieved at these superhigh frequencies.

1.2. Description of a Klystron Amplifier.—A cutaway view of a typical klystron amplifier is shown in Fig. 1–1, and important features of the tube have been labeled. The terminology relating to these tubes is in some cases new; in other cases it has been borrowed from conventional vacuum-tube and circuit practice but used in a modified manner. For this reason, the terms introduced in Fig. 1–1 and later parts of the text are defined when they first occur in the text and also have been included in a glossary in Appendix A.

The electron gun, sometimes referred to as the "cathode assembly," furnishes a beam of electrons. In this illustration, the anode is a part of the metallic tube body, rather than an integral part of the electron gun structure. The electrons in the beam are accelerated toward the anode plane, but this plane consists of a gridlike structure, and most of the electrons continue past the anode plane along the axis of the tube, forming an

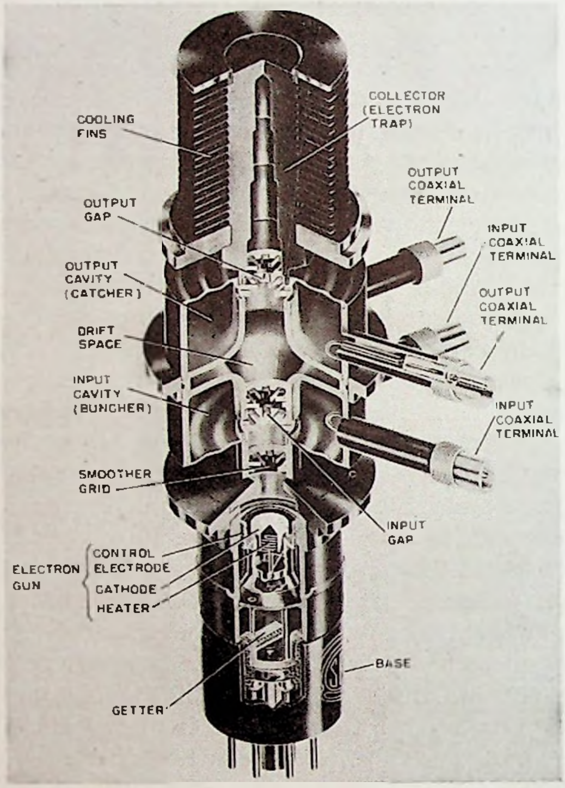


FIG. 1–1.—Isometric view of a Type 3K30/410R klystron cutaway to show internal construction.

electron beam. Since the grid structure at the anode plane shields the electron beam from the electric field of the electron gun, the electrons in the beam will continue along the axis of the tube with an average velocity which is determined by the voltage between the cathode and anode. This voltage is usually referred to as the "acceleration voltage," or "beam voltage."

Cavity resonators are used in practically all klystron designs. There are two cavity resonators in Fig. 1–1, corresponding to the tuned circuits used at lower frequencies. The resonators are inductively coupled to the electron beam; *i.e.*, their fields act upon the electron velocity and absorb power from the electron beam without any necessity for the beam current itself to flow in the circuit. This interaction between the radio-frequency circuit and the electron beam takes place at the input and output gaps. These gaps are formed by closely spaced grid structures which are part of the resonators. The distance between the input and output gaps is known as the "drift space." The electron bunching takes place in this drift space as the electrons travel from the input resonator to the output resonator. One or more coaxial lines, known as "coaxial output terminals," are coupled to the magnetic field within the resonators by small loops. These coaxial output terminals permit radio-frequency power to be supplied to the input resonator and allow the output of the klystron amplifier to be connected to a useful load.

All of the energy in the electron beam is not converted into radio-frequency power; the electrons continue beyond the output gap and give up their residual energy in the form of heat. Note that in a klystron the equivalent of plate dissipation takes place in a part of the tube that can be cooled conveniently, and is not a part of the radio-frequency circuit.

1.3. Comparison with Triodes.—There are three fundamental differences between klystrons and the vacuum tubes used at lower frequencies. First, the radio-frequency circuits are an integral part of the tube; in many cases they are actually a part of the vacuum envelope. The usual concept of a tuned circuit connected by short leads to the electrodes of a vacuum tube is no longer applicable in the superhigh-frequency region. This distinction between low-frequency circuits and microwave circuits is not limited to klystron tubes but applies as well to other types of microwave vacuum tubes.

Second, the electron source in a klystron is completely independent of the radio-frequency circuit. The electrons in the beam attain a high velocity before they reach the region where the radio-frequency field controls the beam. As a result of this high electron velocity, the spacing of the input gap of a klystron may be rather large without introducing serious transit-time difficulties. In contrast, the control grid of a triode is in a region of very low electron velocity and spacings must be quite small if the tubes are intended for operation at very high frequencies.

The manner in which the beam is controlled by the radio-frequency voltage is the third difference between the two types of tubes. Transit time is actually utilized in the drift space of a klystron to convert an electron beam which has only velocity modulation into a beam with an alternating-current component. This dependence upon transit time in a klystron introduces a number of electrical characteristics for these tubes which are quite different from the characteristics of the more familiar triodes. Therefore an understanding of the principles of electron bunching in velocity modulation tubes is essential for an explanation of the characteristics of these tubes.

1.4. Electron-bunching Action.—A radio-frequency field will be built up across the input gap if power of the correct frequency is introduced into the input resonator. Electrons in the beam that pass the input gap when the field is negative will be decelerated and will continue along the drift space with a velocity that is less than the average velocity. Electrons leaving later in the cycle when the field is positive will be speeded up, or accelerated, and will travel along the drift space with greater than average velocity. The transit time across the resonator gap can be a small fraction of a cycle even at superhigh frequencies, but the transit time in the drift space is usually a large number of cycles. The small changes of

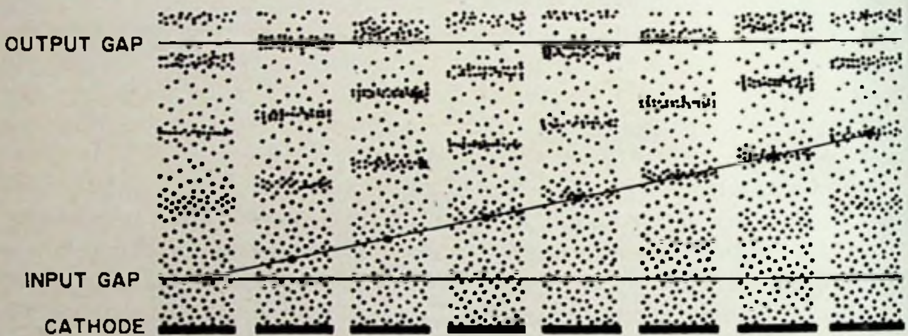


FIG. 1-2.—Electron distribution in a klystron for eight different times, showing the bunching action as the electrons progress along the drift space.

velocity introduced by the input gap permit the electrons that left later in the cycle, with higher than average velocity, to overtake the electrons that were slowed down but passed the resonator gap earlier in the cycle. As a result, the beam is converted from a uniform current to a bunched or pulsating current.

This bunching action takes place in the drift space, which is free of any radio-frequency field and may be completely field-free; *i.e.*, no additional voltages are usually applied to this part of the tube structure. The transit time is again quite small in the region where the beam is acted upon by the radio-frequency field in the output resonator. The electron bunches pass the output gap when the radio-frequency field is negative and the electrons in the bunch are slowed down by the action of the field. This reduction in velocity is equivalent to a reduction in the kinetic energy of the electrons, and the kinetic energy lost by the electrons is transferred to the energy of the electromagnetic field in the resonator. The excess energy in the beam is dissipated in the form of heat at a suitable electron trap beyond the output gap.

Figure 1–2 shows the distribution of electrons in the beam as the beam moves along the drift space. This illustration is equivalent to "stopping" the motion of the electrons at eight different times. As the electrons progress along the drift space between the input and output gaps, they become more closely grouped. Follow the group of electrons that are partly formed in the first column of Fig. 1–2. In the second column these electrons have moved toward the output gap, and the bunch is more compact. The bunching is sharpest in the fourth column. As the bunch continues to move along the drift space, the bunch becomes slightly wider but more electrons become a part of this wider bunch. The latter degree of bunching is most efficient for a klystron amplifier.

1.5. Applegate Diagrams.—A straight line can be drawn through the position of the same electron in each of the columns of Fig. 1–2 because the electron has a constant velocity and moves an equal distance during each equal interval of time. It is easily seen that such a line represents the position of that particular electron for any value of time as well as for the eight different times illustrated in Fig. 1–2. This type of space-time diagram, showing the position of an electron along the drift space for a number of electrons leaving at different times, was proposed by Applegate[1] in order to describe electron bunching phenomena and is known as an "Applegate diagram." A similar space-time diagram has been used by Brüche and Recknagel.[2]

In an Applegate diagram (see Fig. 1–3), time is measured along the horizontal axis, and the position of electrons along the drift space is plotted as the vertical coordinate. The lines represent electrons leaving the cathode at uniform time intervals. The slope of a line is the distance divided by time; *i.e.*, the slope of a line represents the velocity of a particular electron. In this analysis the electrons are assumed to leave the cathode with zero velocity and are accelerated toward the anode. Beyond the anode plane the electron velocities are identical and equal to the average electron velocity until the electrons reach the position of the input resonator gap. The input gap voltage, indicated by the sine-wave variation superimposed upon the acceleration voltage at the bottom of the diagram, varies the velocity of the electrons depending on the amount and direction of the voltage at the time an electron passes the input gap. The transit time across the resonator gaps has been assumed negligible in order to simplify the construction.

As a result of the velocity variations superimposed upon the average velocity of the electron beam at the input gap, the electron velocities are

[1] L. M. Applegate prepared the specification in U.S. Patent No. 2269456 issued to W. W. Hansen and R. H. Varian on an "Electron Beam Oscillator."

[2] E. Brüche and A. Recknagel, Phase Focusing of Electrons in Rapidly Fluctuating Fields, *Zeit. Phys.*, 108: 459–482, March, 1938.

again constant but are no longer the same for the different electrons, and the slopes of the lines have been modified in accordance with the expression

$$v = v_0 + v_1 \sin \omega_1 t_1 \tag{1-1}$$

where v is the velocity of a particular electron passing the input gap at time t_1, v_0 is the average electron velocity corresponding to the acceleration voltage of the electron gun, v_1 is the peak amplitude of the velocity modulation, and ω_1 is the angular frequency of the oscillation in the input resonator.

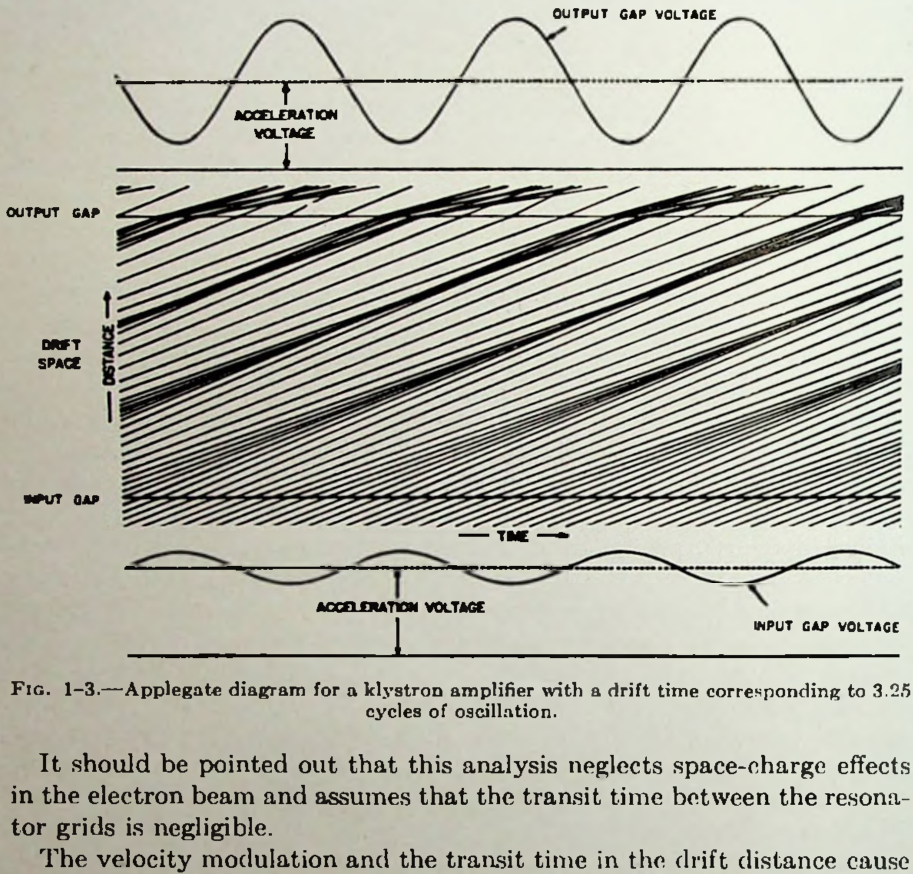


FIG. 1-3.—Applegate diagram for a klystron amplifier with a drift time corresponding to 3.25 cycles of oscillation.

It should be pointed out that this analysis neglects space-charge effects in the electron beam and assumes that the transit time between the resonator grids is negligible.

The velocity modulation and the transit time in the drift distance cause the faster electrons to overtake the slower ones and produce electron bunching. An approximation for the current distribution at any point along the drift distance at any time may be obtained by inspection of the diagram, since the distance along the time axis between successive lines on an Applegate diagram is inversely proportional to the instantaneous current in the beam. It is apparent that the electron density is uniform and that the beam is direct current at the position of the input resonator, but an alternating component of current is superimposed on the direct current as the velocity modulation produces a bunching of the electron beam.

Maximum radio-frequency energy will be transferred to the output resonator if the output gap is placed at the position shown in Fig. 1–3. In an actual tube it is not necessary to vary the position of the output resonator along the drift space; the correct bunching for maximum energy transfer may also be obtained by varying the amount of velocity modulation, *i.e.*, the input gap voltage, or by changing the average electron velocity by changing the anode voltage.

Electrons in the bunches are decelerated by the output resonator field and leave the output gap at very low velocities corresponding to almost horizontal slopes. Electrons passing the output gap during the other half of the cycle are accelerated and absorb energy from the radio-frequency field. However, the energy absorbed by the beam during the wrong half of the cycle is small because there are very few electrons in the beam during this part of the cycle. Considerably greater power is transferred by the electrons in the bunch than is absorbed by the few electrons that become speeded up by the field, and the net result is the conversion of d-c beam power into radio-frequency power.

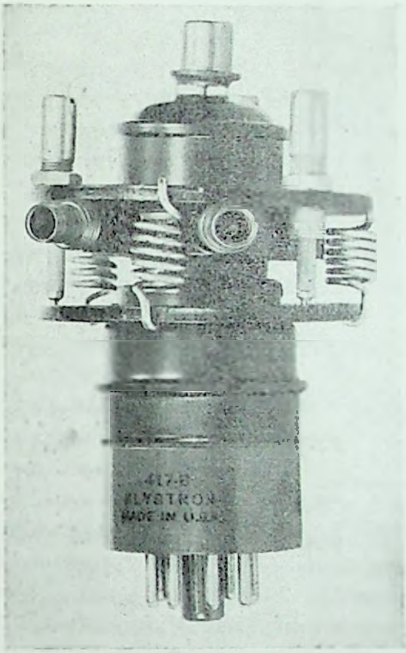


FIG. 1–4.—Type 417B klystron. The resonator and tuning mechanism are integral parts of the tube.

KLYSTRON CONSTRUCTION

1.6. Methods of Construction.—The brief description of a typical klystron amplifier in Sec. 1.2 served as an introduction to the principles of klystron operation, but it does not describe the different parts in sufficient detail or suggest alternative designs which are quite useful. The component parts and structural details are described more completely in the sections that follow.

Two different kinds of mechanical construction used in the manufacture of klystrons are illustrated in Figs. 1–4 and 1–5. Both tubes are reflex oscillators. The Type 417A shown in Fig. 1–4 is of metal construction. A disk seal tube, the Type 707A, has a glass envelope with two disk electrodes extending through the glass. The cavity resonator for Type 707A is not shown in Fig. 1–5 and is clamped to the tube outside of the vacuum envelope. Many variations of these two methods of constructing tubes are possible, of course.

1.7. The Electron Gun.—The electron gun consists of an indirectly heated cathode button, usually oxide-coated, and a focusing ring. An aperture or control grid for varying the beam current is provided in some designs. The control electrode and focusing ring are often combined into a single unit which also serves as a heat shield for the cathode. Although the metal shell of the klystron is ordinarily the anode, there may be exceptions to this design. Additional focusing of the electron beam, either magnetic or electric, may be applied after the beam has been accelerated, but the complication often outweighs the gains obtainable.

FIG. 1–5.—Type 707A klystron showing disk-seal construction. An external resonator must be connected to the copper disks which extend through the glass envelope. (*Courtesy of Bell Telephone Laboratories.*)

A noninductive cathode heater is usually essential for a-c operation, as any varying magnetic field in the vicinity of the electron gun will affect the path of the beam, introducing current variations that cause frequency modulation as well as amplitude modulation of the output. Although the hum from this source has been minimized by the use of a noninductive heater coil, a rectified a-c supply or battery may be used to heat the cathode if the hum caused by alternating current in the heater is objectionable. Other methods of heating a unipotential cathode surface are used for special klystron designs, but the details of such cathode construction will be described in a later chapter.

Various methods of adjusting the average beam current can be used if such control is desired. A mesh grid requiring positive control voltages is used in some tubes. Other designs using large apertures require negative control voltages and have the advantage that no power is required by the control electrode. However, they require considerably greater voltage variations for the same beam current control.

1.8. Cavity Resonators.—The fact that the resonators, which replace the parallel *LC* circuits used at lower frequencies, are an integral part of the construction of a klystron is often overlooked. It is true that the resonators may not be included entirely in the vacuum system, as in Fig. 1–5, and in such cases they may be removable, but the capacity section of the resonators must be acted upon by the electron beam and is usually built within the tube itself. The electrons must cross the gap formed by the capacity section of the resonator in less than one half cycle if the electric field is to modify the electron velocity efficiently or absorb power from the beam. Resonators of the form illustrated in Fig. 1–1 furnish a satisfactory design for use in klystrons.

Electrons must traverse the resonator gap but should not be intercepted, since the beam must continue along the drift space and transfer energy to the output resonator. The capacity section is therefore either an aperture or has a grid structure. An aperture usually has an effective gap width much greater than the actual spacing. Since the gap width is a limitation due to the requirement of a transit time less than a half cycle,

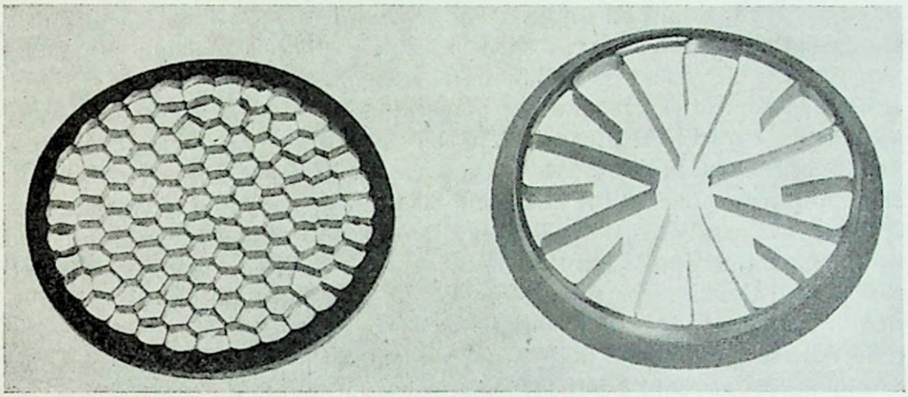


Fig. 1–6.—Two types of grid structures that are used in klystron tubes.

grid or mesh structures, as illustrated in Fig. 1–6, are usual in klystron resonators. Radial metallic fins are used frequently for higher power tubes. Fine wire mesh is often used for low-voltage designs.

1.9. Coupling to the Resonator—Some means of obtaining output from the tube must be provided and, in amplifier klystrons, an input coupling is also necessary. The radio-frequency electric field is strongest between the klystron grids, but coupling to the electric field in the resonator is not convenient. The magnetic field about the axis of the resonators offers a practical solution to the form of coupling; therefore, a loop is used which lies in a plane including the axis of the resonator. The loop size is usually chosen to load the resonator properly.

The loop that is coupled to the magnetic field in the resonator is connected to an output terminal by a short length of coaxial line. When the resonator is part of the vacuum envelope, this coaxial line must be closed with a glass-to-metal seal. The coaxial output terminal is occasionally called an "antenna seal." If the resonator is not a part of the vacuum envelope, as in tubes of disk-seal construction, the coupling loop and the output line may be insulated from the resonator. An insulated output line may also be used if a glass bulb extending into the magnetic field is provided when the resonator forms a part of the vacuum tube. The amount of coupling may be changed readily when the output line does not require a vacuum seal.

1.10. Tuning.—Several methods of tuning the resonators may be employed. A convenient tuning arrangement used in many klystron designs utilizes a flexible diaphragm as a part of the resonator wall to permit variation of the gap spacing between the resonator grids. This construction allows a tuning range of several per cent. Changing the volume of the resonator by inserting metal plungers also furnishes a means of tuning over a limited range. The resonator may be made external to the vacuum tube; in such cases it is possible to choose the tuning range by substitution of resonators of different size.

Two factors affect the tuning range of a klystron using variable grid spacing to control frequency. The electrons must traverse the electric field between the klystron grids in less than one half cycle. This transit-time limitation governs the minimum electron velocity and maximum grid spacing that can be used. The other limit to the tuning range is imposed by the fact that the losses become quite high if the grids are spaced too closely and the klystron fails to oscillate. An additional practical limitation is introduced by the fact that tuning becomes too critical before the tube refuses to oscillate. These factors limit the practical tuning range to about 20 per cent of the average wavelength. Considerably greater tuning range is easily obtainable at reduced output.

1.11. The Drift Space.—There are two resonators in the klystron amplifier illustrated in Fig. 1–1: an input resonator and an output resonator. The gaps formed by the resonator grids are separated by a drift space. This is the region of the tube in which the electron bunching action takes place. The electron beam must continue along this drift space without diverging to the walls of the tube. A field-free region is desirable for this purpose. In tubes of metal construction, the drift space is field-free if no more electrodes are added. However, in tubes with glass walls, it may be necessary to add a metal shield around the drift space inside of the glass envelope, or use some form of magnetic focusing, in order to prevent the accumulation of electron charges on the glass walls, which will destroy the focus of the electron beam.

In reflex klystrons, the functions of the input resonator and the output resonator are combined in a single resonator and the drift space is replaced by the reflecting field between the resonator gap and a reflector electrode. Other types of klystrons may require three or more resonators, and the different resonators may be separated by drift spaces; in some cases the distances between adjacent resonator gaps may be as small as possible so that no additional bunching action takes place. These special designs will be discussed in more detail in later sections.

1.12. Heat Dissipation.—Since klystrons are not one hundred per cent efficient, all of the energy in the electron beam is not converted into r-f power and it is necessary to dissipate the residual energy in the form of heat. It is also desirable to prevent unwanted electrons from returning through the resonator grids and possibly subtracting power from the resonator or causing undesired feedback. A stepped trap with fins on the outside surface, as shown in Fig. 1-1, provides a convenient method of dissipating the heat and reducing the number of secondary electrons that might return to the resonators.

CHAPTER 2

CAVITY RESONATORS

2.1. Limitations of Conventional Circuits.—It is the combination of cavity resonators with velocity modulation which has made klystron tubes practical at superhigh frequencies. This type of circuit is so important in the microwave region that it will be considered in some detail. Cavity resonators are used not only in klystron tubes; this type of resonant circuit is also necessary for magnetrons and special triode-type tubes when they are used at superhigh frequencies. The best method of presenting the circuit problems at microwave frequencies is to review the limitations of ordinary lumped constant circuits, consisting of an inductance and a capacitance in parallel.

One of these problems at the higher frequencies is the necessity for short leads between resonant circuit and the electrodes of the vacuum tube. It is obvious that the coupling between the vacuum tube and the circuit will not be satisfactory if the length of the leads to the resonant circuit becomes an appreciable fraction of a wavelength. The leads will act as a transmission line, and the voltage at the tube will be less than the voltage at the resonant circuit. Since a quarter wavelength at 3,000 megacycles is only 1 in., the difficulty is easily visualized. Another problem is introduced by the interelectrode tube capacities when the frequency is increased. This capacity may become the major part of the total capacity in the resonant circuit. In addition, the losses in a lumped constant circuit become excessive as the size of the circuit elements becomes smaller and smaller.

2.2. Development of a Cavity Resonator.—Most of these circuit difficulties may be avoided by making the resonant circuit an integral part of the vacuum tube, utilizing the tube electrodes as the capacity for the resonant circuit, and allowing the leads to become the inductance. Consider the single wire loop and the circular plate condenser shown in Fig. 2–1*A*. As the size of the loop is decreased to reduce the wavelength, the losses become tremendous and the loop practically vanishes before the desired wavelength is achieved by this method of approach. The alternative is to start from the single loop and condenser, with the capacity section modified as in Fig. 2–1*B* to permit the electron beam to pass, and reduce the resonant wavelength by adding a large number of loops in parallel to decrease the inductance. This would lead to the shape shown in Fig. 2–1C, which resembles the resonators actually used in klystrons. As an alternative, it would be possible to analyze a klystron resonator as a capacity loaded concentric transmission line. The latter method could easily lead to erroneous conclusions, however, when the dimensions involved are large compared to the wavelength.

It should therefore be reiterated that the ordinary conceptions of lumped constant circuits, if not used with care, may lead to fallacies when the dimensions involved are of the same order of magnitude as the wavelength.

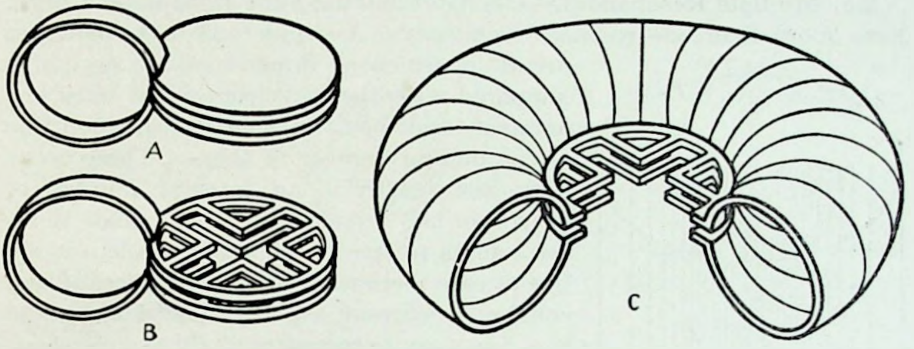


FIG. 2–1.—Development of a toroidal cavity resonator from a lumped constant circuit.

2.3. Cylindrical Cavity Resonators.—The warning above is even more important when cylindrical cavities are considered instead of the reentrant form of cavity resonator. Cylindrical cavities will be used to illustrate the principles underlying resonators.

Consider a circular metallic cylinder, closed at both ends with conducting plates. Assume that at some particular instant electrons flow away from the center of one end toward the outer diameter; at the same time electrons move toward the center of the opposite end. Allow the current along the cylinder to have any distribution required to fulfill this assumption. The current will soon die out, then reverse and flow in the opposite direction. Oscillation will continue until damped out by the losses in the conducting walls. Continuous oscillation can be maintained by supplying power at the required frequency from an external source. Although it is not obvious, it can be shown that, when the resonator is excited in this manner, the resonant frequency depends only upon the radius of the cylinder.

2.4. Field Patterns in a Cylindrical Cavity.—The charge distribution described in Sec. 2.3 indicates that the electric field would be parallel to the axis of the cylinder. A current must flow along the cylindrical wall of the cavity; therefore, a magnetic field concentric with the axis will exist.

These fields are illustrated in Fig. 2–2. This method of exciting a cylindrical cavity represents only one of the modes possible in a cylindrical cavity resonator. This particular mode has a resonant wavelength λ which is independent of the length of the cylinder, and is given by

$$\lambda = 2.61a \tag{2–1}$$

where a is the radius of the cylinder in centimeters and λ is the resonant wavelength in centimeters.

2.5. Multiple Resonances.—Cavity resonators, like transmission lines, have more than one resonant frequency. A simple coil and condenser circuit, in which all dimensions are negligible compared with the wavelength, has only one resonant frequency. A long transmission line has an infinite number of them. These occur when the length is an integral number of quarter or half wavelengths, and a single number suffices to specify the order of a harmonic. In this case there will be nodes, or places of zero voltage or current, equally spaced along the line. In a cavity resonator all three dimensions are in general comparable with a wavelength, and nodes can exist in three different directions. This complicates the situation somewhat. Not only are three numbers now required to specify the order of a certain "harmonic," but the higher resonant frequencies are no longer integral multiples of the lowest one.

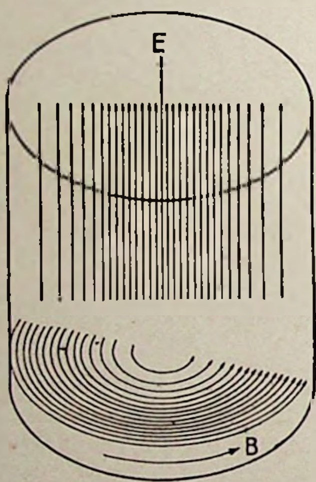


FIG. 2-2.—Electric and magnetic fields in a cylindrical cavity resonator.

It is possible to divide the fields within a cylindrical cavity resonator into two main types: one type has the electric field E parallel to the axis as in Fig. 2–2, and the magnetic field is in a transverse plane; the second type is a similar family with the electric and magnetic fields interchanged.

The transverse magnetic field is usually known as a TM mode, and the modes with a transverse electric field are indicated by the abbreviation TE. These field relations are analogous to the fields in wave guides. The usual wave-guide terminology for the different modes describes the field pattern and subscripts give the number of modes in the field in the circumferential and radial directions. For example, the field illustrated in Fig. 2–2 has a transverse magnetic field without any nodes around the circumference; *i.e.*, the magnetic field is circular. The magnetic field is zero at the axis; therefore, one node exists in the radial direction, and the designation TM_{01} is given to this mode.

In a discussion of cavity resonators it may be convenient to replace the usual wave-guide notation by the Bessel function describing the boundary conditions in the cavity. The introduction of Bessel functions occurs because these functions appear in the solutions of problems with circular symmetry. The strength of the electric field in Fig. 2–2 is proportional to the zero-order Bessel function J_0. This function appears in the equation given in Sec. 2.6 for the resonant wavelength of a cavity, and it will be convenient to refer to this mode of resonance as a J_0 mode because that designation allows the resonant wavelength to be determined readily.

2.6. Calculation of Resonant Wavelength.—The derivation of the resonant wavelength of a cavity resonator is rather complicated and will not be given here; it may be found in texts on ultrahigh-frequency waves.[1] If b is the length of the cavity and a is the radius, both in centimeters, the resonant wavelength λ for the TM modes is given by

$$\frac{1}{\lambda^2} = \left(\frac{p}{2b}\right)^2 + \left(\frac{k}{2\pi a}\right)^2 \tag{2–2}$$

where k is the value that makes the Bessel function J_n equal to zero. The multiple resonant frequencies for the cavity correspond to the different values permitted for n and p as shown below:

$$J_n(k) = 0$$

$$n = 0,1,2,3, \text{ etc.}$$

and

$$p = 0,1,2,3, \text{ etc.}$$

For the TE modes the resonant wavelength is determined by Bessel function derivatives.

$$\frac{1}{\lambda^2} = \left(\frac{p'}{2b}\right)^2 + \left(\frac{k'}{2\pi a}\right)^2 \tag{2–3}$$

$$J_n'(k') = 0$$

$$n = 0,1,2,3, \text{ etc.}$$

$$p' = 1,2,3, \text{ etc., } i.e., \ p' \neq 0$$

It can be seen that Eq. (2–1) is a special case of Eq. (2–2) for the first zero of the J_0 Bessel function.

$$p = 0$$

$$n = 0$$

$$k = 2.405$$

[1] R. J. Sarbacher and W. A. Edson, *Hyper and Ultra-High Frequency Engineering*, John Wiley & Sons, Inc., New York, 1943.

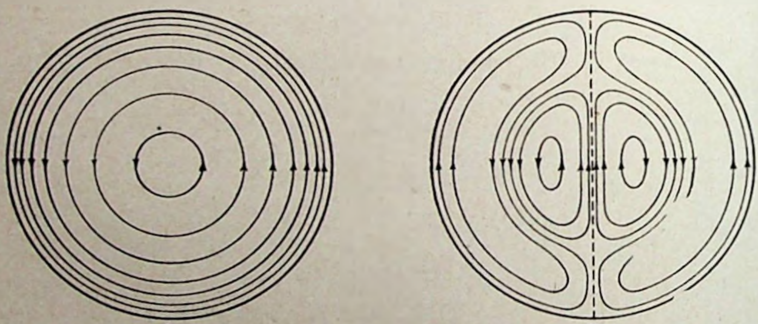
Note that p must be equal to zero if the resonant wavelength is to be independent of the length of the cavity. Also note that a whole family of modes, corresponding to higher order Bessel functions, may be independent of length. The higher order modes all correspond to shorter wavelengths, but the relation between these wavelengths is not harmonic.

A list of roots, or values of k that make the Bessel functions equal to zero, is given in Table I, which includes only the first two zeros for each Bessel function.

TABLE I.—BESSEL FUNCTION ROOTS

Bessel function	First root	Second root
J_0	2.405	5.520
J_1	3.832	7.016
J_2	5.136	8.417
J_0'	3.832	7.016
J_1'	1.841	5.331
J_2'	3.054	6.706

2.7. Field Pattern for a Higher Mode.—The magnetic field pattern in Fig. 2–2 is circular. An end view of it is repeated in Fig. 2–3A. In order to simplify the illustration for the higher modes, p is chosen equal to zero. This choice of p corresponds to no nodes being present in the



A B

FIG. 2–3.—Magnetic fields for higher modes in a cylindrical cavity resonator.

field along the length of the cylinder. The value of p, which is an integer, indicates the number of nodes along the axis of the cylinder. The magnetic field pattern for the mode corresponding to the second zero of the J_1 Bessel function is shown in Fig. 2–3B for the same size of cylindrical cavity. The value of n equal to unity indicates that there will be one node around the circumference, shown by a dotted line. The fact that the mode corresponds to the second zero of this Bessel function requires two patterns along the radius of the cavity.

It is apparent that this mode of excitation of a cavity resonator would be expected to have a smaller resonant wavelength, owing to the smaller size of the fields. Substituting values in Eq. (2–2) gives the resonant wavelength in centimeters.

$$\lambda = 2.61a \tag{2–1}$$

for the mode illustrated by Fig. 2–3A, and

$$\frac{1}{\lambda^2} = \left(\frac{7.016}{2\pi a}\right)^2 \tag{2–4}$$

or

$$\lambda = 0.89a \tag{2–5}$$

for the mode in Fig. 2–3B. In each case, a is the radius of the cavity in centimeters.

The wavelength referred to in Secs. 2.4 to 2.7 is the wavelength in centimeters in free space. The actual length between nodes within the cavity may be quite different. This difference is introduced by the velocity of propagation of the wave within the cavity being different from the velocity of light, and the situation is analogous to the variation of the velocity of propagation of a wave within a wave guide.

2.8. Definition of Circuit Parameters.—When the dimensions of the circuits become comparable to the wavelength involved, as in cavity resonators, many of the concepts of conventional circuits fail to apply. As an example of this statement, we shall consider various definitions of the ordinary circuit constants when applied to cavity resonators. Inductance L may be defined by two independent equations:

$$L = \frac{2 \times \text{energy stored}}{I^2} \tag{2–6}$$

$$L = \frac{\text{flux linkages}}{I} \tag{2–7}$$

where I is the current flowing in the circuit. Both definitions give an identical value for the inductance in a lumped constant circuit. However, when the dimensions of the circuit elements become almost as large as the wavelength, the two definitions give different values for the inductance because the definitions of current become ambiguous.

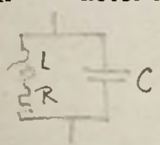
It is also possible to excite a cavity resonator so that the electric field pattern is always concentric with the axis of the resonator and capacity ceases to possess any usual interpretation. Measurement of current and voltage is difficult or even impossible. Therefore the idea of a resistance has little meaning, although it is often useful to think of an equivalent shunt resistance as defined below.

The measurement of radio-frequency power, resonant frequency or wavelength, and the band width of a resonator is easily made. We shall therefore describe a resonator with three parameters which are ordinarily considered as derived from L, C, and R, and future discussion of cavity resonators will refer to the following definitions. All these parameters depend upon the size and shape of the resonator.

$$\lambda = \text{resonant wavelength in centimeters} \tag{2-8}$$

$$Q = 2\pi \frac{\text{energy stored}}{\text{energy loss/cycle}} \tag{2-9}$$

Q is a loss factor and replaces the usual definition:



$$Q = \frac{\omega L}{R} \tag{2-10}$$

$$R_s = \text{shunt resistance of the resonator} \tag{2-11}$$

The shunt resistance of a resonator may be compared to the resistance represented by a parallel LC circuit at resonance.

2.9. Factors Affecting Resonator Characteristics.—Shape and size obviously determine the resonant wavelength of a cavity. Other factors that influence the resonator characteristics are not equally apparent. The ratio of volume to surface area determines Q as well as the losses in the conducting surface of the cavity, so that, in general, a shape that provides an increased ratio of the volume to the surface area improves the Q of the cavity. It is also true that the same shape with larger dimensions and longer resonant wavelength has a higher value of Q. A sharp reentrant point in a resonator may increase the current concentration tremendously and lower Q by a large factor.

Losses in a resonator are not only introduced by the resistivity of the conducting material of which it is constructed, but may be introduced by circuits coupled to the resonator and by losses in the load and input coupling loops. Losses in the conducting material are not proportional to the first power of the resistivity of the conductor, since r-f resistance depends on the depth of penetration of the current, or the skin depth, as well as on the resistivity itself. Actually, losses in a cavity resonator of nonmagnetic material are proportional to the square root of resistivity. A resonator of brass with a resistivity four times that of copper may have a Q one-half as great if no other losses are present. Losses in magnetic materials are usually much higher so that the Q of resonators of magnetic material may be very much lower than that indicated by the resistivity itself.

2.10. Shunt Resistance.—The shunt resistance of a resonator is also determined by a shape factor, but the dependence upon size and shape differs from the factors that determine Q. For the same size and shape, the shunt resistance is reduced by losses to the same degree that the losses reduce Q. The shunt resistance is a particularly important factor in the reentrant type of cavity that is required for klystron tubes. In this type of resonator, the shunt resistance determines the radio-frequency voltage that appears across the resonator gap when power is supplied to the resonator. This relation may be stated in the familiar form:

$$\text{Power} = \frac{(\text{voltage})^2}{R_s} \tag{2-12}$$

Since the shunt resistance is an important factor in determining the minimum current required to sustain oscillations in a klystron, its importance in klystron resonator design is evident. Decreasing the shunt resistance increases the starting current. Resonators with low Q are sometimes desired for wide frequency band applications. Shunt resistance is sacrificed when the Q is reduced by heavy loading, and the efficiency of low Q klystrons is decreased if other factors remain unchanged.

2.11. Beam Loading.—The presence of an electron beam in the gap of a klystron resonator affects the Q and also the shunt resistance. The resonant frequency may also be changed because the electron beam has an effect that is equivalent to changing the dielectric constant of the resonator gap. If the gap transit time is large, the electrons that pass the gap during the positive half of the cycle absorb more energy from the resonator field than will be given back to the resonator by the electrons that are slowed down during the negative half of the cycle. This means that the beam removes energy from the resonator and represents a loss which has the same effect as loading the resonator with a resistance. This loss, due to beam loading, decreases the Q of the resonator and also reduces the effective shunt resistance.

Since the beam loading depends upon both the current and the transit time or electron velocity, both the beam current and the acceleration voltage affect the amount that the Q and shunt resistance are reduced by the presence of an electron beam in the resonator gap. If the transit time across the gap is small, the beam loading due to transit time is negligible; however, the presence of secondary electrons in the gap may introduce a considerable loss. This beam loading due to secondary electrons will exist even when the transit time across the resonator gap is small.

2.12. Reentrant Shapes.—Cavity resonators with reentrant shapes are of particular interest because this type of resonator is used for klystron tubes. The simplest type is the quarter-wave coaxial line, although this type of resonator is not used in klystrons. A quarter-wavelength coaxial line, open at one end and shorted at the other, is quite resonant. The resonator may be made self-shielding by extending the outer conductor some distance beyond the end of the inner conductor and closing the end. If the closed end is sufficiently far from the open end of the inner conductor, there will be no appreciable capacity loading, and the resonant wavelength of such a cavity will be approximately four times the length of the inner conductor.

Varying the length of the inner conductor of a quarter-wave line is a convenient method of changing the resonant wavelength. The relationship between the length of the inner conductor and the resonant wavelength is quite linear over a wide range, although the ratio of proportionality is not four to one because fringing of the field at the open end of the inner conductor introduces a small error. Decreasing the length of the inner conductor to zero obviously does not decrease the resonant wavelength to zero because the cylindrical cavity formed by the outer conductor will be resonant.

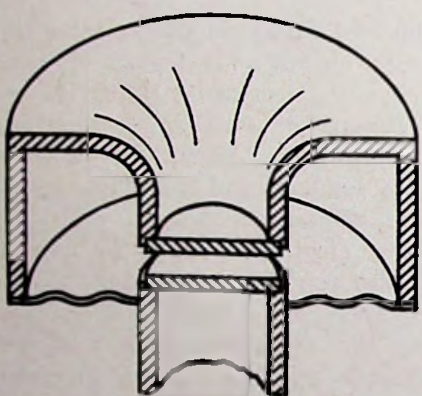


FIG. 2-4.—Reentrant cavity resonator of the type used in most klystron designs.

2.13. Capacity Loading.—Quarter-wavelength reentrant cavities and nonreentrant cylindrical cavities have very high shunt resistances but are not satisfactory as klystron resonators because they do not have a region of strong electric field that can be traversed with a short transit time. The type of reentrant cavity illustrated in Fig. 2–4 furnishes a strong uniform electric field over a considerable area, the gap width is small, and the shunt resistance is satisfactorily high, although less than some of the other resonators described previously. This type of resonator may be considered a capacity-loaded coaxial line. In the same way that the parallel line can be made shorter and tuned with a variable capacity, the quarter-wave coaxial line may be decreased in size by bringing the end of the cavity close to the inner conductor to add capacity loading.

This construction furnishes the strong, uniform electric field that is required for interaction between the resonators and the electron beam of the klystron tube. The capacity section of a klystron resonator is usually made in the form of grids so that the beam will be acted upon by the resonator but will not be intercepted. Two types of grid construction are illustrated in Fig. 1–6 in Sec. 1.8.

2.14. Calculation of Resonance.—The calculation of the resonant frequency of reentrant cavities is not simple as in the case of circular cylindrical cavities, because the fields and boundary conditions are not the same in different regions of the cavity. However, it is possible to make a *crude* estimate of the resonant frequency of a capacity-loaded resonator by assuming that the resonator consists of a lumped capacity which can be determined from the formula for a parallel plate condenser, and a lumped inductance which is given by the transmission-line equations. The capacitance C in farads would be

$$C = 0.088 \times 10^{-12} \frac{A}{d} \tag{2-13}$$

where A is the area in square centimeters and d is the gap spacing in centimeters.

The inductance of a short coaxial line may be obtained from the transmission-line equations. If the length l, in centimeters, of the inner conductor is very much less than a quarter wavelength, the inductance L in henries for a short coaxial line is

$$L = 3.9 \times 10^{-9}\, l \log_{10} \frac{a_2}{a_1} \tag{2-14}$$

where a_2 is the radius of the outer conductor and a_1 is the radius of the inner conductor. Equation (2-14) does not apply to transmission lines that are used in reentrant cavities and must be multiplied by a correction factor,

$$\frac{\tan (2\pi l/\lambda)}{2\pi l/\lambda}$$

to obtain the inductance of a long transmission line. The inductance of a long coaxial transmission line becomes

$$L = 3.9 \times 10^{-9} \frac{\lambda}{2\pi} \tan \frac{2\pi l}{\lambda} \log_{10} \frac{a_2}{a_1} \tag{2-15}$$

An approximate value of the resonant frequency can then be obtained from the familiar expression

$$f = \frac{1}{2\pi\sqrt{LC}} \tag{2-16}$$

The resonant frequency computed from Eqs. (2-13), (2-15), and (2-16) may be in error by as much as 30 per cent. This inaccuracy is introduced by the assumption that the resonator can be treated as a lumped constant circuit. If more exact assumptions for the field distribution within the cavity are made, including the effects of fringing of the fields, the calculation becomes rather difficult. The results of exact calculations, based on these more accurate assumptions, are presented in Charts I to IX in Appendix B.

2.15. High-frequency Resonances.—Coaxial cavity resonators also have higher resonant frequencies. The quarter-wavelength line, for example, may also be excited at any odd number of quarter wavelengths. This type of resonator would resonate not only at the fundamental frequency but also for the third, fifth, and other odd harmonics. When capacity loading exists, however, the higher resonances no longer occur in harmonic relation.

2.16. Tuning Cavity Resonators.—Most types of cavity resonators can be tuned readily; this statement applies to nonreentrant as well as to reentrant cavities. The mode illustrated in Fig. 2–2, which depends only upon the radius of the cylinder, is a notable exception. Other modes in cylindrical cavities which depend upon the length b as well as upon the radius may be tuned by varying the position of a large plunger which acts as one end of the cylinder. Cylindrical cavities may also be tuned a small amount by inserting small plungers into the cavity, or by rotating a small metal vane in the field through an angle of 90 deg.

Quarter-wavelength coaxial line resonators may be tuned by changing the length of the inner conductor. This effect was discussed in some detail in Sec. 2.12.

Klystron resonators similar to the design shown in Fig. 2–4 may be tuned by one or both of two basic methods. Varying the spacing of the gap between the resonator grids is often a convenient method of changing the resonant frequency. This method of tuning corresponds to tuning a conventional circuit by changing the capacity of a variable condenser. Decreasing the spacing of the resonator gap increases the capacity and tunes the resonator to a lower frequency. The spacing of the gap is usually varied by distorting the end wall of the resonator, shown as a flexible diaphragm in Fig. 2–4. This motion will also change the volume of the resonator slightly, but the capacity change is the predominant effect in tuning the resonator.

Tuning may also be accomplished by changing the volume of the cavity. This method of tuning is equivalent to changing the inductance in a resonant transmission line. If the klystron is designed for resonators external to the vacuum envelope, the methods of changing the volume are almost unlimited. When the cavity is constructed like a coaxial line, a sliding plunger may be used to change the length of the line and vary the resonant frequency over a very wide range. Or small plungers may be inserted from the sides of a small, fixed-volume type of external resonator, to change the frequency over a smaller range which might be compared to band-spread tuning. The latter method might be considered equivalent to changing the effective diameter of the coaxial line.

CHAPTER 3

ELECTRON-BUNCHING THEORY

3.1. Velocity Modulation.—Although klystron tubes are quite generally known as "velocity-modulation tubes," the term is not accurately descriptive of their operating principle. The velocity-modulation process actually refers only to the principle of superimposing a periodic variation of velocity upon the average velocity of the electrons in a beam. This velocity modulation may then be utilized in various ways. Some tubes operate upon the principle of velocity selection at one of the electrodes. The term "klystron" is applied to a vacuum tube that utilizes the velocity modulation by allowing a relatively long transit time to convert the velocity-modulated beam into a beam with periodic variations of current density. The process has become known as "electron bunching" in this country and is frequently called "phase focusing" in Europe. Some tubes, klystron detectors, for example, may utilize both electron bunching and velocity selection.

The introductory discussion of electron bunching in Chap. 1 does not adequately explain the electrical characteristics of klystron tubes, particularly their dependence on voltage, which is one of the outstanding differences between these tubes and more conventional triode vacuum tubes. A complete analysis of electron bunching will be required. Several methods of analysis may be used, one of the simplest of which considers the electrons in the beam as particles, acted upon by forces that affect their motion. This method of analysis gives a physical picture of the electron-bunching process which is helpful in understanding the operation of these tubes. The basic theory is illustrated best by an analysis of a klystron amplifier; the ideas developed in this manner can then be extended to explain the behavior of more complicated klystron tubes such as oscillators, frequency multipliers, and other designs.

3.2. Electron Ballistics.—A brief review of electron ballistics will describe the motion of the electrons and allow their transit times to be calculated. The electron-bunching equations, which relate the output current to the input gap voltage and the beam voltage, are derived from these transit times. This analysis does not include all the factors that are important in klystron design, but it is a sufficiently good approximation so that the most important considerations can be included. A complete analysis becomes so complicated that the simpler relations may become lost.

The acceleration of the electrons by the electron gun will give a velocity v_0, which is determined by the acceleration voltage E_0. The emission velocity at the cathode will be assumed equal to zero for all electrons. The acceleration voltage is also known as the "beam voltage" and is the voltage between the cathode and the anode plane of the klystron. The relation between the average electron velocity v_0 and the acceleration voltage may be obtained from the fact that the kinetic energy gained by an electron of mass m and charge e is equal to the potential energy that accelerates the electron. This relation may be stated

$$\frac{1}{2} mv_0^2 = E_0e \tag{3-1}$$

or Eq. (3-1) may be rewritten in the form

$$v_0 = \sqrt{\frac{2e}{m} E_0} \tag{3-2}$$

The constants in Eq. (3-2) may be evaluated in the proper units so that the velocity is in centimeters per second and E_0 is expressed in volts; then

$$v_0 = 6 \times 10^7 \sqrt{E_0} \tag{3-3}$$

All electrons have the same velocity when they enter the input resonator gap, but the radio-frequency input to the amplifier produces a voltage across the input gap. As a result, the electron beam is velocity-modulated after passing the gap. The velocity of an electron in the drift space between the input and output resonators will depend upon the time that electrons passed the input gap. E_1 is the peak value of the radio-frequency voltage across the input gap, f_1 is the frequency of oscillation, ω_1 is the angular frequency corresponding to f_1, and t_1 is the departure time of an electron. The instantaneous voltage across the input gap will be

$$E = E_1 \sin \omega_1 t_1 \tag{3-4}$$

The effective voltage acting upon the electron will be $E_0 + E$ and the velocity of the electron departing at the time t_1 will be

$$v = \sqrt{\frac{2e}{m} (E_0 + E_1 \sin \omega_1 t_1)} \tag{3-5}$$

3.3. Assumptions in the Bunching Analysis.—Several important assumptions are involved in this analysis. The amount of velocity modulation is considered quite small, and adequate drift time is assumed so that any degree of bunching may occur. Also, the transit time across the resonator gaps is assumed to be negligible, and space-charge forces have been neglected. The effect of violating these assumptions, and methods for correcting the analysis when these assumptions are not valid, are discussed in later sections.

3.4. Transit Time in the Drift Space.—A sectional view of a klystron amplifier, indicating the voltages and dimensions that are important in this analysis, is shown in Fig. 3-1. The transit time T between input and

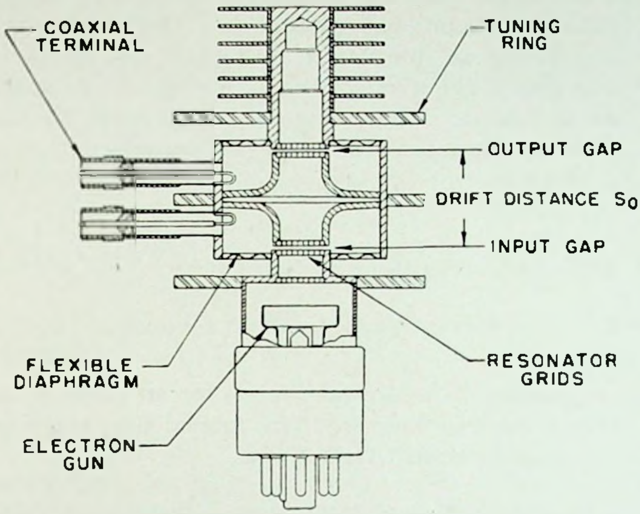


FIG. 3-1.—Sectional view of a klystron amplifier.

output gaps will be required in the calculation of the bunched beam current. This transit time is determined by the drift distance s_0 and the velocity v. The expression for the velocity in Eq. (3-5) may be rewritten

$$v = \sqrt{\frac{2e}{m} E_0} \sqrt{1 + \frac{E_1}{E_0} \sin \omega_1 t_1} \tag{3-6}$$

or, in the approximate form,

$$v \cong v_0 \left(1 + \frac{E_1}{2E_0} \sin \omega_1 t_1\right) \tag{3-7}$$

since the ratio E_1/E_0 has been assumed to be quite small. Then the transit time T will be given by

$$T = \frac{s_0}{v} = \frac{s_0}{v_0\left(1 + \dfrac{E_1}{2E_0} \sin \omega_1 t_1\right)} \tag{3-8}$$

Another mathematical approximation converts Eq. (3-8) into a more useful form:

$$T = T_0\left(1 - \frac{E_1}{2E_0} \sin \omega_1 t_1\right) \tag{3-9}$$

where T_0 is the transit time corresponding to an electron with average velocity.

3.5. Bunched-beam Current.—The relation between the bunched-beam current I_2 and the transit time can be derived by considering the number of electrons passing the input gap at time t_1. The number of electrons crossing the gap during an interval dt_1 will be $I_0\, dt_1$, since the current passing the input gap is the average beam current and I_0 is the number of electrons per unit time. These same electrons reach the output gap at a time t_2. This same number of electrons will pass the output gap during an interval dt_2; therefore

$$I_2\, dt_2 = I_0\, dt_1 \tag{3-10}$$

Equation (3–10) may be rewritten

$$I_2 = I_0 \frac{dt_1}{dt_2} \tag{3-11}$$

In order to evaluate I_2, the arrival time t_2 for an electron leaving the input gap at time t_1 must be known. This arrival time is the sum of the departure time t_1 and the transit time T; *i.e.*,

$$t_2 = t_1 + T = t_1 + T_0 - T_0 \frac{E_1}{2E_0} \sin \omega_1 t_1 \tag{3-12}$$

Differentiating Eq. (3–12) gives

$$dt_2 = dt_1 - \omega_1 T_0 \frac{E_1}{2E_0} \cos \omega_1 t_1\, dt_1 \tag{3-13}$$

The transit time T_0 corresponds to a certain number of oscillation cycles during the time an electron of average velocity is traveling along the drift space. This number will be designated N and is merely another way of expressing the transit time from the input gap to the output gap for an electron with average velocity. It is not necessarily an integer and may have any value in a klystron amplifier, but N is restricted to certain values that satisfy the proper phase relations if the klystron tube is used as an oscillator. The value of N should not be confused with the number of times the electrons may become bunched in the drift space. In an over-bunched amplifier, the bunch may form, then separate, and reform again farther along the drift tube. Oscillators do not become overbunched in this manner, and the electrons are formed into a bunch only once, although a number of bunches may be in the process of formation at the same time if transit of the drift space requires more than one cycle. Reference to an Applegate diagram (Fig. 1–3 in Sec. 1.5) will show that N also corresponds to the number of bunches in the process of formation at any instant in time.

The term $\omega_1 T_0$ in Eq. (3–13) is a transit angle, and in some cases this average value of the transit angle in the drift space will be indicated by the symbol τ_0. It will be convenient to replace the transit angle in Eq. (3–13) by its equivalent in terms of N.

$$\tau_0 = \omega_1 T_0 = 2\pi N \tag{3–14}$$

Equations (3–13) and (3–14) may then be substituted in Eq. (3–11) to obtain

$$I_2 = \frac{I_0}{1 - \pi N(E_1/E_0) \cos \omega_1 t_1} \tag{3–15}$$

or

$$I_2 = \frac{I_0}{1 - x \cos \omega_1 t_1} \tag{3–16}$$

The quantity x is known as the bunching parameter and may be evaluated from the relation

$$x = \pi N \frac{E_1}{E_0} \tag{3–17}$$

3.6. Arrival-time Curves.—Equation (3–16) does not furnish all of the information required for evaluating I_2 because it relates the instantaneous current at the output gap to the time the electrons passed the input gap. The relation between the arrival time t_2 and the departure time t_1 is given by Eq. (3–12). This relation may be rewritten in terms of the bunching parameter x to obtain

$$t_2 = t_1 + T_0 - \frac{x}{\omega_1} \sin \omega_1 t_1 \tag{3–18}$$

Since it is not possible to solve Eq. (3–18) for t_1 in terms of t_2, it is convenient to plot t_2 as a function of t_1. Such a curve for x equal to 1.84 is shown in Fig. 3–2 with t_1 as the vertical coordinate. Then I_2 may be evaluated by obtaining the value of t_1 corresponding to the time t_2, and substituting this value of t_1 in Eq. (3–16).

The bunched-beam current may also be obtained directly from Fig. 3–2 by measuring the slope of the curve and substituting the value of dt_1/dt_2 into Eq. (3–11). When the bunching parameter x is greater than unity, there may be three values of t_1 which correspond to the same time t_2. This means that electrons that left the input gap at three different times arrive at the output gap simultaneously; therefore the sum of the slopes must be used to obtain the total current. The slope dt_1/dt_2 will be negative for some values of t_1. This negative sign merely indicates that electrons that left the input gap later have passed other electrons in the drift space and arrive at the output gap first. For this reason, the sum of the absolute values of the slopes is used, and the negative sign is ignored in computing the bunched current.

If the value of the bunching parameter x is greater than unity and cos $\omega_1 t_1$ is positive, the value of I_2 given by Eq. (3–16) may become negative. The term $1 - x \cos \omega_1 t_1$ is the reciprocal of the slope dt_1/dt_2; therefore, the absolute magnitude must also be used in computing the current contribution with this equation.

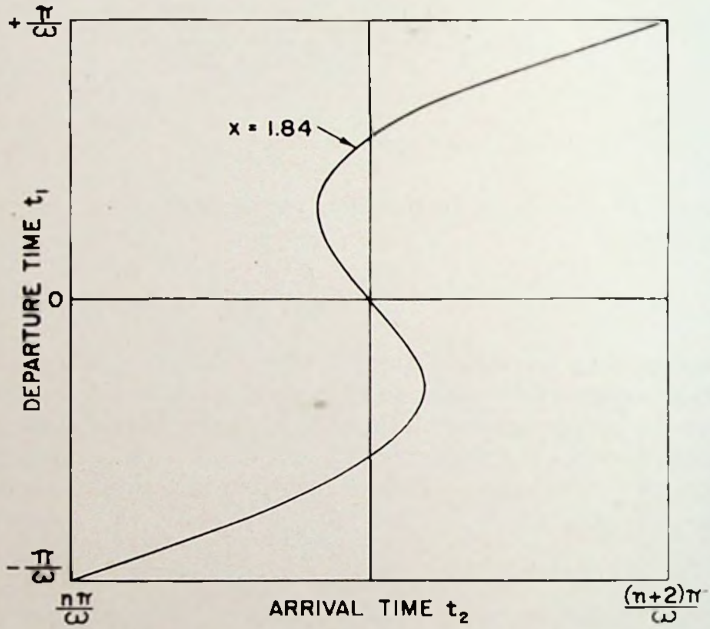


FIG. 3-2.—Electron arrival-time curve for a value of $x = 1.84$ for the bunching parameter. One complete cycle is shown.

3.7. Current Distribution.—Curves of instantaneous beam current at the output gap I_2/I_0 are plotted in Fig. 3-3 for three values of the bunching parameter: $x = 0.50$, $x = 1.00$, and $x = 1.84$. Two complete cycles are shown; the time corresponding to one cycle is $2\pi/\omega_1$. When the bunching parameter x is less than unity, the instantaneous current is finite, but an indication of a sharp peak is evident. When x equals unity, the current peak becomes infinite. For values of x greater than unity the infinite peak separates as indicated in Fig. 3-3.

3.8. Current Curves from Cycloids.—The expression for the bunched-beam current in Eq. (3–16) is not very convenient for plotting the instantaneous current. However, a graphical method is available because the denominator in Eq. (3–16) is one of the parametric equations for a cycloid and it is possible to modify Eq. (3–18) to obtain the second parametric equation which relates t_1 and t_2. Therefore the reciprocal of the instantaneous current can be obtained from a cycloid with a rolling circle of unit radius and a generating circle with a radius equal to x, the bunching parameter defined by Eq. (3–17). The beam current is proportional to the reciprocal of the value obtained from the cycloid curve when it is single-valued, and the sum of the reciprocals of the absolute magnitude of the values from the cycloid when x is greater than unity. The basis for neglecting the negative sign was explained in Sec. 3.6.

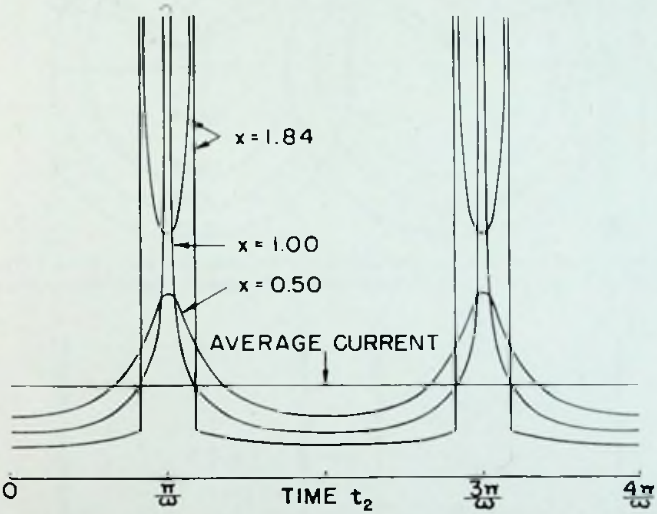


FIG. 3–3.—Current distribution for the first peak in bunching. Two complete cycles are shown. Three values of the bunching parameter are illustrated.

A series of cycloids representing six different values of the bunching parameter is shown in Fig. 3–4, and corresponding current-distribution curves computed from these cycloids are illustrated by Figs. 3–3, 3–5, and 3–6.

Figure 3–3 includes the current-distribution curves for three values of the bunching parameter: $x = 0.50$, $x = 1.00$, and $x = 1.84$. $x = 1.84$ is the value required for maximum output. A value of unity for the bunching parameter gives a single infinite current peak, but this value does not correspond to optimum bunching. More energy can be extracted from the bunched beam if the infinite current peak is allowed to diverge slightly, since the energy in the bunch is increased and fewer electrons remain in the half of the cycle that subtracts energy from the radio-frequency field in the output resonator.

3.9. Effect of Overbunching.—If the bunching process is allowed to continue, the current peaks will diverge until there is no conversion of energy from the beam. The current distribution for this case is shown in

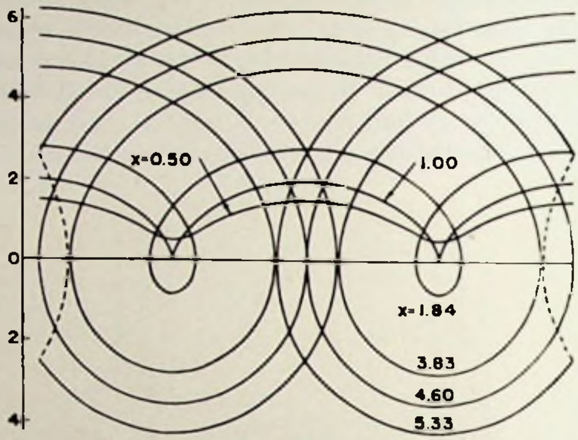


FIG. 3-4.—Family of cycloids for obtaining current-distribution curves. The bunching parameter x corresponds to the ratio of the radius of the generating circle to the radius of the rolling circle.

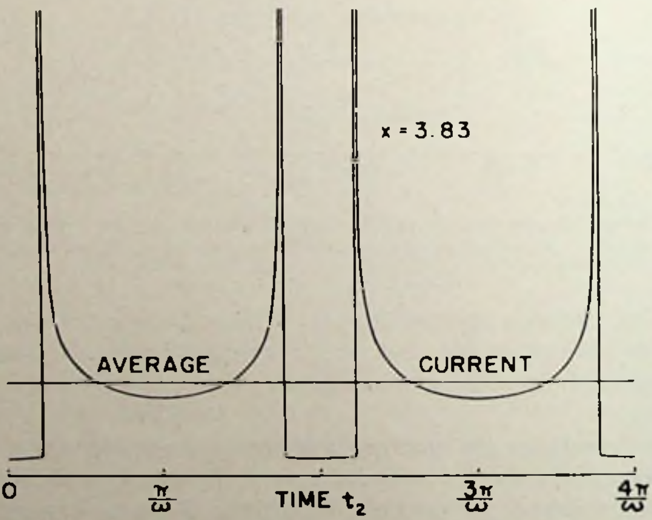


FIG. 3-5.—Current distribution giving zero output corresponds to a value of $x = 3.83$ for the bunching parameter.

Fig. 3-5 and occurs when the bunching parameter has a value of 3.83. Note that the beam has not become direct current, but that two current peaks still exist. These two current peaks are not 180 deg. out of phase, as might be expected, but actually 86 deg., as shown by Fig. 3-5. The basis for zero output is not obvious from inspection of Fig. 3–5, but depends upon the fact that the integrated effect over a complete cycle is zero.

If overbunching is increased, the two infinite current peaks will merge and form a single infinite current peak that is 180 deg. out of phase with the first infinite peak when the bunching parameter equals unity. This second single infinite peak corresponds to a value of 4.60 for the bunching parameter. Further increase of the bunching until $x = 5.33$ produces a second maximum in the output. These current-distribution curves are shown in Fig. 3–6.

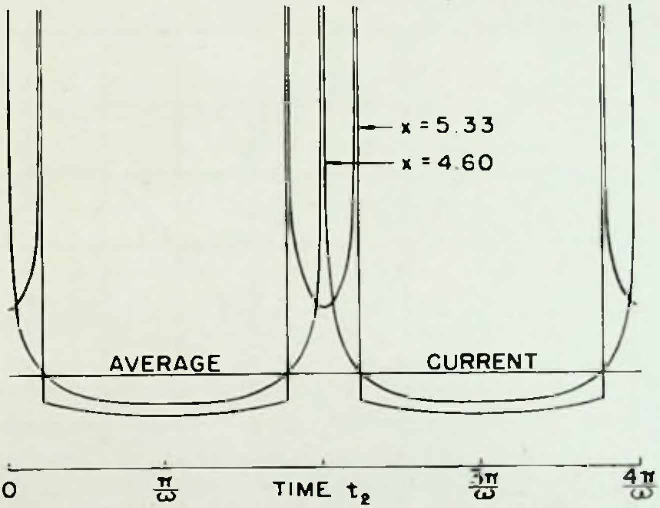


FIG. 3–6.—Current distribution for the second peak in bunching when the bunching parameter x is large.

3.10. Harmonic Analysis of Bunched Current.—This sharply peaked current distribution is rich in harmonics. It is possible to evaluate the instantaneous current as a Fourier series, and the amplitudes of the harmonics are given by Bessel functions. The expression for the bunched current becomes

$$I_2 = I_0 [1 + 2J_1(x) \sin (\omega_1 t_2 - 2\pi N) + 2J_2(2x) \sin 2(\omega_1 t_2 - 2\pi N) + \dots + 2J_n(nx) \sin n(\omega_1 t_2 - 2\pi N)] \quad (3\text{–}19)$$

Only the second term is of interest in a klystron amplifier. It is the fundamental component of the radio-frequency current and will be designated i_2.

$$i_2 = 2I_0 J_1(x) \sin (\omega_1 t_2 - 2\pi N) \quad (3\text{–}20)$$

3.11. Output Current Characteristic.—The peak value of the radio-frequency current is proportional to the J_1 Bessel function as indicated by Eq. (3–20). The magnitude of the fundamental component is plotted in Fig. 3–7 as a function of the bunching parameter x. In order to make the illustration more general, the expression for x given in Eq. (3–17) has been modified by a "beam-coupling" coefficient, β, which must be introduced when the transit time across the input gap is not negligible. The correct value of the bunching parameter is

$$x = \beta\pi N \frac{E_1}{E_0} \tag{3–21}$$

This effect is discussed in Sec. 3.12.

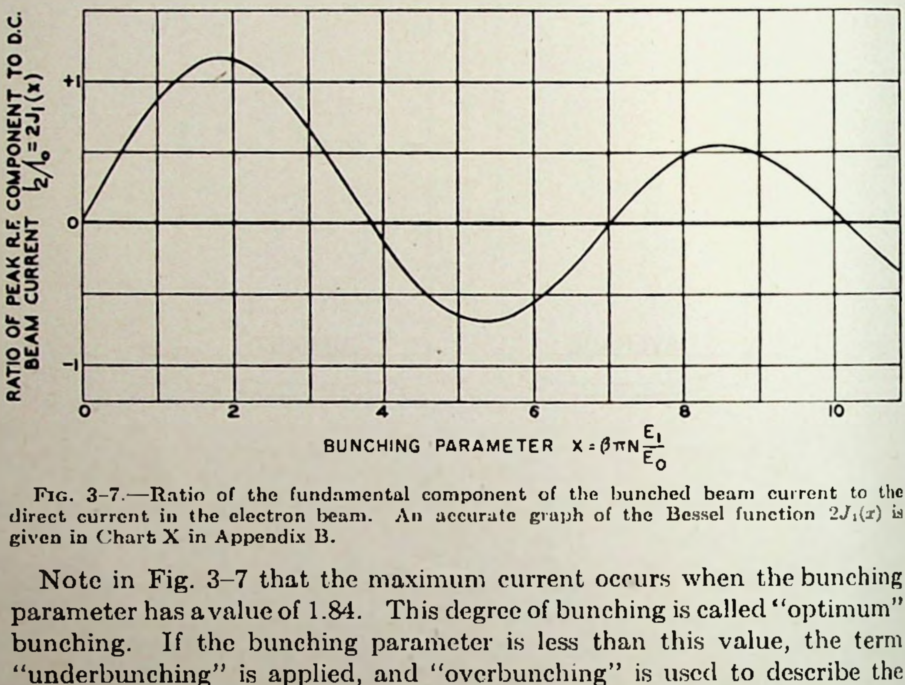


FIG. 3–7.—Ratio of the fundamental component of the bunched beam current to the direct current in the electron beam. An accurate graph of the Bessel function $2J_1(x)$ is given in Chart X in Appendix B.

Note in Fig. 3–7 that the maximum current occurs when the bunching parameter has a value of 1.84. This degree of bunching is called "optimum" bunching. If the bunching parameter is less than this value, the term "underbunching" is applied, and "overbunching" is used to describe the region where x has a value greater than that required for optimum bunching. Note that the current is zero when the bunching parameter has a value of 3.83, but that the current increases again for larger values of x. The negative value of the Bessel function indicates that the phase of the radio-frequency current has changed 180 deg. This phase change is apparent in the illustrations of current distribution for an overbunched beam described in Sec. 3.9.

This bunching characteristic of klystron tubes is analogous to the I_p vs. E_g characteristic for conventional vacuum tubes. The analogy is quite convenient and furnishes the basis for the definition of a transconductance for a klystron tube. However, it is important to remember two basic differences between the bunching characteristic in Fig. 3–7 and the triode analogy. Increasing the input gap voltage in a klystron will eventually reduce the output to zero owing to the effect of overbunching. This behavior is quite a contrast to the saturation characteristic of a triode or pentode. The second important difference occurs in the definitions of transconductance. The transconductance of a triode is the slope of a line tangent to the curve at the operating value of grid bias. The operating point on the klystron bunching characteristic represents the peak value of the radio-frequency voltage and is more nearly analogous to ΔE_g than to E_g. For this reason, the transconductance of a klystron is proportional to the slope of a line drawn from the origin to the operating point on the curve.

3.12. Effect of Transit Time at the Input Gap.—The previous discussion has ignored the effect of the transit time of the electrons in the resonator gaps although a correction factor β was introduced in Eq. (3–21). If an electron crosses the gap in a small fraction of an oscillation cycle, then the change in kinetic energy will be determined by the potential difference across the gap at that instant and no correction factor will be needed. However, if the electron requires a full cycle to traverse the resonator gap, it will be accelerated during half of the cycle and decelerated during the remainder of it. As a result, the net change in kinetic energy will be zero if the gap voltage is very small compared to the beam voltage. This effect may be included in the analysis by introducing the beam-coupling coefficient mentioned in Sec. 3.11.

It is necessary to know the transit time across the gap in order to evaluate β. If the gap spacing is d, an electron with average velocity v_0 will cross the gap in time equal to d/v_0. The gap transit angle δ will be the product of the angular frequency ω_1 and the gap transit time.

$$\delta = \omega_1 d/v_0 \tag{3–22}$$

The interchange of energy between the resonator and the beam will depend upon the average value of the gap voltage during the time the electron is crossing the gap. If the expression for the gap voltage given in Eq. (3–4) is averaged over a period corresponding to the gap transit angle δ, the result will be

$$\begin{aligned}\frac{1}{d/v_0}\int_{t-d/2v_0}^{t+d/2v_0} E_1 \sin \omega_1 t\, dt &= \frac{E_1}{\omega_1 d/v_0}\left[\cos \omega_1\left(t - \frac{d}{2v_0}\right) - \cos \omega_1\left(t + \frac{d}{2v_0}\right)\right] \\ &= \frac{2E_1}{\omega_1 d/v_0} \sin \omega_1 t \sin \frac{\omega_1 d}{2v_0} \\ &= \frac{\sin\ (\delta/2)}{\delta/2} E_1 \sin \omega_1 t \end{aligned} \tag{3–23}$$

The beam-coupling coefficient β is therefore given by the relation

$$\beta = \frac{\sin (\delta/2)}{\delta/2} \tag{3-24}$$

When the gap transit time is small, the value of β approaches unity and Eqs. (3-21) and (3-17) for the bunching parameter become identical. In practice, β is always less than unity and Eq. (3-21) must always be used for the bunching parameter.

3.13. Effect of Transit Time at the Output Gap.—A similar conversion loss occurs at the output gap and may be included in the analysis by considering that the effective radio-frequency component of the bunched-beam current is βi_2 instead of i_2. If the output gap voltage is large and

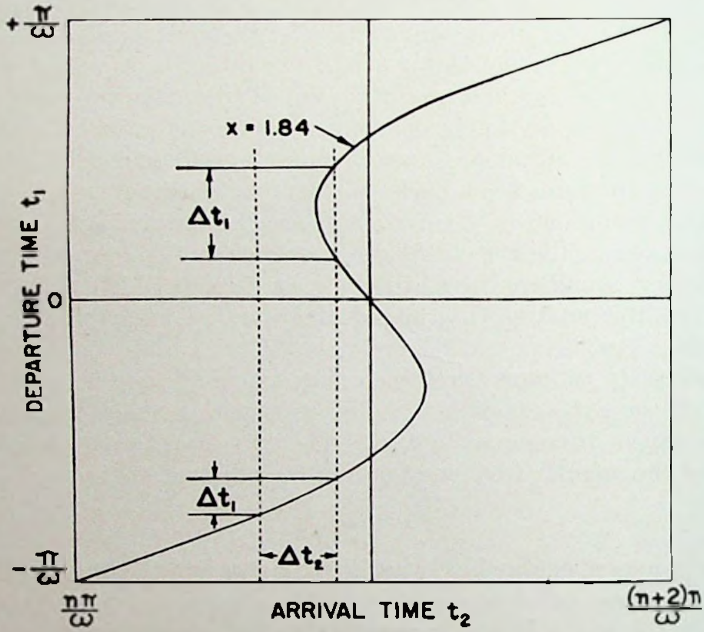


FIG. 3-8.—Electron arrival-time curve for a value of $x = 1.84$ for the bunching parameter. The effect of finite transit time at the resonator gap is illustrated.

comparable to the beam voltage, the value of β for the output gap will be different from the value for the input gap because the electrons will not be traveling with average velocity. This factor will not be important in many tubes, particularly in reflex oscillators; any further analysis of specific tube types will assume that β is equal at the input and output gaps. It is also important to remember that the expression for β in Eq. (3-24) is not valid when the velocity variation is large compared to the average velocity; *i.e.*, when the output gap voltage is comparable to the beam voltage.

An interesting graphical analysis of the effect of transit time at the output gap may be obtained from a t_1 vs. t_2 diagram. Figure 3–2 in Sec. 3.6 is redrawn in Fig. 3–8 to show the effect of a gap transit time corresponding to a transit angle of $\pi/4$ radians. A value of 1.84 is used for the bunching parameter. This value of x is correct for optimum bunching under the assumptions of negligible gap transit time and small input gap voltage.

In the derivation of the expression for the bunched-beam current in Eq. (3–16), the assumption of negligible transit time at the output gap permitted the use of the derivative dt_1/dt_2. However, when the gap transit

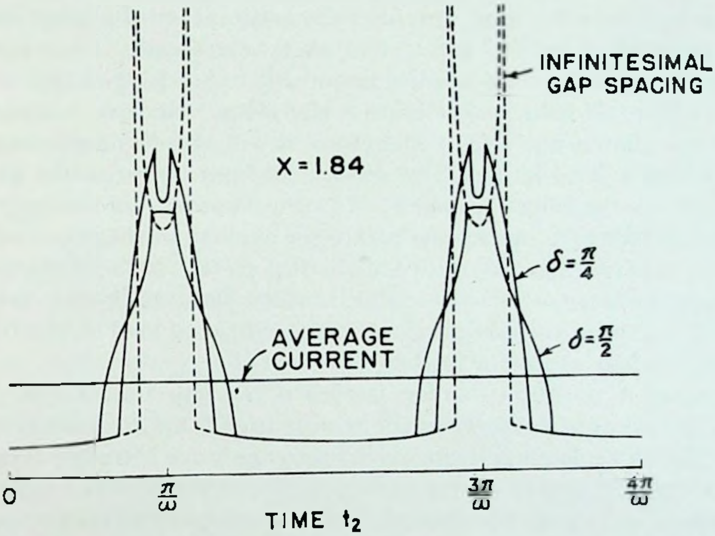


FIG. 3–9.—Current-distribution curves for a value of $x = 1.84$ and gap transit times corresponding to one-eighth and one-quarter of a cycle.

time is large, the incremental ratio $\Delta t_1/\Delta t_2$ must be used. The computation of the current corresponding to time t_2 equal to $(n + 3/4)\pi/\omega_1$ is illustrated in Fig. 3–8. The increment Δt_2 is shown equal to $\pi/4\omega_1$ and can be considered equivalent to unit distance for the graphical computation of current. Then the sum of the two increments Δt_1, measured by the same scale, gives the instantaneous beam current. For the case illustrated,

$$\sum \frac{\Delta t_1}{\Delta t_2} = 0.49 + 1.13 = 1.62 \tag{3–25}$$

Curves of bunched-beam current, including the effect of gap transit time, are shown in Fig. 3–9 for gap transit times corresponding to transit angle of $\pi/4$ and $\pi/2$. The theoretical curve for infinitesimal transit time is shown dotted for comparison. All curves correspond to a value of 1.84

for the bunching parameter. These curves may also be obtained by averaging the current given by the dotted line, or the similar curve in Fig. 3-3, over a period corresponding to the transit angle δ. This averaging process is equivalent to converting dt_1/dt_2 into $\Delta t_1/\Delta t_2$. Note that the double-peaked form of current distribution disappears when the gap transit time becomes greater than the spacing between the two current peaks in the theoretical case for infinitesimal transit time.

3.14. Beam Loading.—It was stated in Sec. 3.12 that the net change in kinetic energy of an electron would be zero if the gap transit time was equivalent to one complete cycle. This is true regardless of the time when the electron enters the gap, provided the assumption of a very small gap voltage is satisfied. If the gap transit angle corresponds to one half cycle, however, the entering time is quite important. An electron that traverses the gap when the field is accelerating the beam will have a transit time slightly less than a half cycle; therefore, it will absorb more energy from the field than will be returned by an electron that traverses the gap when the field is decelerating the beam. In other words, more energy will be absorbed by the beam during one half of the cycle than the beam will transfer to the field during the other half of the cycle. As a result, power is required to produce a velocity variation when the transit time is one half cycle. This power loss is equivalent to an additional load on the resonator and is referred to as "beam loading."

The transit-time beam loading is zero if the gap transit time is zero, and has a maximum value when the transit time is approximately one half cycle. The beam loading decreases to zero again for a transit time of one complete cycle and then becomes negative. This negative beam loading is analogous to the negative resistance characteristic of an oscillating diode. The magnitude of the beam loading is not only determined by the gap transit time, and therefore dependent upon the beam voltage, but it is also proportional to the beam current.

Secondary electrons may be another source of beam loading. If secondary electrons are formed in the resonator gap, the greater number of secondaries will have very low velocities when introduced into the gap and will leave the gap with a larger velocity regardless of the direction of the radio-frequency field. Therefore all these secondary electrons will absorb energy from the field and introduce a power loss. This form of beam loading may be expected to be proportional to the beam current, but there is no direct relation between the magnitude of the beam loading introduced by secondary electrons and the beam voltage applied to the tube.

3.15. Space-charge Debunching.—The effect of the forces between the electrons in the beam was neglected in order to simplify the analysis; however, these space-charge forces are quite important and must be considered in a klystron design. If the beam is uniform and is in a field-free drift space, there will be sufficient positive ions formed by molecular collisions, even in a good vacuum, to neutralize the space charge of the beam. This condition does not apply to a beam that is bunched, because the mobility of the heavy positive ions is quite low, and space-charge forces can exist. These forces oppose the formation of the electron bunches and can be resolved into two components. One component tends to destroy the focus of the beam and is termed "transverse" debunching. The other component repels the electrons that should become a part of the bunch and reduces the bunching action. This component is called "longitudinal" debunching.

Transverse debunching reduces the efficiency of a klystron because many of the electrons from the bunch become lost on the walls of the tube. There is also an excess of positive space charge in the region between bunches. This positive space charge may force electrons, which would normally be lost owing to divergence of the beam, back into the beam during the wrong part of the cycle when they would subtract additional energy from the output resonator. Therefore the effect of transverse debunching is not limited to the loss of electrons from the bunch itself.

Transverse debunching can be eliminated by using a strong longitudinal magnetic field to prevent divergence of the beam, but the use of a magnetic focusing field adds considerably to the complexity of a system and is seldom used.

Longitudinal debunching reduces the bunching action and therefore reduces the gain of a klystron. Within certain limits this effect may be overcome by increasing the drift distance. If the drift distance becomes too great, the debunching forces act during a longer interval, and it may be impossible to bunch the electron beam. For this reason, there is an optimum drift distance and a maximum gain which can be obtained with a klystron amplifier.

3.16. Bunching in a Short Drift Space.—All the curves in Figs. 3–2 to 3–9 illustrate bunching that is symmetrical; *i.e.*, these curves represent the conditions when the input gap voltage is small and there is sufficient drift time to permit bunching to occur. Under these conditions the current distribution is a symmetrical function of time. These conditions are satisfied if the drift time is an interval corresponding to 10 or more oscillation cycles. Drift times of this magnitude are obtained occasionally in tubes using low acceleration voltages; however, the usual conditions of operation correspond to drift times which are very much shorter.

It is apparent from Eq. (3–21) that a small number of oscillation cycles during the bunching interval require a large voltage E_1 at the input gap. This condition causes a distortion of the bunched current distribution. Figure 3–10 shows the distortion of the arrival-time curve corresponding to $x = 1.84$ when N has a value of $3\frac{1}{4}$ cycles. These conditions correspond to a ratio of $E_1/E_0 = 0.18$.

There are two reasons for the unsymmetrical current distribution when the drift time is short. A large ratio of input gap voltage to acceleration voltage is required to obtain bunching in a short time; as a result, the velocity distribution is not sinusoidal and the electrons that are decelerated are slowed down more than the electrons that are accelerated are speeded

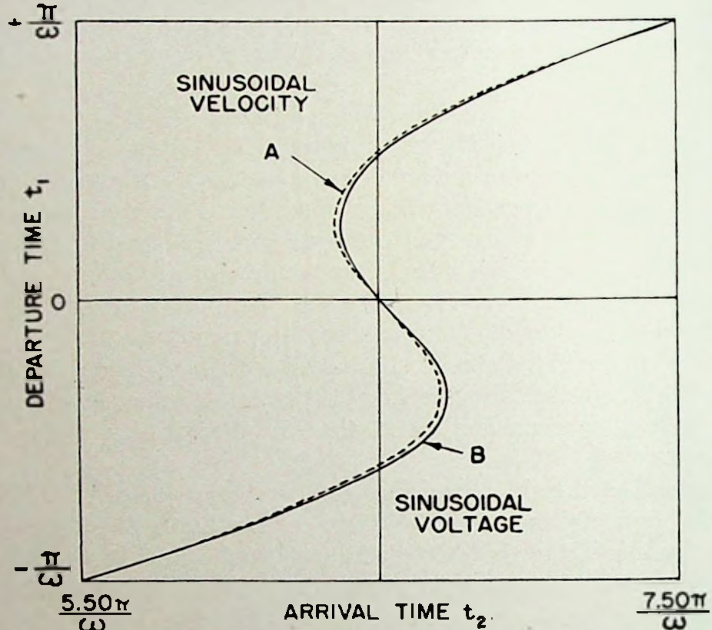


FIG. 3-10.—Electron arrival-time curves showing second-order effects when the velocity modulation is large but sinusoidal (A) and when the input gap voltage is sinusoidal (B).

up. In addition, the slowest and fastest electrons cannot be considered as traveling with the average velocity when the velocity variations are large. This factor causes the slower electrons to be comparatively overbunched while the faster electrons are somewhat underbunched. The nonsinusoidal velocity distribution and the distortion due to large variations of velocity produce an additive effect on the current distribution in the bunch.

The two effects are shown separately in Fig. 3-10. The dotted curve A has been computed for an assumed sinusoidal velocity variation. The difference in transit time between a slow electron and the average transit time is greater than the difference between the faster electrons and the average. As a result, the dotted curve is shifted slightly to the right in comparison to Fig. 3–2. The total distortion due to both factors is included in the solid curve B for a sinusoidal voltage variation. The additional distortion caused by the nonsinusoidal velocity variation when the bunched voltage is large compared to the acceleration voltage is shown by the difference between A and B in Fig. 3–10.

3.17. Phase Shift Caused by Distortion of the Bunch.—There is a phase shift introduced by the distortion of the bunched current. The electron that passes the input gap when the voltage is zero and changing

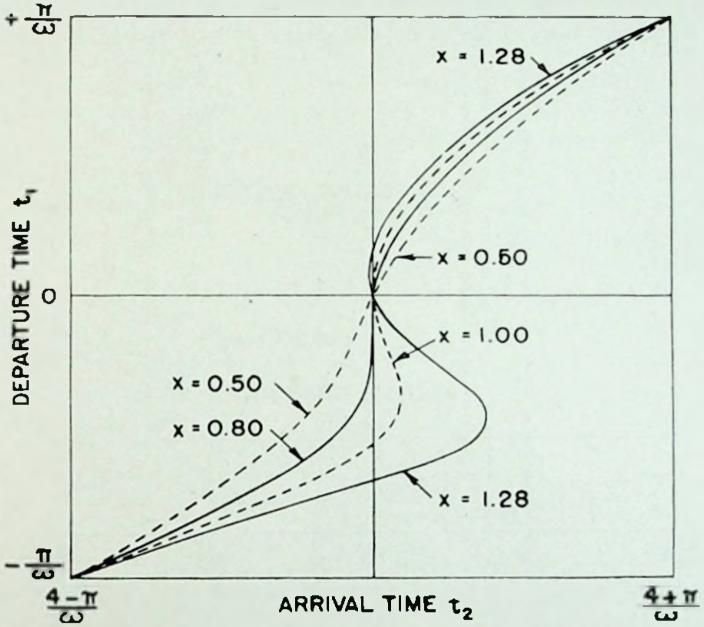


FIG. 3–11.—Electron arrival-time curve for an extremely short drift distance. The phase shift caused by the distortion of the electron bunch increases as the input gap voltage is increased.

from deceleration to acceleration does not become the center of the bunch. A drift time of $2/\pi$ cycles has been represented in Fig. 3–11 in order to emphasize this phase shift. Bunching parameter values of $x = 0.50$, $x = 0.80$, $x = 1.00$, and $x = 1.28$ have been used for computing the t_1 vs. t_2 curves in Fig. 3–11. Corresponding current-distribution curves are given in Fig. 3–12. Note that the current curve for $x = 0.50$ does not differ greatly from the corresponding curve in Fig. 3–3. The distortion becomes greater as the input gap voltage is increased; *i.e.*, the phase shift is a function of the input gap voltage.

The choice of a drift time of $2/\pi$ cycles means that the four values of the bunching parameter correspond to E_1/E_0 ratios of 0.25, 0.40, 0.50, and 0.64. The larger values of input gap voltage cause a nonsinusoidal variation of velocity; as a result, the average velocity is somewhat less than the velocity corresponding to the acceleration voltage, and the expression for the bunching parameter in Eq. (3–21) is no longer valid. The slower electrons become bunched sooner, and an infinite current peak occurs when the value of x computed from Eq. (3–21) is only 0.80. A value of $x = 1.00$ gives a decided double peak in the current distribution, and optimum bunching corresponds to a value of 1.28 instead of 1.84. This deviation from the first-order theory indicates that the output characteristic of an amplifier may differ from the curve shown in Fig. 3–7.

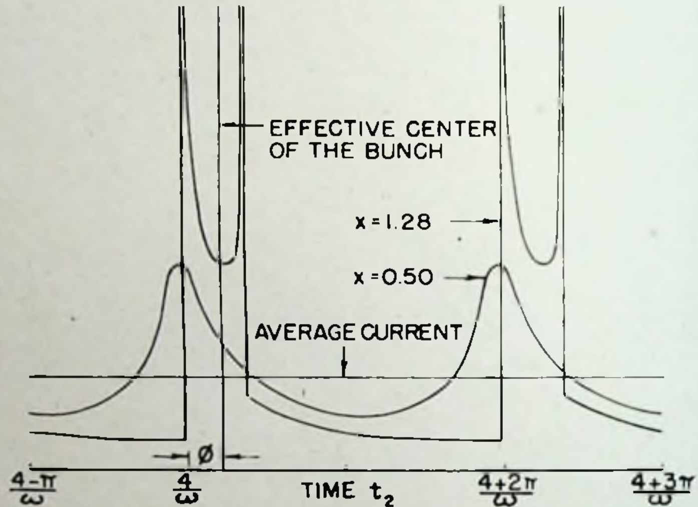


FIG. 3–12.—Bunched-current distribution corresponding to the short drift distance illustrated by Fig. 3–11. Two complete cycles are shown.

The previous discussion of phase shift caused by the distortion of the bunch has been qualitative. When the current distribution is not distorted badly, the simple assumption that the center of the bunch corresponds to the median time between the two infinite current peaks is quite satisfactory. Note that this assumption is not accurate for the current distribution in Fig. 3–12 when $x = 1.28$, and the curve is quite unsymmetrical. It is usually satisfactory to assume that the bunching is optimum when the two infinite current peaks have separated by a time interval equal to that predicted for $x = 1.84$ in the small-signal theory. The latter assumption was used in estimating the value of x required for optimum bunching in Fig. 3–11.

3.18. Power Transferred to Output Resonator.—An exact graphical analysis may be used to compute the energy transferred to the output resonator, and the phase shift and the value of bunching for optimum output may be obtained by cut-and-try methods if the assumptions above are unsatisfactory or if a quantitative answer is desired. The energy transferred from the beam to the output resonator will be given by

$$W = \frac{\omega}{2\pi} \int_0^{2\pi/\omega} I_2 E_2 \cos(\omega t_2 - \tau_0 - \phi)\, dt_2 \qquad (3\text{–}26)$$

E_2 is the peak value of the output gap voltage. The angle τ_0 is the transit angle of the electron that passed the buncher grids at a time $t = 0$. The designation of average transit angle is not applicable in this case, but the angle τ_0 still represents the transit time of an electron traveling with a velocity corresponding to the beam voltage. This definition means that $\omega t_2 - \tau_0$ is zero for the electron corresponding to $t_1 = 0$ and is equivalent to shifting the t_2 coordinate in the diagrams of the arrival time, or t_1 vs. t_2 diagrams. The phase angle ϕ is determined by the tuning of the output resonator and by the distortion of the bunched-beam current. This value of ϕ will be zero if the output resonator is tuned to resonance with the input frequency and there is no distortion of the bunch. If distortion of the bunch does occur, the phase will shift in order to transfer the maximum energy to the output resonator. Equation (3–10) may be substituted in Eq. (3–26) in order to avoid the difficulty of evaluating the energy contribution of the infinite current peaks. Since I_0 is a constant, Eq. (3–26) may be rewritten

$$W = \frac{\omega}{2\pi} I_0 \int_0^{2\pi/\omega} E_2 \cos(\omega t_2 - \tau_0 - \phi)\, dt_1 \qquad (3\text{–}27)$$

Convenient limits are chosen for the integration since it is necessary to integrate only over one complete cycle. It is necessary to replot E_2 cos $(\omega t_2 - \tau_0 - \phi)$ on the t_1 time scale before integration. This may be done conveniently by transferring the value of $E_2 \cos(\omega t_2 - \tau_0 - \phi)$ at a given time t_2 to the corresponding time t_1 given by the t_1 vs. t_2 diagram. This process is illustrated by Fig. 3–13, which corresponds to an output gap voltage with a peak amplitude of $0.7E_0$ replotted with the aid of line B of Fig. 3–10. A phase shift of 8 deg. has been assumed.

A value for the conversion efficiency of the tube can be obtained from the ratio of the effective energy transferred to the resonator, which is given by the difference between the shaded areas below and above the line BC representing the acceleration voltage, to the area of the rectangle $ABCD$ which represents the d-c energy in the beam. The value of efficiency obtained from Fig. 3–13 is 40 per cent. This is a theoretical efficiency and includes the power absorbed by ohmic losses in the resonator. The output efficiency would depend upon the ratio of power delivered to the load to the power losses in the resonator. If the value of E_2 had been chosen 1.0 instead of 0.7, the efficiency computed from this graphical integration would have been 40/0.7, or approximately 58 per cent, the theoretical efficiency of a klystron amplifier.

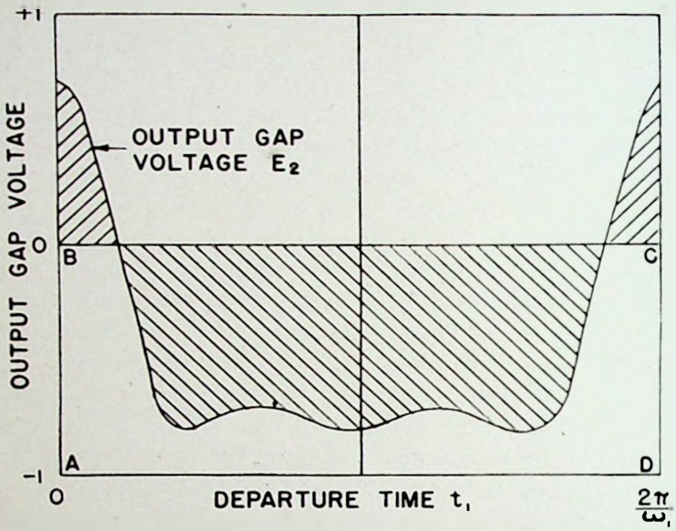


FIG. 3-13.—Energy diagram for computing the efficiency of a klystron from a graphical integration. The efficiency corresponds to the ratio of the difference of the two shaded sections to the area of the rectangle *ABCD*.

3.19. Phase Shift from a Graphical Integration.—It is not necessary to estimate the phase shift introduced by the distortion of the bunch, or obtain a value by cut-and-try methods from calculations of the efficiency using a number of arbitrarily chosen phase angles. Both the phase angle and the maximum efficiency for the given value of x may be computed exactly from two graphical integrations. If one integration is performed, using Eq. (3-27) with ϕ equal to zero,

$$W_1 = \frac{\omega}{2\pi} I_0 \int^{2\pi/\omega} E_2 \cos (\omega t_2 - \tau_0)\, dt_1 \tag{3-28}$$

and a second integration is made, with the voltage E_2 shifted 90 deg. in phase as indicated by the expression below:

$$W_2 = \frac{\omega}{2\pi} I_0 \int_0^{2\pi/\omega} E_2 \sin (\omega t_2 - \tau_0)\, dt_1 \tag{3-29}$$

then the phase angle ϕ and the maximum value of the energy transferred, W_{max} may be computed from the relations

$$\tan \phi = W_2/W_1 \tag{3-30}$$

and

$$W_{max} = \sqrt{W_1^2 + W_2^2} \tag{3-31}$$

The phase shift for various ratios of E_1/E_0, corresponding to the curves in Figs. 3–11 and 3–12, have been computed from Eq. (3–30) and are shown in Table II. These data verify the qualitative conclusion, stated in Sec. 3.15 in the discussion of Figs. 3–11 and 3–12, that the phase shift increases as the input gap voltage is increased.

TABLE II.—PHASE SHIFT INTRODUCED BY DISTORTION OF THE BUNCH

Bunching parameter	E_1/E_0	Drift time, cycles	Phase shift, degrees
0.50	0.25	$2/\pi$	6.4
0.80	0.40	$2/\pi$	12.0
1.00	0.50	$2/\pi$	18.3
1.28	0.64	$2/\pi$	42.0

There is no convenient graphical method for determining the conditions for optimum output, although the efficiency may be computed for a number of choices of input gap voltage E_1 or bunching parameter x, and the bunching parameter required for optimum output may be estimated by inspection of the curve of efficiency vs. bunching parameter.

CHAPTER 4

KLYSTRON AMPLIFIERS

4.1. Types of Klystron Amplifiers.—The usual distinction between types of amplifiers depends on whether the object is to develop the maximum voltage at the output or as much power as possible in the load impedance. The choice is achieved by the selection of the input- and output-circuit constants and the values of grid and plate voltages. A klystron amplifier has input and output circuits (resonators) which have approximately the same impedance. Therefore the power gain of a klystron amplifier is equal to the square of the voltage gain. Since there are no klystron amplifiers with "infinite" input impedance, there are no klystron voltage amplifiers in the usual sense of the definition.

It is more satisfactory to classify klystrons on the basis of input level. The input power will be proportional to the square of the voltage E_1 at the input gap; therefore, the input power is proportional to the square of the bunching parameter x in Eq. (3–21). The output power delivered to a constant impedance load will be proportional to the square of the radio-frequency component of current in the bunched electron beam, which is given by Eq. (3–20). This means that a theoretical curve of output power vs. drive power can be obtained by squaring the values of both coordinates in Fig. 3–7. A curve computed in this manner is shown in Fig. 4–1.

A low-level amplifier is operated in the underbunched region near zero drive power where the characteristic curve is linear and the output is proportional to the input. The gain of the amplifier is a maximum in this region. More output can be obtained by sacrificing gain and operating the amplifier at the point of optimum bunching. This point corresponds to the peak of the first maximum in the output curve in Fig. 4–1, and an amplifier operated at these conditions is termed a "high-level" amplifier. It is common practice to refer to high-level klystron amplifiers as "power amplifiers." Low-level amplifiers may be called "voltage amplifiers" because they are operated in the region of maximum gain, but this classification does not agree with the usual distinction between these two types of amplifiers.

4.2. Interaction between a Resonator and an Electron Beam.—Analogies between klystrons and conventional vacuum tubes are quite useful and will be used frequently in the following chapters. Some of these analogies are obvious extensions of well-known concepts. One of the most useful analogies, the equivalent circuit for a klystron which will be introduced in the next section, deserves a special explanation. The velocity modulation introduced at the input gap of a klystron amplifier is not difficult to understand; the electron is a charged particle and it will be accelerated or decelerated by the electric field in the input gap. This field exists because radio-frequency power is supplied to the resonator. The

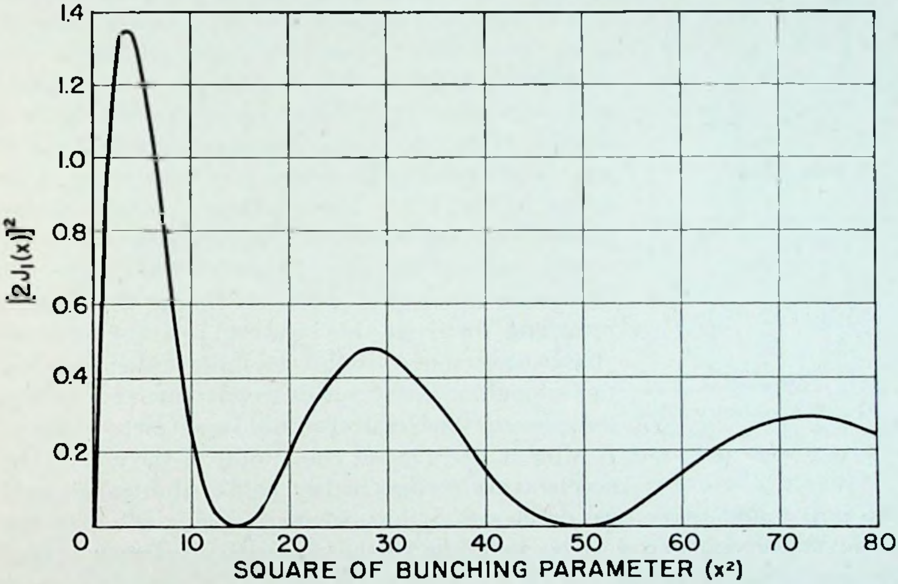


FIG. 4-1.—Theoretical curve of the output power vs. the square of the bunching parameter.

explanation of a transfer of energy from the bunched electron beam to the output resonator is not so obvious, and the problem is frequently avoided by assuming that a steady-state alternating field exists at the output gap and that the electron beam gives up some of its kinetic energy to maintain this field.

It is possible to show that an a-c component in the electron beam will build up the steady-state field by giving up energy to a resonator when the only voltage present is caused by thermal agitation. This effect is quite small, however, and a thermal voltage is not required for the explanation of the energy transfer. A bunched electron beam will build up the field in a resonator in the same way that an alternating current flowing in a resonant circuit will build up a voltage across the circuit.

Consider a resonator in which no radio-frequency field exists and assume that, initially, no beam current is passing the resonator gap. Then allow the beam current to flow for an interval that is very short compared to the time corresponding to one cycle of oscillation at the resonant frequency of the cavity. While this current pulse is in the resonator gap there will be a field in the gap, since a field always accompanies an electron charge. This field will react upon the moving electrons so that they leave the gap with a slightly reduced velocity. The energy lost by the electrons in the current pulse remains in the resonator as a residual field, and the build-up process has started.

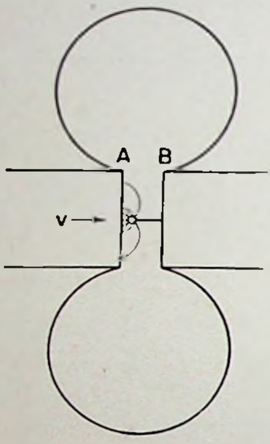


Fig. 4–2.—Electric field introduced by the presence of an electron in the resonator gap.

A satisfactory physical picture of this action can be obtained with the aid of a simple analogy. For simplicity, a single electron passing the resonator gap at the same time during each cycle will be considered. This electron enters the gap with a velocity v and travels across the gap from A to B, as shown in Fig. 4–2. There will be a positive charge induced on the walls of the resonator when the electron enters the gap. Most of the charge will be concentrated at A, and this charge will exert a retarding force on the electron. If the induced charge could move around the walls of the resonator fast enough, *i.e.*, if A and B were connected by zero impedance, the electron would be attracted toward B after it had passed the middle of the gap. The acceleration during the last half of the transit past the gap would cancel the deceleration that occurred during the first half of the transit and the electron would leave the gap with a velocity v equal to its entering velocity.

The induced charge at A cannot move rapidly to B because this motion of the charge corresponds to a current that sets up a magnetic field. As a result, the walls of the resonator connecting surfaces A and B will act like an inductance which prevents the induced charge from moving from A to B instantaneously. The electron will continue to be retarded by the positive charge remaining at A, and there will be less attraction toward the surface B. The velocity that it loses in the first part of its transit through the gap will not be regained and it will leave the gap with less velocity than v, its entering velocity. The electron has lost kinetic energy to the field it induced, and this energy remains in the resonator as a residual field corresponding to the induced charge at A.

A current will build up in the resonator walls as the charge at A tends to equalize on both sides of the capacity represented by the surfaces A and B. Since the walls of the resonator correspond to an inductance, this current will continue to flow after the charge has been equalized. One half cycle later, the charge will be concentrated at B. Then the current will reverse and the surface A will again be charged positively one cycle later when the next electron arrives. The process will be repeated and the field in the resonator will build up rapidly, since more energy is transferred by the electron when it is decelerated by a stronger field. The field in the resonator will continue to build up until the energy transferred during each cycle is just equal to the losses in the resonator and the energy absorbed by a load that may be coupled to it.

4.3. Equivalent Circuit for a Klystron Amplifier.—This explanation of the energy transfer to a resonator indicates that there is little difference between the case when the electron current is inductively coupled to the circuit and the well-known case of a current that is actually conducted by the circuit. The resonator and electron beam may therefore be represented by a conventional circuit, and an equivalent circuit for a klystron amplifier is shown in Fig. 4–3. A multiple grid tube is shown to emphasize

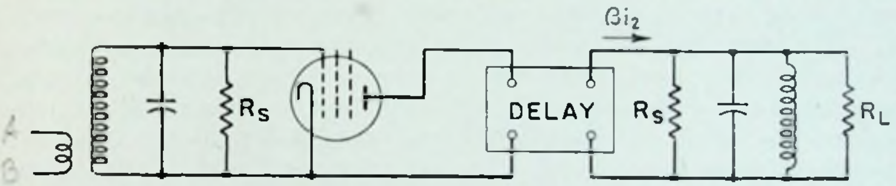


FIG. 4–3.—Equivalent circuit of a klystron amplifier.

the fact that only electron coupling exists in a velocity-modulated amplifier, and that the input and output circuits are completely isolated. The "delay" shown in the equivalent circuit is introduced by the transit time of the electrons in the drift space and is a function of the acceleration voltage and the length of the drift space. This delay does not affect the operation of a klystron as a simple amplifier but is important when a klystron amplifier is phase-modulated.

The input and output resonators are represented by identical RLC circuits. Inductive coupling to the input circuit is shown; similar coupling is used to obtain power from the output circuit, but the equivalent circuit has been simplified by showing the equivalent shunt load resistance R_L instead of the inductive coupling to the actual load resistance. The resistors R_s represent the shunt resistance of the cavity resonators and include beam loading effects as well as the resonator losses. The losses in the two resonators may not be equal in practice, but the assumption that they are equal does not introduce any serious errors.

4.4. Electron-bunching Relationships.—If there is a radio-frequency voltage of peak value E_1 at the input resonator, there will be a radio-frequency current βi_2 at the output as the result of the electron-bunching process. The relations among the input voltage E_1, to beam voltage E_0, the beam current I_0, the radio-frequency component of current i_2, and the number of oscillation cycles during transit of the drift space, designated by N, were derived in Chap. 3 and will be repeated here.

$$i_2 = 2I_0J_1(x)\sin\omega_1t_2 \tag{4-1}$$

$$x = \beta\pi NE_1/E_0 \tag{4-2}$$

Equation (4–1) is similar to Eq. (3–20) in Sec. 3.10 and Eq. (4–2) is identical to Eq. (3–21). The factor β is the beam-coupling coefficient described in Sec. 3.12 and is defined by

$$\beta = \frac{\sin(\delta/2)}{\delta/2} \tag{4-3}$$

where δ is the transit angle at the input gap.

4.5. Transconductance of a Klystron.—These bunching equations may be used to define a transconductance for a klystron. This transconductance is the ratio of the peak value of the effective output current βi_2 to the peak input voltage E_1. This definition of the transconductance is essentially the same as the definition for conventional vacuum tubes, since the peak values used in the definition for a klystron would be analogous to the increments used in the conventional definition. However, the bunching characteristic for a klystron given in Fig. 3–7 is plotted in terms of the peak values; therefore the transconductance is the slope of a line from the origin to the operating point on the curve, instead of the slope of the curve. This distinction is necessary because the bunching curve does not correspond to the i_p vs. E_g curve for a conventional vacuum tube.

It is also necessary to introduce a new tube constant, the "large-signal transconductance," since the usual definition is for a small-signal transconductance. Klystron power amplifiers operate in a nonlinear region of the curve where the transconductance is reduced from the small-signal value.

Equation (4–2) for the bunching parameter may be rewritten to give an expression for E_1:

$$E_1 = \frac{E_0x}{\beta\pi N} \tag{4-4}$$

The value of the transconductance g_m can then be determined from the ratio of the peak value of βi_2 to the peak voltage E_1.

$$g_m = \frac{2\beta I_0J_1(x)}{E_1} \tag{4-5}$$

Substituting Eq. (4–4) in Eq. (4–5) gives

$$g_m = \frac{\beta^2\pi NI_0}{E_0}\,\frac{2J_1(x)}{x} \tag{4-6}$$

For small values of x, $J_1(x)$ is equal to $x/2$ and the ratio $2J_1(x)/x$ is unity. Therefore the small-signal transconductance g_{ms} is given by

$$g_{ms} = \frac{\beta^2 \pi N I_0}{E_0} \tag{4-7}$$

4.6. Voltage Gain of an Amplifier.—The output voltage of the klystron amplifier will be the product of the effective current βi_2 and the loaded shunt resistance of the circuit, *i.e.*, the resistance of R_s and R_L in parallel. If the output circuit is not tuned to resonance with the input frequency, there will be a phase angle ϕ_2 between the current i_2 and the output voltage E_2, and the voltage will be reduced by a factor $\cos \phi_2$. This output voltage is given by

$$E_2 = 2\beta I_0 J_1(x) \frac{R_s R_L}{R_s + R_L} \cos \phi_2 \tag{4-8}$$

The phase angle ϕ is determined by the ratio of the loaded shunt resistance and the shunt reactive component X_s.

$$\tan \phi_2 = \frac{R_s R_L}{(R_s + R_L) X_s} \tag{4-9}$$

When the output circuit is at resonance with the input frequency, there is no reactive component; *i.e.*, X_s is infinite and the phase angle is zero.

Substituting Eq. (4–5) into (4–8) and dividing by E_1 gives an expression for the voltage gain:

$$\text{Voltage gain} = \frac{E_2}{E_1} = g_m \frac{R_s R_L}{R_s + R_L} \cos \phi_2 \tag{4-10}$$

Equation (4–10) is similar to the relation for the voltage gain of a grid-driven triode or pentode. This similarity is quite important and is added evidence of the usefulness of the analogy between klystrons and conventional vacuum tubes.

It should be pointed out that this discussion assumes that the output voltage E_2 is always less than the beam voltage E_0, and that electrons are not turned back or reflected by the field at the output gap. The importance of this assumption is discussed in Sec. 4.12, in the analysis of the effect of the beam current on the output load required for maximum output.

4.7. Power Output.—An equation for the power delivered to the load can be obtained from Eq. (4–8). The total power delivered by the bunched-beam current will be divided between the resonator losses represented by R_s and the equivalent load resistance R_L. Reference to the equivalent diagram in Fig. 4–3 shows that the peak radio-frequency voltage E_2 appears across the two resistances which are in parallel; therefore, the power delivered to the load will be one-half the square of the peak voltage

E_2, divided by the load resistance R_L. If P_L is used to represent the power delivered to the load,

$$P_L = \frac{E_2^2}{2R_L} = \left[2\beta I_0 J_1(x)\right]^2 \frac{R_s^2 R_L}{2(R_s + R_L)^2} \cos^2 \phi_2 \qquad (4\text{–}11)$$

It is apparent from Eq. (4–11) that the power output is proportional to the square of i_2, the radio-frequency current component of the bunched-beam current. This result is also obvious from the usual relations between

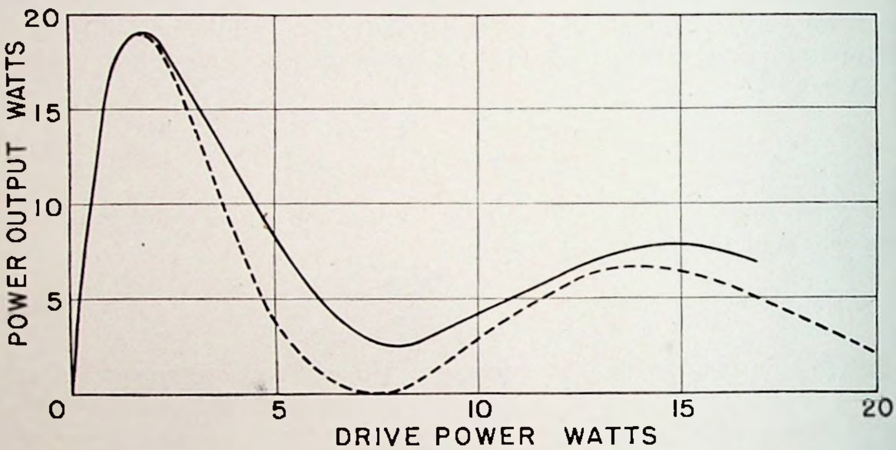


FIG. 4–4.—Power output of a klystron amplifier as a function of the drive power. The theoretical curve from Fig. 4–1 is shown as a dotted line for comparison.

current, voltage, and power, and justifies the reasoning used in Sec. 4.1 to explain the theoretical curve of power output in Fig. 4–1. The theoretical output curve is repeated as a dotted line in Fig. 4–4 for comparison with a typical curve for a Type 410R klystron. The theoretical curve has been matched in amplitude and position with the first peak of the experimental curve.

The actual amplifier characteristic differs from the theoretical prediction in two distinctive details. Perhaps the most striking difference is the fact that the power output does not become zero when the input power is sufficient to overbunch the electron beam. This difference cannot be attributed to experimental error, since the existence of a finite power output instead of zero output is unmistakable. There is also a discrepancy in the position of the maximum and minimum power points when the electron beam is overbunched. All these discrepancies may be explained by the fact that the input gap voltage is not a small fraction of the beam voltage in a practical klystron amplifier. Probably space-charge debunching effects are also present, and other second-order effects, which are not included in the simple theory, contribute to the difference between the actual and the theoretical curves.

Another output characteristic which indicates the effect of overbunching is shown in Fig. 4–5. In this case the drive power is maintained constant but the beam voltage of the klystron is varied. As the beam voltage E_0 is decreased, the number of oscillation cycles during the transit through the drift space will increase owing to the reduced velocity of the electrons

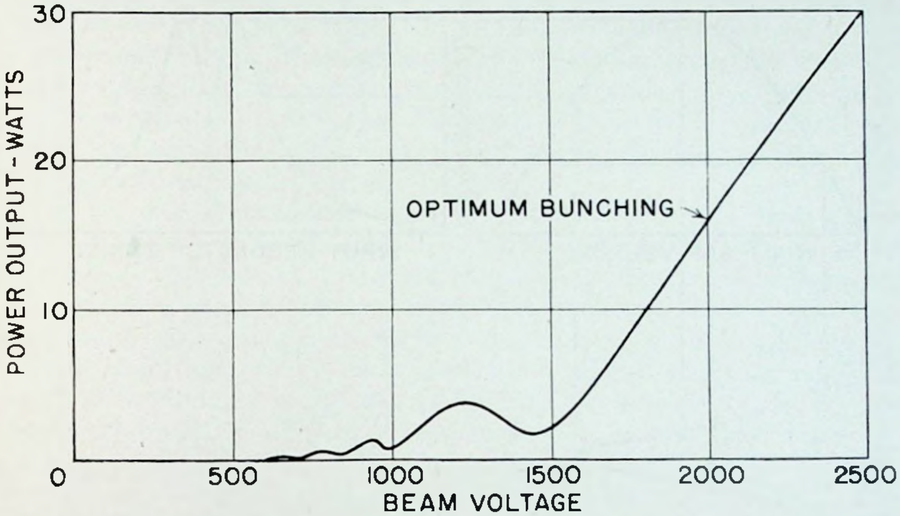


FIG. 4–5.—Power output of a klystron amplifier as a function of the beam voltage. The drive power is maintained constant.

in the beam. As a result, the value of the bunching parameter x given by Eq. (3–21) will increase as E_0 is decreased, provided the gap transit time is small and the coupling coefficient β does not decrease very rapidly as the electron velocity is reduced. This means that the output from a klystron amplifier may have several maxima and minima at low beam voltages, as indicated in Fig. 4–5. The output of these overbunched maxima will not be large, since the reduction of beam voltage reduces the power input to the amplifier considerably.

4.8. Overbunching.—It is not necessary to change the source of drive power or the beam voltage on the klystron amplifier in order to observe this overbunched characteristic. If sufficient input power is available for overbunching the electron beam when the input cavity is tuned to resonance, the drive power may be reduced by detuning the input resonator. When the tuning has been changed until the voltage at the input gap is reduced to the correct voltage for optimum bunching, the output of the amplifier will be a maximum. The input tuning may be changed to either side of resonance to reduce the input gap voltage, and the output curve will have two peaks as the input tuning is varied.

This behavior can be illustrated by a diagram similar to the construction frequently used to explain the operation of a Class A triode amplifier, using an i_p vs. E_g curve to transfer points from a sine curve representing input voltage to a similar sine curve showing the plate current. Figure

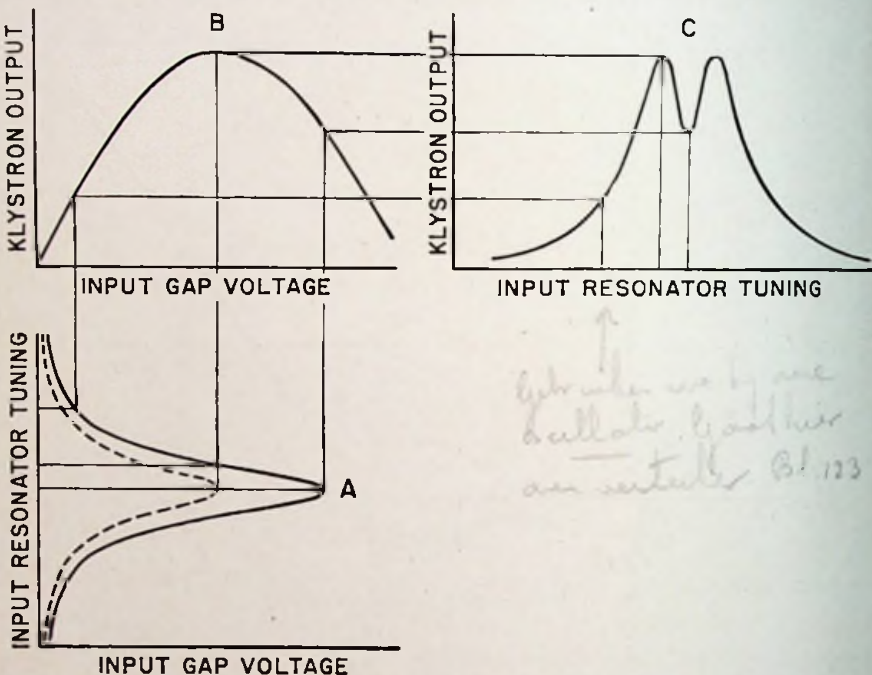


Fig. 4–6.—The effect of changing the input tuning of an overbunched klystron amplifier is related to the klystron bunching characteristic.

4–6 is a variation of this construction showing the relation between the input gap voltage and the tuning of the input resonator of the klystron, instead of the sine curve of voltage as a function of time. A bunching characteristic similar to Fig. 3–7, showing output voltage or current as a function of input gap voltage, replaces the triode static characteristic. The curve that results when points are transferred from curve A to curve C shows the typical double-peaked output characteristic of an overbunched klystron amplifier when the input tuning is changed. This double peak in the output of an overbunched klystron should not be confused with the double peak in the bunched-beam current curves in Figs. 3–3, 3–5, and 3–6, which are functions of time, but is a static characteristic and is related to the shape of the output characteristic curves in Figs. 3–7, 4–1, and 4–4.

Referring to Fig. 4–6, the construction of curve C may be explained as follows: When the input circuit is tuned far on one side of resonance, the input gap voltage is quite small and the electron beam is underbunched. As a result, the output gap voltage is also small. Tuning the resonator in the direction of resonance will increase the voltage at the input gap in accordance with the familiar resonance curve shown in Fig. 4–6A. The output gap voltage will also increase until the point representing optimum bunching in curve B is reached. Since the input gap voltage is still less than the maximum value when the circuit is resonant, additional tuning will decrease the output voltage and give a minimum in curve C when the circuit is tuned to resonance. Tuning the circuit to the other side of resonance will again give the input gap voltage required for optimum output and the curve will be symmetrical with a dip at the point corresponding to resonance of the input circuit.

It is not desirable to operate a klystron amplifier with the input circuit detuned to prevent overbunching. This adjustment not only would be wasteful of drive power but does not represent a stable tuning adjustment. The operating point for optimum output is on the steep side of the resonance curve where a small tuning change can produce a serious change in the output. Reducing the drive power, until the adjustment for resonance would give optimum output, would improve the over-all stability of the amplifier because the operating point would then be in a region where the output resonator current is fairly constant. The resonance curve for this reduced value of drive power is shown dotted below curve A in Fig. 4–6. Additional suggestions for the utilization of the overbunching characteristic in obtaining the proper adjustment of klystron amplifiers are given in the discussion of operating suggestions in Chap. 12.

4.9. Output Resonator Tuning.—The tuning of the output circuit does not affect the bunching of the beam at the output gap. It is true that the voltage at the output gap modifies the electron velocities but, if the gap transit time is small, the radio-frequency current at the output gap is determined entirely by the drift time and the input gap voltage. Therefore the variation of output as the second resonator is tuned will follow the usual resonance curve. The degree of bunching merely determines the maximum output that can be obtained when the output circuit is tuned to resonance.

4.10. Output Coupling.—An equivalent parallel load resistance R_L has been used in Fig. 4–3 and the analysis in Secs. 4.6 and 4.7. This simplification implies that the load connected to the coaxial output line is transformed into a pure resistance as seen by the output resonator. This relation could be true, but in general the load will not be a pure resistance but will have a reactive component; also, the coaxial line and coupling loop would not transform a purely resistive load into an equivalent resistance unless the electrical length of the output line was an integral number of quarter wavelengths.

A complete circuit for the output resonator and load is shown in Fig. 4–7. The output resonator is coupled to a coaxial output line by a coupling loop. These elements are represented by the *RLC* circuit which is inductively coupled to a short length of coaxial line. An impedance transformer is shown between the output line of the klystron and another coaxial line which is terminated by a load impedance Z_{load}. It is quite

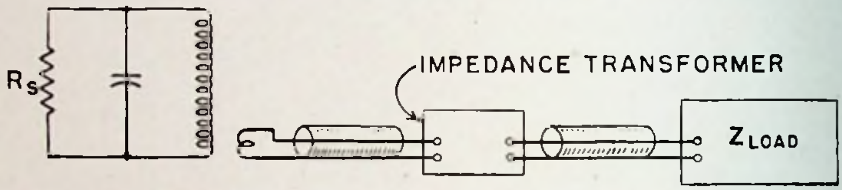


FIG. 4–7.—Circuit of the output resonator and a load.

important to consider all radio-frequency connections as transmission lines at these microwave frequencies. The impedance transformer may be a simple variable length of line, a parallel shorted line of variable length, or a combination of two or more of such elements.

The net result of the complicated array of lines, load, and coupling loop is an equivalent load impedance in parallel with the *RLC* circuit representing the cavity resonator. The equivalent circuit is shown in Fig. 4–8. The only addition to Fig. 4–3 is an equivalent load reactance X_L, which may be either inductive or capacitive, of course, and may be infinite if the load is purely resistive.

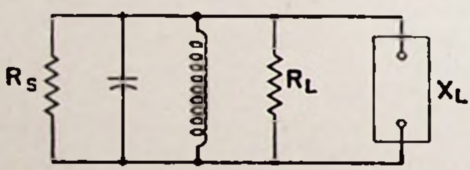


FIG. 4–8.—Equivalent circuit corresponding to Fig. 4–7.

4.11. Size of Coupling Loop.—Although the size of the coupling loop is important, it is not the only factor in the transformation ratio between the load and the resonator. Smaller coupling represents a lighter load, *i.e.*, a higher value of parallel resistance. The very high values of Q and shunt resistance of a cavity resonator require large impedance ratios when coupled to coaxial lines, which are ordinarily designed to have a characteristic impedance in the range of 50 to 100 ohms. As a result, the size of the coupling loops is usually quite small.

If the equivalent load resistance R_L is to be quite small compared to the shunt resistance of the cavity, a larger value of coupling will be required. It may not be possible to increase the coupling sufficiently by increasing the size of the loop. As the loop is made larger and its dimensions approach those of a wavelength, its self-inductance may increase so rapidly that increasing the size of the loop does not increase the effective coupling. Under these circumstances, special precautions to reduce the self-inductance of the coupling loop are necessary.

4.12. Effect of Varying the Load.—A klystron amplifier may be considered a constant-current source. Referring to the equivalent circuit in Fig. 4–3, it is apparent that the usual circuit relations apply and that the maximum power output will occur when the equivalent load resistance

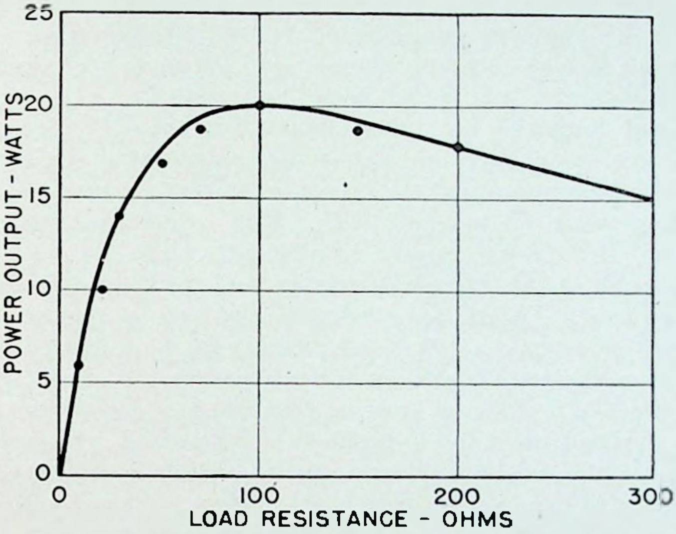


FIG. 4–9.—Power output of a klystron amplifier as a function of the load resistance.

R_L is equal to the cavity shunt resistance R_s. This relation may also be derived by differentiating Eq. (4–11) with respect to R_L. The actual load resistance is much smaller than the value of the equivalent load resistance R_L. The correct value of load resistance will depend upon the amount of coupling to the resonator, as discussed in Sec. 4.11. When this load resistance is connected to the coaxial output terminal, maximum power transfer to the load will occur.

As an example, a load of 100 ohms connected to the output line may correspond to a value of equivalent resistance R_L, equal to the cavity shunt resistance R_s. Maximum power output will be transferred to this load resistance of 100 ohms. A lower value of load resistance will decrease the effective shunt resistance of the cavity and the resonator will not absorb so much power from the beam. A greater percentage of the power absorbed from the electron beam will be delivered to the load as the load resistance is decreased, but the power transferred by the beam to the output resonator becomes zero when the loaded shunt resistance of the resonator is zero.

When the load resistance is increased, the loaded shunt resistance of the resonator can never become greater than the shunt resistance representing the cavity losses and, since the percentage of power delivered to the load decreases when the load resistance is increased, the power output drops off for high values of load resistance.

The behavior of a klystron amplifier is illustrated by the theoretical curve in Fig. 4–9 for a constant-current generator with losses corresponding to a shunt resistance of 100 ohms. The peak value of this theoretical curve has been matched to the peak power output obtainable from a klystron amplifier which transfers maximum power to a 100-ohm load. Values of power output for other load resistances are indicated by the experimental points that lie fairly close to the theoretical curve.

4.13. Load Required for Large Beam Current.—The previous discussion of load matching, and most of the analyses that appear later, are based on the assumption that the voltage at the output gap is less than the beam voltage when R_L is equal to R_s. This assumption was mentioned in Sec. 4.6. If the beam current is sufficiently large, the output voltage E_2 indicated by Eq. (4–8) may be greater than the beam voltage E_0 when R_L is equal to R_s. When this condition exists, the simple analysis is incorrect, and more output will be obtained if R_L is decreased until E_2 is equal to E_0 and electrons are not turned back at the output gap. It is therefore possible to design a klystron that will not transfer the maximum power to the load when the impedances are matched. In fact, in order to obtain the theoretical maximum efficiency from a klystron amplifier it is necessary to have a very large beam current so that R_L is very much less than R_s and most of the power is delivered to the load.

4.14. Efficiency of a Klystron Amplifier.—For maximum efficiency, the output circuit would always be tuned to resonance and $\cos^2 \phi_2$ in Eq. (4–11) would be unity. The power delivered to the load can be written

$$P_L = E_2 \beta I_0 J_1(x) \frac{R_s}{R_s + R_L} \tag{4–12}$$

and the efficiency is given by

$$\text{Efficiency} = \frac{P_L}{E_0 I_0} = \beta \frac{E_2}{E_0} J_1(x) \frac{R_s}{R_s + R_L} \tag{4–13}$$

If $R_s \gg R_L$ and βE_2 is equal to E_0, the efficiency is equal to the maximum value of the J_1 Bessel function. Therefore the theoretical efficiency of a klystron amplifier is 58 per cent. When the beam current is small and R_L must be equal to R_s to obtain maximum output, half the power is dissipated in the resonator losses and the efficiency can only be 29 per cent. For still smaller beam currents, E_2 will be less than the beam voltage E_0 and the efficiency will be reduced further by the ratio E_2/E_0. All these figures for the efficiency must be decreased if the focusing of the beam is poor or the beam is partly intercepted, and the emission current from the cathode is greater than the actual current in the beam at the output resonator.

Several factors make the realization of the ideal efficiency rather difficult. The current density that can be obtained from most cathode surfaces may not be sufficient to allow the klystron to be loaded heavily enough to transfer most of the power to the load. This difficulty is primarily an electron-gun design problem.

Space-charge effects also limit the useful increase of beam current. Beam loading, mentioned in Sec. 3.14, reduces the shunt resistance of the resonators and reduces not only the gain of a klystron amplifier but also the efficiency. Heavier loading is required, and the output voltage is therefore less. Debunching was discussed in Sec. 3.15 and limits the maximum current that can be used because an increase in the beam current may actually reduce the radio-frequency component of current.

Most klystron designs do not have sufficient beam current to require heavy loading to prevent turning back electrons at the output gap. This means that maximum output is usually obtained when the equivalent shunt load resistance is equal to the shunt resistance of the resonator. Efficiencies from 6 to 15 per cent are typical for klystron amplifiers operating at a frequency of 3,000 megacycles. Higher efficiencies have sometimes been obtained. Higher efficiencies can be obtained in the lower frequency designs, and it is difficult to obtain good efficiencies as the frequency range becomes higher. These figures do not represent an inherent limitation in klystron design, and future tubes may not be limited in this manner.

4.15. Rieke Diagrams.—In practice, the output load is not always purely resistive but may have a reactive component. The effect of this reactive component is equivalent to detuning the output resonator and therefore reduces the power output. In order to illustrate this effect, it is convenient to use the complex plane to represent the resistive and reactive components of the load. An impedance $R + JX$ is represented by a point with a horizontal coordinate determined by the resistance R, and the vertical or imaginary coordinate is determined by the reactance X. Any point on this plane will correspond to some value of load impedance, and this load will give a certain output. Points of equal output can be connected by a line which is a contour of constant power output. This type of representation is known as a Rieke[1] diagram.

[1] A polar diagram was used originally by F. F. Rieke at the Radiation Laboratory, Massachusetts Institute of Technology, to describe the load characteristics of magnetron oscillators. However, the term has been adopted for other microwave tubes and is used for a load diagram showing contours of constant power or constant frequency on any convenient impedance plane.

Either impedance or its reciprocal, admittance, may be represented by the coordinates of a Rieke diagram. Shunt circuits as used in the equivalent circuit of the output resonator (Fig. 4–8) are best expressed by admittances because parallel admittances are added directly. Therefore the Rieke diagram in Fig. 4–10, which shows the power output characteristic of a klystron amplifier as the load is varied, has conductance as the coordinate of the horizontal real axis, and susceptance as the vertical or imaginary coordinate. A large load resistance is equivalent to a light

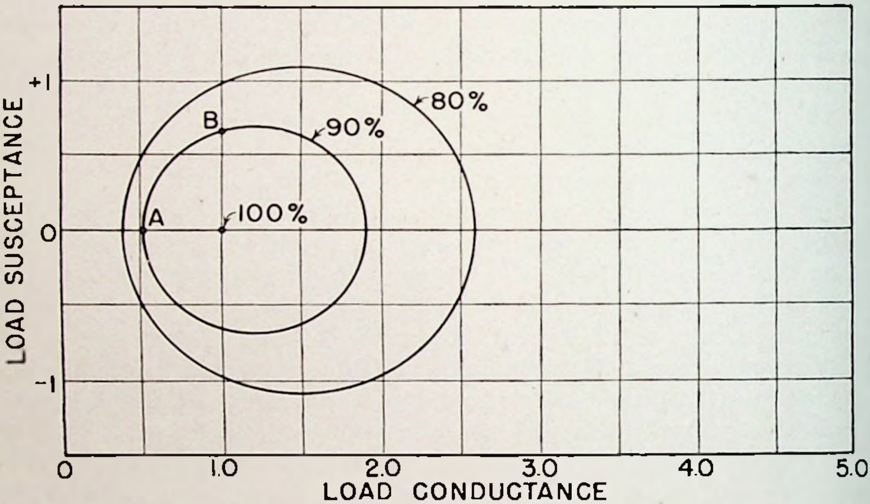


FIG. 4–10.—Theoretical Rieke diagram showing the power output of a klystron amplifier as a function of the load admittance.

load. Expressed as a conductance, it has a small value; *i.e.*, zero load corresponds to zero conductance. Similarly, an infinite reactance corresponds to the absence of a reactive component in a parallel circuit and is equivalent to zero susceptance. Normalized coordinates are used; *i.e.*, the conductance representing the load giving maximum output has been labeled unity.

In order to correlate the power output and the coordinates of the Rieke diagram, Eq. (4–11) will be rewritten in terms of the conductances:

$$P_L = \left[2\beta I_0 J_1(x)\right]^2 \frac{G_L}{2(G_s + G_L)^2} \cos^2 \phi_2 \qquad (4\text{–}14)$$

The load conductance as represented by G_L and G_s is the shunt conductance of the output resonator. The use of normalized coordinates simplifies the analysis because G_s is considered unity and G_L has a unit value when the load conductance is the correct value for maximum output. This choice of coordinates eliminates the transformation ratio of the coupling loop from the analysis, but means that the assumption of an output gap voltage smaller than the beam voltage is necessary. This assumption is true for most klystron designs.

The phase angle ϕ_2 can also be defined in terms of the susceptances and conductances of the circuit and load. The tangent of the phase angle is the ratio of the sum of the susceptances to the sum of the conductances.

$$\tan \phi_2 = \frac{B_s + B_L}{G_s + G_L} \tag{4-15}$$

The cavity susceptance B_s is zero when the output circuit is tuned to resonance. The following discussion will assume that B_s is zero and that the load susceptance B_L is the only component and is equal to the imaginary coordinate of the Rieke diagram in Fig. 4-10.

If the klystron amplifier is lightly loaded, for example, by a pure conductance with a value of 0.52, the phase angle ϕ_2 in Eq. (4-14) will be zero. The power output obtained by substituting these values in Eq. (4-14) will be 90 per cent of the maximum value when G_L is unity. This point on the 90 per cent contour on the Rieke diagram is indicated by A in Fig. 4-10. When the load has a reactive component, for example, at the point $1.00 + 0.71j$, the value of $\cos^2 \phi_2$ becomes 0.9 and locates point B on the 90 per cent contour. The locus of this contour is a circle, but the proof of this relation will not be given. A circle contour of 80 per cent power output has also been indicated on the Rieke diagram.

4.16. Rieke Diagram When Resonator Is Retuned.—Reference to Eq. (4-15) indicates that the phase angle ϕ_2 may be maintained zero by detuning the resonator to introduce a susceptance B_s which may be opposite in sign and cancel the load susceptance B_L. If the output resonator is retuned for maximum output each time the load is changed while taking data for a Rieke diagram of a klystron amplifier, the contours will become vertical lines of constant conductance. A theoretical Rieke diagram for this case is illustrated in Fig. 4-11. In actual practice, the contours are not straight lines and do not extend to infinity, because losses in the output line and the coupling loop limit the power output. In this case the contours may close together when the load susceptance becomes quite large.

4.17. Typical Rieke Diagram for a Klystron Amplifier.—The admittance of a load is ordinarily measured with a standing wave detector, which measures the standing wave ratio and the position of the minimum or maximum voltage along a transmission line. These coordinates are superimposed on an admittance diagram in Fig. 4-12. Circles representing constant-voltage standing wave ratio (often abbreviated VSWR) have their centers along the real axis. The orthogonal family of circles indicates the position of the minimum voltage in the transmission line, measured in fractions of a wavelength from the load.

A load admittance that gives maximum output may not correspond to a unity standing wave ratio. This may be true because the klystron may not have been designed for operating into the characteristic impedance of the line used to measure the standing wave ratio. In addition, the load required for optimum output may vary somewhat as the power input to the klystron is changed. As a result, the contours of constant power output may be displaced with respect to the new unit coordinates which are based on the characteristic impedance of the line used as a standing wave detector.

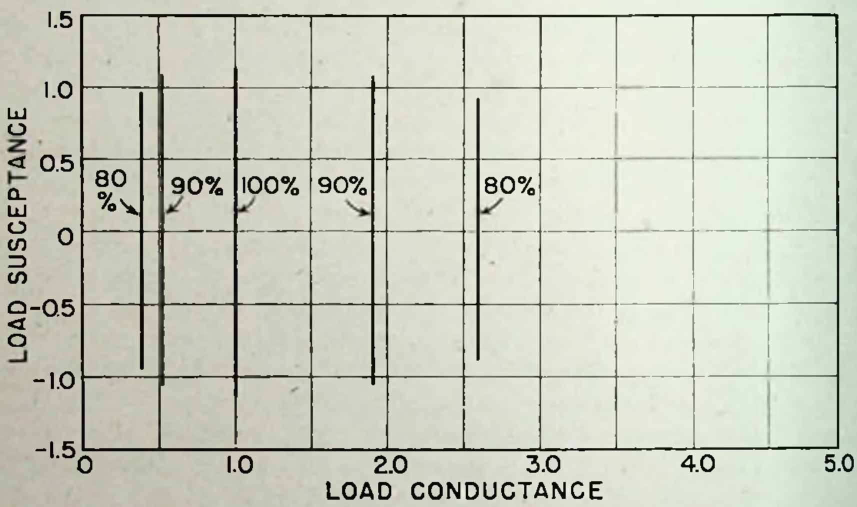


FIG. 4–11.—Theoretical Rieke diagram showing the power output of a klystron amplifier as a function of the load admittance when the output resonator is retuned to give maximum output for each adjustment of the load.

Actual power output in watts for a Type 410R klystron amplifier is indicated by the Rieke diagram in Fig. 4–12. A comparison with Fig. 4–10 emphasizes the similarity between a theoretical Rieke diagram and a typical operating characteristic. In Fig. 4–12, unity standing wave ratio and unit admittance correspond to the admittance of a transmission line with a characteristic impedance of 75 ohms, since a line with this impedance was used in taking the data.

4.18. Low-level Klystron Amplifiers.—Although the previous analysis applies equally well to power amplifiers and low-level amplifiers, it is primarily of interest in the application of power amplifiers. The problem in a low-level amplifier is not obtaining efficiency, or even maximizing gain, but it is important that an amplifier that is to be used as a radio-frequency amplifier in a receiver have a satisfactory signal-to-noise ratio. A brief discussion of the factors involved will explain why velocity-modulation tubes are usually considered unsatisfactory as receiver amplifier tubes.

In a klystron amplifier, the voltage appearing across the input gap depends upon the shunt resistance of the input resonator, the signal power, and an equivalent noise power which is introduced by random fluctuations of the beam current. These current fluctuations are known as "shot noise," and this noise level may be many times greater than the level of

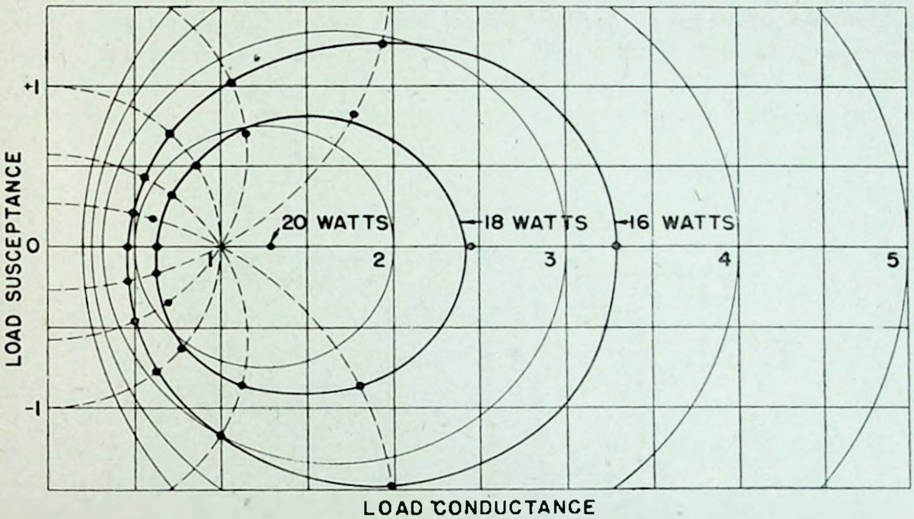


FIG. 4-12.—Actual Ricke diagram for a Type 3K30/410R klystron amplifier. The standing wave ratio caused by the load admittance is superimposed on the diagram.

"thermal noise." The current fluctuations introduce a noise voltage at the input gap, and the velocity modulation superimposed on the beam includes this noise component. As a result, the noise is amplified by the klystron and appears in the output in the same ratio as the signal and noise voltages at the input gap. This means that the ratio of the signal to the shot noise at the input resonator is a limiting factor in the use of the tube as a low-level amplifier.

There are some applications in which a large signal-to-noise ratio might be tolerated if the advantages of a radio-frequency amplifier are desired. In general, however, the noise figure of a klystron amplifier will reduce the sensitivity of a receiver, and a crystal rectifier used as a superheterodyne mixer without radio-frequency amplification is more satisfactory. Present klystron types have a noise level which is approximately 20 db higher than thermal noise.

CHAPTER 5

KLYSTRON FREQUENCY MULTIPLIERS

5.1. Frequency Multiplication.—As the frequency of oscillation of a source of radio power becomes higher, the problem of maintaining the frequency stability is increased. The use of a stable oscillator at a low frequency, followed by a series of Class C stages operating at harmonics of the oscillator frequency, is a familar solution to the stability problem

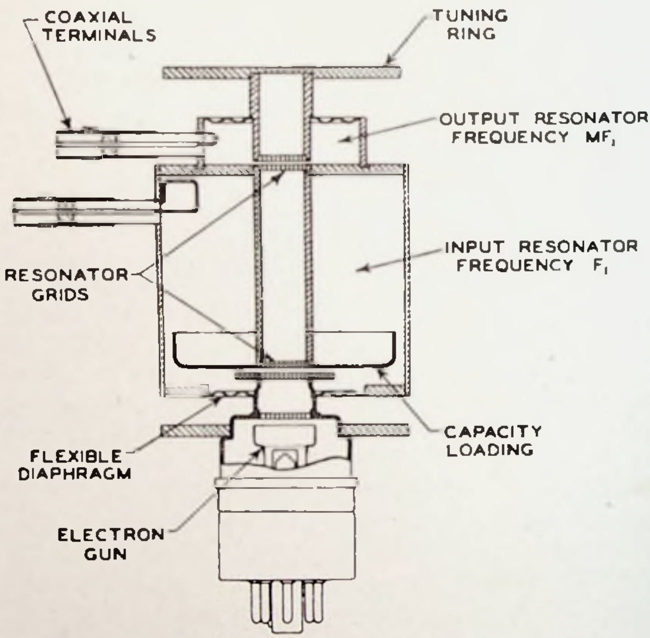


FIG. 5-1.—Sectional view of a klystron frequency multiplier.

in the very high-frequency region. Klystron tubes may be used in a similar manner and offer one method[1] of obtaining good frequency stability in the microwave region above 1,000 megacycles. A klystron frequency multiplier is merely a modified version of a klystron amplifier in which the output frequency is a harmonic or multiple of the input frequency. An illustration

[1] Automatic frequency control of microwave oscillators is also used to improve the frequency stability, but a description of the circuits involved has not been included in this text, which is primarily concerned with klystron characteristics.

of a klystron frequency multiplier is given in Fig. 5–1. This type of klystron has an output resonator similar to the resonators in the amplifier shown in Fig. 3–1, but the input resonator is much larger and tunes to a lower frequency.

5.2. Harmonics in a Bunched Electron Beam.—The electron beam in a klystron frequency multiplier is velocity modulated at a frequency f_1 by the field in the input resonator. The electron bunching action occurs at this same frequency f_1, but the harmonic content of the bunched-beam current is quite high and considerable output can be obtained from the second resonator when it is tuned to some multiple Mf_1 of the input frequency.

A first-order analysis of a klystron frequency multiplier can be obtained from the relation for the bunched-electron-beam current derived in Sec. 3.10. Only the first alternating component was used in the analysis for a klystron amplifier, but the other components are important in a frequency multiplier. The expression for the bunched-beam current given by Eq. (3–19) is repeated here in a modified form in order to show the harmonic content in the electron beam.

$$I_2 = I_0 [1 + 2J_1(x) \sin (\omega_1 t_2 - 2\pi N_1) + 2J_2(2x) \sin 2(\omega_1 t_2 - 2\pi N_1) + \cdots + 2J_M(Mx) \sin M(\omega_1 t_2 - 2\pi N_1)] \tag{5–1}$$

The theoretical harmonic content is given by the maximum values of the Bessel functions $J_M(Mx)$. These values and the ratios of the maximum of the harmonic to the maximum value of the fundamental are given in Table III, which shows that the theoretical ratio of the harmonic to the fundamental component of current does not decrease very rapidly as the multiplication factor is increased. In practice, the harmonic efficiencies indicated by the Bessel function maxima are not obtained. Space-charge debunching forces that oppose the formation of an electron bunch affect

TABLE III

Harmonic	Bessel function, maxima	Ratio of harmonic to fundamental
1	0.58	1.00
2	0.48	0.83
3	0.43	0.75
5	0.37	0.64
10	0.30	0.52
15	0.27	0.47
20	0.24	0.42

the higher harmonics more seriously than the fundamental component and the lower harmonics. However, klystrons using frequency multiplication ratios as high as twenty do give useful power output. Klystrons with smaller multiplication ratios are much more efficient, and tubes with a multiplication ratio of only two or three may have efficiencies that approach the efficiency obtained with a klystron amplifier.

5.3. Suppression of Adjacent Harmonics.—The presence of a large number of harmonics of almost equal intensity in the bunched electron beam might lead one to expect the output of a klystron frequency multiplier to contain all these frequencies simultaneously. However, this is not the case because the adjacent harmonics are suppressed by the high Q of the output resonator. Any simple resonant circuit has a band width that is inversely proportional to the Q of the circuit. The response to an adjacent harmonic will be small if the frequency of the harmonic differs from the resonant frequency of the cavity by an amount that is much greater than the band width of the cavity. This response may be evaluated from a formula giving the ratio of the impedance of the resonator at a frequency far from resonance to the impedance at resonance. A modification of a relation derived by Terman[1] is given below.

$$\frac{Z_{M+1}}{Z_M} = \frac{1}{Q\left(\frac{M+1}{M} - \frac{M}{M+1}\right)} \tag{5-2}$$

Equation (5–2) states that the adjacent harmonic $M + 1$ is supressed by a factor inversely proportional to Q and the factor $\left(\frac{M+1}{M} - \frac{M}{M+1}\right)$ when the resonator is tuned to the frequency of the Mth harmonic of the input frequency of the klystron. As an example, consider a klystron with a multiplication factor of ten. The eleventh harmonic will be suppressed by a factor of $1/0.2Q$. A typical value of Q for a resonator would be 1,000 and only one-half of 1 per cent of the eleventh harmonic would be present in the output of a klystron frequency multiplier tuned to the tenth harmonic of the input frequency. This degree of harmonic suppression would be satisfactory for most purposes.

5.4. Bunching Characteristic of a Multiplier.—A klystron frequency multiplier has a bunching characteristic which is similar in many respects to the characteristic for a klystron amplifier. In fact, the amplifier characteristic illustrated in Fig. 3–7 and discussed in Sec. 3.11 is a special case of the more general relationship that includes frequency multipliers. Bunching characteristics for multiplication ratios of 3 and 10 are shown

[1] F. E. Terman, *Radio Engineers' Handbook*, p. 144, McGraw-Hill Book Company, Inc., New York, 1943.

in Fig. 5–2, and an amplifier characteristic has been included for comparison. The vertical coordinate of Fig. 5–2 represents the Mth harmonic component divided by the beam current I_0, and the values of the points on the curves are, therefore, given by $2J_M(Mx)$. The bunching parameter x is the horizontal coordinate. Its value is given by the familiar relation

$$x = \beta_1 \pi N_1 \frac{E_1}{E_0} \tag{5-3}$$

The subscripts have been added to β_1 and N_1 to denote that the transit times refer to the input frequency of the multiplier.

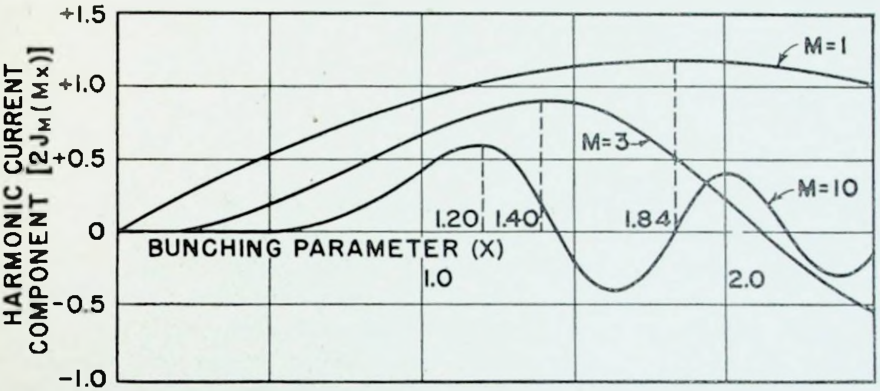


FIG. 5–2.—Bunching characteristics for a constant input frequency and several multiplication ratios.

Optimum bunching for a multiplier requires a smaller value of x than is required for an amplifier with the same input frequency. This result is evident from an inspection of Fig. 5–2. As a result, a frequency multiplier with the same input frequency and drift distance as an amplifier requires less drive power for the same operating voltage. The wider, double-peaked electron bunch which gives optimum output for an amplifier is not satisfactory for a multiplier because the current peaks may be out of phase at the harmonic frequency. As the multiplication becomes greater, the current distribution required approaches closely to the single peaked bunch in Fig. 3–3 corresponding to a value of unity for the bunching parameter.

This discussion of the value of the bunching parameter that is required for a frequency multiplier can be misleading if the implications of some of the assumptions are overlooked. For example, the drive power required by a multiplier may be considerably greater than that of an amplifier if the input frequency of the two tubes being compared is different. If the input frequency is decreased without a corresponding increase of the drift distance, the number of oscillation cycles at the input frequency during the transit of the drift space will be decreased; *i.e.*, the value of N_1 in Eq. (5–3) will be reduced. As a result, a larger value of input gap voltage E_1 will be required to maintain the correct value of the bunching parameter.

Equation (5–3) may be rewritten in a form that is more suitable for klystron frequency multipliers as they are usually designed. If the drift distance and output frequency are the design factors that remain constant, then N_1 varies as the multiplication ratio M is changed. However, the number of oscillation cycles at the output frequency will be constant. Since the relation between the number of cycles, referred to the two frequencies, is given by

$$N_2 = MN_1 \qquad (5\text{–}4)$$

Eq. (5–3) may be rewritten

$$x = \beta_1 \pi \frac{N_2}{M} \frac{E_1}{E_0} \qquad (5\text{–}5)$$

Therefore the input gap voltage E_1 must usually be increased to maintain optimum bunching when the multiplication ratio M is increased.

This discussion of bunching ignores the effects of space charge. Such an analysis is useful in describing the output characteristics of a multiplier as the input gap voltage is varied, but the magnitude of the output is seriously affected by the debunching phenomena. The discrepancies between the first-order analysis and the actual output characteristics will be considered in Sec. 5.6.

5.5. Effect of Overbunching.—A frequency multiplier is more sensitive to changes in input gap voltage than an amplifier. Reference to the curve in Fig. 5–2 for a multiplication ratio of 10 shows that the output is negligible until the input gap voltage is half the value required for optimum output. Then the output increases rapidly to a maximum, and overbunching causes the output to decrease, theoretically to zero, quite rapidly. Additional input voltage causes the output to reappear, and a number of output peaks occur, similar to the behavior of an overbunched amplifier illustrated in Fig. 3–7, but the output maxima are closer together in the case of a multiplier klystron.

The fact that the output of a klystron frequency multiplier decreases rapidly as the input gap voltage is reduced makes this type of klystron quite sensitive to small changes in the tuning of the input resonator. This effect becomes more serious as the multiplication ratio is increased. Compare the curves for $M = 1$ and $M = 10$ in Fig. 5–2. If the bunching parameter is reduced from 1.20 to 1.01, the radio-frequency current in the tenth harmonic becomes 71 per cent of the maximum value. The same change in the tuning of the input resonator of an amplifier will reduce the fundamental component of the bunched-beam current only 3 per cent. Since the sharpness of tuning is usually determined by the Q of the circuit, as mentioned in Sec. 5.3, this effect in a klystron frequency multiplier is equivalent to an increase in the Q of the input resonator.

5.6. Output Characteristic.—The discrepancies between the first-order theory for a klystron frequency multiplier and actual conditions are most apparent in the magnitude of the power output. The theoretical curve

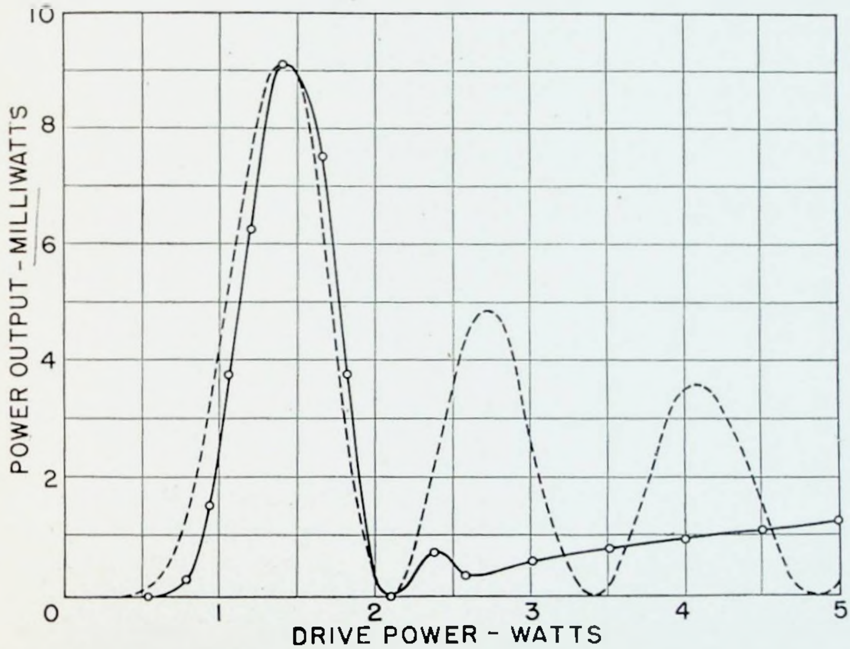


FIG. 5-3.—Power output of a klystron frequency multiplier as a function of drive power. A theoretical curve is also shown.

for the power output as a function of input power can be obtained by squaring both coordinates of the curve for $M = 10$ in Fig. 5-2. This construction was used in computing a similar theoretical curve for an amplifier which is shown in Fig. 4-1. The output power from a typical klystron with a multiplication ratio of 10 has been plotted in Fig. 5-3 as a function of drive power to the input resonator. A theoretical curve is shown in the same illustration as a dotted line and has been matched to the experimental curve at the point of maximum output. The two curves do not differ seriously when the drive power is small, but the multiple output peaks predicted by the first-order theory for the overbunched region beyond the first maximum of the output are missing or considerably reduced in amplitude.

An explanation for the failure of the simple theory when the electron beam in a frequency multiplier is overbunched can be supplied by a consideration of the debunching forces caused by the space charge of the electrons in the bunches. This effect was discussed in Sec. 3.15. The forces of replusion between the electrons in the bunch oppose the formation of an electron bunch. As a result, the current distribution is not so sharp as indicated in Fig. 3–3. The spreading of the electron bunch may not affect the magnitude of the fundamental component appreciably, but the harmonic content will be reduced considerably. This effect is one of the factors that prevent the realization of the theoretical efficiency of a klystron frequency multiplier. Additional discussion of debunching is given in Sec. 5.9.

The debunching effect becomes more pronounced when the electron beam is overbunched. The harmonic content predicted by the simple theory and indicated by the series of output peaks in the dotted curve in Fig. 5–3 is almost completely destroyed when the electron beam is overbunched. Some of the effects of debunching are undoubtedly hidden because the amplitude of the theoretical curve has been matched to the experimental curve arbitrarily at the first output peak. However, the discrepancy between the theoretical curve and the actual output characteristic is quite apparent in Fig. 5–3. Comparison with Fig. 4–4 will show that the second output peak for an overbunched klystron amplifier agrees much more closely with the theoretical curve.

5.7. Beam Voltage Variations.—Another interesting output characteristic showing the effect of varying the beam voltage on a klystron frequency multiplier is illustrated in Fig. 5–4. In this case the drive power is maintained constant. As the beam voltage is decreased from the value that gives optimum bunching, the number of oscillation cycles during the transit through the drift space will increase owing to the reduced velocity of the electrons in the beam. As a result, the value of the bunching parameter x in Eq. (5–3) will increase. When the electron beam is overbunched, corresponding to the values of x greater than 1.20, the output will first be reduced to zero; then successive output maxima corresponding to the overbunched region in Fig. 5–2 will occur. The output of these overbunched maxima at low voltages will not be large, since the reduction of beam voltage decreases the beam power input to the tube considerably.

This characteristic for a frequency multiplier differs from the similar characteristic for an amplifier shown in Fig. 4–5, owing to the nature of the J_{10} Bessel function. Increasing the beam voltage decreases the bunching if the drive power is constant. The increased beam current accompanying an increase in beam voltage does not maintain the output from a multiplier at a high level because the value of $J_{10}(10x)$ decreases rapidly as the bunching is reduced, and the curve in Fig. 5–4 shows a definite maximum just beyond the point corresponding to optimum bunching. The output peaks for low beam voltages in a frequency multiplier are more closely spaced than the successive peaks in a klystron amplifier because the value of $J_{10}(10x)$ fluctuates rapidly for values of x greater than unity (see Fig. 5–2).

Figure 5–4 is a theoretical curve and does not include the effects of space-charge debunching. The multiple output peaks are present in an actual characteristic for a klystron frequency multiplier, but their amplitude is very much reduced.

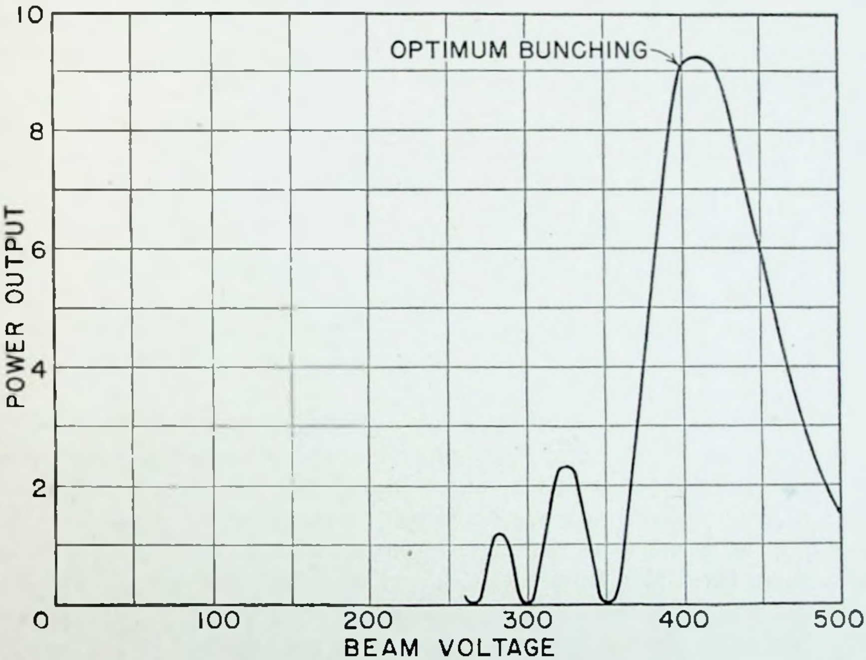


FIG. 5–4.—Theoretical curve of the power output of a klystron frequency multiplier as a function of the beam voltage when the drive power is constant.

In addition to the amplitude variations, the frequency of the output will differ from an integral harmonic of the input frequency if the beam voltage is varied rapidly. The transit time between input and output gaps will vary and there will be phase modulation introduced because the frequency of the output depends upon the rate of change of the beam voltage.

5.8. Drive Power Requirements.—Not only is it necessary to know the the effect of changing the drive power on the output of the tube; the factors in the design of a klystron frequency multiplier which govern the amount of drive power required are equally important. The statement in Sec. 5.4 that larger multiplication ratios require smaller values of the bunching parameter is likely to lead to the erroneous assumption that less drive power might be required. It was pointed out previously that this conclusion would be true only if the increased multiplication ratio was obtained by maintaining the input frequency and drift distance constant, and changing the output resonator to a higher frequency. However, debunching considerations usually dictate that the drift distance be decreased as the output frequency is increased.

For constant output frequency and drift distance, the input gap voltage must be increased in proportion to the multiplication ratio. This relation was derived in Eq. (5–5) and is repeated below. N_2 represents the number of cycles of drift time referred to the output frequency, and M is the multiplication ratio.

$$x = \beta_1 \pi \frac{N_2}{M} \frac{E_1}{E_0} \tag{5–5}$$

Since the optimum value of the bunching parameter x decreases only slightly as M is increased, the voltage E_1 required at the input gap is approximately proportional to M, the multiplication ratio. Since the drive power is proportional to the square of the input gap voltage, the power requirements may become quite high when large multiplication ratios are used.

5.9. Space-charge Debunching.—An obvious but unsatisfactory answer to the problem of obtaining large multiplication ratios without increasing the input gap voltage would be an increased drift length. Two different types of debunching make such a design impracticable. Transverse debunching causes a loss of the electron beam owing to the repulsion of the electrons in the bunch toward the side walls of the drift space. This loss is appreciable, but longitudinal debunching introduces a more serious difficulty. The loss of harmonic content in the bunched beam may be great enough so that an increase of current through the output gap, not including the electrons lost by transverse debunching, may actually reduce the output. The maximum output will occur when the output gap voltage is very much less than the beam voltage and the multiplier efficiency will be extremely small.

These debunching effects are greatly reduced by decreasing the drift distance. Since the amount of debunching that can be tolerated depends on the output frequency, the drift distance is usually determined by the output frequency rather than the input frequency of a klystron multiplier.

5.10. Large Input Voltages.—A typical multiplier tube might have an input frequency of 300 megacycles, a multiplication ratio of 10, and a drift distance of 4.5 cm. If the beam voltage were 625 volts, the electron velocity, given by Eq. (3–3) in Sec. 3.2, would be 1.5×10^9 cm per sec. The transit time in the drift space T_0 is determined by the drift distance s_0 and the average electron velocity v_0.

$$T_0 = \frac{s_0}{v_0} \tag{5-6}$$

Therefore, the transit time for the electrons with average velocity would be 3×10^{-9} sec. The number of cycles corresponding to the transit time in the drift space is proportional to the input frequency f_1 and the transit time T_0.

$$N_1 = f_1 T_0 \tag{5-7}$$

For the tube design being considered, N_1 has a value of 0.9 cycle.

It was pointed out in Sec. 3.14 that short drift distances corresponding to values of N_1 less than unity require large voltages at the input gap to produce the electron bunching. Also, Fig. 3-11 in Sec. 3.17 shows that the bunching parameter defined by Eqs. (3-21) and (5-3) does not apply when the input gap voltage is a large fraction of the beam voltage. However, it is convenient to define the bunching parameter in this manner although the first-order analysis is not applicable.

5.11. Transit-time Effects.—The necessity for large input voltages in a klystron frequency multiplier also means that the average velocity of the electrons in the beam is less than the velocity corresponding to the acceleration voltage. For example, a resonator gap voltage equal to the acceleration voltage would increase the velocity of the electrons by only 41 per cent when the field was positive, but would completely stop the electrons when the field was opposing their motion. Since many electrons would have almost zero velocity, the average velocity would become less than the velocity of an unaccelerated electron.

Many electrons travel much slower than the average velocity; therefore, the transit time across the output gap will not be the same for all of the electrons. Although this wide variation of transit time makes the simplified analysis of the beam coupling coefficient given in Sec. 3.12 of doubtful value, the general nature of the result applies to frequency multipliers as well as to amplifiers. The transit time at the output gap, based on the average beam velocity, must be kept reasonably small or a serious loss of output will occur.

5.12. Output Load Characteristics.—Since the harmonic component of the beam current remains constant if the beam voltage and the drive power of an amplifier or a frequency multiplier are unchanged, either type of klystron can be considered as a resonant circuit excited by a constant-current source. Therefore, the differences between the bunching characteristics of a multiplier and an amplifier do not affect the behavior of the output when the load is varied. Similarly, the transit time affects only the magnitude of the harmonic component. As a result, the discussion of load characteristics, resonator coupling, and Rieke diagrams in Secs. 4.9 to 4.11 and Sec. 4.14 applies equally well to a klystron frequency multiplier.

The probability of attaining an output gap voltage as large as the beam voltage in a multiplier is rather remote, particularly if the multiplication ratio is large; therefore, the maximum output will be obtained when the equivalent load resistance is equal to the shunt resistance of the output resonator.

CHAPTER 6

REFLECTION BUNCHING

6.1. Reflex Klystron Oscillators.—Any amplifier will oscillate if sufficient energy is fed back to the input circuit in the proper phase. It might seem logical to apply that principle to a two-resonator klystron amplifier to develop the theory of a klystron oscillator. However, it will be simpler to consider first a type of klystron that has only one resonant circuit and combines the functions of input and output circuits in the same resonator. Feedback in this single resonator oscillator is obtained by reflecting the electron beam so that it passes through the resonator gap a second time. This type of klystron is called a "reflex oscillator." The analysis of a reflex klystron is relatively simple because two coupled circuits are not involved.

In order to explain the operation of these reflex oscillators, an understanding of the electron bunching that occurs while the electrons are in the reflecting field will be required. A qualitative explanation of reflection bunching will be given as an introduction; then a ballistic analysis of a reflex klystron similar to that used in Chap. 3 will be derived. The similarity between reflection field bunching and bunching in a field-free drift space will be shown. These relations will be applied in Chap. 7 to an analysis of the electrical characteristics of a reflex oscillator.

6.2. Operating Principles of a Reflex Klystron.—A sectional view of a reflex klystron is shown in Fig. 6–1. An electron gun furnishes a beam of electrons which are accelerated toward the anode plane by the beam voltage E_0. The beam travels along the axis of the tube beyond the anode plane with a uniform velocity v_0 corresponding to E_0, the acceleration voltage, until it reaches the resonator gap. A radio-frequency voltage across the resonator gap will modify the velocity of the electrons in the beam. Some electrons will be speeded up when the field has a direction that will accelerate the beam. Other electrons will be slowed down during another part of the radio-frequency cycle, and the velocity of some electrons will not be changed because they pass the gap when the resonator voltage is zero. The velocity variation will be assumed to be small, and the average velocity of the electrons in the beam will be identical to the velocity corresponding to the acceleration voltage, since an equal number of electrons will be slowed down and speeded up during one radio-frequency cycle.

Beyond the resonator gap, the electrons encounter a retarding electric field produced by the potential between the reflector and the anode

$E_0 + E_r$. This reflecting field brings the electrons to rest and returns them to the cavity resonator. The shape of the reflector electrode is designed to preserve the focus of the beam. The beam current is constant when the beam leaves the resonator gap, but electron bunching takes place while the electrons are in the reflection space, and the beam is density modulated when it returns to the cavity resonator.

6.3. Applegate Diagram.—If space-charge effects and the focusing action of the reflector shape are neglected, the bunching action is analogous to the motion of objects in a gravitational field. An Applegate diagram, in Fig. 6–2, is a convenient method of illustrating the bunching action. This

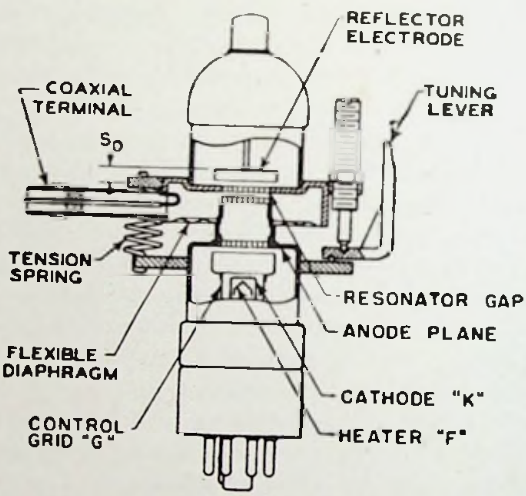


Fig. 6–1.—Sectional view of a reflex klystron.

diagram represents the resonator gap voltage as a function of time, and plots the position in the reflection space of a number of electrons that pass the resonator gap at selected intervals during a complete cycle. The opposite action of the radio-frequency field on the electrons leaving the resonator and those returning to the resonator after bunching has been shown on the diagram.

An electron that has been speeded up by the action of the radio-frequency field will travel farther into the reflecting field and will take longer than the average time to return to the resonator. This behavior is similar to throwing a ball into the air; the harder the ball is thrown, the longer it takes to return to the ground. Reference to Fig. 6–2 will show that an electron that passes the resonator gap early in the cycle at a time t_a is accelerated and requires a longer time to return than an electron leaving at time t_b when the radio-frequency field is zero. The electrons that leave at time t_c later in the cycle require less than the average transit time and all these electrons return to the resonator in a bunch at time t_r. Bunching of the electron beam is the result, and the uniform flow of beam current is converted into an equivalent direct current with a superimposed alternating component.

6.4. Phase Conditions for Oscillation.—The arrival time t_r of a group of electrons returning to the resonator depends upon the physical dimensions of the klystron and also upon the acceleration voltage and the reflector voltage. In general, the transit time for the electron with average velocity, leaving at time t_b, may correspond to any number of cycles of the radio

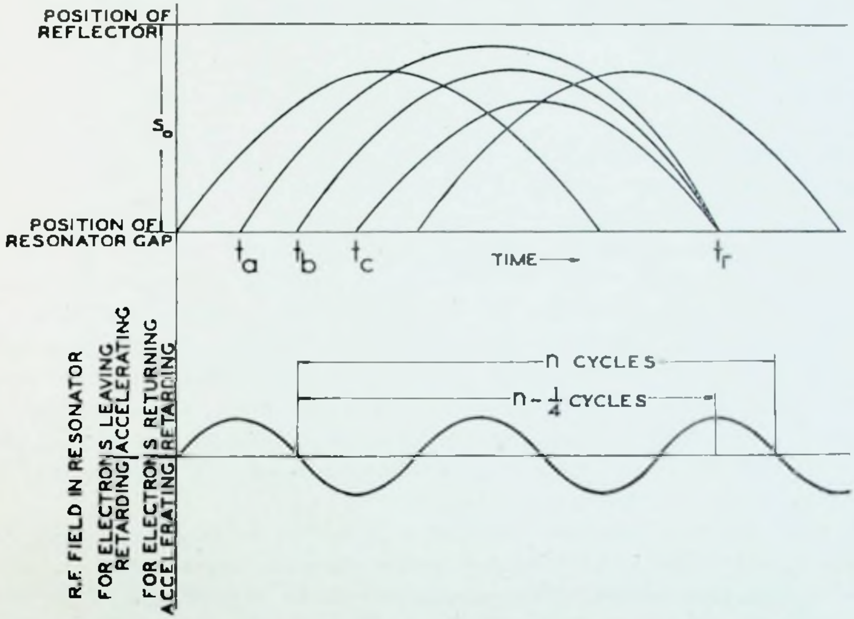


FIG. 6–2.—Applegate diagram for a reflex klystron oscillator.

frequency, and this number need not be an integer. But in order to sustain oscillations, the electron bunch must arrive during the time when the radio-frequency field is retarding the returning electrons, so that the electron velocity is reduced and some of the kinetic energy of the electrons is transferred into electromagnetic energy in the cavity resonator field.

The electron that is to become the center of the bunch leaves at the time t_b when the radio-frequency voltage is zero and changing from acceleration to deceleration. At an integral number of cycles later, the radio-frequency voltage will again be zero but, for the returning electron, the field will be changing from deceleration to acceleration. This time is indicated by n cycles in Fig. 6–2. Since a maximum retarding field is required for maximum energy transfer from the bunched electron beam, the transit time for an electron that enters the reflecting field with average velocity must correspond to one quarter cycle less than an integral number of cycles. This transit-time requirement may be verified by inspection of Fig. 6–2.

6.5. Multiple Transits.—Most of the electrons are collected by the metal walls of the tube after they have given up energy to the resonator field. Other electrons may have been lost by interception by the grid structures. A few electrons may survive these chances of getting collected and will be decelerated near the cathode surface, then reaccelerated with the newly emitted electrons. Upon reentering the reflection space, these electrons will behave differently from the electrons that are going through the round-trip cycle for the first time. These electrons that make multiple transits may produce undesirable effects, but in most cases their effect may be neglected. More important factors (such as space-charge debunching forces) will be neglected in order to simplify the analysis.

6.6. Electron Ballistics.—It was mentioned previously that the electrons that pass the resonator gap when the radio-frequency voltage is zero enter the reflecting field without any change in velocity and are defined as electrons with average velocity. Electrons that pass the resonator gap at a time t_b (see Fig. 6–2), when the radio-frequency field is changing from acceleration to deceleration, become the center of the bunch. The electrons in the bunch have different velocities, which are continually changing during the time the electrons are in the reflection space; however, it is convenient to consider that the bunch moves as a unit along a path determined by the electron that is to become the center of the bunch. Note that the lines in Fig. 6–2 representing electrons leaving at times t_a, t_b, and t_c appear to converge about the center of the bunch.

A brief review of electron ballistics will derive the equations that are useful in determining the relationships between the transit time and the tube design parameters. The calculation of the transit time from the tube voltages and the reflector electrode spacing will not be accurate because the effects of space charge have been neglected. Although the effects of space charge are quite important, the assumption simplifies the analysis considerably, and the result is quite useful.

The average electron velocity v_0 is determined by the acceleration voltage E_0. The relation is the same as that given in Eq. (3–1) and is repeated below.

$$\tfrac{1}{2}mv_0^2 = E_0 e \qquad (6\text{–}1)$$

Equation (6–1) is then rewritten in the form,

$$v_0 = \sqrt{\frac{2e}{m} E_0} \qquad (6\text{–}2)$$

Evaluating the constant in the proper units so that v_0 is expressed in centimeters per second and E_0 is in volts gives

$$v_0 = 6 \times 10^7 \sqrt{E_0} \tag{6-3}$$

6.7. Transit Time in the Reflection Space.—Other laws of motion of particles may be used to determine the transit time. If the deceleration is denoted by a, then the position of a particle as a function of time is given by

$$s = v_0 t - \tfrac{1}{2}at^2 \tag{6-4}$$

When t is equal to the average transit time T_0, the electron has returned to the resonator, the electron velocity is again v_0, but in the opposite direction, and s is equal to zero, *i.e.*,

$$0 = v_0 T_0 - \tfrac{1}{2}aT_0^2 \tag{6-5}$$

There are two solutions to Eq. (6–5). T_0 equal to zero corresponds to an electron that has not traversed the reflection space and is disregarded. The other solution is

$$T_0 = \frac{2v_0}{a} \tag{6-6}$$

The deceleration a may be evaluated from the familar equation for the force acting on a particle. This force is given by the product of the charge on the electron and the gradient of the potential between the anode and the reflector electrode. If the reflector field is assumed to be uniform, the gradient is simply the sum of the voltages on the reflector electrode divided by s_0, the reflector spacing. Therefore,

$$F = ma = e\,\frac{E_0 + E_r}{s_0} \tag{6-7}$$

Substitution of Eqs. (6–7) and (6–2) in Eq. (6–6) gives

$$T_0 = \frac{2v_0}{\dfrac{e}{m}\,\dfrac{E_0 + E_r}{s_0}} = 4s_0\,\frac{\sqrt{(m/2e)E_0}}{E_0 + E_r} \tag{6-8}$$

for the average transit time.

It is usually more convenient to express the transit time in terms of a number of oscillation cycles rather than as a time interval. This equivalent number of cycles will be designated N and is defined by

$$N = fT_0 \tag{6-9}$$

where f is the frequency of oscillation. Equation (6–8) may therefore be written

$$N = 4fs_0\,\frac{\sqrt{(m/2e)E_0}}{E_0 + E_r} \tag{6-10}$$

6.8. Voltage Modes in an Oscillator.—If oscillation is to be at maximum strength, the number of cycles during the transit time in the reflection space must satisfy the relation mentioned in the discussion of Fig. 6–2, *i.e.*,

$$N = n - \tfrac{1}{4} \tag{6–11}$$

where n is an integer greater than zero. Oscillation at the same frequency will occur for a number of values of N, and each value of N may be provided by the proper choice of the acceleration voltage and the reflector voltage. A series of curves showing the reflector voltage required to give constant

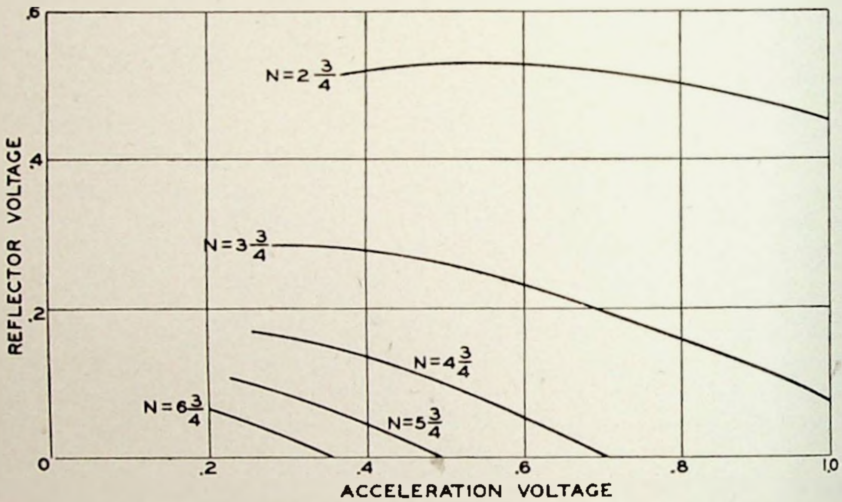


FIG. 6-3.—Family of theoretical curves showing voltage modes in a reflex oscillator.

frequency for any value of acceleration voltage are shown in Fig. 6–3. Each curve represents a different value of N. The value of N may be estimated from the frequency, reflector spacing, and voltages involved. These transit times are an important factor in the behavior of reflex klystrons, and the importance of transit time will be discussed in greater detail in the sections that follow. In practice, transit times corresponding to values of N between $1\frac{3}{4}$ and $10\frac{3}{4}$ cycles are typical.

6.9. Electron Bunching Relationships.—It is obvious that electron bunching must occur in a reflex klystron because the velocity variation introduced by the resonator voltage produces a variation of the transit times of electrons that pass the resonator gap at different times during a cycle. This variation of transit time may be expected from Eq. (6–6), which may be rewritten in terms of a varying velocity instead of the average velocity, and becomes

$$T = \frac{2v}{a} \tag{6–12}$$

when T and v are varying quantities. The current distribution in the bunched beam is similar to the bunching in a two-resonator klystron, but the manner in which the electrons become grouped is different and there is a phase difference of 180 deg between the two types of bunching.

These differences between reflection field bunching and field-free bunching are introduced because the transit time is proportional to the electron velocity in a reflex klystron, while the transit time in the field-free drift space between the resonators in a two-resonator klystron is inversely proportional to the velocity. As a result, the electron bunch in a reflex klystron is formed around the electron which passed the resonator gap when the radio-frequency voltage was changing from acceleration to deceleration. In contrast, the bunch in a two-resonator klystron forms around the electron that passed the input resonator gap when the radio-frequency field was changing from deceleration to acceleration.

6.10. Bunched-beam Current.—An analysis of the bunching process in a reflex klystron may be made, following the method used for the two-resonator type of klystron. Negligible transit time across the resonator gap will be assumed in the preliminary analysis, and the factors that must be modified when this assumption is invalid will be discussed in a later section.

The electrons approach the resonator gap with an average velocity v_0, which is determined by the acceleration voltage E_0 as shown in Eq. (6–2). The velocity of the electrons will be modified by the radio-frequency voltage at the resonator gap, and after passing the gap the velocity will be

$$v = \sqrt{\frac{2e}{m}}\,\sqrt{E_0 + E_1 \sin \omega t_1} \qquad (6\text{–}13)$$

where E_1 is the peak value of the radio-frequency voltage at the resonator gap, ω is the angular frequency and equal to $2\pi f$, and t_1 is the time an electron passes the resonator gap. The transit time of an electron will be given by Eq. (6–12) and may be rewritten in a form similar to Eq. (6–8).

$$T = 4s_0 \frac{\sqrt{(m/2e)E_0}}{E_0 + E_r} \sqrt{1 + \frac{E_1}{E_0} \sin \omega t_1} \qquad (6\text{–}14)$$

Equation (6–8) may be substituted in Eq. (6–14) to give

$$T = T_0 \sqrt{1 + \frac{E_1}{E_0} \sin \omega t_1} \qquad (6\text{–}15)$$

When the ratio of E_1/E_0 is small an approximate form of Eq. (6–15) may be used:

$$T \cong T_0\left(1 + \frac{E_1}{2E_0} \sin \omega t_1\right) \qquad (6\text{–}16)$$

Returning electrons will arrive at the resonator gap at a time t_2, which will be the sum of the transit time T and the departure time t_1.

$$t_2 = t_1 + T_0\left(1 + \frac{E_1}{2E_0}\sin \omega t_1\right) \tag{6-17}$$

The number of electrons that return to the resonator during a time interval dt_2 will be equal to the product of the instantaneous beam current in the reverse direction I_2 and the time interval dt_2. This same number of electrons originally passed the resonator gap during an interval dt_1, when the beam current in the forward direction was equal to I_0, the d-c beam current. If these expressions for the number of electrons are equated,

$$I_2\, dt_2 = I_0\, dt_1 \tag{6-18}$$

and the instantaneous bunched current is given by

$$I_2 = I_0 \frac{dt_1}{dt_2} \tag{6-19}$$

Differentiating both sides of Eq. (6–17) gives

$$dt_2 = dt_1\left(1 + \omega T_0 \frac{E_1}{2E_0} \cos \omega t_1\right) \tag{6-20}$$

or

$$dt_2 = dt_1\left(1 + \pi f T_0 \frac{E_1}{E_0} \cos \omega t_1\right) \tag{6-21}$$

Substituting Eq. (6–9) in Eq. (6–21) gives

$$dt_2 = dt_1\left(1 + \pi N \frac{E_1}{E_0} \cos \omega t_1\right) \tag{6-22}$$

which may be rewritten

$$dt_2 = dt_1(1 + x \cos \omega t_1) \tag{6-23}$$

The quantity x is known as the "bunching parameter" and is defined by

$$x = \pi N \frac{E_1}{E_0} \tag{6-24}$$

6.11. Gap Transit Time.—The effective gap voltage will be decreased if the transit time across the gap is not negligible, as pointed out in Sec. 3.12. The value of the bunching parameter will be less, and Eq. (6–24) must be modified to include the beam coupling coefficient β.

$$x = \beta \pi N \frac{E_1}{E_0} \tag{6-25}$$

The value of β was derived in Sec. 3.12 and is given by

$$\beta = \frac{\sin(\delta/2)}{\delta/2} \tag{6-26}$$

6.12. Reflection plus Field-free Bunching.—The existence of a field-free bunching space in addition to the reflection space requires a modification of this analysis. An Applegate diagram similar to Fig. 6–2, but including more electron paths during a cycle, is shown in Fig. 6–4. Compare this illustration with the Applegate diagram in Fig. 1–3, Sec. 1.5, showing the bunching in a field-free drift space. Note that the current distribution in the reflected bunch is similar to that in the field-free case, but that the manner in which the electrons reach this distribution is decidedly different.

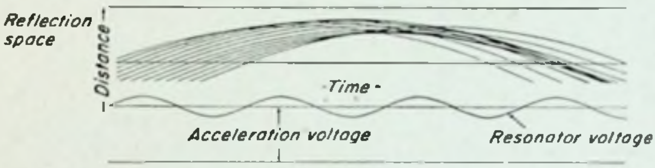


FIG. 6–4.—Applegate diagram showing electron bunching in a reflecting field.

In the field-free case, an infinite current peak exists before the bunched beam reaches the output gap. The electrons then pass each other; *i.e.*, there is a "crossover" of electrons traveling along the beam. The infinite current peak separates and a large but finite current exists between two diverging infinite current peaks. This current distribution is illustrated in Fig. 3–3, Sec. 3.7. A similar current distribution occurs in a reflex tube, but there is no single infinite current peak formed prior to the double infinite current peak.

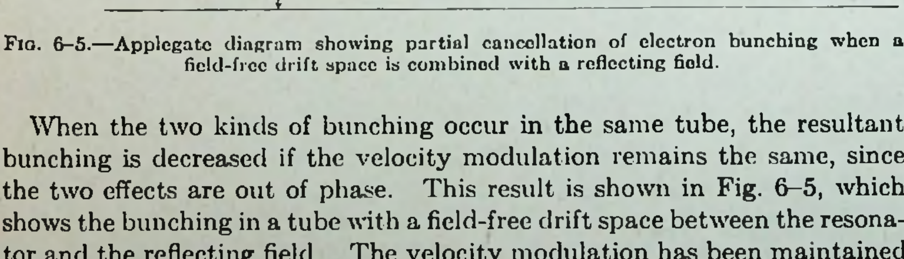


FIG. 6–5.—Applegate diagram showing partial cancellation of electron bunching when a field-free drift space is combined with a reflecting field.

When the two kinds of bunching occur in the same tube, the resultant bunching is decreased if the velocity modulation remains the same, since the two effects are out of phase. This result is shown in Fig. 6–5, which shows the bunching in a tube with a field-free drift space between the resonator and the reflecting field. The velocity modulation has been maintained equal to that shown in Fig. 6–4 and the electron beam is underbunched owing to the effect of the field-free drift distance.

6.13. Effective Bunching Time.—The equation for the bunching parameter can be modified to include the combination of field-free and reflection bunching. Since the two types of bunching tend to cancel each other, Eq. (6–25) becomes

$$x = \beta\pi N_r \frac{E_1}{E_0} - \beta\pi N_{ff} \frac{E_1}{E_0} \tag{6–27}$$

where N_{ff} is the number of cycles during both trips in the field-free drift space and N_r is the number of cycles during the round trip in the reflecting field. Since N_r is usually greater than N_{ff}, we can write

$$N' = N_r - N_{ff} \tag{6–28}$$

and Eq. (6–27) becomes

$$x = \beta\pi N' \frac{E_1}{E_0} \tag{6–29}$$

The phase of the returning electrons depends upon the total transit time N_t, and for maximum output

$$N_t = N_r + N_{ff} = n - \tfrac{1}{4} \tag{6–30}$$

where n is an integer.

It is apparent from these relations that the beam must be considerably overbunched in the reflection space if proper bunching is to be obtained when the two types of bunching are combined in the same tube. The two types of bunching do not cancel exactly, since there is more distortion of the bunch in a field-free drift space when the input gap voltage is large.

If the reflection field is not uniform, and this is usually the case, the effective bunching time may be greater than the actual transit time. For this reason, an equivalent bunching time should be used, and its value may be either greater or less than the value of N, depending upon the design of the klystron. In order to simplify the analysis, a uniform reflecting field will be assumed, and the effective value of bunching time may be substituted for N in the final result.

6.14. Electron Arrival Time.—Substituting Eq. (6–23) in Eq. (6–19) gives

$$I_2 = \frac{I_0}{1 + x \cos \omega t_1} \tag{6–31}$$

Equation (6–31) is similar in form to the expression for the bunched current in a two-resonator klystron.

The equations for the instantaneous current express this current as a function of t_1, the departure time of the electrons when they enter the reflecting field. It is more desirable to know the relation between the instantaneous current and t_2, the arrival time of the returning electrons.

This relationship is easily obtained if a curve of t_1 vs. t_2 is available, and a family of such curves is illustrated in Fig. 6–6 for several different values of the bunching parameter x. This graphical representation of the relationship is necessary because Eq. (6–17) cannot be solved explicitly for t_1. Rewriting Eq. (6–17) in terms of the bunching parameter x gives a form that is convenient for computation of the curves in Fig. 6–6. Equation (6–17) then becomes

$$t_2 = t_1 + T_0 + \frac{x}{\omega} \sin \omega t_1 \tag{6–32}$$

A phase difference of 180 deg, indicated by the formation of the bunch around the electrons that depart at a time equal to π/ω, is the only signifi-

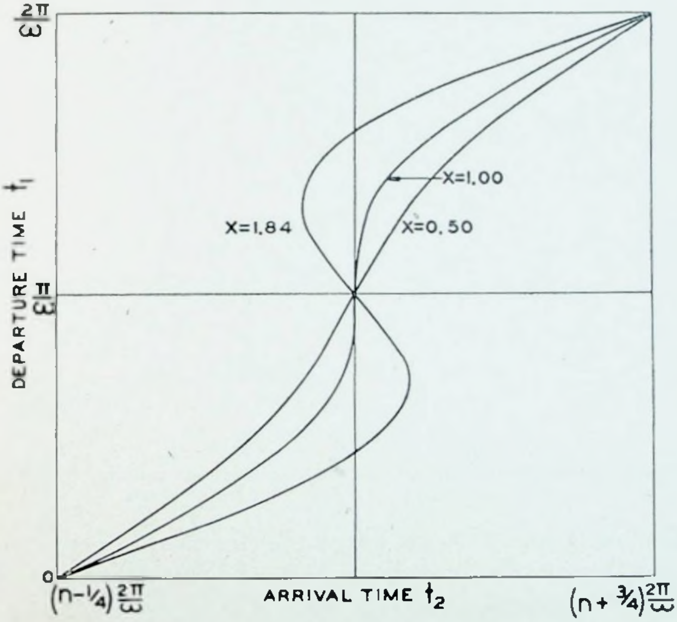


FIG. 6–6.—Electron arrival-time curves for three values of the bunching parameter.

cant difference between Fig. 6–6 and the similar curves for an amplifier given in Sec. 3.6 (Fig. 3–2). However, the transit time indicated by Fig. 6–6 is $(n - \frac{1}{4})$ cycles and the transit time for the amplifier corresponds to an integral number of cycles. The phase relation required between the input and output of an oscillator is not necessary in an amplifier.

The discussion in Sec. 3.6 of the significance of a negative slope for the arrival-time curve also applies to a reflex klystron and the curves in Fig. 6–6. The instantaneous current is the sum of the absolute magnitudes of the slopes for any value of the arrival time t_2.

6.15. Radio-frequency Current in Beam.—Curves of instantaneous current, corresponding to the t_1 vs. t_2 curves in Fig. 6–6, are shown in Fig. 6–7. The current peaks when the bunching parameter is unity, or greater, are not infinite if the transit time across the resonator gap is finite. However, it is convenient to treat the gap length as infinitesimal in deriving the expression for the radio-frequency component of the beam current. Then the correction factor β, known as the "beam-coupling coefficient" and derived in Sec. 3.12, can be applied when the transit time across the resonator gap is appreciable.

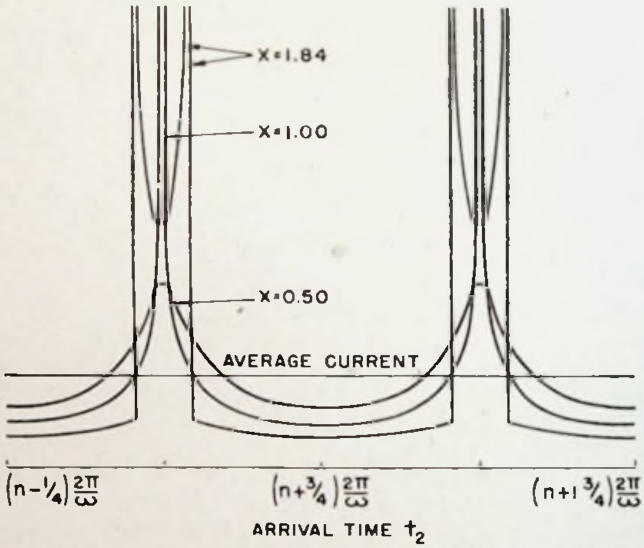


FIG. 6–7.—Instantaneous beam current. Two complete cycles are shown and three values of the bunching parameter x are represented.

Since the instantaneous beam current is identical to that for the field-free case, the current may be expressed by a Fourier series with coefficients that are Bessel functions of the first kind.

$$I_2 = I_0 \, [1 + 2J_1(x) \sin (\omega t_2 - 2\pi N) + 2J_2(2x) \sin 2(\omega t_2 - 2\pi N) + \cdots + 2J_n(nx) \sin n(\omega t_2 - 2\pi N)] \qquad (6\text{–}33)$$

Only the second term is of particular interest in an oscillator, and the fundamental component of the radio-frequency current in the beam, which will be designated i_2, is given by

$$i_2 = 2I_0J_1(x) \sin (\omega t_2 - 2\pi N) \qquad (6\text{–}34)$$

The higher harmonics are unimportant because reflex klystrons are usually designed to operate with a high effective Q.

Figure 6–8 shows the peak value of the radio-frequency component of the bunched-beam current as a function of the bunching parameter. The peak value has been divided by I_0 so that the ordinates of the curve are

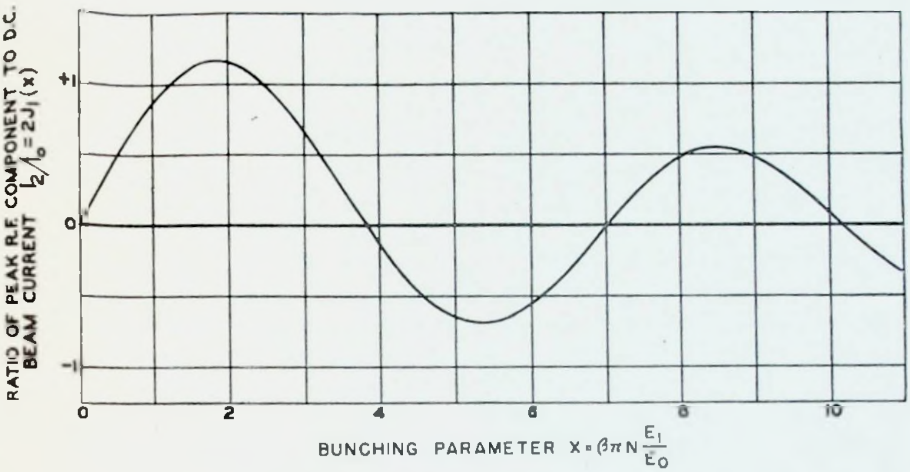


FIG. 6–8.—Ratio of the radio-frequency component of the bunched-beam current to the direct current in the electron beam. An accurate graph of the Bessel function $2J_1(x)$ is given in Chart X in Appendix B.

equal to $2J_1(x)$. This Bessel function output curve is characteristic of klystron tubes, and Fig. 6–8 is similar to Fig. 3–7 for a two-resonator klystron.

CHAPTER 7

REFLEX OSCILLATORS

7.1. General Oscillator Theory.—Reflex klystrons are a simple form of oscillator in which a single resonator acts as both input and output circuit and, for this reason, these tubes furnish a convenient introduction to oscillator theory. There are two principal differences between reflex oscillators and circuits using conventional vacuum tubes. Velocity-modulation tubes, in fact, all transit-time devices including triodes operated at ultrahigh frequencies, introduce a phase angle between the input voltage and the output current. This means that a transconductance is no longer suitable for describing the tube characteristic and must be replaced by a transadmittance. The second difference between klystrons and triodes is introduced by the overbunching characteristic, which is decidedly different from the saturation characteristics of conventional low-frequency vacuum tubes.

The conditions for stable oscillation in any circuit are that the power supplied by the tube is equal to the losses in the circuit, including the coupled load, and that the total phase angle around the feedback circuit is zero. It is convenient to restate this principle by saying that the frequency of oscillation and the strength of oscillation will adjust themselves until the transadmittance of the tube is equal in magnitude to the total admittance of the circuit and load, but of opposite sign. Another familiar form of the latter statement is that a circuit will oscillate if a negative resistance equal to the circuit resistance is introduced.

The effect of the reflected beam in a reflex klystron oscillator can be explained quite easily by assuming that the radio-frequency component of the bunched beam introduces an admittance Y_2 in parallel with the resonant circuit. This method reduces the analysis to a simple circuit problem in which a change in the value of Y_2 may change the resonant frequency or losses in the circuit. The results are correct; in fact, it can be shown that this and various other methods of analysis are mathematically identical. The advantage of the method to be used here is primarily convenience in visualizing the problem, since the effect of varying components in a circuit is often more easily understood than the effect of varying the parameters in an equation.

7.2. Equivalent Circuit.—An equivalent circuit for a reflex klystron oscillator, based on the method outlined above, is shown in Fig. 7–1. The cavity resonator and its coupled load are represented by the parallel *RLC* circuit. The copper losses and other resonator losses, such as loading caused by the beam itself or secondary electrons, are represented by an equivalent shunt resistance R_s, and the coupled load or output circuit is considered as another parallel resistance R_L. Then the effective resistance R_{sL} will be given by the expression for two resistances in parallel.

$$R_{sL} = \frac{R_s R_L}{R_s + R_L} \tag{7-1}$$

The equivalent capacity C represents the capacity of the resonator gap. The value of this capacity can be estimated to a satisfactory approximation

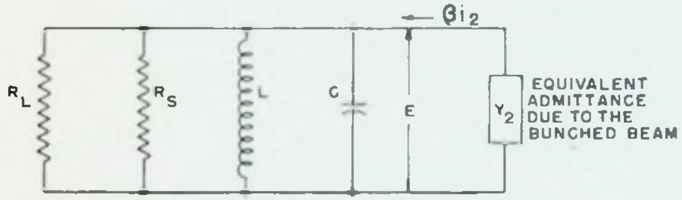


FIG. 7–1.—Equivalent circuit for a reflex klystron oscillator.

from the formula for a parallel plate condenser, using the area and spacing of the resonator grids forming the gap for the dimensions of the condenser. The value of the equivalent inductance L is chosen to make the resonant frequency of the equivalent circuit equal to the resonant frequency of the cavity.

If the reflex klystron is oscillating, or if energy is coupled into the cavity resonator from an external source, then a voltage will exist across the resonator gap. This voltage is represented by the voltage E across the capacity C in the equivalent diagram in Fig. 7–1, and the value of E is given by

$$E = E_1 \sin \omega t \tag{7-2}$$

where E_1 is the peak value of the voltage across the resonator gap, and ω and t represent the angular frequency of oscillation and time.

The bunching action produces a radio-frequency current i_2, which depends upon the beam current I_0 and the bunching parameter x, as shown in Eq. (6–34.) This equation is repeated for convenience.

$$i_2 = 2I_0 J_1(x) \sin (\omega t - 2\pi N) \tag{7-3}$$

N represents the number of oscillation cycles during the time an electron is in the reflection space. The effect of a nonuniform reflecting field will be neglected, and N will be used to represent both the actual transit time and the effective bunching time in order to simplify this analysis (see Sec. 6.13).

A current βi_2 is shown flowing out of the "fictitious" admittance Y_2, which represents the effect of the bunched-beam current in the equivalent diagram. This direction for the current is chosen because Y_2 represents the source of power. The beam coupling coefficient β is introduced in order to include the effect of the decreased energy transfer from the beam to the resonator when the gap transit time is large. This factor must be included in each step of the derivation in which it should appear; but a value of unity, corresponding to negligible gap transit time, will be assumed in most cases in order to simplify the discussion of this analysis.

7.3. Circuit Admittance.—Admittances are used in this discussion because they can be added when considering parallel circuits. The total admittance of the resonator Y_s would include the susceptance terms for the inductance and capacity as well as the conductance terms representing the losses and load.

$$Y_s = \frac{1}{R_s} + \frac{1}{R_L} - \frac{j}{\omega L} + j\omega C \qquad (7\text{–}4)$$

This form will be quite convenient in the analysis of a reflex oscillator because real and imaginary terms may be considered separately.

7.4. Equivalent Beam Admittance.—An evaluation of the admittance Y_2 which is added to the resonator admittance may be obtained from the fact that a voltage E must cause a current βi_2 to flow. The magnitude of Y_2 will be determined by the ratio of the peak value of βi_2 and the peak voltage E_1.

$$|\,Y_2\,| = \frac{2\beta I_0 J_1(x)}{E_1} \qquad (7\text{–}5)$$

The phase of Y_2 is determined by the transit time in the reflection field If the transit time corresponds to $(n - \frac{1}{4})$ cycles, where n is an integer, then the electrons in the bunch will be retarded, and the beam will transfer energy to the radio-frequency field in the resonator. This relation was explained in the discussion of Fig. 6–2. Under these conditions Y_2 will be a pure negative conductance. A transit time of $(n + \frac{1}{4})$ cycles corresponds to a transfer of energy from the radio-frequency field to the electron beam, and in this case Y_2 is a positive conductance; *i.e.*, the beam represents an additional loss in the circuit.

Other values of transit time cause Y_2 to be complex, since the radio-frequency component of the bunched-beam current will not be in phase with the resonator voltage. The phase angle of i_2 will be represented by ϕ, and ϕ will be considered zero when the transit time in the reflection field corresponds to $(n - \frac{1}{4})$ cycles. The expression for i_2 in Eq. (7–3) may be rewritten

$$i_2 = 2I_0J_1(x) \sin [\omega t - 2\pi(n - \tfrac{1}{4}) + \phi] \tag{7–6}$$

Comparison of Eqs. (7–3) and (7–6) shows that the phase angle ϕ is defined by

$$\phi = 2\pi(n - \tfrac{1}{4}) - 2\pi N \tag{7–7}$$

N may have any value and is determined by the transit time in the reflection space, but n is always an integer. If the transit time is correct for maximum output, then the phase angle ϕ is zero, and N is given by

$$N = n - \tfrac{1}{4} \tag{7–8}$$

Decreasing either the beam voltage or the reflector voltage increases the transit time in the reflection space and therefore increases the value of N.

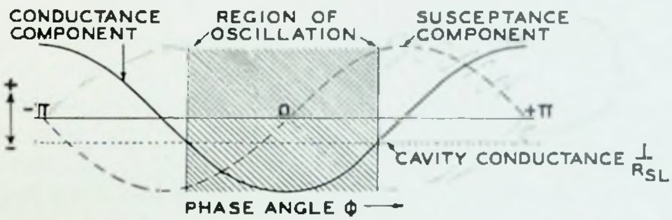


FIG. 7–2.—Conductance and susceptance components of the electron-beam admittance.

This means that the electron bunch returns to the resonator gap later than the correct time for maximum output. This arrival time corresponds to a negative phase angle ϕ, as indicated by Eq. (7–7).

It will be convenient to express i_2 in the vector form instead of the sinusoidal form in Eq. (7–6). The phase will be referred to the time corresponding to the optimum adjustment of the voltages so that the term $2\pi(n - \frac{1}{4})$ may be dropped. Then Eq. (7–6) may be written

$$i_2 = 2I_0J_1(x) (\cos \phi + j \sin \phi) \tag{7–9}$$

Since Y_2 is a negative admittance when ϕ is equal to zero, as defined in the discussion below Eq. (7–5), the complex admittance is

$$Y_2 = \frac{-\beta i_2}{E_1} = \frac{2\beta I_0 J_1(x)}{E_1} (-\cos \phi - j \sin \phi) \tag{7–10}$$

Both components of the admittance are plotted in Fig. 7–2. The conductance, which is the real term in Eq. (7–10), is shown as a solid line, and the susceptance is a dash line. The vertical scale in Fig. 7–2 is purely arbitrary since I_0, $J_1(x)$, and E_1 are unspecified.

7.5. Simplified Theory of Oscillation.—A qualitative analysis of a reflex oscillator may be obtained from inspection of Fig. 7–2. As the phase angle is increased from a negative value to zero, the conductance changes from a positive value, indicating a loss, to a negative value representing a source of power. Oscillation will occur when the negative conductance is equal in magnitude to the conductance of the cavity; *i.e.*, when the source of power is just sufficient to supply the losses in the resonator and the load. The magnitude of the circuit conductance is indicated by the horizontal dotted line in Fig. 7–2. The shaded portion shows the region in which oscillation will occur.

When ϕ is equal to zero, corresponding to the transit time for maximum output, the beam susceptance is zero and the tube will oscillate at the natural frequency of the resonator. Note that the equivalent capacity of the resonator corresponds to a positive susceptance in Eq. (7–4). Decreasing the voltage applied to the klystron will increase the transit time and make ϕ negative. A negative value of $\sin \phi$ in Eq. (7–10) makes the imaginary term positive and introduces an additional positive susceptance in parallel with C, and the frequency of oscillation becomes less than the natural frequency of the resonator. A negative susceptance might be considered a negative capacitance which decreases the effect of C, or it might be viewed as an inductance in parallel with L. Either viewpoint indicates that the oscillation frequency of the system will be increased when ϕ is positive.

7.6. Experimental Verification of Theory.—The value of this analysis can be demonstrated by experimental verification of the theory. If the beam current is kept quite small so that oscillation does not occur, the magnitude of the beam conductance and susceptance components will be sinusoidal, as shown by Fig. 7–2, and the effective Q and resonant frequency of the cavity will vary as the phase of the feedback is changed by varying the reflector voltage. These changes were measured; the results of the experiment are shown in Fig. 7–3 and agree quite closely with the theoretical prediction.

A casual inspection of Figs. 7–2 and 7–3 might suggest that the tuning effect becomes small for large values of the phase angle ϕ near the points where oscillation fails to occur, because the sine function is not changing rapidly. This behavior is correct for the conditions represented by Fig. 7–3, but when the beam current is large enough to maintain oscillation, *i.e.*, when the beam current is much greater than the starting current, the sinusoidal variation of frequency does not occur. Actually, the scale in Fig. 7–2 depends upon the ratio $J_1(x)/E_1$, and this ratio decreases as the strength of oscillation increases. As a result, the tuning effect decreases rapidly as the transit time in the reflection space approaches the value required to make the phase angle ϕ equal to zero, and the frequency deviation is actually porportional to the tangent of the phase angle rather than to the sine. This effect will be apparent from the quantitative analysis that follows.

7.7. Analysis of Reflex-oscillator Characteristics.—When the shunt resistance of a resonator is independent of frequency, the analysis is simplified because the power output and efficiency relations are obtained by

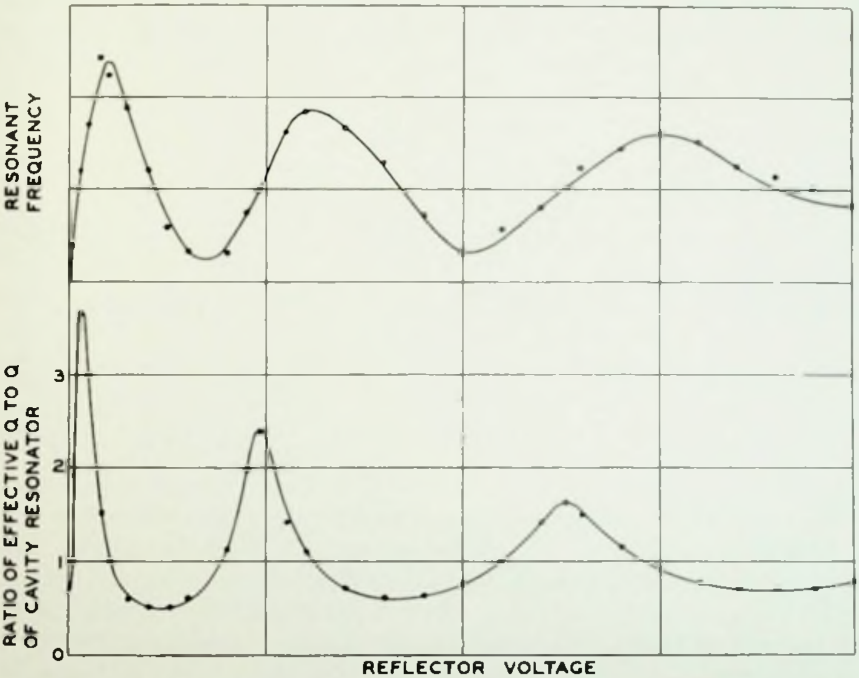


FIG. 7-3.—Experimental curves showing the effect of the electron-beam admittance on the resonant frequency and Q of a resonator.

considering only the conductance component of the beam admittance. After the strength of oscillation has been determined, the frequency of oscillation can be obtained from the magnitude of the beam susceptance. If the beam conductance is greater than the value required to supply the losses in the resonator and its load, the strength of oscillation will increase until the value of the negative beam conductance is reduced to the conductance of the resonator and its load. This means that the conductance of the system is zero when the klystron is oscillating. The sum of the susceptances must also be zero, and this relation determines the frequency of oscillation.

7.8. Starting Current.—The lowest value of beam current I_0 which will allow oscillation to exist is known as the "starting" current. This characteristic is quite important and will be used to illustrate this method of analysis. The sum of the cavity conductance and the beam conductance from Eq. (7–10) must be zero for oscillation to occur.

$$\frac{1}{R_{sL}} - \frac{2\beta I_0 J_1(x)}{E_1} \cos\phi = 0 \tag{7-11}$$

The peak resonator voltage E_1 and x are related, and the analysis is simplified if x is used as the variable. The usual expression for the bunching parameter is

$$x = \beta\pi N \frac{E_1}{E_0} \tag{7-12}$$

and may be rewritten

$$E_1 = \frac{E_0 x}{\beta\pi N} \tag{7-13}$$

Substituting Eq. (7–13) in Eq. (7–11) and rearranging terms gives

$$\frac{x}{2J_1(x)} = \frac{\beta^2\pi N I_0 R_{sL}}{E_0} \cos\phi \tag{7-14}$$

$$\frac{x}{2J_1(x)} = \frac{\beta^2\pi N I_0 R_s R_L}{E_0(R_s + R_L)} \cos\phi \tag{7-15}$$

Weak oscillation corresponds to extremely small resonator voltage, and the bunching parameter x is almost zero under these conditions. The Bessel function $J_1(x)$ is equal to $x/2$ for small values of x; therefore, the left side of the equation will be unity when I_0 is equal to the starting current. The current will be a minimum for the starting conditions only if the phase is correct, *i.e.*, $\cos\phi$ must be a maximum and ϕ is equal to zero, the phase for maximum output. When these conditions are imposed on Eq. (7–14), we obtain an expression for the starting current.

$$I_{\text{start}} = \frac{E_0}{\beta^2\pi N R_{sL}} \tag{7-16}$$

Reasonable values that might be substituted into Eq. (7–16) in order to give some idea of the current required for oscillation are shown below:

$$\begin{aligned} \beta^2 &= 1.0 \\ E_0 &= 300 \text{ volts} \\ N &= 4\tfrac{3}{4} \text{ cycles} \\ R_{sL} &= 20{,}000 \text{ ohms} \end{aligned}$$

Representative values have been chosen and indicate that a beam current of 1 ma will maintain oscillation.

7.9. Reduction of Transadmittance.—If the beam current is increased above the starting value, the transadmittance would normally be increased. However, the klystron will oscillate more strongly and the curvature of the bunching characteristic, shown again in Fig. 7–4, will limit the transadmittance to the value that is required to start oscillation. This condition is in agreement with the general theory of oscillators stated in Sec. 7.1. As a result, any change in the input to the tube or the circuit will cause the strength of oscillation to change until the transadmittance becomes equal in magnitude to the circuit admittance.

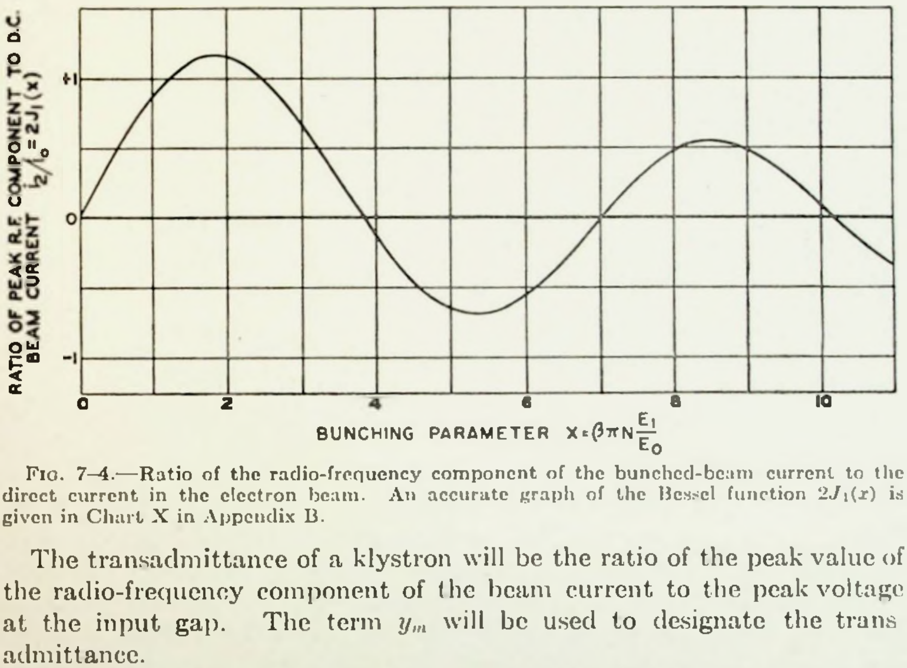


FIG. 7–4.—Ratio of the radio-frequency component of the bunched-beam current to the direct current in the electron beam. An accurate graph of the Bessel function $2J_1(x)$ is given in Chart X in Appendix B.

The transadmittance of a klystron will be the ratio of the peak value of the radio-frequency component of the beam current to the peak voltage at the input gap. The term y_m will be used to designate the trans admittance.

$$y_m = \frac{2\beta I_0 J_1(x)}{E_1} \tag{7–17}$$

Substituting Eq. (7–13) for E_1 in Eq. (7–17) gives

$$y_m = \frac{\beta^2 \pi N I_0}{E_0} \frac{2J_1(x)}{x} \tag{7–18}$$

The small signal transadmittance is defined as the tube transadmittance for small input voltages. Under these conditions the Bessel function $J_1(x)$ is approximately equal to $x/2$. Therefore, the small signal transadmittance y_{ms} is

$$y_{ms} = \frac{\beta^2 \pi N I_0}{E_0} \tag{7–19}$$

The ratio of the small-signal transadmittance to the large-signal transadmittance indicates the reduction that occurs in the transadmittance when the tube oscillates. This basic parameter, which applies to conventional vacuum tubes as well as to velocity modulation types, has been named "transreduction factor."

The term is not limited to the analysis of oscillators but is equally useful in amplifier and other vacuum-tube circuits. When used in an analysis of klystron operation, the value of the parameter has the convenient mathematical equivalent, $x/2J_1(x)$. This result is apparent from Eqs. (7–18) and (7–19).

Since the large-signal transadmittance must be equal to the circuit admittance when the tube is oscillating, the transreduction factor for any oscillator is equal to the ratio of the small-signal transadmittance to the circuit admittance. Owing to the simplification introduced by a single resonator acting as both input and output circuit, it is convenient to treat the real components of the admittances separately. The transreduction factor is also equal to the ratio of the small-signal transconductance g_{ms}, to the circuit conductance $1/R_{sL}$. But the transconductance is the real part of the transadmittance, therefore,

$$g_{ms} = y_{ms} \cos \phi = \frac{\beta^2 \pi N I_0}{E_0} \cos \phi \tag{7–20}$$

This definition of the transreduction factor furnishes another method of deriving Eq. (7–14).

7.10. Power Delivered by the Beam.—Increasing the beam current above the starting current value will greatly increase the output. This can be shown by deriving the expression for the power delivered to the resonator and load. This power will be designated P_2 and is the power delivered by the bunched beam to the shunt resistance R_{sL}. The value of P_2 is given by one-half the product of E_1, the peak resonator voltage, and the peak value of the in-phase component of i_2. This product must be reduced by the beam-coupling coefficient β, in order to include the effect of finite transit time across the resonator gap.

$$P_2 = \tfrac{1}{2} E_1 \beta i_2 \cos \phi = \beta E_1 I_0 J_1(x) \cos \phi \tag{7–21}$$

Substituting the expression for E_1 in Eq. (7–13) in Eq. (7–21),

$$P_2 = \frac{E_0 I_0 \cos \phi}{\pi N} x J_1(x) \tag{7–22}$$

In order to compute P_2, it is necessary to know the dependence of E_1 or x upon the beam current I_0. Equation (7–12) or (7–13) does not furnish this information, but the relation can be obtained indirectly from Eq. (7–14). Values may be substituted in Eqs. (7–14) and (7–15) to obtain the value of the transreduction factor $x/2J_1(x)$ corresponding to the assumed value of the beam current I_0. The relation between the bunching parameter x and $x/2J_1(x)$ can be obtained from a table of Bessel functions, or from Fig. 7–4, which is a curve of $2J_1(x)$ as a function of x. This

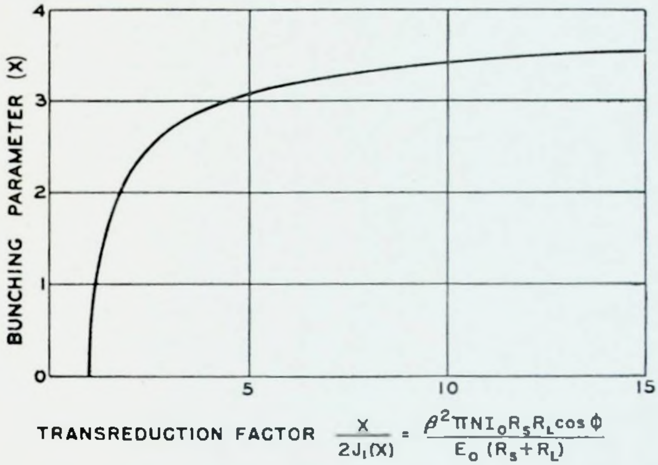


FIG. 7–5.—Bunching parameter x as a function of the beam current and other variables. An accurate graph of x vs. $x/2J_1(x)$ is given in Chart XI in Appendix B.

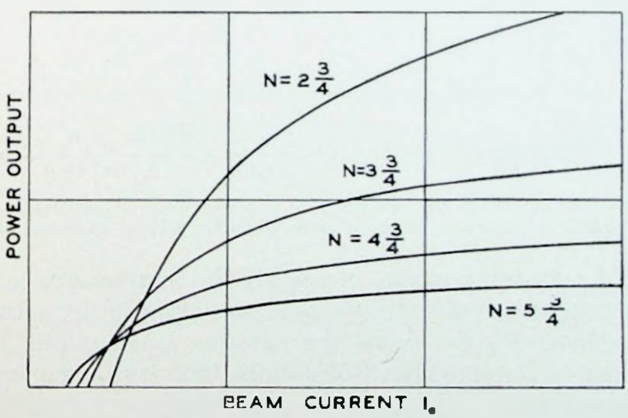


FIG. 7–6.—Theoretical curves of power output as a function of beam current. Several modes corresponding to different transit times are shown.

relation between x and $x/2J_1(x)$ is given in Fig. 7–5 for values of x between zero and 3.83, corresponding to the first zero of the Bessel function. The value of $x/2J_1(x)$ computed from Eqs. (7–14) and (7–15) is used with Fig. 7–5 to obtain values for x and $J_1(x)$, corresponding to the assumed value of the beam current I_0, and the power can then be computed from Eq. (7–22).

Curves of power delivered by the bunched beam as a function of beam current I_0, computed in the manner described above, are shown in Fig. 7–6 for various values of N. These curves not only show the increase of power as the current is increased above the starting value, but also indicate that the maximum power from a reflex oscillator and the starting current are inversely proportional to N, the number of cycles during transit in the

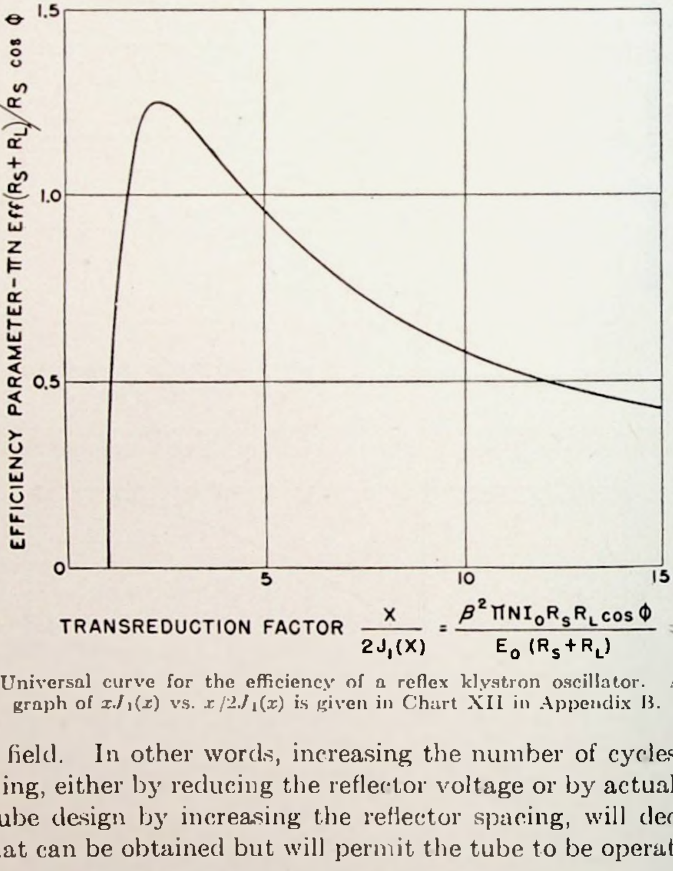


FIG. 7–7.—Universal curve for the efficiency of a reflex klystron oscillator. An accurate graph of $xJ_1(x)$ vs. $x/2J_1(x)$ is given in Chart XII in Appendix B.

reflection field. In other words, increasing the number of cycles required for bunching, either by reducing the reflector voltage or by actually changing the tube design by increasing the reflector spacing, will decrease the output that can be obtained but will permit the tube to be operated with a smaller beam current.

7.11. Universal Efficiency Curve.—If any of the variables other than the beam current are changed, such as the load resistance or the phase angle ϕ, the use of curves to show the effect of each variable becomes quite complicated. Fortunately, all the variables can be combined into dimensionless parameters and the characteristics can be presented in a universal curve as illustrated by Fig. 7–7. The transreduction factor $x/2J_1(x)$ in Eq. (7–14) and Sec. 7.9 is one example of a useful dimensionless parameter, and the efficiency parameter to be derived below is another example.

The power delivered by the bunched beam, defined by Eq. (7–22), is not all useful power since some is absorbed by the resonator losses. We are more interested in the power delivered to the load, which will be designated P_L. Then

$$P_L = \frac{R_s}{R_L + R_s} P_2 = \frac{R_s E_0 I_0 \cos \phi}{\pi N (R_L + R_s)} x J_1(x) \tag{7–23}$$

If we divide the power output by the beam power input, we obtain the efficiency of the klystron oscillator. Equation (7–23) can be rearranged so that the efficiency, abbreviated eff., and the other factors involved are related to a dimensionless efficiency parameter $xJ_1(x)$.

$$xJ_1(x) = \frac{\pi N}{\cos \phi} \frac{R_L + R_s}{R_s} \text{ eff.} \tag{7–24}$$

Figure 7–7 combines these two dimensionless parameters in a single curve which relates the output characteristics of a reflex klystron oscillator to the design factors that may be varied. The vertical coordinate is $xJ_1(x)$ and $x/2J_1(x)$ is the horizontal coordinate.

7.12. Effect of Voltage on Klystron Output.—Most of the output characteristics that are typical of reflex klystron oscillators can be predicted by inspection of Fig. 7–7. Consider the case when the load, beam current, and acceleration voltage remain fixed, but the reflector voltage is varied. Assume that the phase angle ϕ is $-\pi/2$ for zero reflector voltage, *i.e.*, when the reflector electrode is at cathode potential. cos ϕ will be zero, corresponding to an operating point at the origin in Fig. 7–7. Increasing the negative reflector voltage will increase ϕ, and cos ϕ will vary from zero to a maximum of unity and then decrease again. The value of N will also vary, but if N is large this variation is not important in a qualitative analysis, and N will be assumed a constant for the range of each voltage mode.

When cos ϕ is zero, the transreduction factor $x/2J_1(x)$ is also zero, since the value of $x/2J_1(x)$ is determined by Eq. (7–14) or (7–15).

$$\frac{x}{2J_1(x)} = \frac{\beta^2 \pi N I_0 R_{sL}}{E_0} \cos \phi \tag{7–14}$$

Oscillation will not occur until cosϕ has increased until the value of $x/2J_1(x)$ is unity. As cos ϕ increases beyond this point, the output will increase as shown by Fig. 7–7. When cos ϕ is unity, $x/2J_1(x)$ will have its maximum value and the output will also be maximum. This is true for the region when the efficiency curve is decreasing because the cos ϕ term increases faster than the efficiency parameter in Fig. 7–7 decreases. As the reflector voltage is increased beyond the value giving maximum output, the phase angle becomes negative, and cos ϕ decreases until the output is again zero.

7.13. Reflector Voltage Modes.—As the reflector voltage is increased further, the sign of cos ϕ will become negative and the beam conductance term in Eq. (7–10) has a positive value. This positive beam conductance represents an additional loss; therefore, oscillation does not occur. When the transit time has changed by an amount equivalent to one complete cycle, the phase is again correct for oscillation and another output mode

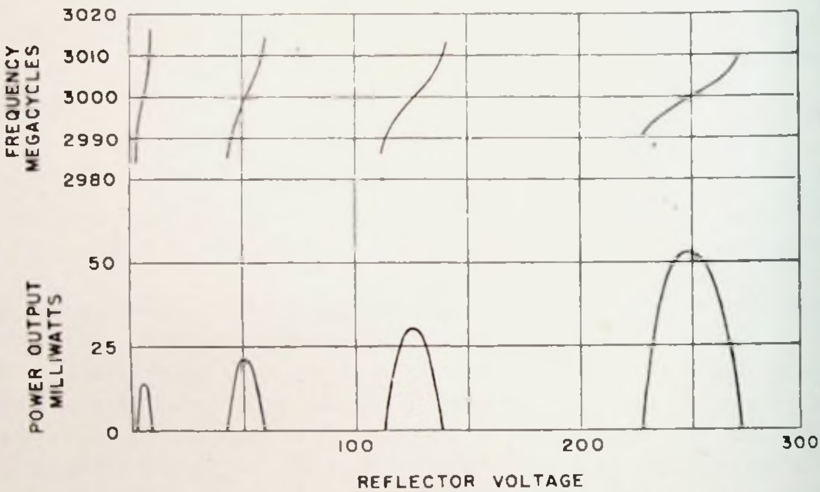


FIG. 7–8.—Power output and frequency characteristics when the reflector voltage of a reflex klystron is varied.

will occur. Normally, there are several of these voltage modes in the normal range of adjustment of the reflector voltage, and oscillation does not occur in the region between modes where the phase angle is incorrect. This behavoir is illustrated in Fig. 7–8.

The higher reflector voltage modes correspond to smaller values of N and the output is greater for two reasons: (1) the ordinate $xJ_1(x)$ in Fig. 7–7 becomes greater as N is decreased, since decreasing N corresponds to moving from right to left on the curve in Fig. 7–7; (2) the efficiency for a particular value on the curve is inversely proportional to N. Eventually it is no longer possible to observe modes with higher reflector voltage because N has become so low that the starting current is greater than the beam current. The last mode observed may have the highest output of the series, or it may have less output than the previous mode. The latter case corresponds to a point in Fig. 7–7 to the left of the maximum of the curve.

7.14. Theoretical Efficiency.—The theoretical efficiency of a reflex klystron oscillator is less than the value for a double resonator oscillator and is inversely proportional to N. The efficiency for any value of N can be calculated from Fig. 7–7. If most of the power is transferred to the load and the phase angle is adjusted for maximum output, then Eq. (7–24) may be rewritten

$$\text{Theoretical efficiency} = \frac{xJ_1(x)}{\pi N} = \frac{1.25}{\pi N} \qquad (7\text{–}25)$$

The assumptions used in this derivation are not valid for small values of N, and theoretical efficiencies between 20 and 30 per cent are indicated when better approximations are made in the computation of efficiency for values of N less than two.

It is interesting to note that the efficiency obtainable for any mode is independent of the beam-coupling coefficient. If the transit time across the resonator gap is large, making the value of β less than unity, then it is theoretically possible to overcome this disadvantage by increasing the beam current. The power output will be greater because the same maximum efficiency requires more power input. If sufficient beam current is available so that the load resistance R_L is small in comparison with the shunt resistance of the resonator R_s, the effect of a small value of β may be counteracted by decreasing the load, *i.e.*, by increasing the value of the shunt load resistance R_L.

7.15. Load Variations.—If the output load impedance is varied (by varying the length of the output line or some other method of impedance transformation) the output will increase to a maximum, then decrease suddenly, and the klystron may refuse to oscillate for certain load impedances. This effect occurs first for the higher reflector voltage modes because the starting current is higher for these modes. When the beam current is constant, the load required for maximum output is different for each mode. Heavier loading is required for maximum output from the modes corresponding to the larger values of N.

This effect can be demonstrated conveniently with a dynamic method of observing the output. An alternating voltage can be superimposed upon the reflector voltage, causing the output to be swept through several modes periodically. The output voltage is applied to a cathode-ray oscilloscope with the sweep synchronized with the reflector voltage modulation. A pattern similar to Fig. 7–8 will be observed. If the klystron is lightly loaded, all the modes will be small, but the output of all modes will increase as the load is increased until the highest reflector voltage mode with the smallest value of N corresponds to the point of maximum efficiency on Fig. 7–7. Increasing the load further will decrease the output from the highest voltage mode until it disappears when the transreduction factor becomes less than unity. The other modes with larger values of N will continue to increase in output, with the modes disappearing successively until the load is so great that the klystron cannot oscillate at any reflector voltage.

7.16. Electronic Tuning.—The qualitative analysis based on Fig. 7-2 predicted that the frequency of oscillation would change as the phase of the bunched beam was varied by changing the acceleration voltage or the reflector voltage. This effect is known as "electronic tuning." The power output and efficiency relationships were obtained by considering only the conductance components of the beam and cavity admittances. Similarly, the electronic tuning analysis requires that the sum of the susceptances is zero. The magnitude of the beam susceptance depends upon the strength of oscillation, however. As a result, the imaginary component of the beam admittance depends upon the magnitude of the real component.

Equation (7-11) may be rewritten

$$\frac{2\beta I_0 J_1(x)}{E_1} = \frac{1}{R_{sL}\cos\phi} \tag{7-26}$$

Then Eq. (7-26) may be substituted in the imaginary term of Eq. (7-10) to obtain the value of the beam susceptance in terms of the phase angle ϕ.

$$-\frac{2\beta I_0 J_1(x)}{E_1}\sin\phi = \frac{-\sin\phi}{R_{sL}\cos\phi} = -\frac{\tan\phi}{R_{sL}} \tag{7-27}$$

Equating all susceptance terms in the resonator and beam admittance to zero gives an expression that may be used to determine the frequency of oscillation.

$$-\frac{1}{\omega L} + \omega C - \frac{\tan\phi}{R_{sL}} = 0 \tag{7-28}$$

Rearranging terms gives

$$(\omega^2 LC - 1)\frac{R_{sL}}{\omega L} = \tan\phi \tag{7-29}$$

If the angular frequency of oscillation for a zero value of the phase angle ϕ is represented by ω_0, then LC is equal to $1/\omega_0^2$. The loaded Q of the resonator Q_L is equal to $R_{sL}/\omega L$; therefore,

$$\left(\frac{\omega^2}{\omega_0^2} - 1\right)Q_L = \tan\phi \tag{7-30}$$

When ω and ω_0 do not differ by more than a few per cent, $(\omega^2/\omega_0^2) - 1$ may be rewritten

$$\left(\frac{\omega^2}{\omega_0^2} - 1\right) = 2\frac{\omega - \omega_0}{\omega_0} = 2\frac{\Delta f}{f_0} \tag{7-31}$$

and Eq. (7–30) becomes the familiar expression for the phase of a parallel resonant circuit.

$$2Q_L \frac{\Delta f}{f_0} = \tan \phi \qquad (7\text{–}32)$$

The term $2Q_L(\Delta f/f_0)$ is a convenient frequency deviation parameter which

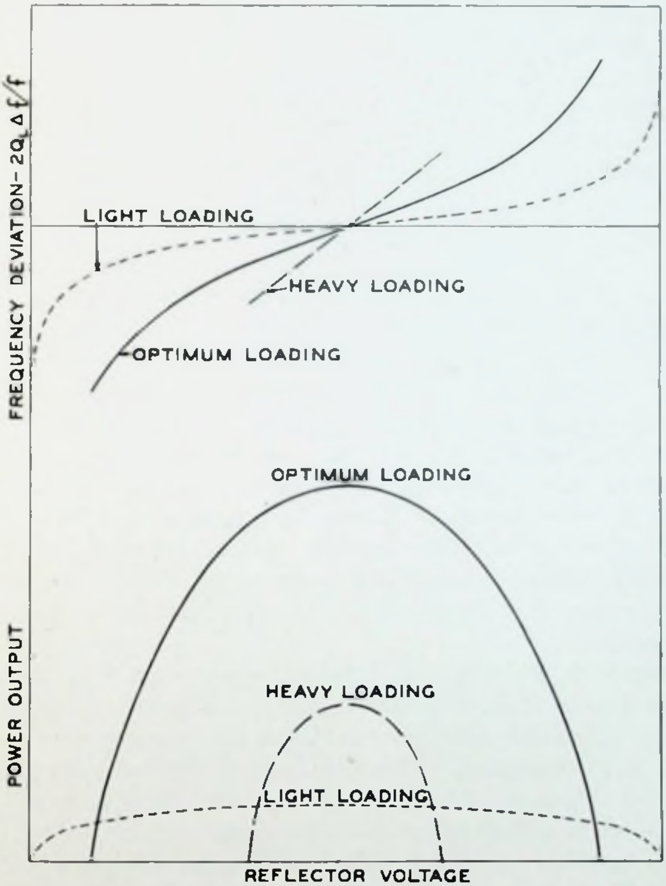


FIG. 7–9.—Power output and frequency characteristics for different loads.

is often used in universal curves for resonant circuits. It relates the actual frequency deviation to the loaded Q of the circuit.

7.17. Output and Frequency Characteristics.—Equation (7–23) and Fig. 7–7 allow the power output to be calculated as a function of the phase angle ϕ, and the frequency deviation from the resonant frequency of the cavity can be obtained from Eq. (7–32). However, it is more useful to know these characteristics as a function of voltage instead of phase. Equation (6–10), repeated below,

$$N = 4fs_0 \frac{\sqrt{(m/2e)E_0}}{E_0 + E_r} \tag{7-33}$$

may be substituted into Eq. (7–7) to obtain a value of ϕ, and this value of ϕ may then be substituted into Eqs. (7–23) and (7–32), giving the output power and frequency characteristics as a function of reflector voltage. Figure 7–2 was obtained in this manner.

Figure 7–9 repeats the characteristics shown in Fig. 7–2, for a single mode and a number of different values of loaded Q. The curves for heavy loading correspond to a load that is almost great enough to prevent oscillation. Curves are also shown for the loading that gives maximum output, and very light loading when most of the power is absorbed by the resonator losses.

A number of interesting conclusions are illustrated by Fig. 7–9. The slope of the linear portion of the frequency characteristic is inversely proportional to the loaded Q of the resonator. This fact is apparent from Eq. (7–32). Increasing the Q by decreasing the load does not decrease the electronic tuning band width as might be expected, since this change will increase the bunching, and the phase angle may be varied over a larger range before the output decreases appreciably. The band width between zero output points actually increases as the loading is decreased, and the band width between half-power points is only decreased slightly. Decreased loading causes the amplitude characteristic to become more uniform over a large range of voltage, but the frequency deviation curve becomes quite nonlinear.

7.18. Frequency Stability.—Although the curves in Figs. 7–8 and 7–9 indicate the nature of the frequency characteristic satisfactorily, they do not furnish a means of obtaining values for the frequency stability under all conditions of operation. This information can be obtained from an expression for the slope of the frequency curve. The term $\Delta f/f_0$ in Eq. (7–32) does not give this information because it represents the difference in frequency between two points; it is therefore the slope of a line connecting these points and not the slope of the curve.

The slope of the curve may be obtained by rewriting Eq. (7–32) and differentiating this equation.

$$2Q_L \frac{f - f_0}{f_0} = \tan \phi \tag{7-34}$$

$$2Q_L \frac{df}{f_0} = \sec^2 \phi \, d\phi \tag{7-35}$$

An expression for $d\phi$ may be obtained by substituting Eq. (7–33) in Eq. (7–7) and differentiating the result.

$$\phi = 2\pi\left(n - \frac{1}{4}\right) - 4fs_0 \frac{\sqrt{(m/2e)E_0}}{E_0 + E_r} 2\pi \tag{7-36}$$

The differentiation will be carried out first for the case when the beam voltage is fixed and the reflector voltage is varied. The frequency can be considered a constant.

$$d\phi = 4fs_0 \frac{\sqrt{(m/2e)E_0}}{(E_0 + E_r)^2} 2\pi dE_r = 2\pi N \frac{dE_r}{E_0 + E_r} \tag{7-37}$$

Substituting Eq. (7–37) in Eq. (7–35) and rearranging terms gives

$$\frac{df}{f_0} = \frac{\pi N}{Q_L} \frac{dE_r}{E_0 + E_r} \sec^2 \phi \tag{7-38}$$

A similar computation gives the result for a varying beam voltage.

$$\frac{df}{f_0} = \frac{\pi N}{2Q_L} \frac{E_0 - E_r}{E_0 + E_r} \frac{dE_0}{E_0} \sec^2 \phi \tag{7-39}$$

Equations (7–38) and (7–39) may also be written in terms of the frequency change per volt. These relations become

$$\frac{df}{dE_r} = \frac{\pi N f_0}{Q_L} \frac{\sec^2 \phi}{E_0 + E_r} \tag{7-40}$$

$$\frac{df}{dE_0} = \frac{\pi N f_0}{2Q_L} \frac{E_0 - E_r}{E_0 + E_r} \frac{\sec^2 \phi}{E_0} \tag{7-41}$$

Note that a reflex klystron is more stable with respect to variations of the beam voltage E_0, also that the frequency is independent of the beam voltage if the reflector voltage and the beam voltage are equal. The latter result was also obvious in Fig. 6–3. Equation (7–41) neglects the tuning effect of the electrons in the gap which may change the resonant frequency of the cavity when the beam current varies owing to changes in the beam voltage.

7.19. Electronic Tuning Band Widths.—The qualitative conclusions in Sec. 7.17 regarding the electronic tuning band width are interesting, but a method of calculating the band widths is more valuable. The desired equations may be obtained by evaluating the phase angle ϕ, for the output being considered, and substituting this value of ϕ in Eq. (7–32). This process will be carried out for the zero-power point and also the half-power point. Equation (7–14) may be rewritten

$$\cos \phi = \frac{E_0(R_s + R_L)}{\beta^2 \pi N I_0 R_s R_L} \frac{x}{2J_1(x)} \tag{7-42}$$

For zero output, the value of $x/2J_1(x)$ is unity; therefore,

$$\cos \phi_0 = \frac{E_0(R_s + R_L)}{\beta^2 \pi N I_0 R_s R_L} \tag{7-43}$$

and

$$\tan \phi_0 = \sqrt{\frac{1}{\cos^2 \phi_0} - 1} \tag{7-44}$$

Note that $\cos \phi_0$, *i.e.*, the cosine of the phase angle when the output is zero, has a value equal to the reciprocal of the transreduction factor $x/2J_1(x)$, for the operating conditions when the phase angle is zero, corresponding to maximum output. Therefore, Eq. (7–44) may be rewritten

$$\tan \phi_0 = \sqrt{\left[\frac{x}{2J_1(x)}\right]^2 - 1} \tag{7-45}$$

The band width between zero-output points is obtained by substituting Eq. (7–44) in Eq. (7–32). However, the frequency deviation $\Delta f/f_0$ is measured from the point of maximum output; therefore, the band width between zero-output points will be twice the value indicated by Eq. (7–32). The term $(2\Delta f/f)_0$ will be introduced to avoid confusion between the band width between the two zero-output points and the frequency deviation from the frequency corresponding to maximum output. Then

$$2Q_L\left(\frac{2\Delta f}{f}\right)_0 = 2 \tan \phi_0 = 2\sqrt{\frac{1}{\cos^2 \phi_0} - 1} \tag{7-46}$$

7.20. Band Width between Half-power Points.—Evaluation of the band width between half-power points is somewhat more complicated and requires the determination of the bunching parameter value which corresponds to one-half the maximum output. The power output for any operating condition is the square of the peak voltage E_1, divided by twice the load resistance R_L.

$$P_L = \frac{E_1^2}{2R_L} = \frac{E_0^2\, x^2}{2\beta^2 \pi^2 N^2 R_L} \tag{7-47}$$

Equation (7–13) has been substituted for E_1 in Eq. (7–47). The value of the bunching parameter x for maximum output can be obtained from Fig. 7–5 with $\cos \phi$ equal to unity. This maximum output does not necessarily correspond to the point of optimum efficiency in Fig. 7–7 but is the maximum output for the given conditions of load and input when the phase angle is zero. These conditions determine the value of $x/2J_1(x)$, and x is then determined from Fig. 7–5. This value of x divided by $\sqrt{2}$ is the value of the bunching parameter which corresponds to the half-power points. Substituting this value of the bunching parameter in Eq. (7–42) gives

$$\cos \phi_{\frac{1}{2}} = \frac{E_0(R_s + R_L)}{\beta^2 \pi N I_0 R_s R_L} \frac{x/\sqrt{2}}{2J_1(x/\sqrt{2})} \tag{7–48}$$

Equation (7–48) may also be written

$$\cos \phi_{\frac{1}{2}} = \frac{2J_1(x)}{x} \frac{x/\sqrt{2}}{2J_1(x/\sqrt{2})} \tag{7–49}$$

A definition for the band width between half-power points, similar to the definition for zero-output conditions, gives

$$2Q_L \left(\frac{2\Delta f}{f}\right)_{\frac{1}{2}} = 2 \tan \phi_{\frac{1}{2}} = 2\sqrt{\frac{1}{\cos^2 \phi_{\frac{1}{2}}} - 1} \tag{7–50}$$

These expressions may appear complicated, but the evaluation of cos $\phi_{\frac{1}{2}}$ from Fig. 7–5 is quite simple. The method can be illustrated by a sample calculation. Assume that $x/2J_1(x)$ equal to 2.30 corresponds to the operating conditions when the phase angle is zero. This corresponds to maximum output from the tube. The bunching parameter x for this value of $x/2J_1(x)$ is 2.40, as indicated by the curve in Fig. 7–5. The value of x for the half-power point would be $2.40/\sqrt{2}$ or 1.70, and corresponds to $x/2J_1(x)$ equal to 1.47. $\cos \phi_{\frac{1}{2}}$ is then 1.47/2.30, or 0.64. Substitution of this value of $\cos \phi_{\frac{1}{2}}$ in Eq. (7–42) gives a value of 2.40 for $2Q_L\ (2\Delta f/f)_{\frac{1}{2}}$.

7.21. Electronic Tuning Curves.—The calculations for band widths between zero output and half-power points have been made, and the results are plotted in Fig. 7–10 as a function of $x/2J_1(x)$, the transreduction factor. A dotted line has been drawn through the origin and tangent to the curve for the band width between half-power points. Since Q_L is proportional to $R_sR_L/(R_s + R_L)$, this dotted line is proportional to Q_L and $(2\Delta f/f)_{\frac{1}{2}}$ will be a maximum at the point of tangency. In other words, the maximum band width between half-power points occurs when the conductance parameter has a value of approximately 2.30, the same as the value required for optimum output from the tube.

It is interesting to note that the band width between half-power points for a single resonant circuit is 2.00 when these coordinates are used for the frequency deviation. The value for a reflex klystron oscillator with the load adjusted for maximum electronic tuning is 2.40, or 20 per cent greater than the band width associated with the loaded Q of the resonator. Increasing the bunching by increasing the beam current, decreasing the loading, or in any other manner which increases the value of the transreduction factor, will increase the value of $2Q_L(2\Delta f/f)_{\frac{1}{2}}$. However, it is not correct to state that the electronic tuning of a reflex klystron is independent of the loaded Q of the resonator. The frequency deviation in the linear region is inversely proportional to Q_L, but increasing Q_L by reducing the load causes overbunching and the tube can oscillate over a wider range of voltage variation. As a result, the half-power point is extended into the nonlinear region of the frequency deviation characteristic, and the actual frequency band width $(2\Delta f/f)_{1/2}$ decreases only slightly from the maximum band width when the oscillator is loaded to give maximum output.

7.22. Rieke Diagrams.—Some of the effects of varying the resistive component of the load have been mentioned in the discussion of Figs. 7–7 and 7–9. The load may also have a reactive component, and the

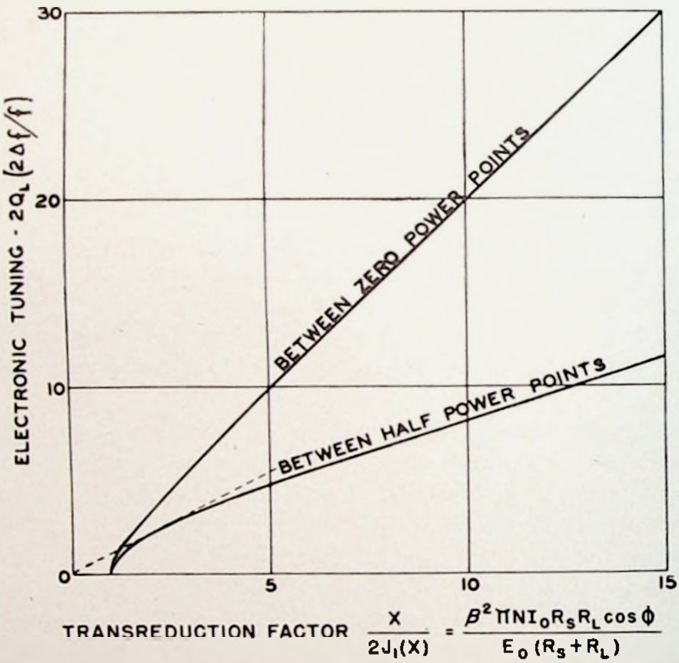


FIG. 7–10.—Universal curves for the electronic tuning band widths of a reflex oscillator.

complete picture can be given only by a Rieke diagram similar to the diagrams used in Secs. 4.14 to 4.16 to show the load characteristics of klystron amplifiers. The equivalent load resistance R_L has a magnitude similar to the shunt resistance of the cavity resonator; *i.e.*, R_L is usually several thousand ohms. The characteristic impedance of the coaxial output line is very much smaller, usually 100 ohms or less for convenient physical dimensions, and the coupling loop must be designed to transform an impedance of perhaps 100 ohms to the required value of several thousand ohms. Rieke diagrams are usually plotted on a scale based on the characteristic impedance of the line.

An admittance diagram will be used in order to represent the load by shunt elements. Zero susceptance then corresponds to a purely resistive load. Normalized coordinates will be used; *i.e.*, unity conductance corresponds to the characteristic impedance of the line; therefore, unity on the conductance axis of the Rieke diagram corresponds to unity standing wave ratio.

A klystron oscillator readjusts its frequency if a reactive component is introduced in the load, since the effect of the parallel reactance is the same as retuning the resonator. For this reason, the contours of constant power output from a reflex oscillator would be expected to be along lines of constant load conductance similar to Fig. 4–11 in Sec. 4.15 for an ampli-

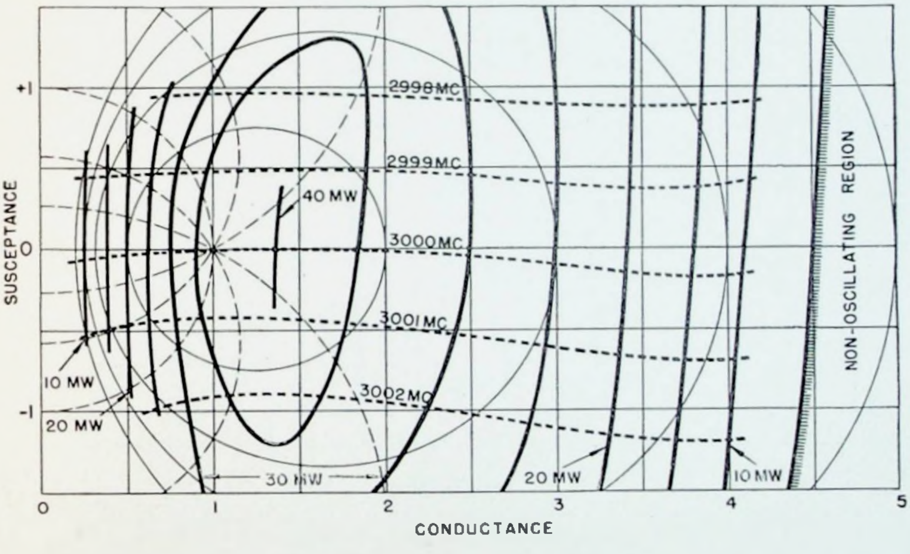


FIG. 7–11.—Rieke diagram for a reflex klystron when the reflector voltage is adjusted for optimum output.

fier with the output resonator retuned. However, losses in the output line and the effect of a line that may be several wavelengths long will distort this ideal pattern.

The frequency changes that accompany load changes are indicated by contours of constant frequency on the Rieke diagram for a reflex oscillator.

7.23. Rieke Diagram for Optimum Phase.—The Rieke diagram for a typical reflex klystron is shown in Fig. 7–11. The reflector voltage has been adjusted to make the phase angle ϕ equal to zero. Note that there is a region where the conductance is large, and the klystron cannot oscillate in this region. This load corresponds to the region below unity in Fig. 7–7 where the beam current is less than the starting current for such a heavy load.

If a variable length of line is used as an impedance transformer, the resistive component of the load can be varied (if there are standing waves in the line), but a reactive component will also be introduced. This reactive component will affect the frequency of oscillation. The analysis of this effect will not be considered in detail, but the importance of the effect is apparent from an inspection of the Rieke diagram. Although a variable length line or trombone is one of the most convenient forms of impedance transformer for an experimental setup, such a device does have a considerable "pulling" effect on the frequency of an oscillator.

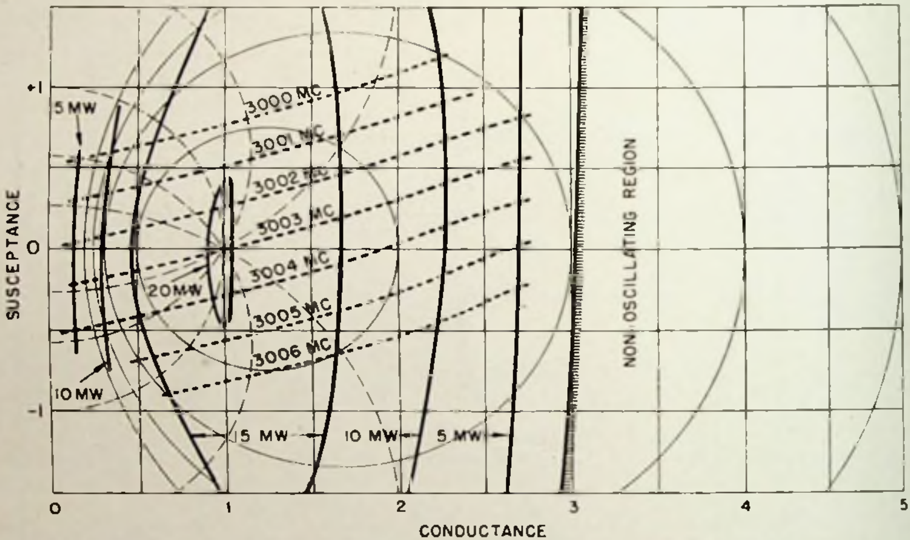


FIG. 7–12.—Rieke diagram for a reflex klystron when the reflector voltage is greater than the value giving optimum output.

7.24. Rieke Diagrams for Incorrect Phase.—If the reflector voltage is adjusted to a value that does not give the optimum phase, either by accident or in order to obtain an electronic tuning effect, the Rieke diagram for the klystron oscillator is considerably different. The case when the reflector voltage is greater than the optimum value is shown in Fig. 7–12. The frequency contours are spaced more closely. Also, the frequency contours are no longer horizontal.

One very important conclusion can be gained from these data. Electronic tuning not only requires a tube with a low Q, which indicates that electronic tuning is merely a form of frequency instability, but the frequency instability due to changes in the load becomes greater when the reflector voltage is changed to produce the electronic tuning. These considerations do not make electronic tuning an undesirable feature, since many applications are simplified considerably by the ability to control the frequency electronically. This feature does introduce certain limitations that should not be overlooked.

Another interesting fact can be observed in Fig. 7–12. Changing the value of a purely resistive load will cause the frequency to change when the reflector is not adjusted for optimum phase. This effect may be noted in Fig. 7–9 when the reflector voltage does not correspond to the adjustment for maximum output and the frequency deviation is not zero. Decreasing the load will decrease the frequency deviation. This effect is also indicated by the magnitude of the beam susceptance in Eq. (7–28). Decreasing the load corresponds to increasing the load resistance R_L, and this change also increases the effective shunt resistance R_{sL}; therefore, decreasing the load will decrease the effective beam susceptance and the frequency deviation will be less. This effect becomes greater when the reflector voltage deviates more from the value required for maximum output.

7.25. Power Losses in the Resonator.—If most of the power is not transferred to the load, then the derivation of the maximum efficiency in Eq. (7–25) does not apply, and the efficiency is dependent upon the load resistance. Actually, R_L must be small compared to R_s if most of the power is to be transferred to the load, and this condition can be obtained only if the beam current available is very much larger than the starting current. The maximum efficiency is less than the theoretical value for practical values of beam current. If the beam current is seven times greater than the starting current, the maximum value of the $x/2J_1(x)$ coordinate in Fig. 7–7 will be 7.0 when R_L is infinite, corresponding to no load. The output will be zero under these conditions and the efficiency will also be zero, since $(R_L + R_s)/R_s$ becomes infinite. As R_L is decreased corresponding to increasing the load, the output will increase.

A family of curves similar to Fig. 7–7 can be plotted to show the effect of power division between the resonator losses and the load. The factor $(R_L + R_s)/R_s$ in the ordinate of Fig. 7–7 is computed for each value of R_L considered, and the ordinates for the revised efficiency curves in Fig. 7–13 are directly proportional to the output efficiency. Each curve corresponds to some chosen value of beam current I_0, and πN times the efficiency is plotted as a function of R_L. The other variables in the transreduction factor are held constant. The phase angle ϕ has been assumed to be zero in this illustration, corresponding to the voltage adjustment for maximum output; therefore, $\cos \phi$ is unity and has not been included in the efficiency coordinate.

If the beam current I_0 were equal to the starting current I_s, the transreduction factor $x/2J_1(x)$ would have a value of unity. The load would be zero, corresponding to an infinite value of R_L. When I_0 is seven times greater than I_s, the value of $x/2J_1(x)$ would be 7.0 if the load resistance

R_L was infinite. The output would be zero, of course. Decreasing R_L would increase the load, and the efficiency would increase until a maximum was reached. Eventually the load would become too great and the tube would fail to oscillate when R_L was reduced until $x/2J_1(x)$ has a value of unity. Similar curves are shown for values of I_0 twelve and twenty times

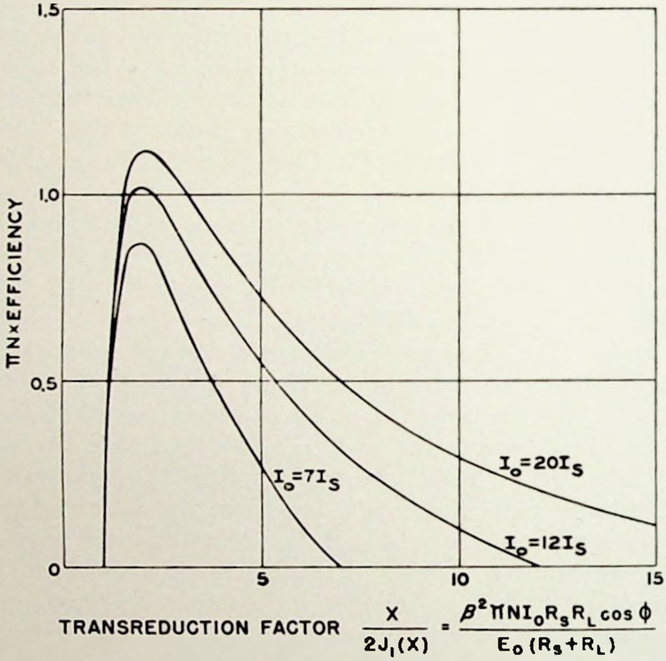


FIG. 7–13.—Efficiency of a reflex oscillator as a function of load. Curves for three values of beam current are shown.

greater than the starting current. Note that the actual efficiency is only 90 per cent of the theoretical efficiency when the beam current is twenty times greater than the starting current.

7.26. Effect of Transit Time on Efficiency.—Figure 7–13 may also be used to compare the efficiencies for different values of N when the beam current remains constant. These conditions can be met by changing the reflector voltage. Consider that the curve in Fig. 7–13 for seven times the starting current corresponds to a value of N equal to 2¾ cycles, and the curve for twelve times the starting current corresponds to the same beam current but a value of 4¾ cycles for N. Then the actual efficiency for optimum loading would be $0.87/2.75\pi$ or 10.1 per cent for N equal to 2¾ cycles, and $1.01/4.75\pi$ or 6.8 per cent for N equal to 4¾ cycles. Although the loading required for maximum output is less for the mode with the shorter transit time, and therefore a larger proportion of the total power is dissipated in the resonator losses, the improved conversion efficiency for the shorter transit time allows the output efficiency to be greater.

7.27. Reflex Klystron Design Considerations.—Most of the previous discussion has been used to predict or explain the electrical characteristics of reflex klystron oscillators when operating voltages, current, and loading were the only variables. It is interesting to consider the effect of varying the design of the tube itself, although it is necessary to remember that the relation between the lumped constants used in the equivalent circuit and the physical dimensions of the cavity resonator is not clearly defined. However, considering the effect of changing these constants can be quite useful in a qualitative analysis of the factors which are important in the design of klystrons.

Reference to the equivalent circuit in Fig. 7–1 will indicate that increasing the ratio of the small-signal beam admittance to the circuit capacitance will increase the amount of electronic tuning. This ratio may be increased by increasing the beam current I_0, by increasing the transit time in the reflection space (increasing the value of N), or by decreasing the circuit capacitance. Decreasing the capacitance by increasing the resonator gap spacing may not be satisfactory because the transit time across the gap may become excessive. This change would reduce the beam-coupling coefficient, which has the same effect as reducing the beam current. Therefore we shall consider only reducing the capacitance by decreasing the area of the resonator gap.

Either increasing the beam current without changing the capacitance, or reducing the area of the gap without changing the current, corresponds to increasing the current density. Therefore the problem of increasing the electronic tuning in a klystron design becomes a problem of increasing the current density. This conclusion assumes that N is already large and that additional transit time in the reflection space will not increase N appreciably.

7.28. Effect of Changing Q.—It is equally interesting to analyze the factors affecting electronic tuning from the viewpoint that increased electron bunching permits heavier loading of the oscillator, and therefore increases the electronic tuning because the loaded Q has been reduced. Reference to Figs. 7–5 and 7–7 will emphasize the fact that the bunching parameter x has a value of 2.40 when the oscillator is adjusted for maximum output. If the beam current is increased, with no design change in the resonator, the resonator voltage E_1 will be increased and the value of the bunching parameter will increase. The magnitude of E_1 is determined by the radio-frequency current i_2 and the loaded shunt resistance R_{sL}.

$$E_1 = 2I_0 \, R_{sL} \, J_1(x) \tag{7–51}$$

Since E_1 must be constant if x remains constant, an increase in I_0 must be accompanied by a decrease in the loaded shunt resistance R_{sL}. Therefore the increased beam current permits the oscillator to be operated with a greater load, and reducing the Q of the loaded circuit increases the electronic tuning.

The effect of decreasing the capacitance may also be related to the loaded Q of the resonator. One of the relations giving the Q of a circuit is

$$Q_L = \omega C R_{sL} = \omega C \frac{R_s R_L}{R_s + R_L} \tag{7-52}$$

The unloaded Q of the circuit will be

$$Q = \omega C R_s \tag{7-53}$$

Therefore Eq. (7–52) may be rewritten

$$Q_L = Q \frac{R_L}{R_s + R_L} \tag{7-54}$$

Decreasing the circuit capacitance by reducing the resonator gap area without changing the gap spacing does not change the unloaded Q appreciably, but does increase the shunt resistance R_s. This change will not affect the loaded shunt resistance R_{sL}, since R_L is usually much smaller than R_s; therefore, the oscillator will operate with the same degree of bunching if the beam current and the load resistance R_L are unchanged. However, Eq. (7–54) indicates that the loaded Q will decrease when the shunt resistance is increased, and the electronic tuning will be increased.

Note that the changes discussed in all the preceding sections correspond to increasing the current density in the electron beam. The various explanations of the electronic tuning are merely different ways of looking at the problem.

7.29. High-power Reflex Klystrons.—The design of an efficient, high-power reflex klystron oscillator would require a different approach. The important design factor would be the transit time in the reflection space; therefore N must be small. As pointed out in the discussion of Eq. (7–25), the analysis is not valid for small values of N, but the trend is indicated correctly. Decreasing N increases the starting current and, if the beam current is already as large as permitted by a practical design, then the load required for optimum output cannot be very great and the electronic tuning will be small. It is also apparent that the theoretical efficiency will not be attained if a large part of the total power goes into the resonator losses. In spite of this factor, however, the efficiency will be greater than that of a reflex klystron designed for a larger value of N. If it were possible to increase the beam current sufficiently so that most of the power could be transferred to the load, then the klystron would have as much electronic tuning as a design with a larger value of N and smaller beam current.

CHAPTER 8

TWO-RESONATOR OSCILLATORS

8.1. Oscillator Circuits.—Although a two-resonator klystron with an external (or internal) feedback line might appear to be a simple oscillator circuit because it resembles a tuned plate–tuned grid oscillator superficially, actually the analysis of this oscillator is more complicated than that of a

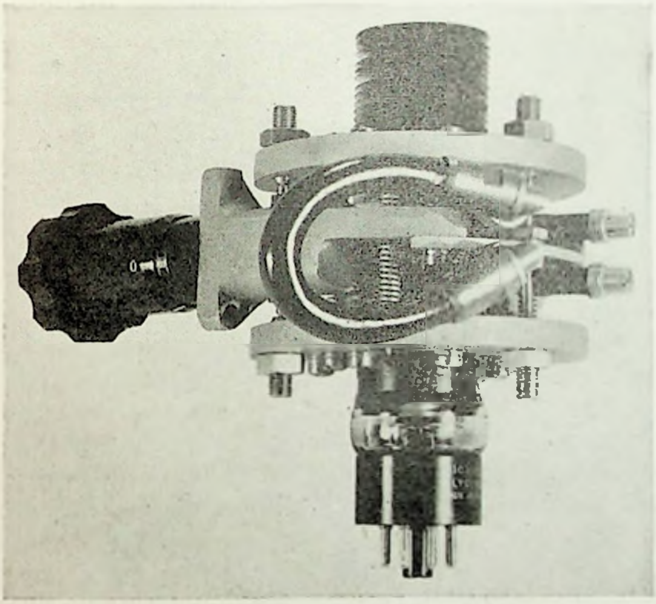


FIG. 8–1.—Type 3K30/410R klystron oscillator mounted in a tuner.

reflex oscillator because the theory of coupled circuits is involved. The type of tube to be discussed is a two-resonator klystron of the construction illustrated in Fig. 3–1, Sec. 3.4, with a field-free drift space separating the input and output resonators. Feedback is obtained by adding a coaxial line from the output resonator to the input resonator. Figure 8–1 shows such a tube complete with its coaxial feedback line and a mechanical tuner for changing the spacing of the input and output gaps.

8.2. Oscillator Theory.—It will be more convenient to treat this type of oscillator by considering the transfer impedance of the coupled-circuit system as a function of the phase angle, since the usual sources of data[1] present the information in this form. The conditions for oscillations in this terminology become: The circuit transfer impedance is equal to the reciprocal of the tube transadmittance; and the phase angle around the complete feedback loop must be zero.

8.3. Equivalent Circuit.—The addition of a feedback line to the equivalent circuit of a klystron amplifier is all that is required for an analysis of the tube as an oscillator. This circuit is shown in Fig. 8–2. A multigrid tube is shown in the equivalent circuit to indicate that the input and output circuits are completely isolated unless some means of feedback is supplied. The peak voltage across the input gap is E_1, which is also indicated as the

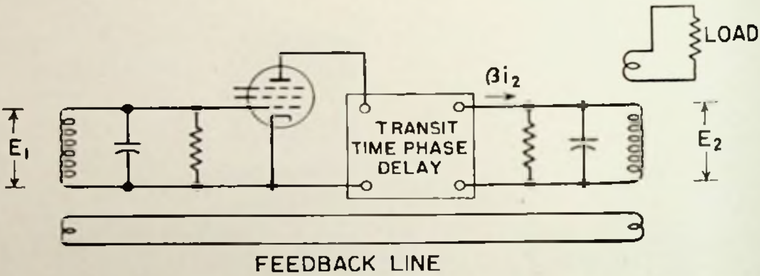


FIG. 8–2.—Equivalent circuit of a klystron oscillator.

voltage across the input circuit of the equivalent diagram. A voltage E_2 is shown across the output circuit and represents the peak voltage at the output gap.

A current βi_2 is shown flowing into the output circuit. This current represents the radio-frequency component of the bunched-beam current and is related to the input voltage E_1 by the usual bunching relationships

$$x = \beta\pi N \frac{E_1}{E_0} \tag{8-1}$$

and

$$\beta i_2 = 2\beta I_0 J_1(x) \tag{8-2}$$

These voltages and currents are sinusoidal quantities but are conveniently represented by their peak values.

A four-terminal network is used to represent τ, the transit-time phase angle. An output load is also shown coupled to the circuit. This load will affect the Q of the output resonator and will be considered in the analysis by assuming that the Q of the output resonator, which is the primary of the coupled-circuit system, is one-half as great as the Q of the secondary. The length of the feedback line will be considered negligible.

[1] F. E. Terman, *Radio Engineers' Handbook*, McGraw-Hill Book Company, Inc., New York, 1943.

8.4. Phase Relations.—Oscillation will occur when the sum of the phase angles around the complete loop is equal to some integral number times 2π radians, provided the magnitude of the feedback is sufficient to maintain oscillation. These phase angles and the relations among them are illustrated in Fig. 8–3. The electrons that passed the input gap at zero phase (*i.e.*, when E_1 was zero and changing from deceleration to acceleration) become the center of the bunch. These electrons correspond to a

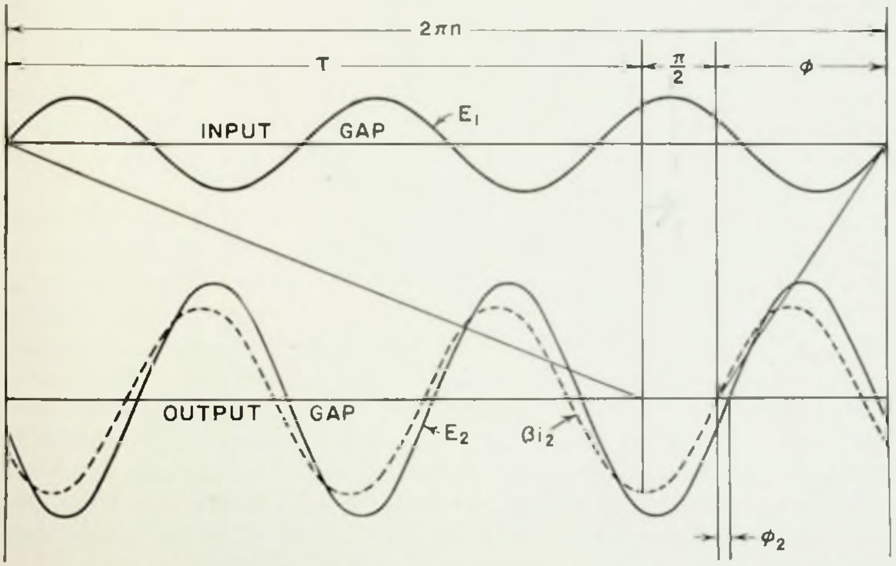


FIG. 8–3.—Phase relations in a klystron oscillator.

maximum value of the electron current and arrive at the output gap after a phase delay equal to τ. This phase delay corresponds to slightly less than 2¼ cycles in Fig. 8–3 and is indicated by the oblique line AB connecting the zero axes of the E_1 and i_2 curves.

In order to transfer maximum energy to the resonator field, the electrons in the bunch must pass the output gap when the resonator field is a maximum and in a direction that will decelerate the electrons. The current in the equivalent circuit must be in phase with the voltage if maximum power is to be transferred. Since electrons are negative charges, i_2 has a maximum negative value when the electron current is a maximum. Therefore, the transit-time phase angle τ corresponds to the phase between a zero value of E_1 and a negative peak of i_2, and the phase angle between E_1 and i_2, introduced by the transit-time and bunching considerations, is equal to $(\tau + \pi/2)$.

A phase angle ϕ_2 between βi_2 and the output voltage E_2 is shown because E_2 need not be in phase with βi_2. However, this phase angle is not used in this analysis of klystron oscillator theory. The angle ϕ represents the phase between βi_2 and E_1, introduced by the coupling between the two circuits. The phase relation which must be satisfied for oscillation can be written

$$\tau + \frac{\pi}{2} + \phi = 2\pi n \tag{8-3}$$

where n must be an integer. The value of n in Fig. 8–3 is 3 cycles.

If the transit-time phase angle τ is changed by varying the acceleration voltage E_0, a corresponding change in ϕ must occur in order to maintain the relation in Eq. (8–3). The phase angle ϕ between βi_2 and E_1 can change only if the frequency of oscillation is changed; therefore a change in the acceleration voltage is accompanied by a change in the frequency of oscillation.

8.5. Coupled-circuit Theory.—The equivalent circuit in Fig. 8–2 shows that a klystron oscillator may be considered as two parallel-resonant coupled circuits, fed by a current source βi_2. Note that the output current of the klystron βi_2 is the primary current of the coupled-circuit system, and that the input gap voltage E_1 is the secondary voltage. The analysis of coupled circuits to obtain the phase and magnitude of the voltage across the primary and secondary can be obtained, with some slight modifications, from radio engineering handbooks and other sources. This information about the phase and magnitude of the feedback can then be combined with the bunching characteristic of a klystron tube to obtain a prediction of the output characteristics of a klystron oscillator as the acceleration voltage is varied.

Certain assumptions will be made to simplify the analysis. The resonant frequencies of both circuits will be considered identical; this assumption will make the characteristics symmetrical with respect to frequency. A feedback line of zero length will be assumed so that the analysis for lumped constant circuits at low frequencies can be used without modification. The Q of the two circuits will not be equal because the output resonator of the klystron is connected to a load. The Q of the output circuit, which corresponds to the primary of the two coupled circuits in Fig. 8–2, will be considered half as great as the Q of the input circuit, which corresponds to the secondary. Sample calculations wi l be carried out for the case of an undercoupled klystron with a coupling coefficient one-half as great as the value for critical coupling. Curves will also be included for a klystron with coupling ten times greater than the critical value.

8.6. Characteristic Curves for Coupled Circuits.—Curves of primary and secondary voltage as a function of frequency are usually given in the reference texts. The curves of Fig. 8–4 show the typical characteristics when the coupling is one-half the critical value. The phase angle ϕ between the primary current βi_2 and the secondary voltage E_1 is also shown. Ratios of $E_1/\beta i_2$ and $E_2/\beta i_2$ are plotted so that unit coordinates may be used, and the maximum possible ratio of $E_1/\beta i_2$ is indicated by a value of unity. Increasing the value of βi_2 will increase the value of voltage across the circuit but will not affect the ratio.

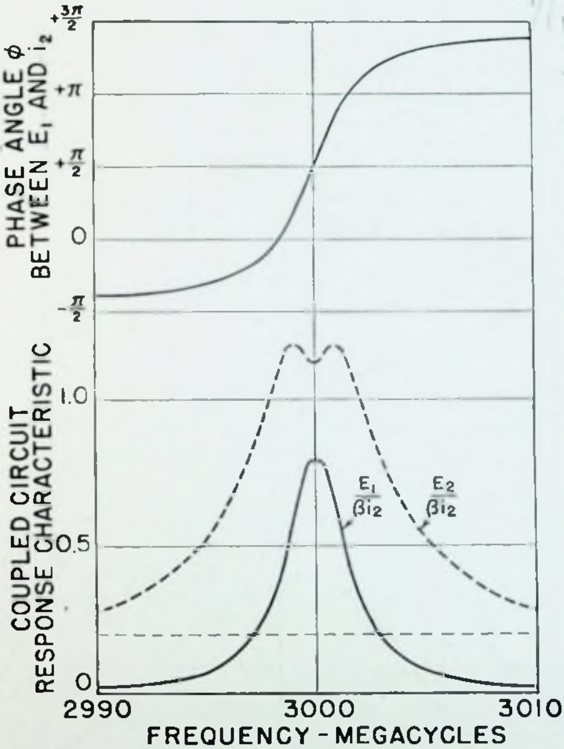


FIG. 8-4.—Phase and voltage relations in two coupled resonant circuits when the coefficient of coupling is one-half the value required for critical coupling. The Q of the primary circuit has been assumed half as great as the Q of the secondary circuit.

The form of the curves in Fig. 8-4 is not ideally suited to the analysis of a klystron oscillator where the frequency is dependent upon the beam voltage as well as upon the tuning. To examine the behavior of such an oscillator as the acceleration voltage is changed, it will be convenient to replot the frequency as a function of voltage. This step is illustrated by the frequency vs. voltage curve in Fig. 8-5.

The relation between voltage and frequency is obtained by correlating the phase angle ϕ in Fig. 8-4 with the drift space transit angle τ, using Eq. (8-3) to evaluate τ. The value of the acceleration voltage E_0 corresponding to this value of transit angle can then be obtained from the relation

$$\tau = \frac{2\pi f s_0}{v_0} \tag{8-4}$$

Equation (3-3) may be substituted for v_0 in Eq. (8-4) giving

$$\tau = \frac{2\pi f s_0}{6 \times 10^7 \sqrt{E_0}} \tag{8-5}$$

E_0 is the acceleration voltage, or beam voltage, s_0 is the drift distance in centimeters, and f is the frequency in cycles per second.

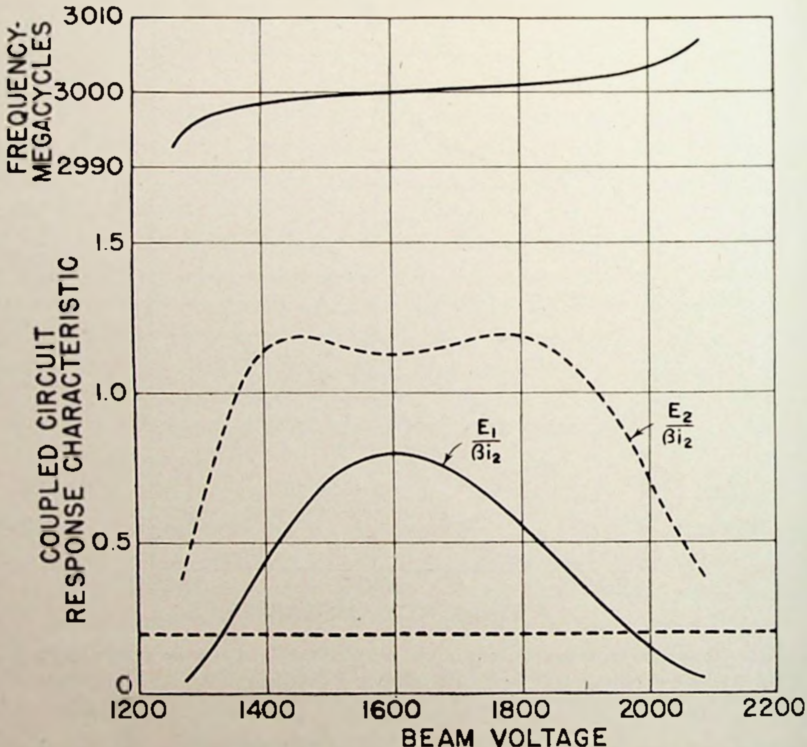


FIG. 8-5.—The data in Fig. 8-4 have been replotted as a function of the beam voltage on the klystron oscillator.

As a typical example, consider a klystron with a drift distance of 2.8 cm operating at 3,000 megacycles per sec. Assume that n in Eq. (8-3) is equal to 3. Since ϕ is equal to $\pi/2$ for zero frequency deviation in Fig. 8-4, τ would have a value of 7π. Equation (8-5) can be rewritten

$$E_0 = \left(\frac{2\pi f s_0}{6 \times 10^7 \tau}\right)^2 \tag{8-6}$$

Substitution of the above values gives a voltage of 1,600, which corresponds to zero frequency deviation. Voltage values for other values of ϕ are computed in the same manner. Then these data are replotted to give the frequency vs. voltage curves in Fig. 8–5. The values of $E_1/\beta i_2$ and $E_2/\beta i_2$ are transferred from Fig. 8–4 to Fig. 8–5, using this f vs. E_0 characteristic.

8.7. Conditions for Oscillation.—One of the requirements for oscillation, discussed in Sec. 8.2, stated that the circuit transfer impedance $E_1/\beta i_2$ must be equal to the reciprocal of the tube transadmittance. Since the transreduction factor $x/2J_1(x)$, discussed in Sec. 7.9, is equal to the ratio of the small-signal transadmittance $\beta^2\pi NI_0/E_0$, to the tube transadmittance, this relation may be restated

$$\frac{E_1}{\beta i_2} = \frac{E_0}{\beta^2\pi NI_0}\,\frac{x}{2J_1(x)} \tag{8–7}$$

Equation (8–7) may also be derived by rewriting Eq. (8–1) in the form

$$E_1 = \frac{E_0 x}{\beta\pi N} \tag{8–8}$$

and substituting Eqs. (8–2) and (8–8) in the ratio $E_1/\beta i_2$.

Equation (8–7) may also be written

$$\frac{x}{2J_1(x)} = \frac{\beta^2\pi NI_0}{E_0}\,\frac{E_1}{\beta i_2} \tag{8–9}$$

The small-signal transadmittance $\beta^2\pi NI_0/E_0$ can be replaced by the symbol y_{ms}.

$$\frac{x}{2J_1(x)} = y_{ms}\,\frac{E_1}{\beta i_2} \tag{8–10}$$

$$\frac{x}{2J_1(x)} = \frac{y_{ms}}{\beta i_2/E_1} \tag{8–11}$$

Equation (8–11) is another way of stating that the transreduction factor for an oscillator is equal to the ratio of the small-signal transadmittance to the circuit transfer admittance.

8.8. Bunching Characteristic.—It is necessary to know i_2 in terms of the transreduction factor $x/2J_1(x)$ in order to compute E_1 or E_2. The usual bunching characteristic, for example, Fig. 3–7 in Sec. 3.11, shows $2J_1(x)$ as a function of x. It will be convenient to replot $2J_1(x)$, which is proportional to i_2, as a function of $x/2J_1(x)$. This modification of the bunching characteristic makes it possible to obtain the value of i_2 from a value of $E_1/\beta i_2$, since Eq. (8–9) shows that $x/2J_1(x)$ is proportional to

$E_1/\beta i_2$. Figure 8–6 shows such a characteristic. It was obtained by computing $x/2J_1(x)$ for each point on the Bessel function curve in Fig. 3–7 and by replotting these data.

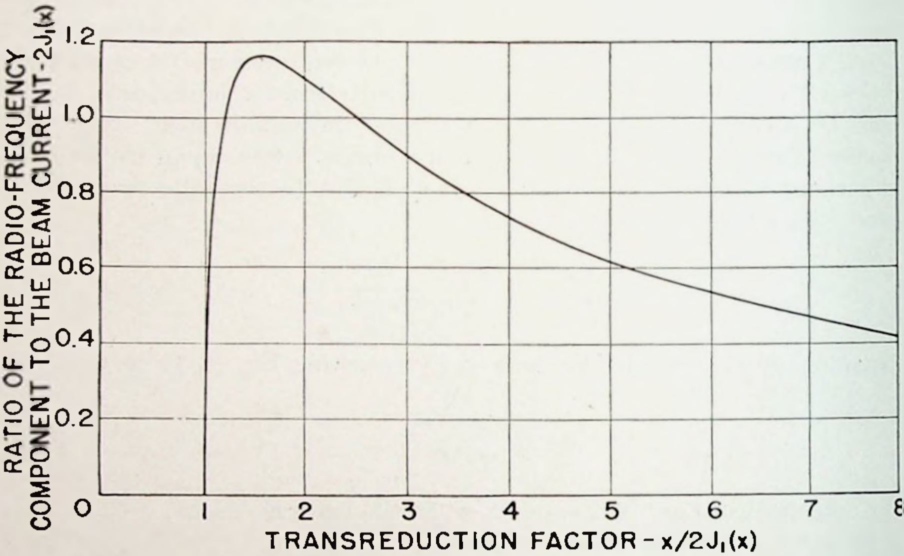


FIG. 8–6.—Ratio of the radio-frequency component of the bunched-beam current to the direct current in the electron beam, replotted as a function of the transreduction factor. An accurate graph of $2J_1(x)$ vs. $x/2J_1(x)$ is given in Chart XIII in Appendix B.

8.9. Oscillator Characteristics.—The method of obtaining the power output characteristics of an oscillator will be illustrated by a sample calculation, using Figs. 8–5 and 8–6. Typical operating conditions might be

$$E_0 = 1{,}600 \text{ volts}$$
$$I_0 = 60 \text{ ma}$$
$$N = 3.5$$
$$\beta^2 = 0.8$$
$$y_{ms} = 332 \text{ micromhos}$$

Figure 8–5 gives the value of $E_1/\beta i_2$ in unit coordinates; therefore, the small-signal transadmittance must be converted to these coordinates before substitution in Eq. (8–10). Assume that a beam current of 15 ma corresponds to the starting current for the oscillator. The transadmittance for this starting current would be 83 micromhos. The beam voltage E_0 would be adjusted for maximum feedback, corresponding to the maximum of the $E_1/\beta i_2$ curve in Fig. 8–6; therefore, the value of $E_1/\beta i_2$ for the starting condition would be 0.8. Since $x/2J_1(x)$ is unity for the starting conditions, Eq. (8–10) shows that the small-signal transadmittance y_{ms} in unit coordinates will be 1.25 for the starting current. This condition establishes the relation between the unit coordinates and the actual operating conditions. The operating beam current is four times greater than the starting current; therefore, the value of y_{ms} for the operating conditions with 60 ma beam current is equivalent to 5.0, and

$$\frac{x}{2J_1(x)} = 5.0\,\frac{E_1}{\beta i_2} \tag{8-12}$$

expresses the required relationship corresponding to Eq. (8–10) for the unit coordinate system used in Fig. 8–5.

A value of $E_1/\beta i_2$ equal to 0.2 makes $x/2J_1(x)$ in Eq. (8–12) equal to unity; therefore, this value of $E_1/\beta i_2$ corresponds to the starting conditions for the oscillator when the operating beam current is 60 ma. This value is indicated by a dash line in Fig. 8–5. The acceleration voltage represented by the intersection of the dash line and the $E_1/\beta i_2$ curve in Fig. 8–5 corresponds to the starting conditions for the oscillator and the output for this value of acceleration voltage is zero.

The output for other values of acceleration voltage may be computed in the following manner. A higher value of $E_1/\beta i_2$ for some other acceleration voltage is substituted in Eq. (8–12) to obtain a value of $x/2J_1(x)$, and a corresponding value of $2J_1(x)$ is determined from Fig. 8–6. The beam current I_0 is computed from the $\frac{3}{2}$ power law and the acceleration voltage. The value of βi_2 is then obtained by substitution of β, I_0, and $2J_1(x)$ in Eq. (8–2). E_2 is calculated from βi_2 and the value of $E_2/\beta i_2$.

Values of $E_2/\beta i_2$ in Fig. 8–5 must be converted from unit coordinates to conventional units before they can be used to compute the output. The conversion relation can be obtained from Eq. (8–10). The transreduction factor $x/2J_1(x)$ is a dimensionless ratio, and any consistent set of units may be used in the right side of the equation. The product of y_{ms} (mhos) and $E_2/\beta i_2$ (ohms) must equal the product of these two factors in the unit coordinate system. Since 83 micromhos corresponds to an admittance of 1.25 in unit coordinates,

$$83 \times 10^{-6}\left(\frac{E_2}{\beta i_2}\right)_{\text{ohms}} = 1.25\left(\frac{E_2}{\beta i_2}\right)_{\text{unit coordinates}} \tag{8-13}$$

A more useful form of the relation would be

$$\left(\frac{E_2}{\beta i_2}\right)_{\text{ohms}} = 15{,}100\left(\frac{E_2}{\beta i_2}\right)_{\text{unit coordinates}} \tag{8-14}$$

When the output gap voltage E_2 has been computed, the power delivered to the load is obtained from the relation

$$P_L = \frac{E_2^{\,2}}{2R_L} \tag{8-15}$$

where R_L is the equivalent load resistance (see Fig. 8–2). These calculations have been made for the case illustrated in Fig. 8–5, assuming a value of 140,000 ohms for R_L. The corresponding theoretical power output curve is shown in Fig. 8–7. The frequency curves are transferred directly from Fig. 8–5 to Fig. 8–7.

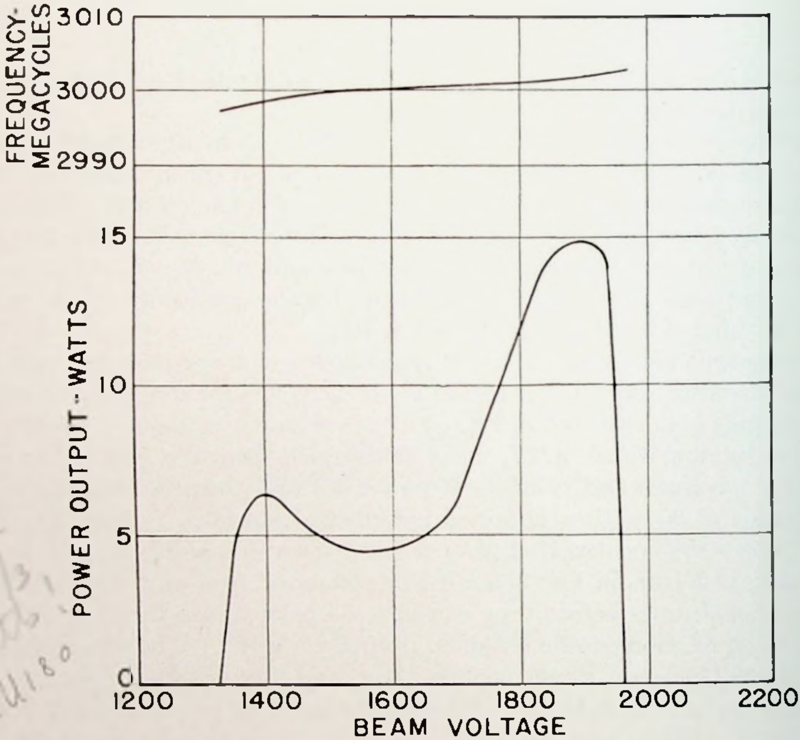


FIG. 8-7.—Theoretical curves of power output and frequency for an undercoupled klystron oscillator.

8.10. Overcoupled Klystron Oscillators.—The undercoupled oscillator analyzed in the previous section has a fairly uniform amplitude characteristic and a linear frequency deviation characteristic for a large portion of the voltage range. If maximum output is desired instead of a linear frequency characteristic, the coupling between the two circuits should be greater. Most of the early klystron designs were overcoupled for this reason. As a result, many people are familiar with the output characteristic of an overcoupled klystron oscillator.

A series of curves corresponding to Figs. 8–4, 8–5, and 8–7 are shown in Figs. 8–8 to 8–10 to illustrate the characteristics of an overcoupled klystron oscillator. A value of feedback coupling ten times greater than critical coupling has been chosen. The method of obtaining the curves is similar to the procedure described in Sec. 8.9 and will not be repeated. A discussion of the differences between the two degrees of coupling is given in the next section.

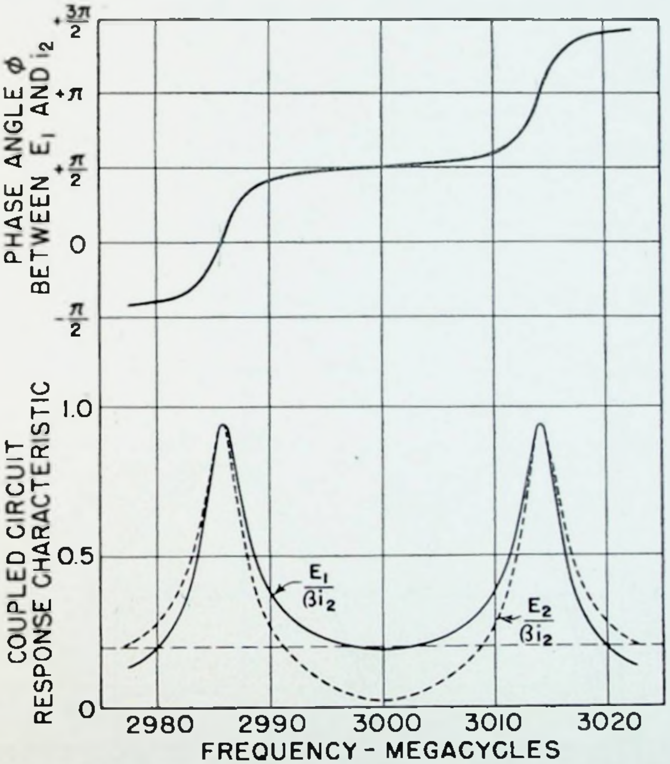


FIG. 8-8.—Phase and voltage relations for two coupled resonant circuits when the coupling coefficient is ten times greater than the critical value.

8.11. Voltage Modes.—There are a number of acceleration voltages corresponding to different values of drift transit angle τ which satisfy the phase relations in Eq. (8-3). A series of voltage modes exist, and oscillation does not occur between them. This behavior is quite similar to the characteristics of the reflex oscillators described in Chap. 7.

The voltage modes occur in pairs when the two resonators are tightly coupled. The maxima of the two modes occur when the frequency has values corresponding approximately to $\phi = 0$ and $\phi = \pi$ in Fig. 8-8. If these values of ϕ are substituted in Eq. (8–3), the value of τ is given approximately by

$$\tau = 2\pi(n \pm \tfrac{1}{4}) \qquad (8\text{–}16)$$

Equation (8–16) is often stated in the form "The number of cycles of

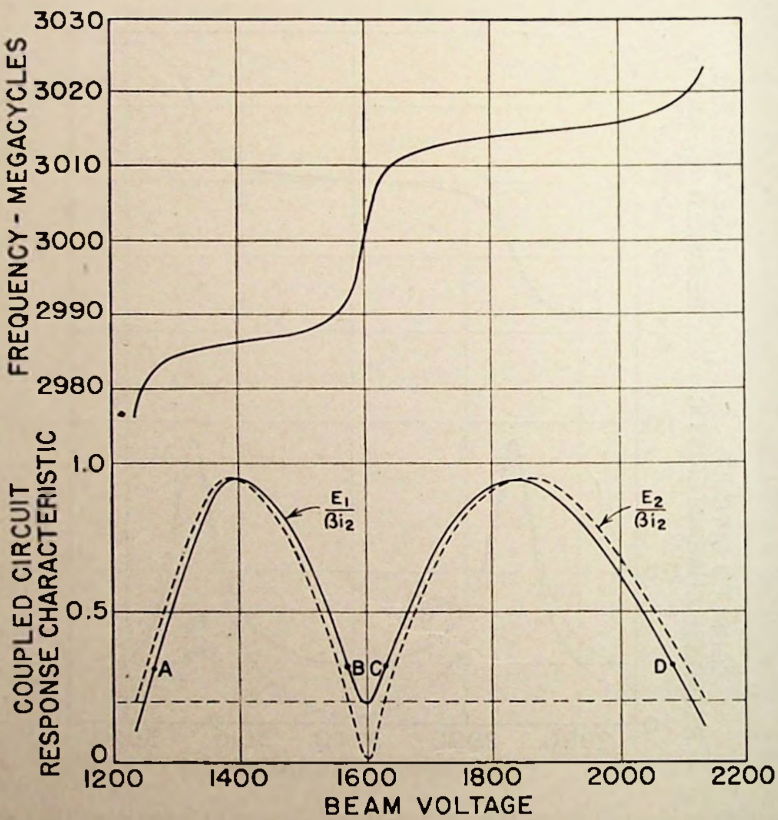


Fig. 8-9.—The data in Fig. 8–8 replotted as a function of the beam voltage on the klystron oscillator.

oscillation during the transit of an electron from the input gap to the output gap is an integer plus or minus one-quarter."

It should be noted that this statement is a special case which does not apply to all klystron designs. For example, τ is equal to $(n + \frac{1}{2})$ for the undercoupled klystron discussed in Sec. 8.9 because the two modes are merged into one. The latter characteristic is typical of the Type 3K30/410R klystron oscillator and differs from overcoupled types which exhibit two separate families of voltage modes.

8.12. Typical Output Characteristics.—Actual data for a Type 3K30/410R klystron oscillator are shown in Figs. 8–11 and 8–12. These curves were obtained by a dynamic method which eliminated tuning changes due to thermal variations caused by changing the power input to the tube. The similarity of Fig. 8–11 to the theoretical curve in Fig. 8–7 indicates

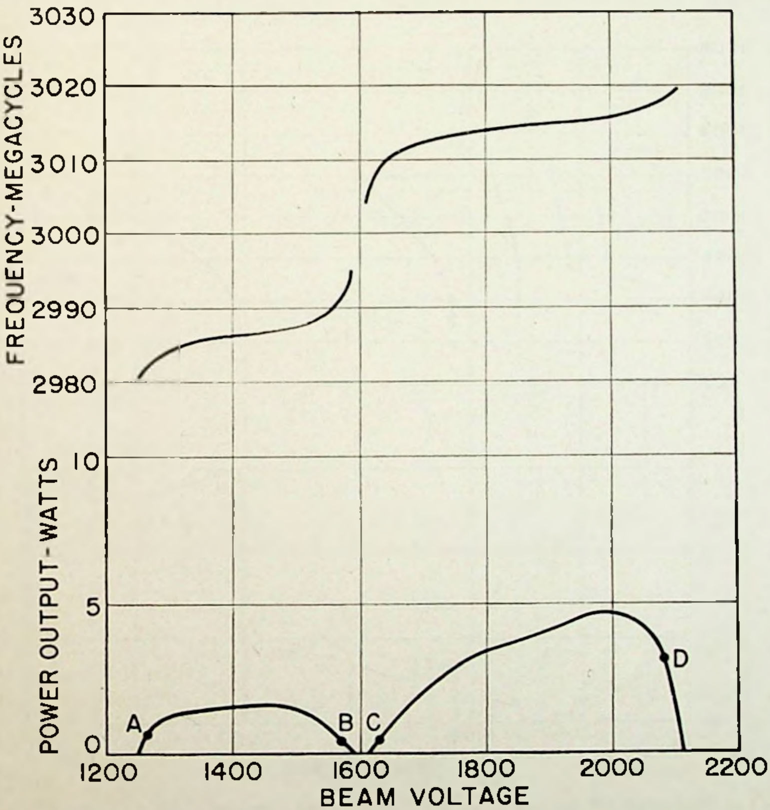


FIG. 8–10.—Theoretical curves of power output and frequency for an overcoupled klystron oscillator.

that the assumptions used in obtaining the theoretical curve are reasonably good. The equal height of the two peaks of each mode in Fig. 8–11 was obtained by purposely detuning the tube to produce the flattest characteristic, and does not represent a deviation from theory. Detuning the tube in the opposite direction emphasizes the higher voltage peak and produces maximum power output and efficiency. This effect is illustrated in Fig. 8–12.

Note that the frequency deviation characteristic in Fig. 8–11 is almost linear with acceleration voltage over the useful range of output. Because of this characteristic, the Type 3K30/410R klystron is quite useful as an oscillator in frequency-modulation circuits. The frequency modulation is obtained by varying the acceleration voltage. However, considerable modulation power is required, and power output must be sacrificed to obtain this wide-band frequency characteristic. (Compare Figs. 8–11 and 8–12.)

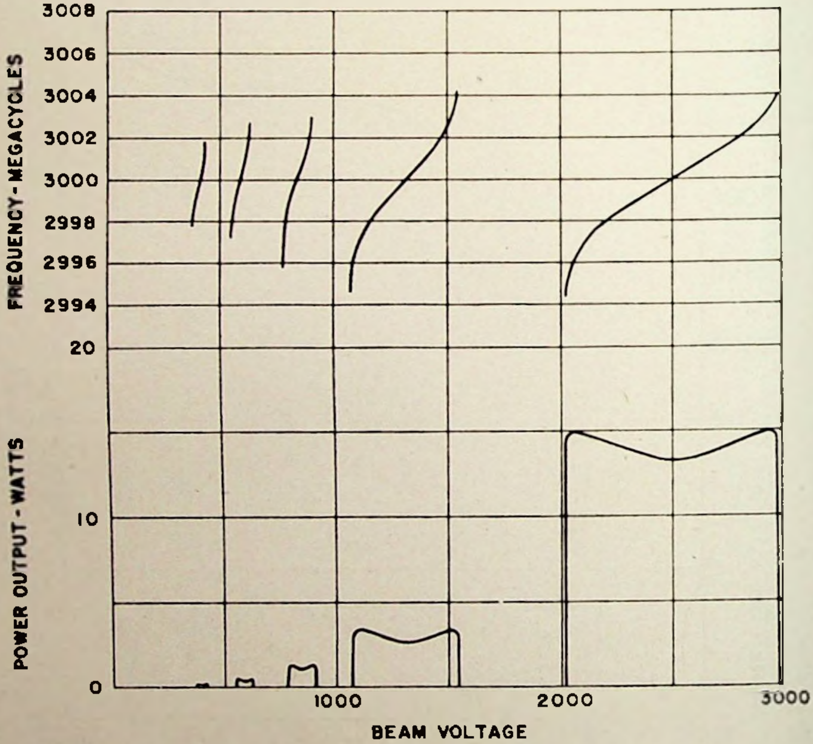


FIG. 8–11.—Experimental curves of power output and frequency for a Type 3K30/410R klystron oscillator.

8.13. Overbunching.—An analysis of the effect of detuning a klystron oscillator requires consideration of the klystron characteristic known as "overbunching." This characteristic was discussed in Secs. 3.9, 3.11, and 4.7; however, overbunching in an oscillator does not necessarily reduce the output as it does in an amplifier. Optimum bunching in Fig. 8–6 corresponds to the maximum in the curve when $x/2J_1(x)$ is equal to 1.59. This value of $x/2J_1(x)$ corresponds to a value of 0.318 for $E_1/\beta i_2$. Larger values of $E_1/\beta i_2$ cause overbunching.

Note in Fig. 8–9 that the ratio $E_2/\beta i_2$ is increasing rapidly at the acceleration voltages corresponding to points A, B, C, and D, which represent the ratio of $E_1/\beta i_2$ required for optimum bunching. Increasing the ratio of $E_1/\beta i_2$ above the optimum value does not reduce i_2 very rapidly, since i_2 is proportional to $2J_1(x)$, and Fig. 8–6 shows that this factor decreases slowly beyond the point of optimum bunching. The increased $E_2/\beta i_2$ ratio allows the output to increase in the region between A and B, also in the region between C and D. Corresponding points have been marked on Fig. 8–10 to illustrate this effect.

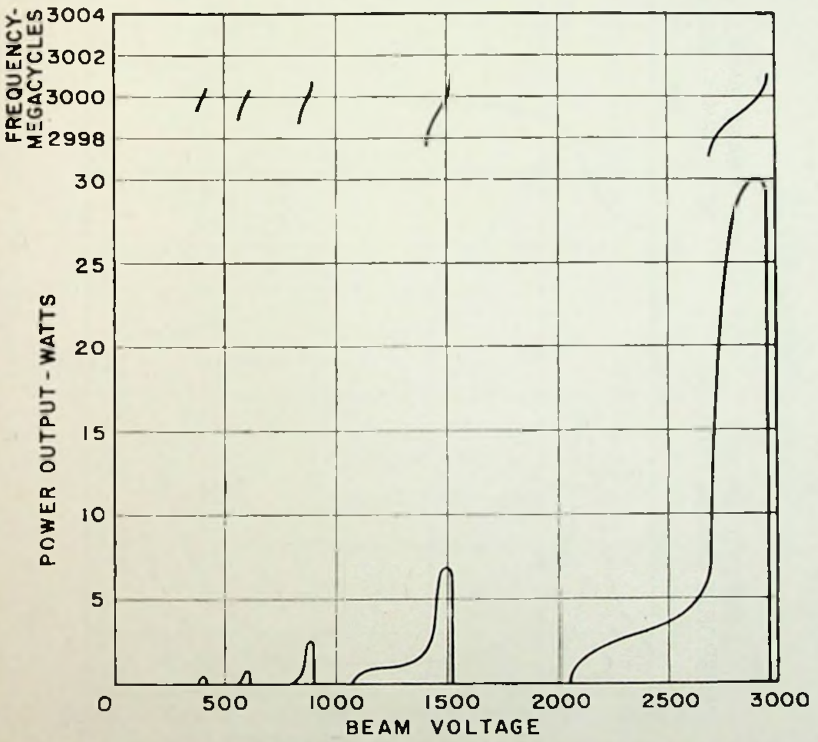


FIG. 8–12.—Experimental curves of power output and frequency for a Type 3K30/410R klystron oscillator tuned to maximize the power output.

8.14. Effect of Tuning the Klystron.—Any deviation from identical tuning will decrease one peak in the $E_1/\beta i_2$ curve for an overcoupled klystron when the losses in the two resonators are unequal. The other peak may remain approximately the same, or it may increase slightly, depending upon the ratio of the losses in the two resonators. The effect of detuning the input resonator to a lower frequency, leaving the output resonator tuned to the frequency corresponding to the identical tuning represented in Fig. 8–8, is illustrated in Fig. 8–13 for the case when the losses in the output resonator are twice as great as the losses in the input resonator. Both patterns are moved in the direction of lower frequency, and they are no longer symmetrical. The higher frequency peak in $E_1/\beta i_2$ is reduced and the corresponding peak in $E_2/\beta i_2$ is increased.

As a result of the detuning, the degree of overbunching in the high-frequency mode will be less and βi_2 will increase. Since $E_2/\beta i_2$ has also been

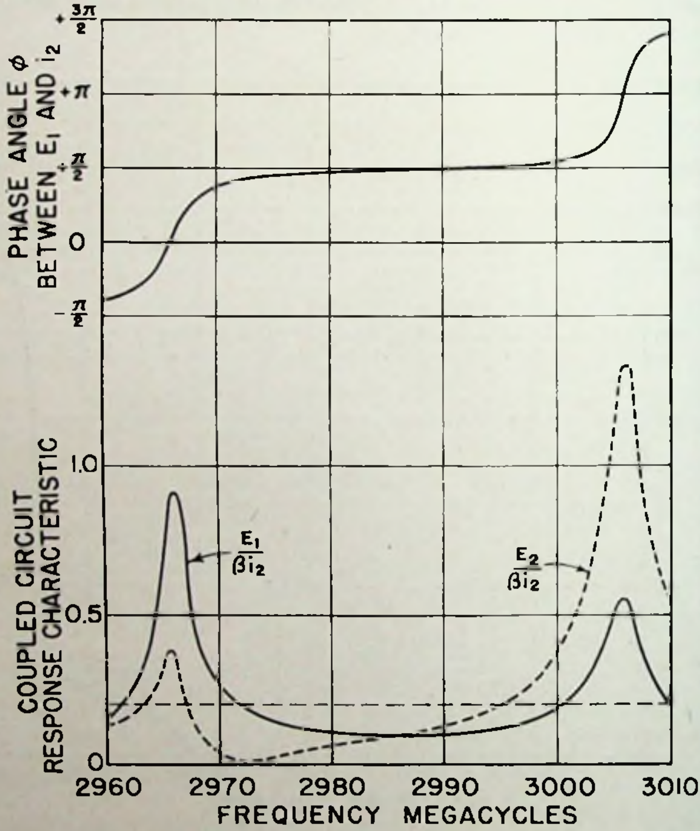


Fig. 8-13.—Phase and voltage relations for two overcoupled resonant circuits which are tuned to two different frequencies.

increased by the detuning, the output of this mode will increase considerably. In contrast, the overbunching remains about the same in the lower frequency mode, but the ratio of E_2/i_2 is decreased; therefore the output of this mode decreases slowly as the input resonator is tuned to a lower frequency. The data in Fig. 8–13 have been converted into a theoretical output characteristic for the detuned case as shown in Fig. 8–14. Compare this illustration with Fig. 8–10.

The ratio of $E_1/\beta i_2$ for the higher frequency mode can be reduced to the value corresponding to the starting conditions while the lower frequency peak in $E_1/\beta i_2$ remains greater than this value. This relation means that continued detuning of the input resonator to a lower frequency causes the higher frequency mode to increase at first, then disappear before the lower frequency mode is eliminated. The frequency of the higher frequency mode approaches the frequency of the tuning of the output resonator. The lower frequency mode can be accentuated by detuning the input resonator to a higher frequency than the tuning of the output resonator.

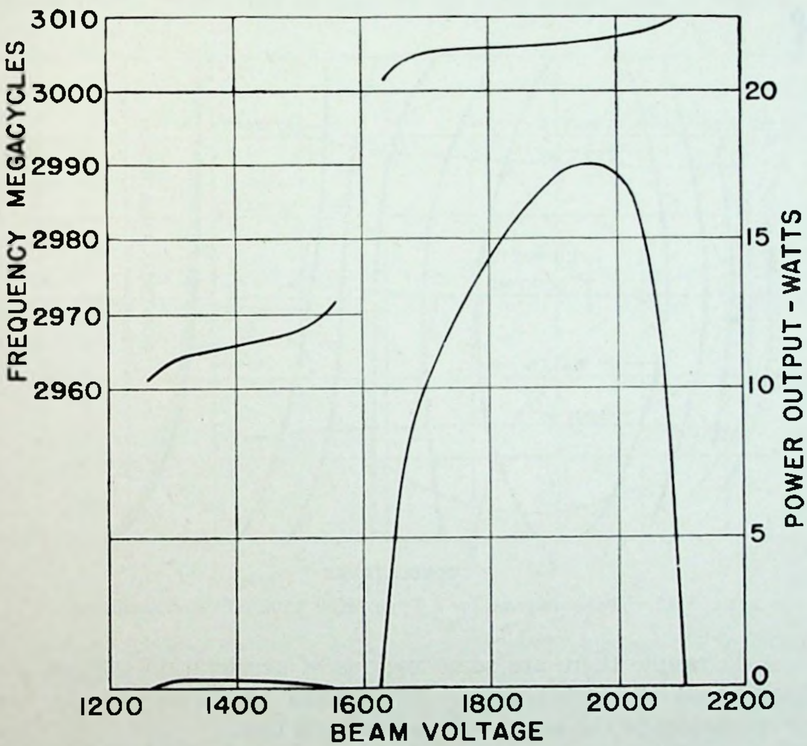


FIG. 8–14.—Theoretical curves of power output and frequency for an overcoupled klystron oscillator when the input resonator is detuned to a lower frequency than the output resonator.

8.15. Effect of Length of Feedback Line.—The previous discussion has been based upon a zero length of feedback line. This assumption obviously does not apply to a Type 3K30/410R klystron, since an external feedback line of considerable electrical length must be used. The problem is complicated when the line is not terminated with its characteristic impedance. However, it is fairly satisfactory to assume that the phase shift in the line can be represented by an angle ϕ_L and to rewrite Eq. (8–3) in the form

$$\tau + \frac{\pi}{2} + \phi + \phi_L = 2\pi n \tag{8–17}$$

The logical conclusion from Eq. (8–17) would be that the acceleration voltage, which determines τ, could be arbitrarily chosen by adjusting the length of the feedback line. Some choice of operating voltage is permitted by such an adjustment. However, the feedback line is not usually matched, and certain line lengths may prevent operation at some frequencies within the tuning range of the tube. Therefore, it is not always possible to correct for changes in the transit time by changes in the length of the feedback

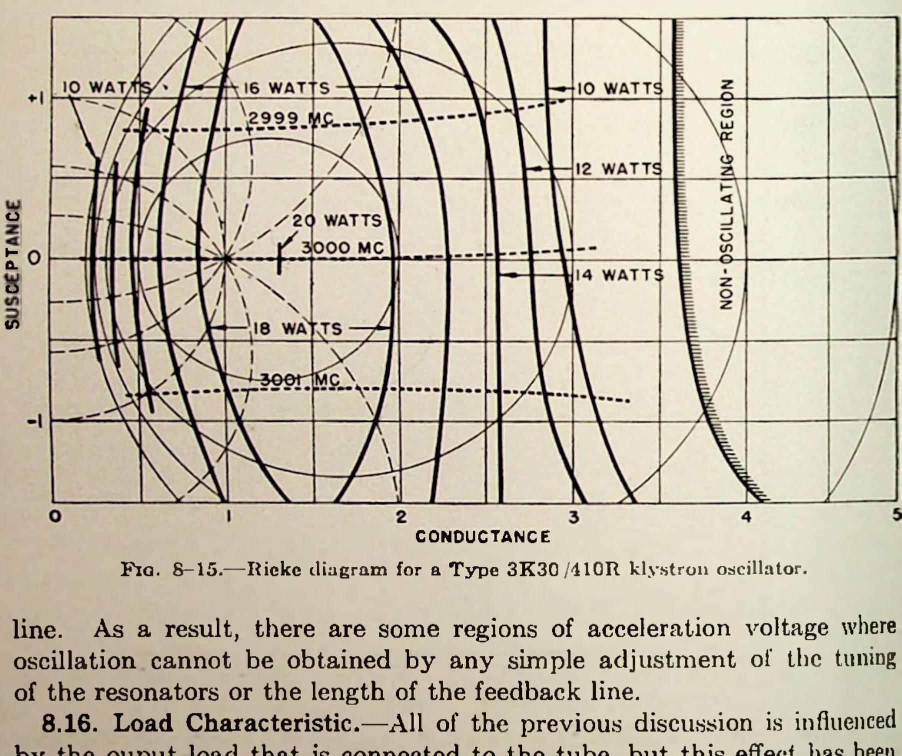


FIG. 8–15.—Rieke diagram for a Type 3K30/410R klystron oscillator.

line. As a result, there are some regions of acceleration voltage where oscillation cannot be obtained by any simple adjustment of the tuning of the resonators or the length of the feedback line.

8.16. Load Characteristic.—All of the previous discussion is influenced by the ouput load that is connected to the tube, but this effect has been included implicitly in the assumption that the losses in the output resonator were twice as great as the losses in the input resonator. This assumption is equivalent to the statement that the resonators are identical and that the power transferred to the load is equal to the power losses in the output resonator itself.

The Q of the output resonator will be one-half as great as that of the input resonator if the tube is loaded to give maximum output and the output gap voltage is less than the beam voltage. This relation was discussed in Secs. 4.11 and 4.13. Therefore the assumption used in the circuit analysis is satisfactory for the usual operating conditions. However, it is important to know the behavior of the tube as the load is varied. The prediction of the load characteristic from a circuit analysis would be quite tedious, but the characteristic is not greatly different from that of a simple reflex oscillator.

A typical Rieke diagram for a two-resonator oscillator is shown in Fig. 8–15. There is a region of nonoscillation where the load is too heavy, and frequency contours are shown on the diagram as well as contours of constant power output.

8.17. Frequency Stability.—Although the electronic tuning of an oscillator is given by the curves in Figs. 8–7 and 8–10, these data are limited to one particular loading of the tube. It is possible to make an approximate calculation for the frequency stability for any loading. Equation (8–17) may be rewritten with the phase angles that are independent of frequency included in a single constant:

$$\tau + \phi = \text{a constant} \tag{8–18}$$

where τ is the drift-space transit-time angle and ϕ is the feedback phase angle which is determined by the frequency and the relative tuning of the two resonators. Differentiating Eq. (8–18) gives

$$d\phi = - d\tau \tag{8–19}$$

The transit angle τ is proportional to the reciprocal of the square root of the acceleration voltage; therefore,

$$\tau = kE_0^{-1/2} \tag{8–20}$$

$$d\tau = -\frac{1}{2} kE_0^{-3/2} dE_0 = -\frac{1}{2}\tau\frac{dE_0}{E_0} \tag{8–21}$$

If N represents the number of oscillation cycles during bunching, Eq. (8–4) may be rewritten

$$\tau = 2\pi N \tag{8–22}$$

and Eq. (8–21) becomes

$$d\tau = - N\pi\frac{dE_0}{E_0} \tag{8–23}$$

Inspection of Fig. 8–4 shows that the slope of the phase curve is approximately unity. This approximation is also quite satisfactory for the case when the two circuits are overcoupled but tuned to the same frequency, and will be sufficiently accurate when the two circuits are detuned. The change in the phase of the feedback for a change in frequency may therefore be written

$$d\phi \cong 2\sqrt{Q_1Q_2}\,\frac{df}{f} \tag{8–24}$$

Substituting Eqs. (8–23) and (8–24) in Eq. (8–19) gives

$$\frac{df}{f} \simeq \frac{\pi N}{2\sqrt{Q_1Q_2}} \frac{dE_0}{E_0} \tag{8-25}$$

Several important results are evident from Eq. (8–25). An increase of beam voltage is accompanied by an increase in the frequency of oscillation. The effective Q of the oscillator is the geometric mean of the Q of the input resonator and the loaded Q of the output resonator; therefore, the stability of a two-resonator oscillator is higher than that of a reflex oscillator. The stability is increased if the transit time in the drift space is decreased; *i.e.*, operating at a higher voltage mode improves the stability. This does not mean that the total electronic tuning is decreased (see Figs. 8–11 and 8–12).

Equation (8–25) may also be written

$$\frac{df}{dE_0} \simeq \frac{\pi N}{E_0} \frac{f}{2\sqrt{Q_1Q_2}} \tag{8-26}$$

An idea of the frequency stability may be obtained by substituting the following typical operating values in Eq. (8–26):

$$\begin{aligned} E_0 &= 1{,}600 \text{ volts} \\ N &= 3.5 \text{ cycles} \\ f &= 3{,}000 \text{ megacycles} \\ Q_1 &= 2{,}000 \\ Q_2 &= 1{,}000 \end{aligned}$$

The frequency stability would be 0.0073 megacycle per volt, or 7.3 kilocycles per volt. It is apparent from this result that adequately filtered and well-regulated power supplies are required for klystron oscillators. This stability equation is also discussed in Chap. 10 on Modulation of Klystrons.

8.18. Beam-current Variation.—The frequency of oscillation is also affected by the current in the electron beam. The presence of an electron space charge in a resonator gap changes the effective dielectric constant of the gap. Varying the beam current will change the resonant frequency of the cavities and the frequency of oscillation will vary. This form of electronic tuning may also be obtained by using auxiliary electron beams which affect only the frequency of the resonators and do not interact with the resonators to change the radio-frequency energy.

CHAPTER 9

MULTIPLE-RESONATOR TUBES

9.1. Typical Tube Types.—Two-resonator klystron amplifiers, multipliers, or oscillators can be considered multiple-resonator tubes in the true meaning of the term, but this designation will be reserved for klystrons with three or more resonators. Tubes with more than three resonators have been built, usually for multistage amplifiers. However, a discussion of three different types of tubes, each using three resonators, will illustrate the principal advantages of multiple-resonator tubes. The more complicated designs will not be considered in detail.

One type of tube is analogous to an electron-coupled oscillator and is known as an "oscillator-buffer" klystron. A third resonator is excited by the bunched-beam current after it has passed through the second resonator. Part of the energy from the second resonator is fed back to the input resonator to produce oscillation. The load is coupled to the third or buffer resonator. Changes in load impedance are isolated from the oscillator portion of such a tube. This factor is important when an oscillator-buffer klystron is followed by a modulated amplifier, since tuning the input resonator of the amplifier will react upon a conventional klystron oscillator because it is coupled to the input resonator of the oscillator by the feedback line. The beam-current variations in a modulated amplifier will affect the tuning of the input resonator of the amplifier, but these variations will be isolated from the oscillator section if an electron-coupled klystron is used. These tubes are also useful for applications where the radio-frequency power is switched rapidly from one antenna to another, or with other forms of variable load.

Another three-resonator tube of different construction gives high amplification with a single tube by utilizing a single beam for a number of stages of amplification. This cascade amplifier system requires $(n + 1)$ resonators for n stages of amplification, but the beam power consumption is that of a single tube. The same gain with separate klystron stages would require many more resonators and use more than n times the beam power of a cascade amplifier klystron because the gain per stage is one-half as great.

Cascade amplifier klystrons have many advantages in addition to a large gain and illustrate a number of interesting features which are introduced when the velocity modulation is not a simple sinusoidal function. For these reasons, the theory of the operation of cascade amplifiers will be considered in more detail than the other types of multiple-resonator tubes.

The third type of multiple-resonator tube to be described is similar to a cascade amplifier, but the last resonator is tuned to a harmonic of the input frequency. The first and second resonators act as an amplifier at the input frequency and reduce the drive power required for the tube. The second and third resonators function as a klystron frequency multiplier.

OSCILLATOR-BUFFER TUBES

9.2. Theory of Operation.—An oscillator-buffer klystron operates on the principle that the electrons in the bunch are not decelerated to zero velocity

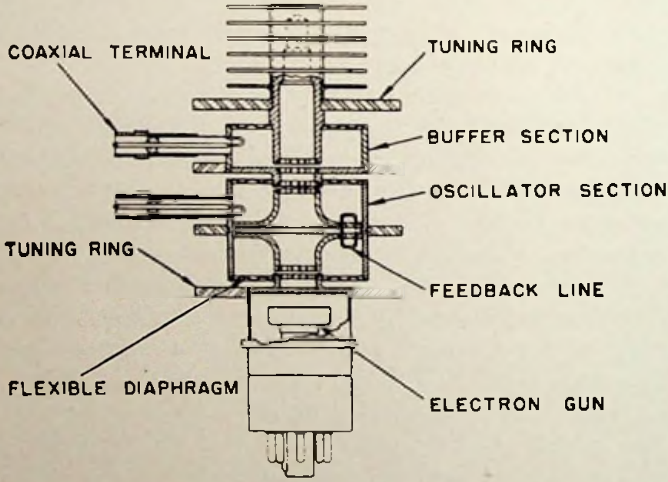


FIG. 9–1.—Sectional view of an oscillator-buffer klystron.

by the voltage at the second resonator gap, but the bunch continues its motion along the axis of the tube and can give up additional energy to a third resonator. This implies that the gap voltage induced by the beam at the second resonator is much less than the beam voltage. This relation is generally true for low-power klystrons. The velocity of the electrons in the bunch will be changed by the voltage at the second gap, and the bunching will be altered. This effect is discussed in Sec. 9.4.

Figure 9–1 shows a sectional view of an oscillator-buffer tube. The first two resonators are similar to a Type 3K30/410R klystron. An internal feedback line is shown, which means that the first two resonators of this particular design may be used only as an oscillator. It is possible to build tubes of this type without the internal feedback line; then the tube may be used as an amplifier or converted to an oscillator by using an external feedback line. The third resonator is the buffer output resonator. The third resonator gap is normally located as close to the second gap as allowed by mechanical-design considerations. This spacing could be reduced by using a modified arrangement of the resonators, but some spacing between the resonators must be maintained to prevent coupling between them.

9.3. Resonator Coupling.—If two resonators are arranged with their gaps adjacent to each other, separated by a single grid structure, the leakage fields through the holes in the grid mesh act as a strong coupling between the two resonators. Obviously such a structure would not be satisfactory for a tube which should isolate the oscillator section from the output. If two grids are used, separated by a short drift tube, the fields will be attenuated rapidly. The drift tube may be considered a wave guide that is too small to transmit energy. For the dimensions involved in a typical tube design, the attenuation of a "wave guide below cutoff" would be 50 db per cm.

Although the attenuation of a short drift tube seems quite high and corresponds to a very low value for the coupling coefficient between the two resonators, the fact that the resonators have a very high Q means that even a very low value of the coupling coefficient may be an appreciable fraction of the critical coupling. Therefore, an attenuation of 50 db, corresponding to a drift distance of 1 cm, may be required to isolate adequately the third from the second resonator.

9.4. Additional Bunching.—A drift distance of 1 cm will be equal to 35 to 50 per cent of the drift distance between the first and second resonators of a tube designed for operation at 3,000 megacycles. Considerable change in the electron bunching will take place, particularly if the voltage at the second gap is high. The beam will become badly overbunched, and the output at the third resonator will be quite small. Therefore the design of an oscillator-buffer klystron must be a compromise between a large separation of the last two resonators to prevent coupling between the output resonator and the oscillator section, and the short distance dictated by the electron-bunching considerations.

9.5. Tuning Considerations.—The effect of the overbunching in an oscillator-buffer tube can be overcome by detuning the oscillator section to reduce the bunching. This adjustment introduces instability and changes the frequency of oscillation slightly, but permits satisfactory operation. If the tube is tuned for maximum output with a load connected to the second resonator, when the load is moved to the third resonator the output will be quite small. Retuning either the first or the second resonator will increase the output from the third resonator by reducing the overbunching. Both must be retuned if the frequency of oscillation is to be accurately maintained.

9.6. Load Characteristics.—An oscillator-buffer klystron is similar to a klystron amplifier, since the load on the third resonator does not affect the frequency of oscillation appreciably or the amount of bunching in the beam. Therefore the output characteristics are identical to those discussed in Secs. 4.11 and 4.14 to 4.16 for a two-resonator klystron amplifier. If the load is connected to the second resonator, the behavior is similar to that of a two-resonator oscillator, and the Rieke diagram would have frequency contours as well as contours of constant power output.

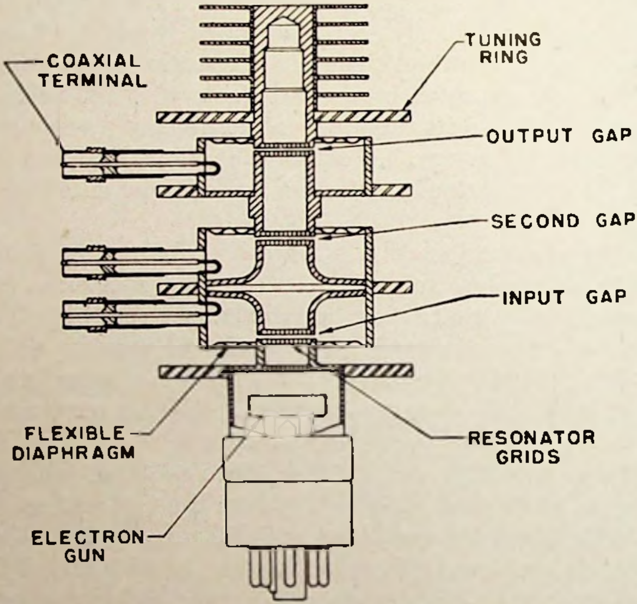


FIG. 9-2.—Sectional view of a cascade amplifier klystron.

CASCADE AMPLIFIER KLYSTRON

9.7. Drift Spaces in a Cascade Amplifier.—An oscillator-buffer tube requires a small drift space between the second and third resonators to prevent overbunching, because the electrons are well bunched at the second gap of an oscillator and additional bunching is undesirable. If a three-resonator tube is used as an amplifier, the additional bunching is desirable because the tube can be operated with very low drive power. The bunching at the second gap will not be complete, but the additional bunching between the second and third gaps will produce optimum bunching at the output resonator. If maximum gain is desired, the second drift space is made equal to the drift distance between the first and second resonators. A typical three-resonator amplifier, known as a "cascade amplifier klystron," is illustrated in Fig. 9-2.

9.8. Cascade Bunching.—A small voltage at the input gap will introduce some velocity modulation, and the beam will be bunched slightly when it passes the second gap. The gain in a velocity-modulation amplifier is relatively independent of the input voltage when the voltage is small, and this gain is greater than the gain of an amplifier with sufficient bunching to give optimum output (see Fig. 4–4). The gain for the first stage will introduce a much larger voltage at the second resonator gap. A large velocity variation is superimposed upon the partly bunched electron beam at the second resonator. This velocity variation is much greater than that introduced by the input gap owing to the voltage gain in the first stage, and this additional velocity variation may give sufficient bunching for optimum output at the third resonator.

This means that a cascade tube may be used as a power amplifier, utilizing the preliminary voltage amplifying stage to reduce the drive power required. A two-resonator amplifier may have a power gain of approximately 15 and require a drive power of 1 watt. In contrast, a cascade amplifier klystron makes possible a gain of several thousand in a single tube. A drive power of 6 mw is sufficient for a cascade amplifier klystron with an output of 15 watts.

9.9. Applegate Diagram.—This qualitative analysis of a cascade amplifier is well illustrated by the Applegate diagram in Fig. 9–3. The amplitudes of the voltages in the three resonators have been chosen to represent the conditions in a typical amplifier. The over-all voltage gain for the two stages is shown as 35; this corresponds to a power gain of 1,225. These figures are purely arbitrary but are not inconsistent with the actual conditions in a typical three-resonator amplifier.

Although a qualitative explanation of a cascade amplifier gives a good picture of the bunching in such a tube and explains the high gain that can be achieved, it does not provide a satisfactory basis for the unusual characteristics that these tubes exhibit. In order to explain the behavior of these tubes, which is described in more detail in Secs. 9.15 to 9.17, it will be necessary to consider the effects that occur when the electron bunching is caused by a complex combination of velocity modulation at several points along the beam. The result not only will depend on the relative amplitudes of the velocity modulation but will be determined by the phase angle between the voltages in the different resonators and the phase relation between the resonator voltages and the bunched-beam current.

9.10. Phase Relations in the Bunching.—There is a phase difference of 90 deg between the bunching in the first and second resonators of a cascade amplifier if the second resonator is tuned exactly to the input frequency. This relation may be verified by referring all phases to the line in the Applegate diagram which represents the position of the electron that passed the input gap at zero time, instead of using absolute time as a reference.

This definition of phase is equivalent to subtracting the average drift transit time.

The electron that leaves the input gap at zero time becomes the center of the partly formed bunch at the second resonator. The radio-frequency

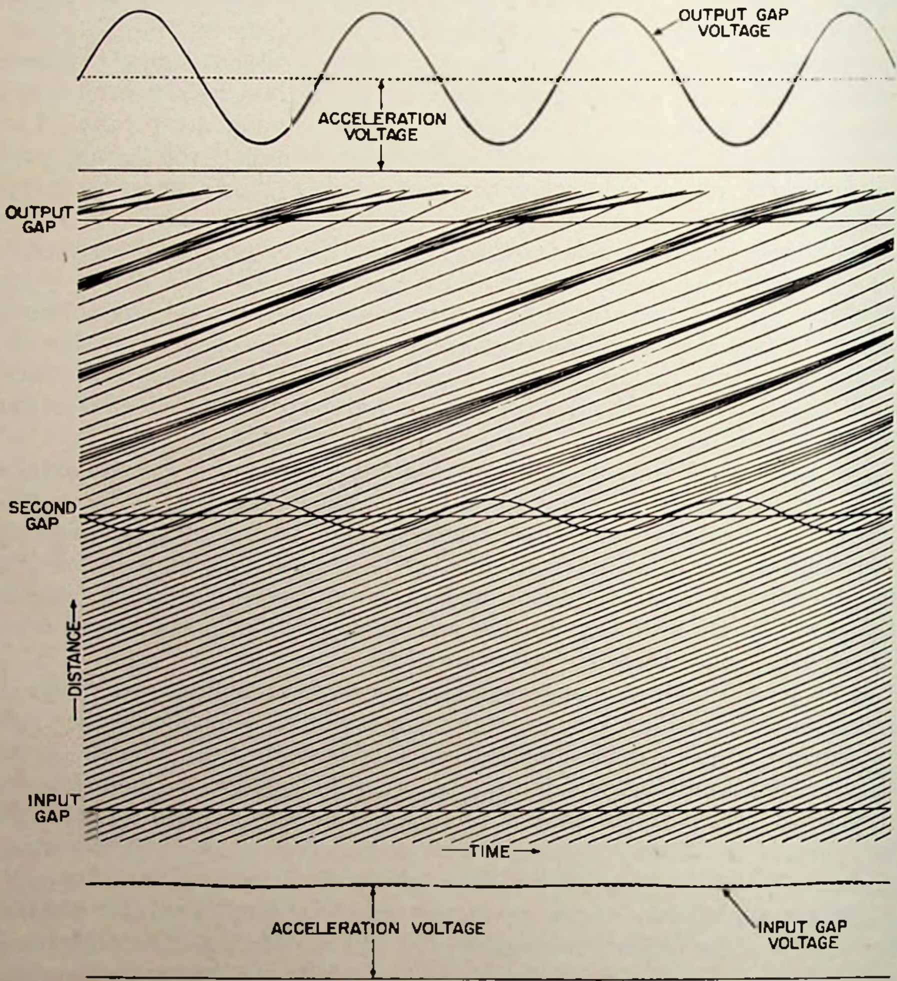


FIG. 9-3.—Applegate diagram for a cascade amplifier klystron with all resonators tuned to the input frequency.

voltage at the second resonator will have a phase that will retard the greatest number of electrons if this resonator is tuned to the input frequency. For this reason, the radio-frequency voltage at the second resonator will lag the voltage at the first resonator by an angle of 90 deg. Figure 9-3 illustrates these phase relations.

9.11. Bunching Parameter Becomes Ambiguous.—Although a value of the usual bunching parameter may be used to describe the bunching in the first drift space, the bunching in the second drift space is the result of the superposition of two velocity variations with different phases. Also, the beam current at the second gap is no longer uniform but has a radio-frequency component. Therefore, the designation of a simple parameter to describe the cascade bunching is impractical. The resultant bunching may be described in terms of a complicated function of several bunching parameters for different parts of the tube, but the same result can be understood more readily with the aid of a graphical analysis of the cascade bunching.

9.12. Arrival-time Diagrams.—The bunched current distribution and the output of a klystron can be computed from the curves showing arrival time as a function of departure time. This principle was discussed in detail in Chap. 3. The relation between the arrival time at the output gap of a three-resonator cascade amplifier, designated t_3, and the departure time t_1 can be obtained by a step-by-step computation. The transit time for the first drift space is tabulated for different values of t_1. A curve of t_1 vs. t_2 is plotted from these data and the phase angle ϕ_2 of the voltage E_2 at the second gap is determined. The modified velocities permit the calculation of the transit times for the second drift space for each electron corresponding to the assumed value of t_1. Then the arrival time is obtained from the relation

$$t_3 = t_1 + T_{12} + T_{23} \tag{9-1}$$

where T represents the transit time and the subscripts refer to the drift spaces between the first and second resonators or between the second and third resonators.

A set of data for a typical case is shown in Table IV. These data were used for constructing the Applegate diagram in Fig. 9–3 and are also plotted in Fig. 9–4. Time is shown in degrees, rather than in seconds or radians, in order to simplify the calculations.

9.13. Efficiency Increased by Cascade Bunching.—Comparison of the Applegate diagram for a two-resonator amplifier shows that there are fewer electrons passing the output gap of a cascade amplifier during the wrong half of the cycle when the field is transferring energy to the beam, and there are more electrons in the bunch. This means that the efficiency of a cascade amplifier is higher than that of a two-resonator klystron. The beam is partly bunched at the second resonator, and the additional velocity modulation causes more electrons from the wrong half of the cycle to become a part of the bunch. This effect can be increased if the second resonator is detuned on the high-frequency side of resonance so that the bunching action of the second resonator is more nearly in phase with the

TABLE IV.—COMPUTATION OF ARRIVAL TIMES FOR A CASCADE AMPLIFIER

Departure time, ωt_1	$\sin \omega t_1$	ωT_{12}	ωt_2	$\sin (\omega t_2 - 1260)$	ωT_{23}	ωt_3
0	0.000	1170	1170	—1.000	1292	2462
30	0.500	1164	1194	—0.914	1270	2464
60	0.866	1160	1220	—0.643	1229	2449
90	1.000	1158	1248	—0.208	1179	2427
120	0.866	1160	1280	0.342	1125	2405
150	0.500	1164	1314	0.809	1086	2400
180	0.000	1170	1350	1.000	1078	2428
210	—0.500	1176	1386	0.809	1094	2480
240	—0.866	1180	1420	0.342	1139	2559
270	—1.000	1182	1452	—0.208	1203	2655
300	—0.866	1180	1480	—0.643	1256	2736
330	—0.500	1176	1506	—0.914	1282	2788
360	0.000	1170	1530	—1.000	1292	2822

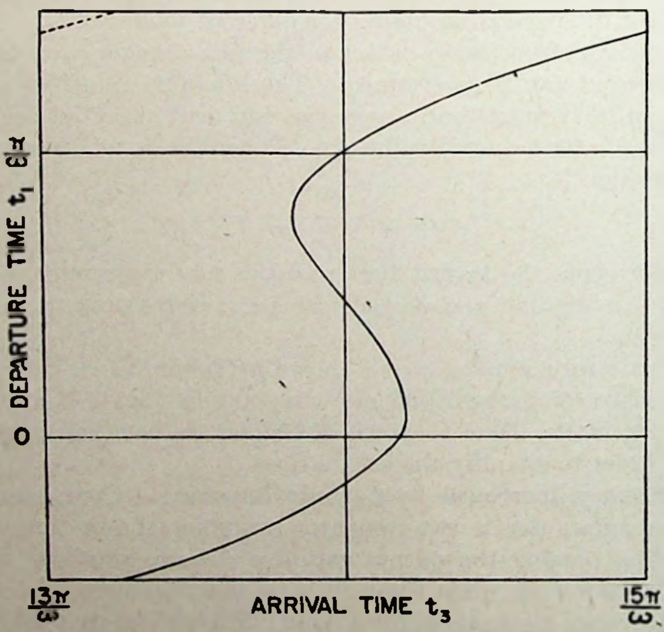


FIG. 9-4.—Arrival-time curve for a cascade amplifier klystron with all resonators tuned to the input frequency.

bunching from the first resonator. Therefore the highest output will occur when the second resonator is slightly detuned to a higher frequency.

9.14. Cancellation of Bunching.—Detuning the second resonator to the low-frequency side of resonance will reduce the voltage induced in the second resonator until it is as small as the input voltage, and will shift the phase until the bunching from the two resonators is almost canceled. The output will vary from a maximum on the high-frequency side of resonance to practically zero as the second resonator is tuned to a lower frequency. Continued tuning to a lower frequency will reduce the voltage at the second resonator until it has no effect upon the beam, and the output will increase asymptotically to the value for the tube as a two-resonator amplifier using only the first and third resonators.

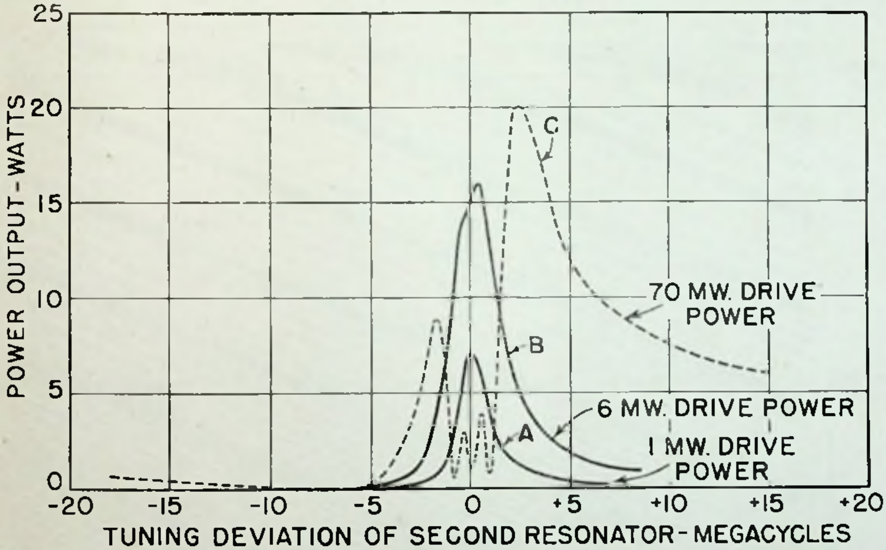


FIG. 9–5.—Experimental curves for a Type 2K35 cascade amplifier klystron showing power output as a function of the tuning of the second resonator. Curves for three values of drive power are shown.

9.15. Effect of Tuning Second Resonator.—Detuning the resonator to the high-frequency side of resonance decreases the voltage and makes the phases more nearly the same, so that the output decreases to the asymptotic value for the tube as a two-resonator amplifier. This characteristic is illustrated by curve *B* in Fig. 9–5. If the voltage gain per stage is much greater than two and the input power to the tube is reduced until the bunching at the second gap is negligible, the efficiency improvement due to the cascade bunching is unimportant and the maximum gain occurs when the second resonator is tuned to a frequency indistinguishably higher than the input frequency. The curve for this characteristic is shown as *A* in Fig. 9–5.

When the input power is increased considerably until the beam can be overbunched, the output vs. tuning characteristic has the form indicated by curve *C*. This curve is shown dotted so that it can be distinguished readily where it crosses the other curves. The point of maximum bunching is indicated by the minimum in the output when the second resonator is tuned to the input frequency (zero tuning deviation). This minimum

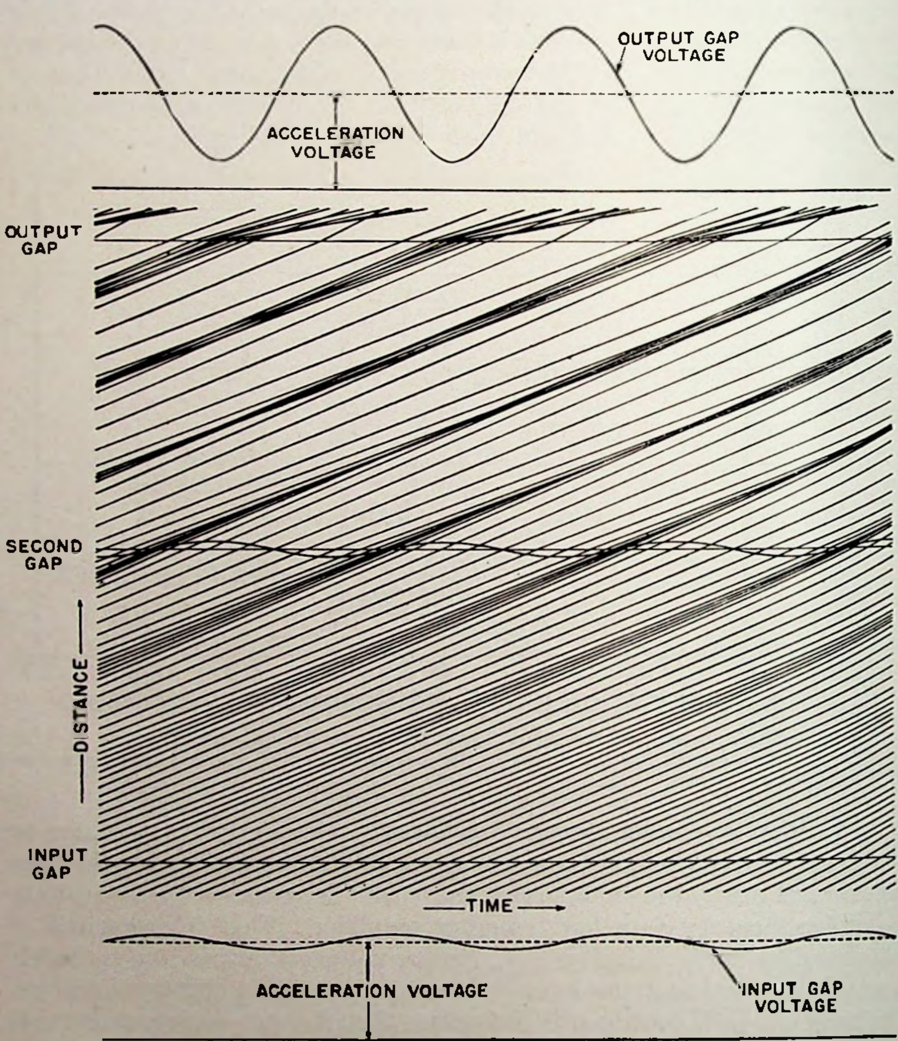


FIG. 9–6.—Applegate diagram for a cascade amplifier with increased drive power when the second resonator is detuned to prevent overbunching.

corresponds to the overbunched case in a two-resonator amplifier when the bunching parameter is approximately 7.02, the value for the second zero in the J_1 Bessel function. As the bunching is decreased by detuning the second resonator in either direction, the output fluctuates in a manner similar to the variations for the two-resonator amplifier illustrated in Fig. 4–4. Then the output increases to two large maximum values.

The output maximum when the second resonator is tuned to a lower frequency than the input is very much smaller than the maximum output on the high-frequency side of resonance. The largest maximum represents a much greater efficiency than that of a two-resonator klystron and is somewhat greater than the efficiency for the case illustrated in Figs. 9–3 and 9–4.

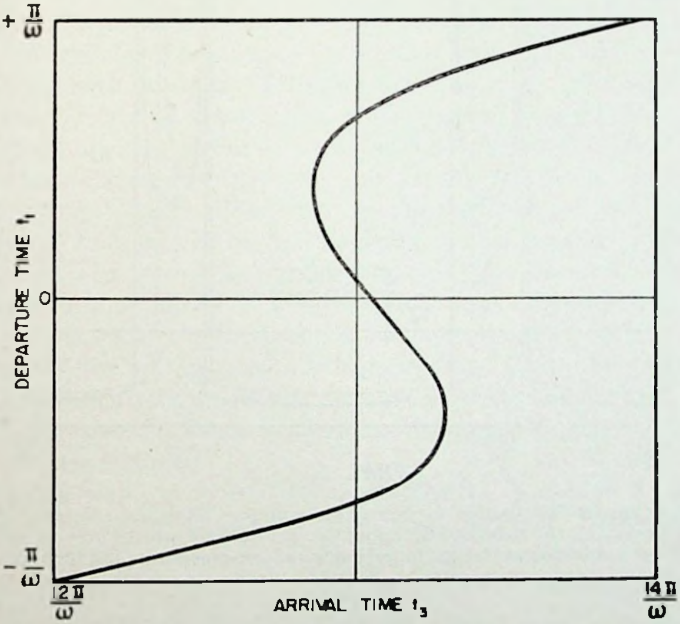


FIG. 9–7.—Arrival-time curve for the detuned cascade amplifier.

9.16. Improved Efficiency of Cascade Amplifiers.—This improvement in the efficiency of a cascade amplifier klystron is caused by the increased bunching at the second gap before the cascade bunching is superimposed on the electron beam. The beam current is no longer uniform at the second gap, and the rebunching action is applied to electrons that would not have become a part of the electron bunch in a single-stage amplifier. As a result, more electrons are removed from the "wrong" half of the cycle and become a part of the electron bunch. This effect can be observed in the Applegate diagram in Fig. 9–6 for the case when the drive power is increased and the second resonator is detuned to avoid overbunching. Detuning on the high-frequency side of resonance improves the phase relationship between the voltage at the second resonator gap and the partly bunched beam.

An arrival-time curve for this case is shown in Fig. 9–7, which appears quite similar to Fig. 9–4 for the case when the drive power is less and the second resonator is tuned to resonance. The only difference that is very apparent is in the phase relations. The electron that left the input gap at zero time ($t_1 = 0$) is not the center of the bunch in Fig. 9–4, but the

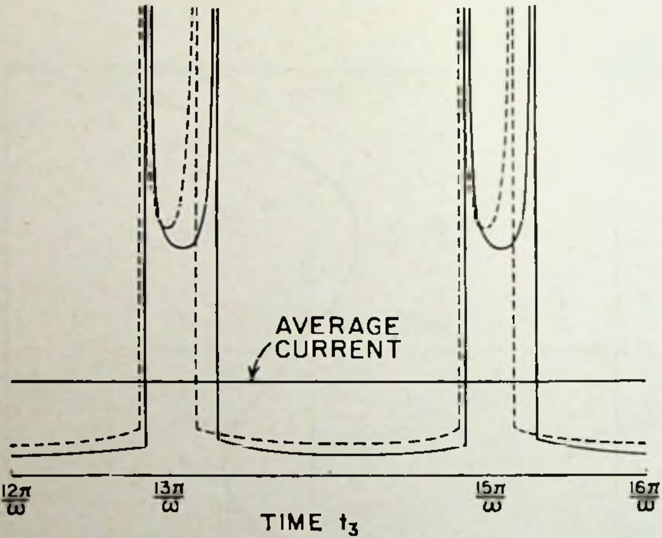


FIG. 9–8.—Current distribution for a cascade amplifier when the efficiency has been improved by overdriving the tube and detuning the second resonator to prevent overbunching. The theoretical curve for optimum bunching in a two-resonator amplifier is shown as a dotted line.

phase relation in Fig. 9–7 is quite similar to the case of a two-resonator klystron as indicated in Fig. 3–2. This phase relationship is also indicated by the Applegate diagram for the detuned case illustrated in Fig. 9–6. Note in Fig. 9–6 that the input gap voltage is greater than in Fig 9–3, but the voltage at the second gap is reduced by the detuning and the bunching from the two resonators is more nearly in phase.

9.17. Bunched Current Distribution.—Adding the magnitudes of the slopes in the arrival-time curve in Fig. 9–7 gives the current distribution in the bunched beam. This result is illustrated in Fig. 9–8. The current distribution giving optimum output with a two-resonator amplifier is shown by a dotted curve for comparison. The principal contribution of the improved efficiency of the cascade amplifier is due to the smaller current in the region between bunches. The theoretical efficiency for the two-resonator case is 58 per cent. A graphical integration similar to that shown in Fig. 3–13 in Sec. 3.18 indicates a theoretical efficiency of 68 per cent for the current distribution in Fig. 9–8.

The actual increase in output will generally be greater than that indicated by the increase in the theoretical efficiency. The maximum efficiency is obtained only if the voltage at the output gap is equal to the beam voltage. When the gap voltage is less than the beam voltage, the load required for maximum output is constant and the voltage at the gap is proportional to the radio-frequency current component in the beam. Therefore the output will be proportional to the square of the radio-frequency component. As a result, increasing the current component from 0.58 to 0.68 will increase the output by a factor $(0.68/0.58)^2$ or 1.38, and the actual efficiency is 38 per cent greater than that of a two-resonator klystron.

9.18. Equivalent Bunching Voltage.—Another viewpoint may be used to explain the improved bunching of a cascade amplifier klystron. Owing to the partial bunching of the electron beam at the time the electrons pass the second gap, the sinusoidal voltage at this gap will act on electrons that left the input gap at very different times during the cycle, and the equivalent bunching voltage will be nonsinusoidal. The Applegate diagram in Fig. 9–6 illustrates this effect. For example, the electron that passes the second gap one-quarter of a cycle before the voltage is zero left the input gap one-third of a cycle before the voltage at the input gap was zero. In other words, the voltage acting on the electrons at the second gap would not be sinusoidal if referred to the time an electron left the first gap, but would have some resemblance to a saw-tooth wave that decreases suddenly and then increases slowly.

A saw-tooth wave produces the most efficient bunching action. Unfortunately, the cavity resonators that must be used at the microwave frequencies cannot produce a voltage with this wave form. However, the equivalent wave form of the two resonators of a cascade amplifier is quite an improvement over a sine-wave voltage.

The analysis proposed above, referring the voltage at the second gap to the time an electron left the first gap, does not give the true picture. The electron does not travel the entire distance with the velocity it acquires at the second gap. It is possible to calculate the voltage at a single gap which would be required to produce the cascade bunching from the data in the arrival-time curve in Fig. 9–7. The transit time for the total drift distance from the input gap to the output gap would be equal to $(t_3 - t_1)$. The equivalent velocity would be proportional to the reciprocal of this transit time, and the equivalent voltage is proportional to the square of the equivalent velocity.

These calculations have been made, and the equivalent bunching voltage for the case illustrated by Fig. 9–7 is shown in Fig. 9–9. The wave form has a strong second harmonic component; also higher harmonics are present;

but the approach to a saw-tooth wave is far from complete. The ideal wave form for bunching with this drift distance is shown as a dotted line. This ideal curve was obtained by assuming that all electrons corresponding to departure times from $-\pi/\omega$ to $+\pi/\omega$ arrived at the output gap simultaneously at a time $t_3 = 13\pi/\omega$. The ideal bunching voltage is not a perfect saw-tooth wave when the drift-space transit time is less than 10 oscillation cycles, although the deviation is not great for this case when the transit time is 6½ cycles.

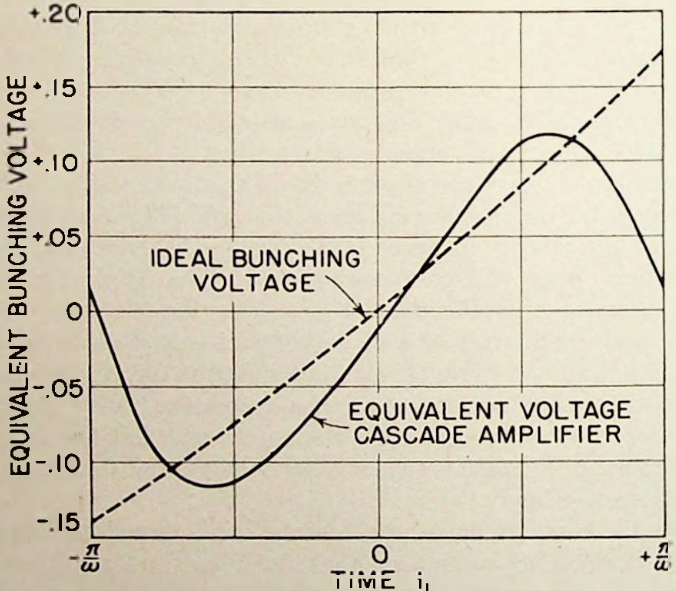


FIG. 9-9.—Equivalent nonsinusoidal voltage at the input gap corresponding to the arrival-time curve in Fig. 9-7.

9.19. Variation of Drive Power.—Although the bunching in a cascade amplifier is quite different from the bunching in a two-resonator tube, the output characteristic for changes in drive power is quite similar for the two types. However, the gain of a cascade amplifier is much greater and is quite dependent on the tuning of the second resonator. If this resonator is tuned to the input frequency, the output curve has a very steep slope corresponding to maximum gain at very low input power; then the electron beam becomes overbunched and the output decreases. Additional drive power produces a second smaller maximum. This curve is labeled 3,000 megacycles in Fig. 9-10. It has the appearance of the squared J_1 Bessel function and resembles the characteristic for a two-resonator klystron as shown in Fig. 4-4, but the gain is very much greater.

Detuning the second resonator 1 megacycle decreases the gain of the

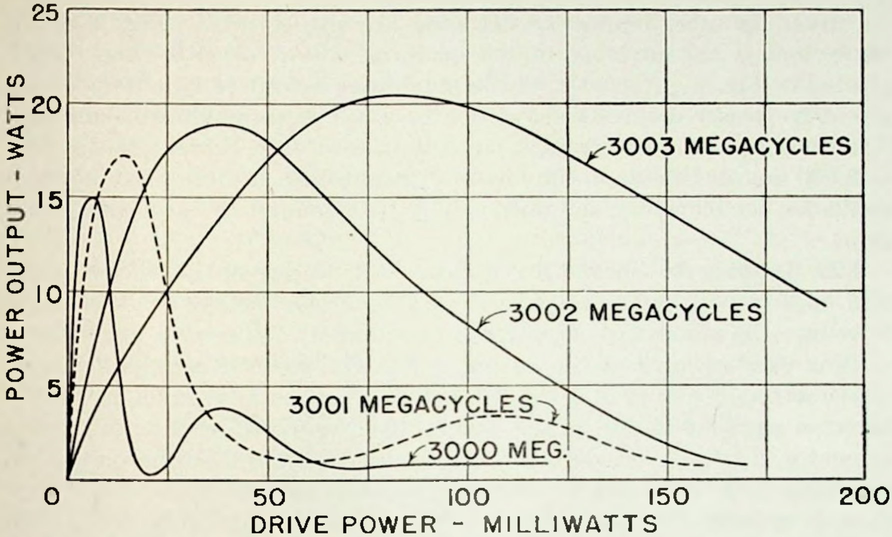


FIG. 9-10.—Experimental curves of power output of a Type 2K35 cascade amplifier as a function of drive power. The four curves represent different tuning adjustments of the second resonator.

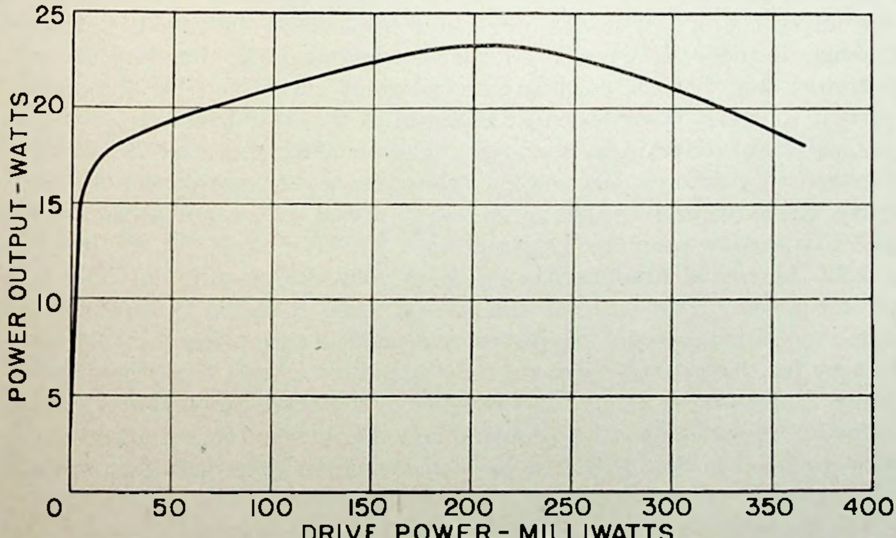


FIG. 9-11.—Experimental curve of the power output of a Type 2K35 cascade amplifier as a function of drive power. The tuning of the second resonator has been adjusted to give maximum output for each value of drive power.

tube, but the improved cascade bunching makes the output greater than is obtained when this resonator is tuned to the input frequency. This curve is a dotted line in Fig. 9–10, and is labeled 3,001 megacycles.

Curves for other degrees of detuning are also shown in Fig. 9–10. A curve that is the envelope of the maxima of the curves in Fig. 9–10 is plotted in Fig. 9–11 and shows the maximum output to be obtained with any drive power if the second resonator is tuned to maximize the power. The tuning is different for each point on the curve. Power gains as high as 3,000 are obtainable in the low-level region and, if more drive power is available, the efficiency and output may be increased by sacrificing power gain.

9.20. Loading the Second Resonator.—An increase of the efficiency may also be obtained by loading the second resonator instead of detuning it to reduce the gain, permitting more bunching at the second gap without causing overbunching at the output gap. The result is almost the same. This method of improving the efficiency is particularly valuable when a klystron amplifier with a wide band width is required, because the second resonator, if not loaded, determines the band width of the tube since it has the highest Q. Loading this resonator increases the band width faster than it reduces the gain because the cascade bunching becomes more efficient.

9.21. Output Load Characteristic.—All the factors discussed previously affect the electron bunching at the third resonator but do not influence the output characteristic as a function of the load on the output resonator. Changes in the load do not affect the electron bunching; therefore, the output resonator may be considered a resonant circuit fed by a constant-current source. The equivalent diagram of the output resonator will be similar to the diagram in Fig. 4–3. The behavior of a cascade amplifier klystron with different loads will be the same as a two-resonator amplifier; and a Rieke diagram similar to Fig. 4–12 in Sec. 4.16 would apply equally well to a cascade amplifier klystron.

9.22. Low-level Amplifiers.—The high gain of a cascade amplifier tube at low-power levels suggests the use of these tubes as radio-frequency amplifiers for receivers. However, the present tube types are not satisfactory for this purpose because they introduce a very large "shot noise" which limits the sensitivity of a receiver. The noise problem in a cascade amplifier is similar to the problem of a two-resonator amplifier, which was discussed in Sec. 4.18, and the conclusions in that discussion are also applicable to cascade amplifier klystrons.

AMPLIFIER-MULTIPLIER TUBES

9.23. Multiplier Drive Power Requirements.—A klystron frequency multiplier usually requires more drive power than an amplifier with the same input frequency because the drift distance is shorter in order to reduce the effects of space-charge debunching. The decreased drift distance requires a larger voltage at the input gap for optimum bunching, and this effect offsets any decrease in voltage permitted because a multiplier requires a smaller value of the bunching parameter than an amplifier. Fortunately, the cascade bunching principle is equally effective in a frequency-multiplier klystron and allows a low drive power to be used.

A typical amplifier-multiplier klystron resembles a three-resonator cascade amplifier, but the third resonator is made smaller and is tuned to a harmonic of the input frequency. The first two resonators act as an amplifier at the input frequency. A few milliwatts of power at the input will produce sufficient voltage at the second resonator to bunch the electron beam adequately in the drift space between the second resonator and the harmonic output resonator. This type of tube may have a power gain; *i.e.*, the output at the harmonic frequency may be greater than the drive power required. However, the output power of a klystron multiplier is not very large, and the output efficiency based on the total power input to the electron beam is low.

9.24. Cascade Bunching Characteristics.—An amplifier-multiplier klystron combines the bunching properties of a multiplier, as described in Chap. 5, with the behavior of a cascade amplifier discussed in the previous sections. The simple Bessel function relations in Chap. 5 do not apply, but the general form of the curves is very similar. This result may be expected from the comparison of the beam-current curves in Fig. 9–8, which shows that the current distribution is not radically different for simple bunching or cascade bunching. Space-charge effects probably modify the characteristics more than the cascade bunching.

Regardless of the effects of space charge on the output of a klystron frequency multiplier, certain generalizations are true when cascade bunching is used. The drive power is reduced. The efficiency is improved and may be increased still more if drive power can be sacrificed and the gain of the amplifier stage reduced by loading or detuning. It is these characteristics of cascade amplifiers which are added to the usual multiplier characteristics when the two types are combined in a single tube using the same electron beam for the amplifier and the multiplier.

A word of caution is necessary, however. The first two resonators must be a satisfactory amplifier or these advantages will not be obtained. Klystron frequency multipliers with input frequencies as low as 300 megacycles usually have resonators that are capacitively loaded to reduce the physical size of the tube. Also, the length of drift space required for a 300-megacycle klystron amplifier would be rather impractical. The combination of a short drift distance and low-impedance resonators makes amplification at this frequency impractical in any klystron of reasonable size. Therefore the amplifier-multiplier klystron is most useful at higher frequencies.

CHAPTER 10

MODULATION OF KLYSTRONS

10.1. Types of Modulation.—Radio-frequency energy may be used to transmit information by superimposing another signal upon the radio-frequency carrier in any one of a number of familiar means of modulation. Perhaps the simplest form of modulation is merely keying the transmitted energy is some manner. If the keying determines whether the carrier is on or off, the system is a form of amplitude modulation. In general, amplitude modulation refers to controlling the amplitude of the carrier envelope in proportion to the amplitude of the modulating signal.

Another modulation system, known as "frequency modulation," does not affect the amplitude of the carrier but changes the instantaneous frequency in proportion to the amplitude of the modulating signal. Phase modulation is a somewhat similar scheme in which the phase of the carrier is proportional to the modulating signal. Phase modulation also introduces a variation of frequency, but the frequency deviation is proportional to the product of the modulation frequency and the amplitude of the modulating signal.

Phase modulation and frequency modulation are produced in a different manner, but phase modulation may be used to produce frequency modulation if a compensating network is introduced so that the frequency deviation is independent of the modulation frequency. The differences between these two types of modulation are discussed in more detail in later sections of this chapter.

Radio energy may also be modulated by keying at a very rapid rate with the modulation information contained in the manner in which the keying pulses are arranged. This method is known as "pulse modulation." It differs from simple keying because the aggregate effect of a large number of closely spaced pulses is used to convey the information, and each pulse is not identified individually as in normal keying.

10.2. Modulation Coefficients.—The magnitude of the different types of modulation is an important consideration in their analysis and is usually expressed by some coefficient that describes the amount of modulation. In amplitude modulation, this coefficient is the percentage of modulation. In frequency modulation, the degree of modulation is determined by the maximum frequency deviation Δf. The frequency deviation is not the only important factor, and a coefficient known as the "modulation index," which is defined as the ratio of the frequency deviation to the modulation frequency $\Delta f/f_m$, is used to describe the amount of frequency modulation present in a radio signal. The modulation index is also useful in analyzing phase modulation and has the same value of $\Delta f/f_m$, although the relation between the modulation index and the modulation signal is different for phase modulation and frequency modulation.

10.3 Multiple Side Bands.—All types of modulation introduce side bands, or frequencies on either side of the carrier frequency. Amplitude modulation produces a single pair of side bands. One side band has a frequency equal to the sum of the carrier frequency and the modulation frequency; the other side band has a frequency equal to the carrier fre-

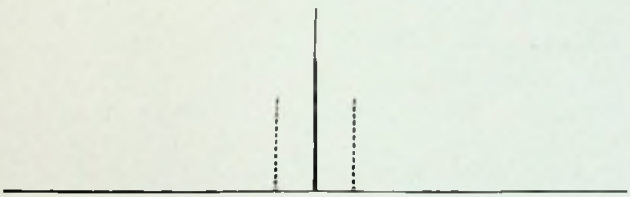


FIG. 10–1.—Carrier and side-band frequencies for an amplitude-modulated signal. The amplitude of the unmodulated carrier is also shown for comparison.

quency minus the modulation frequency. This relation is indicated by Fig. 10–1. The amplitude of the unmodulated carrier is shown for comparison. Other types of modulation have more than a single pair of side bands. The distribution of side bands for a frequency-modulated wave may become quite large. A typical example showing five pairs of side bands is given in Fig. 10–2*A*. This illustration corresponds to a value of 3.0 for the modulation index.

The number of side bands is determined by the modulation index. If the modulation index is the same as illustrated in *A* in Fig. 10–2 but the modulation frequency is half as great, there will be the same number of side bands as in *A* spaced more closely as shown in *B*. However, if the modulation frequency is half as great but the frequency deviation is the same as in *A*, the modulation index $\Delta f/f_m$ will be doubled and the number of side bands will be increased. This result is shown in Fig. 10–2*C* for a modulation index of 6.0. Note that the frequency band occupied by the side bands depends upon the frequency deviation and not upon the modulation index.

Figure 10–2 may also be used to illustrate the difference between frequency modulation and phase modulation. If the modulation voltage is constant, the frequency deviation will be constant in a frequency-modulation system and the modulation index $\Delta f/f_m$ will be inversely proportional to the modulation frequency. Figures 10–2*A* and *C* correspond to a constant modulation voltage in a frequency-modulation system. With phase modulation, the frequency deviation Δf is proportional to the modulation frequency; therefore, the modulation index is constant if the modulation voltage is unchanged, and Figs. 10–2A and B correspond to a constant modulation voltage in a phase-modulation system.

10.4. Combining Two Types of Modulation.—The distribution of side bands is symmetrical in a signal with pure amplitude modulation or pure

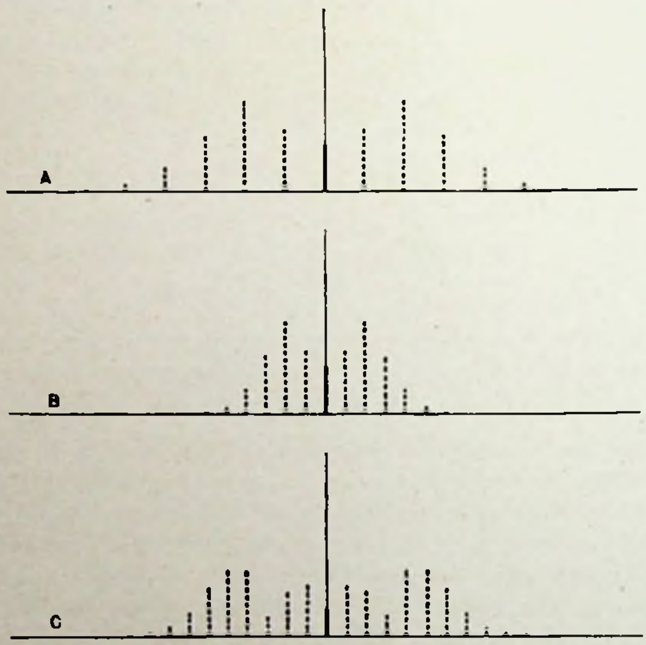


FIG. 10–2.—Carrier and side-band frequencies for frequency-modulated signals, showing multiple side bands.

frequency modulation, as shown in Figs. 10–1 and 10–2. However, if a small amount of frequency modulation were superimposed on an amplitude-modulated signal, the two side bands would not be of equal amplitude, as shown in Fig. 10–1, but would differ in amplitude. Similarly, a small amount of amplitude modulation superimposed on a frequency-modulated signal would cause the patterns in Fig. 10–2 to become unsymmetrical. The effect of the undesired amplitude modulation in a frequency-modulation system can be overcome by using a limiter in the receiver, but in an amplitude-modulation receiver the presence of frequency modulation causes distortion and can be quite undesirable.

10.5. Frequency Modulation in Klystrons.—Two factors make it difficult to obtain pure amplitude modulation with any microwave vacuum tube. This difficulty exists with microwave triodes and magnetrons as well as with klystrons. The first effect is introduced by the presence of the electron beam in the resonator gap. The electron beam changes the effective dielectric constant of the capacity represented by the resonator gap, and variation of the beam current changes the resonant frequency of the cavity. This effect is present in any tube utilizing cavity resonators as an integral part of the tube; therefore, a variation of beam current produces frequency modulation.

Frequency modulation caused by voltage changes in any device that depends upon transit time in its operation is the second factor that makes pure amplitude modulation difficult to obtain. Magnetrons exhibit this characteristic as well as klystrons. It is also true of triodes operated in the microwave region, since a triode is quite dependent on transit time at these frequencies. The difficulty in obtaining pure amplitude modulation with a klystron is not a handicap, since the transit-time dependence simplifies the use of frequency modulation with these tubes. As a result, klystrons can be frequency-modulated as easily as low-frequency triodes can be amplitude-modulated.

Frequency modulation is particularly convenient with reflex klystron oscillators because the modulation voltage can be applied to the reflector electrode. No current is drawn by this electrode and the power requirements are therefore small. These advantages of reflex klystrons, plus their simplicity and ease of tuning, make this type of tube more desirable than microwave magnetrons and triodes for many applications.

10.6. Demodulation.—Klystron tubes may also be used as detectors or mixers. If an amplitude-modulated signal is applied to the input resonator of a klystron, both the carrier and the modulation frequency are present in the bunched electron beam, and a velocity-sorting electrode will detect the modulation frequency. When a klystron tube is used as a mixer, a different radio frequency from a local oscillator may be introduced at a second resonator and the difference frequency will be detected by the velocity-sorting electrode and may be coupled to an intermediate-frequency amplifier.

FREQUENCY MODULATION

10.7. Inherent Characteristics of Klystrons.—The fact that klystrons are easily frequency-modulated but pure amplitude modulation is difficult to obtain serves to emphasize an important factor in superhigh-frequency design. It is not wise to attempt to use familiar techniques merely because they are successful at the lower frequencies. Applications should be modified to utilize the characteristics of new tubes to the best advantage, rather than attempt to adapt the older techniques. In the case of klystron oscillators, the electrical characteristics are ideal for frequency modulation, and it is desirable to design equipment with this factor in mind.

In order to obtain the improvement in the signal-to-noise ratio, which is possible when frequency modulation is used, the modulation index must be much greater than unity. This means that the frequency deviation must be large compared to the highest modulation frequency to be used if the noise-suppression feature of the frequency modulation is to be obtained. Large frequency deviations are readily obtained with klystron oscillators. Reference to the frequency curves in Figs. 7–9 and 8–11 indicates that frequency deviations of several megacycles are possible. Therefore high modulation frequencies are possible with these tubes when frequency modulation with a modulation index greater than unity is used.

It is not necessary to use large values of the modulation index $\Delta f/f_m$ when using frequency modulation. If the modulation index is small there is very little difference between frequency modulation and amplitude modulation in quality or band-width requirements. If frequency modulation is the simplest to obtain, then it is logical to use this type of modulation. The receiver circuit may be somewhat more complicated but the advantages of simplicity when using frequency modulation with klystron oscillators are usually far more important.

10.8. Self-excited Klystron Oscillators.—The basis for the frequency deviation that occurs when the voltage on a klystron oscillator is varied is evident from the phase relations in any self-excited oscillator. The total phase angle of the tube and feedback circuit must be an integral number times 2π radians. The phase angle introduced by a triode at low frequencies is relatively independent of the electrode voltages, but the drift time phase angle of a klystron varies over a wide range if the voltage is changed. Then the frequency of oscillation must change so that the phase shift introduced by the circuit equalizes the phase shift produced by the voltage change.

10.9. Frequency Deviation.—Relations giving the frequency deviation in terms of the voltage variations were derived for reflex klystrons and two-resonator oscillators in Chaps. 7 and 8. These relations apply only to the linear portion of the characteristic, but this region is the one of interest for frequency modulation. If the reflector voltage on a reflex klystron is varied, the frequency deviation can be obtained from a modified form of Eq. (7–40):

$$\Delta f = \frac{\pi N f}{Q_L} \frac{\Delta E_r}{E_0 + E_r} \tag{10–1}$$

Variation of the beam voltage of a reflex klystron also changes the frequency and the relation is obtained from Eq. (7–41).

$$\Delta f = \frac{\pi N f}{2Q_L} \frac{E_0 - E_r}{E_0(E_0 + E_r)} \Delta E_0 \tag{10–2}$$

The relations in Eqs. (10–1) and (10–2) assume that the reflecting field is uniform and neglect space-charge effects. In addition, Eq. (10–2) neglects the tuning effect of the electrons in the resonator gap, which changes the resonant frequency of the cavity when the beam current varies owing to changes in the beam voltage.

A two-resonator oscillator can be frequency-modulated by varying the beam voltage, and the deviation is

$$\Delta f = \frac{\pi N f}{2\sqrt{Q_1 Q_2}} \frac{\Delta E_0}{E_0} \tag{10–3}$$

This relation was derived in Eq. (8–26). Note that the geometric mean

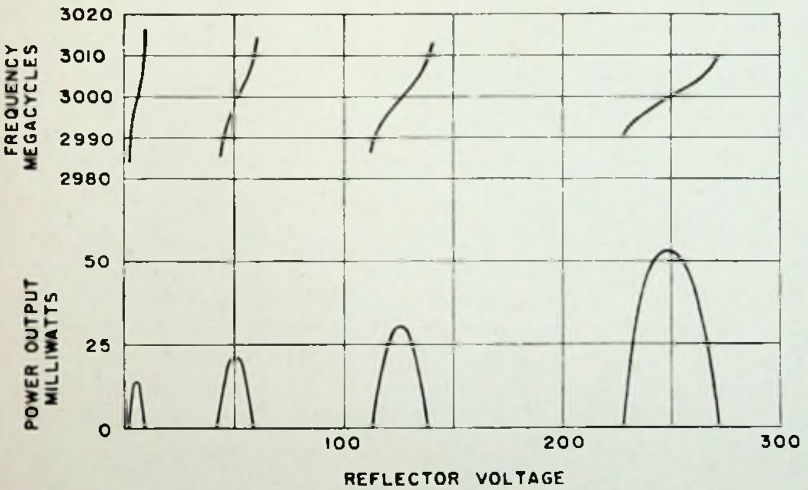


FIG. 10–3.—Frequency-modulation characteristic of a reflex klystron oscillator.

of the Q of the input and output resonators is the effective Q of the oscillator. Since the input resonator is not loaded, the two-resonator oscillator is more stable than a reflex oscillator under most operating conditions. Theoretically, if the reflector voltage is equal to the beam voltage on a reflex klystron, the beam voltage does not affect the frequency. This result is indicated by Eq. (10–2) and is also illustrated by the curve for N equal to 2¾ in Fig. 6–3.

Typical curves showing the frequency-modulation characteristics of klystron oscillators are given in Figs. 10–3 and 10–4. Note that the frequency deviation is not linear over the entire range of oscillation. The curves are similar to tangent curves, but considerable frequency deviation can be obtained in the linear regions of the characteristics, and the corresponding amplitude variations are small.

10.10. Modulation Index.—The ratio of the maximum frequency deviation to the modulation frequency $\Delta f/f_m$ is known as the "modulation index." This factor determines the number of side bands that exist when a carrier is frequency-modulated. If the modulation voltage superimposed upon the reflector voltage is denoted E_m, the modulation index is proportional

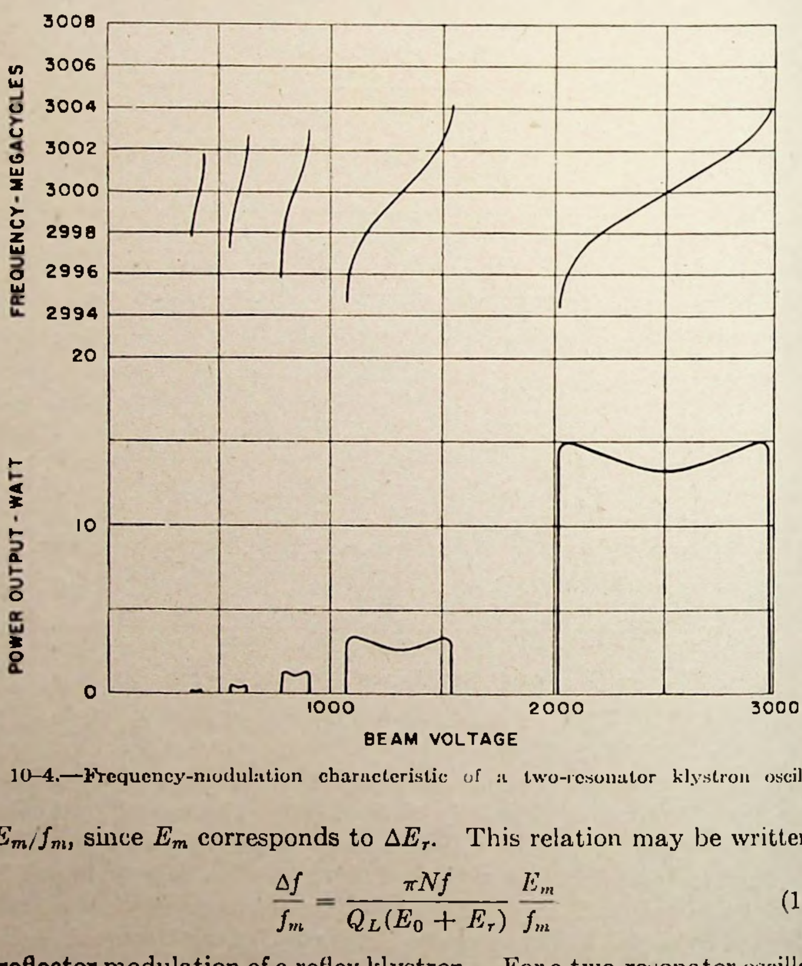


FIG. 10–4.—Frequency-modulation characteristic of a two-resonator klystron oscillator.

to E_m/f_m, since E_m corresponds to ΔE_r. This relation may be written

$$\frac{\Delta f}{f_m} = \frac{\pi N f}{Q_L(E_0 + E_r)} \frac{E_m}{f_m} \tag{10-4}$$

for reflector modulation of a reflex klystron. For a two-resonator oscillator,

$$\frac{\Delta f}{f_m} = \frac{\pi N f}{2E_0\sqrt{Q_1 Q_2}} \frac{E_m}{f_m} \tag{10-5}$$

A constant value of modulation voltage E_m will give the same value of frequency deviation as the modulation frequency is varied. A low modulation frequency will correspond to a large value of the modulation index and will produce many side bands, but these side bands will be closely spaced and will occur within approximately the same band width required by a higher modulation frequency.

10.11. Frequency Multiplication.—Frequency modulation of klystrons is not limited to oscillators but may be used also with klystron frequency multipliers. Some frequency-modulation systems use a very small frequency deviation at a low frequency and depend upon frequency multiplication to increase the frequency deviation. If the frequency deviation is not increased sufficiently by multiplying from 1 to 60 megacycles, for example, the difference frequency between the modulated 60 megacycles and a fixed oscillator with a frequency of 59 megacycles may again be multiplied. In this manner a multiplication of the deviation by a factor of 3,600 may be obtained when the final carrier frequency is 60 megacycles.

Klystron frequency multipliers make possible tremendous ratios of frequency multiplication. For example, an oscillator with a frequency of 1 megacycle may be used to drive a series of frequency multiplier stages with a multiplication ratio of 300 and an output frequency of 300 megacycles. A single klystron frequency multiplier will give an additional multiplication factor of 10 with an output frequency of 3,000 megacycles. This system has a direct multiplication ratio of 3,000, which is adequate for increasing the frequency deviation without resorting to the frequency mixing method required for the lower frequencies.

PHASE MODULATION

10.12. Phase Relations in Klystrons.—Although phase modulation and frequency modulation are closely related, and phase-modulation methods may be used to obtain a frequency-modulated signal if corrective networks are used, the two methods of modulation are quite different. Frequency modulation is an inherent characteristic of klystron oscillators when the beam voltage is varied, while the same variation of voltage produces phase modulation in a klystron amplifier. This difference is caused by a difference in the phase relations between the input and output voltages in the two types of tubes.

Frequency modulation occurs in klystron oscillators because the bunching frequency at the input resonator can vary as the drift time is changed. If the drift time is decreased, for example, the tube will oscillate at a new frequency which will introduce a change in the phase angle of the feedback to compensate for the change in the drift-space transit angle. As a result, a change in the beam voltage of an oscillator produces a change in frequency, and the tube continues to oscillate at this new frequency as long as the voltage change is maintained.

The input frequency of an amplifier cannot change owing to any variation of the beam voltage because the bunching frequency is determined by a separate driver source which does not vary. If the beam voltage of an amplifier is changed to some new fixed value, the phase of the radio-frequency voltage at the output resonator will change with respect to the input gap voltage, but succeeding cycles will have the same phase difference and the output frequency will be identical to the input frequency.

10.13. Rate of Change of Phase Produces Frequency Deviation.—As indicated in the preceding paragraph, a difference in the phase angle does not change the output frequency of an amplifier. It is the instantaneous phase angle that is important in phase modulation, not the phase with respect to a fixed reference. The instantaneous frequency will differ from the average frequency (input frequency) only if the phase angle between successive cycles is greater or less than 2π radians. This instantaneous phase variation occurs only if the phase is changing. Therefore the instantaneous frequency of the output depends upon the rate of change of the phase angle. If the drift time could be decreased continuously at a constant rate, the output frequency would be constant and higher than the input frequency.

10.14. Transit-time Relations.—The phase angle at the output gap depends upon the transit time of the electrons in the drift space. The average transit time T_0 is determined by the drift distance s_0 and the average velocity.

$$T_0 = \frac{s_0}{v_0} = \frac{s_0}{\sqrt{(2e/m)E_0}} \tag{10-6}$$

If we consider only the phase of the electrons that are to become the center of the bunch, the radio-frequency voltage at the input gap will be zero and need not be included in the analysis. This simplification is equivalent to assuming that the phase modulation and the radio-frequency modulation of the beam can be superimposed, and it is accurate for small variations of voltage.

A modulation voltage $E_m \sin \omega_m t$ added to the beam voltage E_0 will cause the transit time through the drift space to vary. This transit time will be designated T and is determined by a relation similar to Eq. (10-6).

$$T = \frac{s_0}{\sqrt{(2e/m)(E_0 + E_m \sin \omega_m t)}} = T_0 \left(1 + \frac{E_m}{E_0} \sin \omega_m t\right)^{-1/2} \tag{10-7}$$

Equation (10-7) may be expanded in the series form, giving

$$T = T_0 \left[1 - \frac{1}{2} \frac{E_m}{E_0} \sin \omega_m t + \frac{3}{8} \left(\frac{E_m}{E_0}\right)^2 \sin^2 \omega_m t + \cdots \right] \tag{10-8}$$

The higher power terms can be neglected when the modulation voltage is small, giving an approximate relation for the transit time.

$$T \simeq T_0 - \frac{T_0}{2}\frac{E_m}{E_0} \sin \omega_m t \tag{10-9}$$

10.15. Frequency Deviation.—The instantaneous angular frequency at the output gap is determined by the rate of change of the phase angle. In order to evaluate the variations of the output frequency it is necessary to relate the phase angle at the output gap and the transit time. The phase angle at the output gap will be determined by the phase angle at the input gap and by the transit time. This relation may also be stated in terms of time, *i.e.*,

$$t_1 = t - T \tag{10-10}$$

In other words, a time t at the output gap is related to an earlier time $(t - T)$ at the input gap.

Since the electrons that become the center of the bunch pass the input gap when the voltage is zero but pass the output gap when the field has its maximum retarding value, there is a phase difference of $\pi/2$ between the input and output voltages. Therefore the output phase angle τ_2 is given by the relation

$$\tau_2 = \omega_1 t - \omega_1 T_0 + \frac{\omega_1 T_0}{2}\frac{E_m}{E_0} \sin \omega_m t - \frac{\pi}{2} \tag{10-11}$$

Equation (10–11) may also be written in terms of N, the number of cycles corresponding to the average transit time.

$$\omega_1 T_0 = 2\pi N \tag{10-12}$$

$$\tau_2 = \omega_1 t - 2\pi N + \pi N \frac{E_m}{E_0} \sin \omega_m t - \frac{\pi}{2} \tag{10-13}$$

An expression for the instantaneous angular frequency at the output gap is obtained by differentiating τ_2 with respect to time.

$$\omega_2 = \frac{d\tau_2}{dt} = \omega_1 + \omega_m \pi N \frac{E_m}{E_0} \cos \omega_m t \tag{10-14}$$

Therefore the instantaneous frequency deviation Δf_2 is determined by

$$\omega_2 - \omega_1 = 2\pi \Delta f_2 = 2\pi f_m \pi N \frac{E_m}{E_0} \cos \omega_m t \tag{10-15}$$

$$\Delta f_2 = f_m \pi N \frac{E_m}{E_0} \cos \omega_m t \tag{10-16}$$

The maximum frequency deviation occurs when

$$\cos \omega_m t = 1 \tag{10-17}$$

and the expression for the maximum frequency deviation Δf is obtained by substituting Eq. (10–17) in Eq. (10–16).

$$\Delta f = f_m \pi N \frac{E_m}{E_0} \tag{10–18}$$

Comparison of Eqs. (10–18) and (10–3) emphasizes the differences between phase modulation and frequency modulation. The frequency deviation is proportional to the modulation voltage in both cases. The frequency deviation obtained with frequency modulation is independent of the modulation frequency but is inversely proportional to the effective Q of the circuits. Phase modulation gives a frequency deviation that is independent of the Q of the circuits, but it is proportional to the modulation frequency. Phase modulation is not entirely independent of the circuits, since the Q will affect the magnitude of the side bands. However, the frequency deviation will be unaffected by the Q of the output circuit if the Q is low enough and none of the side bands are suppressed. Phase modulation is completely independent of the Q of the input resonator.

10.16. Modulation Index.—The ratio of the maximum frequency deviation to the modulation frequency $\Delta f/f_m$ is the modulation index for phase modulation as well as for frequency modulation. Equation (10–18) may be rewritten

$$\frac{\Delta f}{f_m} = \pi N \frac{E_m}{E_0} \tag{10–19}$$

The amplitude of the current at the nth side-band frequency is designated A_n and is determined by the magnitude of the modulation index in accordance with the relation

$$A_n = J_n\left(\frac{\Delta f}{f_m}\right) \tag{10–20}$$

where J_n is the nth order Bessel function of the first kind. $\Delta f/f_m$ must be equal to 1.84 for maximum output at the first side band. Note the similarity between the expressions for the bunching parameter and the modulation index for phase modulation. It is interesting to note that the modulation voltage for maximum output at the first side band is the same as the radio-frequency voltage required for optimum bunching.

10.17. Multiple Side Bands.—A number of pairs of side bands exist simultaneously if the modulation index is large. These multiple side bands were discussed in Sec. 10.3 and illustrated in Fig. 10–2. However, it is not necessary for all of the side bands to be resent in the resonant circuit when phase modulation is used, although all the side-band components are present in the beam. If the modulation frequency is sufficiently high, so that the side bands are widely separated, it is possible to select a single side band with a high Q circuit and suppress the carrier and other side bands. A number of single side-band systems are described in more detail in Sec. 10.18.

The ability of a klystron amplifier to select a single side band and exclude the others gives a striking demonstration of the existence of multiple side bands. If a modulation frequency of 30 megacycles is used with a klystron amplifier operating at 3,000 megacycles, for example, the output resonator may be tuned successively to the carrier frequency of 3,000 megacycles and to side-band frequencies of 2,910, 2,940, 2,970, 3,030, 3,060, and 3,090 megacycles. The Q of the output resonator will be sufficiently high so that only one side band is observed at a time, and the distribution of side bands illustrated in Fig. 10–2 may be experimentally verified.

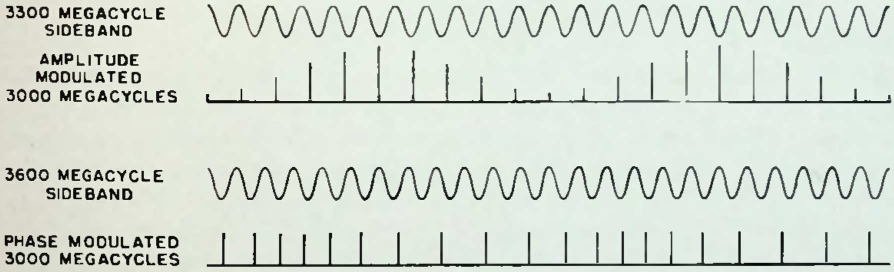


FIG. 10–5.—Transfer of energy from a modulated, bunched electron beam to a side-band frequency. Amplitude modulation and phase modulation are shown.

A simple graphical illustration of the phase relations between the electron bunches in a klystron amplifier can be used to explain why an amplitude-modulated electron beam can give up energy to only one pair of side bands, while a phase-modulated electron beam can excite a resonator at the frequency of a second side band. A carrier frequency of 3,000 megacycles will be assumed and a modulation frequency of 300 megacycles will be chosen for the illustration in order to emphasize the effect. The amplitude-modulated electron beam can transfer energy to a frequency of 3,300 megacycles but not to the 3,600-megacycle side band, while the phase-modulated beam can excite the frequency of the second side band.

Ten electron bunches pass the output gap during one cycle of the modulating voltage. These bunches are represented by 20 vertical lines with different amplitudes in Fig. 10–5, corresponding to two cycles of the modulation frequency. The amount of energy transferred to the output resonator in an amplitude-modulated klystron depends on the phase of the bunch with respect to the radio-frequency voltage as well as to the amount of energy in the electron bunch. The energy in an electron bunch depends upon the current (*i.e.*, the number of electrons) in that electron bunch, which is determined by the modulation voltage and indicated by the amplitude of the vertical line representing the electron bunch.

The high-frequency side band (3,300 megacycles) is shown directly above the amplitude-modulated bunches. When the radio-frequency voltage at the output gap is negative, indicated by the lower half of the sine wave, the output resonator will absorb energy from the bunched beam by slowing down the electrons in the bunch. If the bunches coincide with the upper or positive half of the cycle, they will be speeded up and energy will be transferred from the resonator to the electron beam. Note that the amplitude of the electron bunches is small when the output gap voltage is positive; *i.e.*, very little energy is transferred from the resonator to the electron beam when the phase is incorrect. The energy in the electron bunches is large when they have the correct phase; therefore, the amplitude-modulated electron beam will transfer energy to the 3,300-megacycle side band.

Note in Fig. 10-5 that the amplitude-modulated bunches cannot be phased properly to transfer energy to the 3,600-megacycle side band which is shown below. Amplitude modulation does not produce more than a single pair of side bands. This result was shown in Fig. 10-1.

The same number of bunches is shown for the phase-modulated case, but the amplitude is constant and the phase of the bunches has been chosen to correspond to the conditions for optimum output at the second side band. The output frequency in this case is 3,600 megacycles. A large percentage of the bunches are phased so that the output gap voltage is negative and energy is extracted from the beam. Other bunches are approximately 90 deg out of phase and have no effect. A few bunches pass the output gap during the wrong half of the cycle, but the net effect is a transfer of energy from the beam to the resonator tuned to the second side band on the high-frequency side of the carrier.

10.18. Applications of Phase Modulation.—Although klystron amplifiers can be phase-modulated readily, this type of modulation is not very suitable for audio frequencies without special circuits. If a modulation voltage one-tenth as great as the beam voltage were available, the frequency deviation would be the same order of magnitude as the modulation frequency. Larger modulation voltages are not desirable; therefore, the method is not satisfactory for audio modulation.

It is possible to utilize the phase modulation characteristic of klystron amplifiers with double modulation systems. The low-frequency modulation may be applied to a high-frequency subcarrier, using either amplitude modulation or frequency modulation; then the subcarrier is used to phase-modulate the klystron amplifier. The output resonator is tuned to one of the first pair of side bands. If the subcarrier frequency is high compared to the band width of the resonator, the other components will be suppressed and the side band that is selected will be identical to a different carrier frequency with modulation.

Phase modulation is quite useful with klystron frequency multipliers, particularly with crystal-controlled klystron multipliers for frequency standards. If a submultiple of the input frequency to the multiplier is used to phase-modulate the klystron, output frequencies may be obtained at intervals separated by the submultiple of the input frequency according to the relation

$$f_2 = Mf_1 \pm n\frac{f_1}{k} \tag{10-21}$$

where f_1 and f_2 are the input and output frequencies, k is determined by the submultiple used, and M and n are integers which may have any reasonably small value. The value of n may also be zero.

10.19. Synchrodyne Systems.—Another important application of phase modulation is known as a "synchrodyne" system. A second frequency is synchronized with a control frequency so that the two frequencies are always separated a fixed amount. This type of system is valuable as a means of furnishing a local oscillator frequency which always differs from the transmitter frequency by the intermediate frequency of the receiver. If the transmitter and receiver are at the same location, as in a radar system, the transmitter frequency may vary but the receiver will always be properly tuned without using automatic frequency control. If the transmitter frequency is quite stable, as required for narrow channel communication systems, a phase-modulation scheme similar to the synchrodyne system may be used. The local oscillator frequency in this case would be synchronized with another stable source, perhaps the frequency of the transmitter at the receiving location, but would be mixed with the signal received from a transmitter at a different location to give an intermediate frequency.

10.20. Phase Modulation with Oscillators.—It has been stated that klystron oscillators give frequency modulation because the frequency of oscillation can change to counteract the variations of transit time. If the modulation frequency becomes so high that the Q of the circuits does not permit the oscillation frequency to vary appreciably, the transit-time variations then introduce a phase-modulation component in the electron beam. The side bands will not be observed in such an oscillator because they are suppressed by the narrow band width of the resonant circuit. The only noticeable effect will be a decrease in the output which is caused by the decrease in the amplitude of the carrier when phase modulation is present. Since the amplitude of the carrier is proportional to the J_0 Bessel function, the decrease will not be serious unless the modulation index becomes large.

When a third resonator is available, as in an oscillator-buffer type of klystron, a phase modulation side band may be utilized by using the first two resonators as a self-excited oscillator and tuning the third resonator to one of the side bands. Since the modulation index cannot be very large without suppressing the oscillation, only the first pair of side bands have any appreciable output. This method of operation is satisfactory when the modulation frequency is quite large, for example, 30 megacycles, and has been used for a synchrodyne system combining the transmitting oscillator and the local oscillator for the receiver in a single tube.

AMPLITUDE MODULATION

10.21. Klystron Oscillators.—Although klystrons are better adapted to frequency modulation than they are to amplitude modulation, their amplitude-modulation characteristics are of considerable importance. The properties of these tubes which are best suited to applications of amplitude modulation will be indicated, and the difficulties that are met will be discussed in order to indicate which characteristic should be avoided.

The amplitude variations in an oscillator with modulation of the beam voltage or reflector voltage are greatest when the frequency deviation is a maximum and the tube is quite unstable. This characteristic may be observed in Figs. 7–15 and 8–12. In addition, the modulation characteristic is not linear and will introduce considerable distortion. Modulation of the voltage of a klystron oscillator is therefore not recommended as a method of obtaining amplitude modulation.

As pointed out in Sec. 10.4, varying the beam current of a klystron oscillator will introduce frequency modulation. For example, a 3K30/410R klystron oscillator will change its operating frequency as much as 10 megacycles if the beam current is changed from 100 to 10 ma. However, this effect may be compensated by combining current modulation and voltage modulation in the proper phase. The current modulation will produce considerable amplitude modulation and some frequency modulation. The voltage modulation will affect the output somewhat, but its principal purpose is to introduce a source of frequency modulation that will cancel the frequency deviation caused by the current modulation.

10.22. Klystron Amplifiers.—Current modulation of klystron amplifiers is much more satisfactory than similar modulation of oscillators because the resonator detuning effect introduces phase modulation of the output instead of frequency modulation. Since the frequency deviation in a phase-modulated amplifier is proportional to the rate of change of the phase, the frequency deviation will be quite small if the modulation frequency is low. Audio-frequency modulation will not introduce serious frequency modulation if the amplifier is isolated adequately from the driving source by a buffer klystron.

A typical amplitude-modulation characteristic for a klystron amplifier is shown in Fig. 10–6. The vertical coordinate is plotted as the square root of the power output in order to show the approximate linearity of the modulation characteristic. Beam current is shown as the horizontal coordinate. Ideally, the beam current should be proportional to a control electrode voltage for the best modulation characteristic. However, the relatively simple electron-gun structure used in most klystrons does not

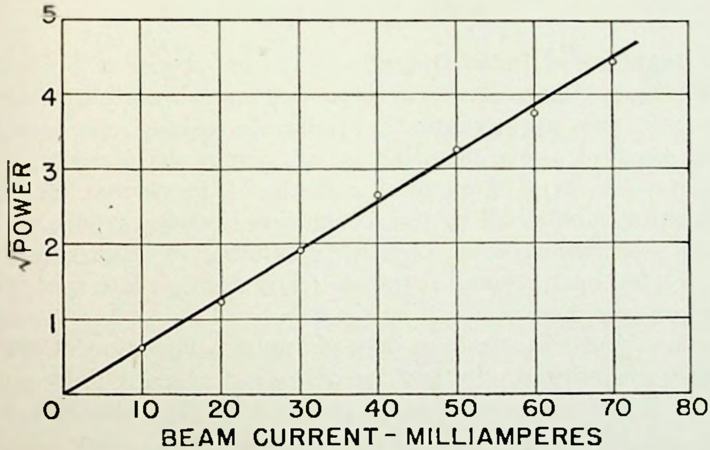


FIG. 10–6.—Amplitude-modulation characteristic of a klystron amplifier.

have a very good control characteristic. The control electrode does not have a linear characteristic, and affects the focus of the beam as well as the emission current. More complicated electron guns have not been designed because other methods of modulation have proved satisfactory.

10.23. Beam Voltage Modulation.—Varying the beam voltage of a klystron amplifier will cause the power output to vary as indicated by Fig. 4–5. The characteristic is nonlinear if the square root of the power output is plotted, and modulation would be satisfactory only if the percentage of modulation was quite small. Phase modulation would also be introduced by the variation of transit time accompanying the voltage variations, but this effect would not be serious for audio-frequency modulation.

10.24. Loss-type Modulation.—Several types of modulation that do not change the beam input but merely introduce a loss to reduce the output are possible, but obviously do not have many advantages. A reactance tube modulator is one example of such a system. If another cavity resonator with a controllable beam is coupled to a klystron amplifier, changes in the tuning or losses in the reactance-tube resonator will affect the output of the amplifier.

Another example of a loss-modulation scheme is the beam-deflection system. Current modulation is obtained by deflecting the beam at right angles to its axis so that only a fraction of the beam passes the output gap of the klystron. Although this method gives results, it is undesirable because the modulation decreases the efficiency of the tube and the peak output is no greater than the unmodulated output of the tube.

PULSE MODULATION

10.25. Definition of Pulse Operation.—Various types of pulse modulation, which have certain advantages in systems using multiplex channels for communication, are particularly applicable to klystrons because they overcome some of the difficulties in obtaining satisfactory amplitude modulation. This type of operation is distinct from normal keying, where the information is obtained by the recognition of coded groups. In pulse modulation, the arrangement of a large number of short pulses with a duration of a few microseconds is translated by a circuit into speech or some other form of information.

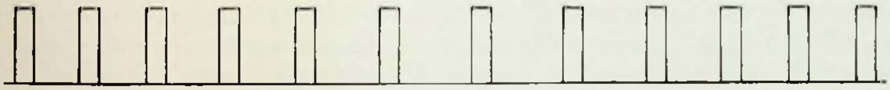
One form of pulse modulation depends upon a variation of the energy in the pulses when the amplitude of the pulses is changed. Pulse-amplitude modulation is not very satisfactory for use with klystrons because these tubes do not have linear amplitude characteristics. This difficulty is easily overcome by using pulse-length modulation, also called "pulse-width" modulation, where the variations in energy are introduced by using pulses of equal amplitude, but the duration of the pulses is controlled by the signal. A third method of varying the energy uses pulses of equal amplitude and duration, but the signal determines the number of pulses that occur during a chosen interval. This form of modulation is known as "pulse number modulation." All these types of modulation correspond to amplitude modulation because the energy is varied by the modulation process.

Pulse time modulation, also called "pulse position modulation," does not vary the amount of energy but uses pulses of equal amplitude and duration. The amplitude of the modulation signal determines the time these pulses occur. In this respect, pulse time modulation is analogous to phase modulation. A somewhat similar type of modulation, using different circuits for controlling the pulses, is known as "pulse frequency modulation."

Illustrations of two of the types of pulse modulation are given in Fig. 10–7. Two multiplex channels are shown in the illustration of pulse length modulation. One channel is shown dotted to help identify the two channels. Channel selection is obtained by using circuits that respond only during a short interval. Only one channel is shown in the illustration of pulse time modulation.

10.26. Band-width Considerations.—If steep-sided pulses are used, as illustrated in Fig. 10-7, the energy will be distributed over a band width approximately equal to the reciprocal of the pulse duration. For example, a pulse with a duration of 1 μsec will correspond to a band width of 1 megacycle. Therefore pulse modulation may be wasteful of the frequency spectrum unless a large number of channels are used with the same trans-

PULSE LENGTH MODULATION - TWO CHANNELS



PULSE TIME MODULATION

FIG. 10-7.—Pulse-length modulation and pulse time modulation.

mitter. For this reason, this form of modulation is most useful in multiplex communication systems.

It is possible to reduce the band width if the pulses are deformed so that the sides are not steep. But pulse-shaping circuits are quite complicated. A simple method of avoiding the difficulties in pulse-shaping circuits consists of a narrow band filter inserted in the output of the pulse-modulated transmitter. Such a filter suppresses the energy in the side bands outside the band width of the filter and has the same effect as modifying the shape of the pulse.

10.27. Pulsed Klystron Oscillators.—Either the reflector voltage or the beam voltage, and in some cases the control grid voltage of a reflex klystron, may be pulsed. Pulsing the reflector voltage requires very little power because the reflector electrode does not normally draw current. This method of pulsing does not decrease the power input to the tube, and it is often preferable to pulse the beam voltage in order to reduce the average input to the tube. The latter form of modulation must be used with two-resonator klystron types if a control grid is not provided, since there is no reflector electrode in that type of tube. A word of caution may be advisable. Most klystrons are designed for c-w operation, and pulsing such tubes at voltages higher than their maximum rating may reduce the life of the tubes, although the average power input may not exceed the rated value.

The build-up time of a pulsed oscillator is an important consideration. Reflex oscillators have unity coupling (single-resonator types only are considered), and these tubes are usually loaded rather heavily, so that the build-up time is quite short, a small fraction of a microsecond. Therefore pulses of very short duration may be used with reflex tubes. Two-resonator klystron oscillators may have a long build-up time (a large fraction of a microsecond), and special designs are required if this type of tube is to be pulsed.

10.28. Klystron Amplifiers.—If a tube that exhibits a long delay in starting as an oscillator is used as a pulsed amplifier, with drive power supplied continuously to the input resonator, the output power will reach its maximum value without any appreciable delay after the beam voltage is applied. Therefore klystron amplifiers may be used satisfactorily with pulses of short duration.

10.29. Other Applications of Pulsed Klystrons.—In addition to applications of pulse modulation for communication, pulsed operation is also useful with signal generators and in measurement work when it is desired to superimpose an alternating component on the microwave power so that audio-frequency amplifiers may be used to increase the level of weak signals until they can be measured easily. Ordinary methods of amplitude modulation would cause undesirable frequency variations, but a flat-topped pulse will provide the audio-frequency component without introducing serious frequency modulation. Long pulses or ordinary keying is often used for this purpose. Short pulses are required when the klystron is used as a pulse generator for testing the response of circuits by observing the effect that the circuits have on the shape of the envelope of the radio-frequency pulse.

Circuits for pulsing the reflector voltage of a reflex klystron are discussed in Chap. 12. These simple circuits are suggested as a means of interrupting the oscillations for test purposes and are not intended as illustrations of pulse modulation. The more complicated circuits required for converting sound or other information into pulse modulation, and multiplexing a number of channels on a single transmitter, present problems that are relatively independent of the modulation characteristics of klystrons; therefore they will not be included.

KLYSTRON DETECTORS

10.30. Velocity Sorting.—Early klystron types included a collector or deflector electrode similar to the reflector electrode in reflex klystrons. This electrode was used as a detector of electron bunching by connecting it to the cathode through a milliammeter. Electrons that are accelerated to a velocity greater than the velocity corresponding to the beam voltage strike the detector electrode and cause a current indication in the milliammeter when the tube is oscillating. Reflex tubes when operated with the reflector electrode near cathode potential exhibit the same detector characteristics, but the method fails when the reflector electrode is sufficiently negative and reflects all electrons.

Unfortunately, the reflector or deflector type of oscillation detector does not give a true indication of oscillation strength. The electrons that strike the detector electrode in a two-resonator klystron are usually electrons that pass the output gap during the interval between bunches and are accelerated by removing energy from the output resonator. The oscillation-detector current is not proportional to the strength of oscillation, and the maximum oscillation-detector current does not correspond to the adjustment of acceleration voltage or tuning for best output. At high voltages the secondary emission from the detector electrode exceeds the primary electron current, and the oscillation detector current is in the reverse direction. The number of secondary electrons is roughly proportional to the number of primary electrons that strike the detector electrode, and the detector behavior is the same regardless of the direction of the detector current.

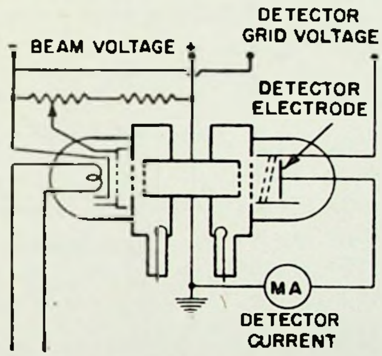


FIG. 10–8.—Circuit diagram for a klystron detector.

10.31. Klystron Detectors.—This oscillation-detector principle is quite useful, however, for a klystron detector or mixer. Figure 10–8 shows the circuit for one design that has been used. The single electrode is replaced by one or more grids connected to a voltage supply which can be made positive or negative with respect to the cathode potential. A collector electrode beyond these grids is connected to the anode and collects all electrons that the grids allow to pass. No electrons reach the collector plate when the detector grids are sufficiently negative with respect to the cathode. Practically all of the beam reaches the collector plate if the detector grids are quite positive. A typical static characteristic for such a detector is shown in Fig. 10–9. Collector current is plotted as a function of detector grid voltage. Extension of this characteristic to the actual case when the detector grid voltage is fixed and the beam voltage is varying is obvious. Dynamic characteristics as sensitive as predicted by the static characteristics have not been attained in practice, however.

10.32. Noise Considerations.—Random fluctuations of the beam current introduce shot noise in klystron detectors. This noise may be 25 or 30 db above the theoretical value for thermal noise. Since crystal rectifiers are considerably better than this figure when used as microwave mixers,

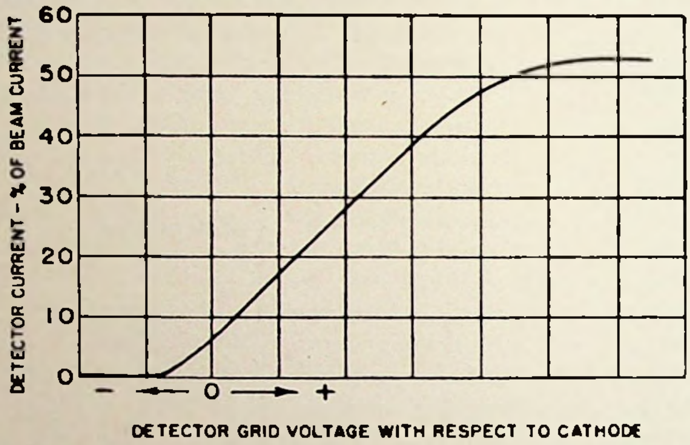


FIG. 10-9.—Static characteristic of a klystron detector.

klystron detectors have not been generally used in receivers. The noise problem in klystron detectors is caused by the same factors that limit the use of klystrons as low-level amplifiers. A successful design for a low-noise amplifier at microwave frequencies would also make klystron detectors practical because the power level in the receiver could then be raised above the noise level of the klystron detector. At present, however, the most sensitive receivers use a crystal mixer without any preamplification, and the necessary gain is obtained in the intermediate-frequency amplifier.

CHAPTER 11

KLYSTRON TUNERS

11.1. Cavity Tuning Methods.—Tuning requirements vary so widely, depending upon the construction of the tube, the number of resonators and the application in which the klystron is used, that the subject will be discussed only in general terms. Specific examples will be introduced to illustrate various points, but they are not intended as a recommendation of a particular method of tuning. Obviously, many variations are possible and cannot all be illustrated.

There are three basic methods of tuning a resonator. The capacitance of the resonator gap may be varied by changing the gap spacing. The capacitance of the resonator may be increased by inserting a movable plunger close to the resonator gap. Or the volume of the resonator may be reduced by inserting a plunger into the resonator at some distance from the gap. The latter method of tuning is analogous to tuning a conventional circuit by varying the inductance. Combinations of more than one of these methods are used in some klystron designs.

11.2. Variable Gap Spacing.—Klystrons that include the resonators as an integral part of the vacuum tube are easily tuned by changing the gap spacing. One wall of the resonator is a flexible diaphragm which permits the gap spacing to be changed. Moving the diaphragm changes the volume of the resonator slightly, but the capacity change introduced by the variation of resonator gap spacing has a greater effect than the volume change and is the principal factor in determining the resonator tuning. The volume of the resonator is decreased when the diaphragm is collapsed, increasing the resonant frequency slightly, but the increased capacity loading due to the reduced resonator gap spacing offsets the decreased volume so that the resonant frequency of a klystron resonator is decreased when the diaphragm is collapsed.

Parallel motion of the resonator grids is desirable but not essential. Wide angularity should be avoided since it prevents good tracking of the tuning of the two resonators with a gang tuner, may waste beam current, reduce output, weakens the flexible diaphragm, and may shorten the life of the tube. The Type 2K41 klystron illustrated in Fig. 11–1 is an example of a klystron tuner designed to give parallel motion of the resonator grids.

Some angularity is permissible, however, and tuners that operate by varying the angle and spacing of the resonator grids are used frequently.

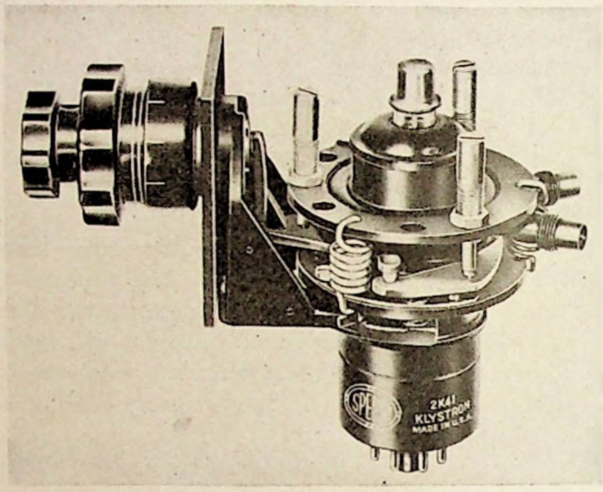
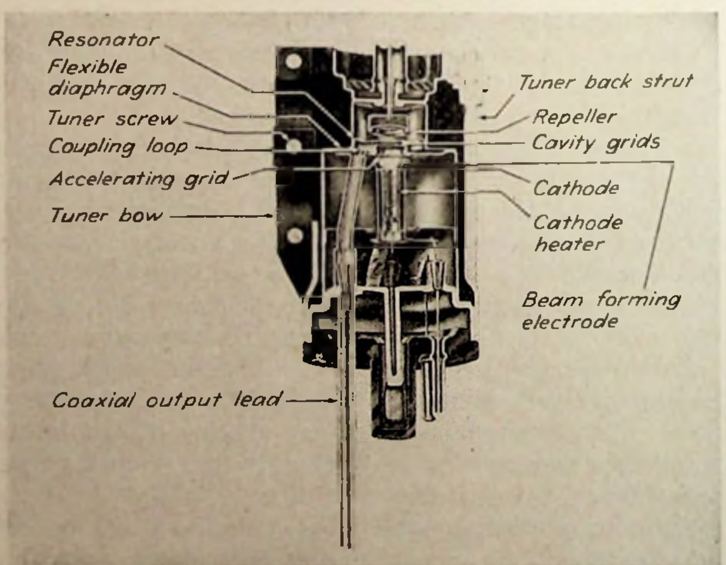


FIG. 11-1.—Type 2K41 klystron. The tuning mechanism is an integral part of the tube and maintains the resonator grids in parallel alignment.



(*Courtesy of Bell Telephone Laboratories.*)

FIG. 11-2.—Cutaway view of a Type 723A /B klystron.

The Type 723A/B reflex oscillator is tuned in this manner (see Fig. 11-2). The flat disk where the diameter of the tube changes, just below the reflector cap, is the flexible diaphragm. The smaller diameter, cylindrical portion of the tube surrounding the reflector electrode is supported on one side by a back strut which acts as a hinge. A right- and left-threaded screw on the opposite side of the tube changes the shape of the "bow" and changes the position of the upper portion of the tube surrounding the reflector electrode. This motion distorts the flexible diaphragm slightly and changes the spacing of the gap of the resonator within the larger diameter part of the tube.

The Model 11C tuner used with the Type 410R klystron is an example

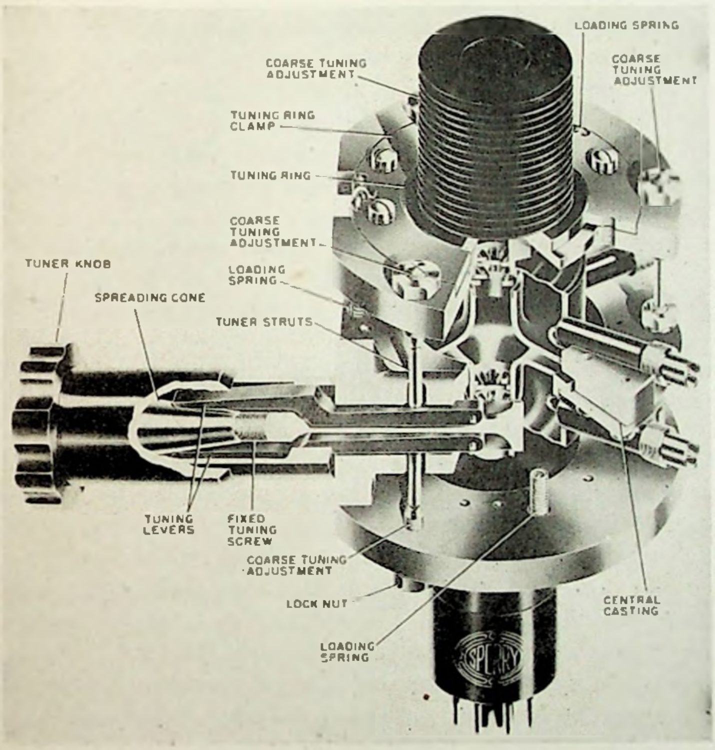


FIG. 11–3.—Isometric view of a Type 3K30/410R klystron mounted in a tuner.

of a gang tuner which operates by changing the angularity between the resonator grids. An isometric drawing of a cutaway tube and tuner is shown in Fig. 11–3. Three adjusting screws for each resonator permit a rough adjustment of the gap spacing and allow the angularity of the clamping rings to be controlled. The fine tuning control varies the gap spacings of the two resonators an equal distance by moving one pair of struts simultaneously while the other two pairs of tuning struts act as the fulcrum. Similar tuners are also designed with a trimmer control on one of the other tuning struts to change the grid spacing of one resonator without affecting the grid spacing of the other. An equal motion of the two struts being moved simultaneously by the fine tuning control should be translated into an equal tuning effect on each resonator. This condition will be met if the two end clamps holding the rough adjustment screws are approximately parallel to the plane of the diaphragms when the fine tuning control is at the center of its range.

Building the tuning mechanism as an integral part of the tube avoids the problems introduced by insecure clamping arrangements and permits presetting the tube to any desired frequency band within the operating range of the tube. This construction is used in newer klystron designs. The tuner becomes merely a mounting bracket for the klystron and a support for the fine tuning control, or the mounting bracket may be an integral part of the tube itself. In these newer designs, the klystron can be removed from the tuner and replaced without affecting the frequency band to which it is tuned.

A fine tuning control is not always required and, for many purposes, particularly transmitter tubes, the tuning rings are a part of the tube. The spacing of the tuning rings is rigidly maintained by three machine screws which are equally spaced around the resonator. The Type 2K35 cascade amplifier is an example of this style of tuning. Adjustment of the tuning is made by loosening two nuts on one strut and changing the spacing between the rings, which determines the gap spacing. All three struts should be adjusted equally if a large change in tuning is required.

The use of variable gap spacing as a tuning method limits the total tuning range to approximately 20 per cent of the center frequency, although this range may be extended if one is willing to accept reduced output. As the gap spacing is decreased, the losses in the resonator become quite high and the output will decrease. The difficulties in maintaining a small spacing accurately are serious, and the tuning of the tube becomes critical when the gap spacing is small because a small change in the spacing gives a large change in the capacity of the gap. Either or both of these factors determine the lower frequency limit of the tube.

If the gap spacing is increased too far, the transit time of the electrons crossing the gap becomes excessive and the output again falls off. As the gap spacing increases, the frequency change produced by a change in the gap spacing decreases, owing to the smaller change in the capacity of the gap. There is a limit to the amount that the diaphragm can be distorted without danger of mechanical failure and loss of vacuum. Therefore, there is an upper limit to the frequency range of the tube.

This method of tuning the resonators by changing the gap spacing is not limited in application to klystrons in which the resonators are a part of the vacuum envelope; it can also be used in cases where part of the resonator is external to the vacuum envelope. Disk-seal tubes usually have the gap spacing fixed by the glass envelope, although tubes of this type can be designed so that one disk is flexible, permitting variation of the gap spacing. However, disk-seal tubes are usually tuned by changing the volume of the external portion of the resonator.

11.3. Volume Tuning Methods.—External resonator tubes are ideal for tuning over wide ranges because the volume can be changed considerably without danger of damaging the tube. One of the most effective schemes for wide-range tuning uses the tube as part of a coaxial line, and the length of the coaxial line is varied by changing the position of a shorting plunger which moves along the axis of the tube and the resonator. Much greater frequency ranges can be covered when a tube is tuned in this way than is possible when the resonator is tuned by changing the gap spacing.

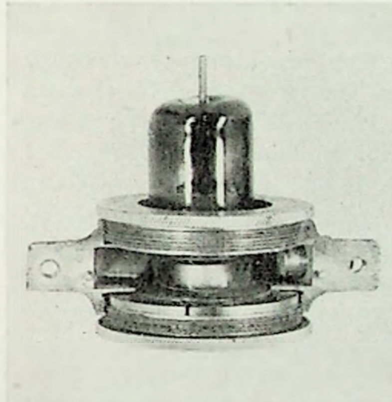


FIG. 11–4.—Two methods of tuning a resonator by changing the volume.

Sliding contacts are a source of trouble in any radio-frequency circuit and can be eliminated if a smaller tuning range is satisfactory. Two simple types of volume tuning arrangements are shown in Fig. 11–4. Both of these methods have a small tuning effect because they affect only a small percentage of the total volume of the resonator. One of the tuning mechanisms in this illustration is a metallic rod which is pushed or screwed into the cavity to reduce the volume. The other tuning member is a flat paddle or vane, which can be rotated 90 deg. The vane has little effect on the tuning of the resonator when the vane lies in a plane perpendicular to the axis of the reentrant portion of the cavity. When the vane is in the position shown, in a plane parallel to the axis, it effectively decreases the volume of the resonator almost as much as the solid plunger. Either of these methods reduces the inductance and therefore increases the resonator frequency.

The two tuning methods described in the previous paragraph are also used with tubes that have resonators as an integral part of the vacuum envelope. In this case some method of distorting the vacuum envelope must be provided, such as a bellows arrangement to permit motion of the plunger, or a glass appendix may be built into the cavity and the plunger can then be rotated within the cavity but outside the vacuum envelope.

11.4. Capacity Loading.—If the plunger or paddle is inserted in a region where the electric field is strong, it will not reduce the inductance appreciably, but the effective capacity of the gap will be increased. This change will decrease the resonator frequency. Therefore, the effect of a tuning plunger depends upon where it is placed within the resonator.

A circular ring surrounding the reentrant portion of the resonator will decrease the inductance and increase the resonant frequency. If this ring is moved toward the gap, the inductance will increase and the capacity loading will also increase. Both of these factors decrease the frequency. Considerable tuning range can be obtained in this manner without any moving contacts. A total tuning range of more than 50 per cent of the center frequency can be obtained with this method. A dielectric ring may also be used, but the tuning range is considerably less than that obtained with a metallic ring of the same dimensions.

11.5. Increased Resonator Losses.—All these volume tuning methods cause an unequal current distribution within the resonator, with the exception of the movable shorting plunger, and therefore increase the losses in the resonator. These extra losses may reduce the Q of the resonator as much as 20 per cent or more, although the effect is not usually that large. These added losses do reduce the efficiency and should be minimized.

11.6. Thermal Effects.—Changes in temperature cause a resonator to expand or contract, and these changes in the dimensions of the resonator affect the resonant frequency. When the resonant circuits are an integral part of the tube structure, as in klystrons and other microwave tubes, the resonators are heated by the beam power supplied to the tube and changes in the power input will affect the temperature of the resonator. Changes in ambient temperature also cause changes in the tuning. Since the capacity effect of the resonator grids is the most important factor in the tuning of a reentrant cavity, changes in the spacing of the resonator gap can produce a large tuning effect. The difference between the expansion of the reentrant portion of the resonator and the external tuning supports may become a large percentage of the gap spacing, and the change in tuning can be very much greater than the percentage of expansion or thermal coefficient of the materials involved.

11.7. Temperature Compensation.—If the resonator and the external tuning struts that maintain the gap spacing were of the same material and at the same temperature, the volume of the resonator, the area of the gap, and the gap spacing would increase uniformly and the frequency would decrease as the temperature increased. The frequency coefficient would be negative and equal to the linear coefficient of expansion of the material for small changes of temperature. However, the use of a strut material with a smaller linear coefficient of expansion, and/or a strut operating at a lower rate of temperature increase than the reentrant part of the resonator, would increase the volume and decrease the grid spacing, causing an increase of the negative frequency coefficient. Both of these effects may exist unless the klystron tuner is carefully designed. It is evident that a tuning strut that expands, under operating temperature conditions, faster than the reentrant portion of the resonator, can increase the resonator grid spacing as temperature increases and counteract the effect of the volume expansion, giving a zero frequency coefficient or causing the frequency coefficient to become positive if the strut expansion is too great.

The amount of compensation required depends upon the linear coefficient of expansion of the resonator material, the temperature gradients in the various parts of the tube and tuner, and the resonator grid spacing. Low-expansion alloys usually have low thermal conductivity as well, causing the temperature inside the resonator to reach a higher value because the heat is not conducted away, but definite gains in frequency stability may be obtained by using Invar and similar alloys in klystron construction. The effect of the gap spacing on the amount of compensation required is readily understood because the same percentage of capacity change requires a greater change in the gap spacing when the gap spacing is large.

11.8. Thermal Effects in Disk-seal Tubes.—Glass has a very low thermal coefficient of expansion; therefore, the change in the gap spacing of a disk-seal tube is determined almost entirely by the expansion of the reentrant portion of the resonator. It is possible to design the shape of this part of the resonator so that the longitudinal expansion is minimized over a wide temperature range. Satisfactory frequency stability for large variations of operating temperature can therefore be obtained.

11.9. Thick Metal Construction.—High-power klystron transmitter tubes are difficult to compensate because the added heat dissipation creates large thermal gradients within the tube. This difficulty can be overcome by discarding the flexible diaphragm type of construction and using thick copper walls for the resonators. Liquid cooling may be used for very large power inputs. Such tubes have a frequency-stability coefficient that is nearly equal to the thermal expansion coefficient of copper, which is quite satisfactory for most applications. Since tuning must be accomplished by some method of volume tuning, the tuning range of these tubes is relatively small.

11.10. Thermal Tuning.—Thermal expansion can be used to advantage if the heat is controllable and is restricted to a tuning member. For example, one of the tuning struts of a klystron can be a small thin-walled tube which is heated by passing current through it. As the expansion of this strut will depend upon the current, the tuning of the klystron can be controlled by changing the current in the thermal strut. Remote control of the klystron frequency is possible as a result of this method of tuning.

Direct conduction as a source of heat has certain disadvantages. Such a strut has a very low electrical resistance and must be operated with a large current and a low voltage. This power is difficult to regulate or control satisfactorily without special circuits. One method of controlling the power input to such a strut depends on heating the strut with a low radio frequency, perhaps 100 kilocycles, using a vacuum tube to generate and control the power. A simple radio-frequency transformer matches the low impedance load to the vacuum-tube oscillator. Stability is obtained by using a regulated voltage on the plate of the oscillator.

Heat losses are quite high if thermal tuning struts are operated in air, and most tubes using thermal tuning have the resonators and the tuning elements built into the vacuum envelope. When this construction is used, it is convenient to heat the tuning strut by electron bombardment. The electron current is easily controlled by a grid electrode between the bombarder cathode and the tuning element which becomes the anode. There is no problem of impedance matching, and the heat is furnished by d-c power at a reasonably high voltage which can be regulated easily. The power required by the control grid can be negligible.

Thermal tuning is most useful in reflex tubes used as local oscillators. The method has one handicap; power is required continuously to keep the tube properly tuned. If the tuning power fails, even momentarily, the frequency of the tube will return to the zero tuning power limit. In addition, there is a lag in the response of thermal tuners when the power input is changed.

11.11. Electronic Tuning.—Tuning by changing the voltage applied to a klystron oscillator was described in Chaps. 7 and 8 on oscillator theory. The term "electrical tuning" has been given to all forms of tuning that can be controlled electrically and includes both thermal tuning and tuning by means of the beam voltage or reflector voltage. The latter method is generally referred to as "electronic tuning."

The electronic tuning band width is defined as the frequency difference, usually in megacycles, between the two points of equal power output. The points that give half the maximum power that can be obtained when the voltage is adjusted to the correct value are frequently chosen. This definition is indicated on Fig. 11–5, which shows the electronic tuning and power output characteristics of a typical reflex klystron oscillator when the reflector voltage is varied. When the load on a reflex oscillator is adjusted for maximum output, the half-power electronic tuning band width is approximately 20 per cent greater than the band width associated with the loaded Q of the resonator. This relation was derived in Secs. 7.19 and 7.20.

Band widths of 30 megacycles or more are obtained easily with reflex klystrons. This property makes these tubes useful as local oscillators because automatic-frequency control can be obtained by applying the d-c component from a frequency discriminator to the reflector voltage.

Electronic tuning is not gained without some sacrifice. It is a form of instability. Power-supply regulation must be very good, and the ripple voltage very small, or the frequency of the output will be modulated. Also, the residual noise in an oscillator is greater when the Q is decreased to obtain electronic tuning, and the residual noise becomes greater when the reflector voltage is changed from the value required for optimum output. This residual noise is too low to be of importance in most local oscillator applications, but it does become important in some radar receivers.

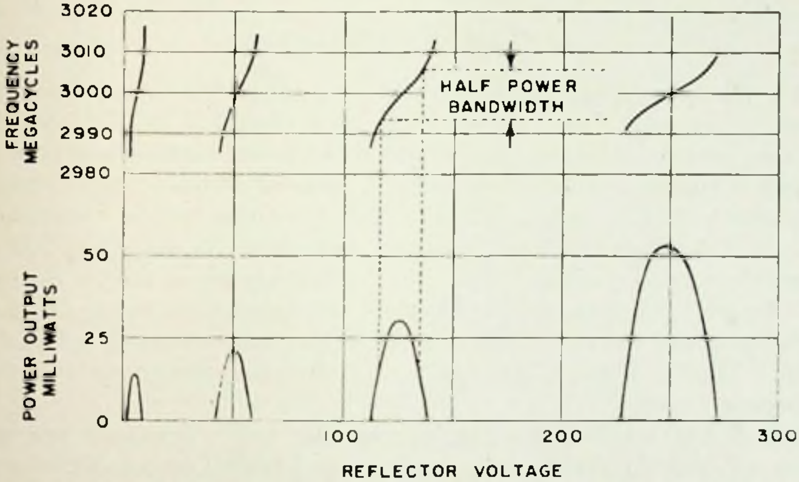


FIG. 11·5.—Electronic-tuning characteristic of a reflex klystron.

These considerations do not detract from the general usefulness of tubes with wide electronic tuning. In addition to the simplicity of automatic frequency control which was mentioned previously, reflex oscillators are ideal for panoramic receivers or spectrum analyzers. The latter term has been rather generally adopted for the type of panoramic receiver developed for use in the microwave region. A spectrum of 30 megacycles or more can be viewed on the oscillograph screen when a reflex klystron is used in such an instrument.

CHAPTER 12

KLYSTRON OPERATION

12.1. Operating Suggestions.—Although most of the factors that affect the operation of klystron tubes have been discussed in the preceding chapters, some of the most important problems will be reviewed here. A number of suggestions intended to aid in using klystrons without previous experience will be included. Considerable repetition will be unavoidable, but it is justified in order to group the suggestions for using the different types in a unified manner. The use of reflex klystrons will be discussed first because these tubes are the simplest and best known type. Then the operation of two-resonator klystron oscillators will be considered in some detail. Finally, klystron amplifiers and frequency-multiplier tubes will be discussed briefly.

12.2. Reflex Oscillators.—Single resonator reflex klystrons are quite simple to operate. If the beam voltage and beam current are adjusted to the correct value and the reflector voltage is varied, oscillation will be observed at some point within the range of adjustment of the reflector voltage, provided that the output load is not too great. The correct load ordinarily corresponds to a small standing wave ratio in a 50-ohm line, although variations between different tubes will occur and an impedance transformer will usually increase the output. A crystal detector connected directly to the output terminal generally gives a satisfactory indication of oscillation without loading the tube too heavily.

If the klystron is tuned by varying the gap spacing, another precaution must be observed. The tube will fail to work if the resonator grids are too close together, and will require high voltage for operation if the spacing is too great and the gap transit time becomes large. It is usually possible to determine whether the spacing is approximately correct by inspecting the flexible diaphragm of the resonator. The diaphragm will appear approximately flat when the spacing is normal. The spacing in disk-seal tubes generally cannot be varied and, in designs such as the Type 723A/B, the spacing can be varied but the tuning mechanism limits the motion in both directions.

When in doubt regarding the gap spacing, the best procedure is to keep it small. Then a crystal detector or some other oscillation indicator is connected to the output terminal of the tube, the beam voltage is applied, and the reflector voltage is varied until oscillation is observed.

12.3. Voltage Modes.—There are a number of combinations of beam voltage and reflector voltage which will give the correct phase for oscillation (see Fig. 6–3). If the beam voltage is fixed and the reflector voltage is varied, oscillation will be observed at several regions within the total range of the reflector voltage. Each region is known as a "voltage mode," and oscillation does not occur in the region between modes. This behavior is typical of all klystron oscillators and is illustrated for a reflex klystron in Fig. 12–1. These voltage modes also occur when the reflector voltage remains constant but the beam voltage is changed.

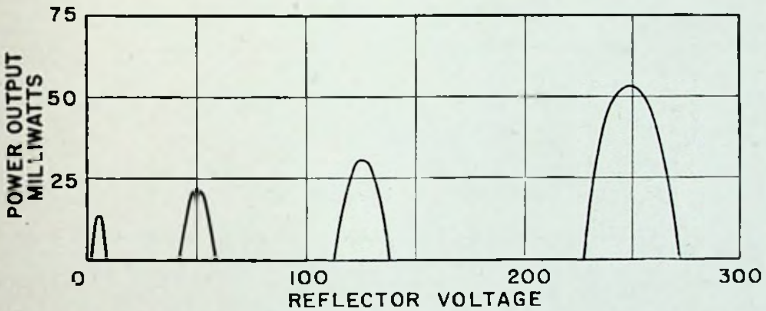


FIG. 12–1.—Voltage modes in the operation of a reflex klystron oscillator.

This dependence between voltage and oscillation in klystron oscillators exists because the phase of the feedback is determined by the transit time of the electrons. A particular value of transit time corresponding to the correct phase for one frequency of oscillation will give a different phase if the frequency is changed. The relation between the phase angle ϕ, the transit time T_0, and the frequency is given by

$$\phi = 2\pi f T_0 \tag{12-1}$$

If the klystron is tuned a considerable amount without a corresponding change in the beam voltage or reflector voltage, the phase angle between the electron bunch and the gap voltage may become so great that the tube will not oscillate. The tuning range that is permissible without changing the voltage depends upon N, the number of cycles during bunching. Differentiating Eq. (12–1) gives

$$d\phi = 2\pi T_0 \, df \tag{12-2}$$

The number of cycles corresponding to the transit time T_0 is

$$N = fT_0 \tag{12-3}$$

Therefore Eq. (12–2) may be rewritten

$$d\phi = 2\pi N \frac{df}{f} \tag{12-4}$$

If N is large, *i.e.*, the beam voltage is low, a frequency change of a few per cent will require changing the voltage. However, tuning over the entire range of a reflex klystron (20 per cent of the center frequency) may usually be accomplished without exceeding the normal range of the reflector voltage or changing to a different mode.

12.4. Circuit for Observing Characteristics.—Obtaining the necessary data for the output characteristic shown in Fig. 12–1 becomes a tedious process if taken point by point. The number of points required becomes tremendous if additional curves are desired for a number of frequencies

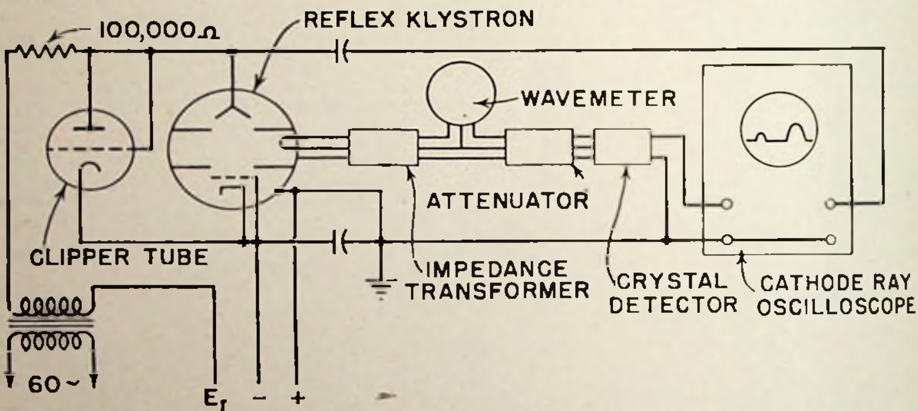


FIG. 12–2.—Circuit diagram for observing the voltage modes of a reflex klystron.

within the tuning range of the tube. Fortunately, a simple circuit can be used to obtain this information dynamically, and the curves can be observed with a cathode-ray oscilloscope. Such a circuit is given in Fig. 12–2.

The reflector voltage is varied periodically by the 60-cycle voltage in series with the lead from the reflector electrode to the reflector voltage power supply. As the voltage is varied, the output will follow the form of the voltage modes. A sample of the reflector voltage is applied to the horizontal deflection plates of a cathode-ray oscilloscope, and the output from the detector is applied to the vertical plates. As a result, a pattern similar to Fig. 12–1 is traced periodically on the oscilloscope screen.

If the peak value of the 60-cycle voltage is greater than the reflector voltage, the reflector electrode will become positive with respect to the cathode during part of the cycle and a current will flow in the reflector circuit. The reflector electrode is not usually designed for operation at a positive potential. Protection for the reflector electrode is provided in the circuit in Fig. 12–2 by a clipper tube between the reflector electrode and the cathode. This tube limits the positive excursion of the reflector voltage. The value of the 60-cycle voltage should always be adjusted so that the clipper tube is not overloaded.

12.5. Frequency Measurement.—An indication of the frequency of oscillation as well as the power output may be obtained if a high-Q cavity resonator that has been calibrated as a frequency meter is coupled to a "tee" in the transmission line between the klystron and the output load. As the reflector voltage sweeps through a mode, the frequency of oscillation changes as indicated by Fig. 7–14. When the frequency of oscillation

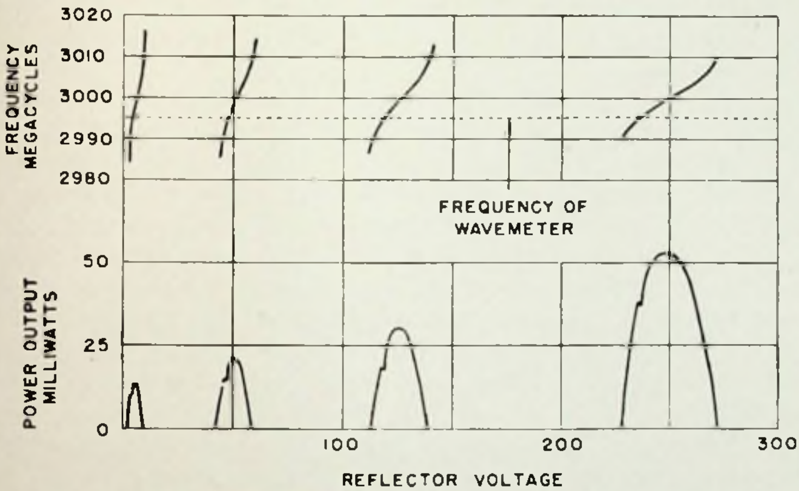


FIG. 12-3.—Wavemeter resonance indication on the voltage mode pattern of a reflex klystron.

corresponds to the tuning of the cavity resonator, some power will be absorbed and a "dip" will occur in the output curve. This effect is illustrated in Fig. 12–3.

The dip in the curve appears in each mode at approximately the same point in the pattern. If the frequency meter cavity is tuned, the dip will move along the curve. The electronic tuning between half-power points may be measured in this manner by noting the calibration point of the resonator when the dip is halfway between the zero line and the peak output, then changing the tuning of the wavemeter until the indication has moved along the curve to the other half-power point It is obviously possible to plot the entire frequency characteristic as a function of reflector voltage by using this technique.

Rather high-Q cavities are required for this application if a sharp indication of frequency is to be obtained. For example, a frequency meter with a Q of approximately 10,000 will usually be necessary. This means that a cylindrical cavity resonator is required. A lower Q cavity similar to the resonators used in the klystrons will merely warp the pattern gradually as the resonator is tuned through the operating frequency of the klystron.

12.6. Effect of Changing the Output Load.—An impedance transformer is indicated in Fig. 12–2 between the klystron and the load. This arrangement permits the load to be changed. If sufficient attenuation is introduced, the crystal detector used to sample the power output will not become saturated. It may therefore be considered a square-law device, and the output indication on the oscilloscope will be proportional to the radio-frequency power. It is not always necessary to use all these refinements and, if the klystron is a low-power tube, the crystal detector may be the only load and the impedance transformer may be a simple trombone or "line stretcher."

If the impedance transformer is adjusted so that the tube is loaded lightly, there will be little power transferred to the load and all the modes will become small and appear to vanish simultaneously. As the loading is increased, the amplitudes on the oscilloscope screen will increase. The mode for the highest reflector voltage will have the most output. As the load is increased further, this high-voltage mode will reach its maximum value, then decrease suddenly, and disappear. The next lower voltage mode will continue to increase until its output reaches a maximum. The modes will disappear in turn until the loading is so heavy that the lowest reflector voltage mode is finally suppressed. This behavior was discussed in Sec. 7.15.

12.7. Detection of Unsatisfactory Tubes.—Dynamic methods of observing klystron characteristics are ideal for determining whether a tube is unsatisfactory. The effect of varying the load can be observed conveniently. If the tube is loaded lightly, there should be a number of modes observable. (Note: Some transmitter reflex designs operate at high voltages and only one mode may be present.) This test furnishes an ideal method of rejecting unsatisfactory tubes. A tube which does not have the usual number of modes, or which does not give the desired output, can be detected with ease.

12.8. Bench Oscillators.—The ease of tuning reflex oscillators makes these tubes particularly suitable as sources of power for experimental tests and measurement work. There is actually little difference between the tubes used as local oscillators in receivers and those used as bench oscillators. A local oscillator is usually operated at the lowest power input that will give satisfactory operation, while the power demands from a bench oscillator are normally much greater. For example, a klystron used as a source of power for checking the standing wave ratio of equipment will be affected by the adjustment of the equipment being tested unless considerable attenuation is inserted between the tube and the test equipment. Both the frequency of oscillation and the power output are dependent upon the load. A variation of either of these factors will introduce errors in the measurement of the standing wave ratio. An attenuation of 20 db may be required for the necessary isolation, although less attenuation will be satisfactory for many applications.

12.9. Pulse Operation.—Many applications of bench oscillators require that the output be modulated. Standing wave detectors may use a high-gain audio amplifier to obtain a sensitivity comparable to a galvanometer. Since amplitude modulation without undesired frequency modulation is

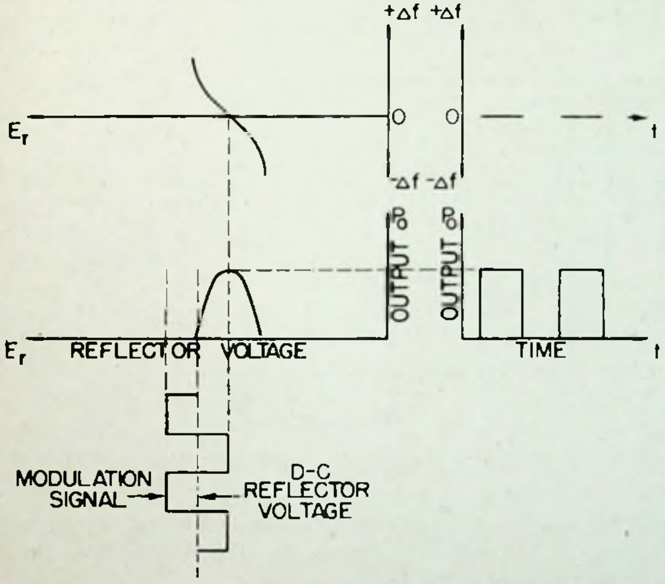


FIG. 12–4.—Pulsed output from a reflex klystron with square wave modulation of the reflector voltage.

difficult to obtain, the klystron is often operated by keying with a square wave. Short pulses of radio-frequency power are often required for testing the response of circuits.

Either the beam voltage or the reflector voltage of a reflex tube may be changed to obtain the pulsed output, or the control grid may be pulsed if this electrode is included in the klystron design. More power is required by the keying circuit when the beam power is cut off, but this method has the advantage that the power input to the klystron is reduced. This simplifies the cooling problem. Pulsing the reflector circuit requires very little power. It is only necessary to change the reflector voltage a small amount in order to shift the voltage from the correct operating point to a value that does not allow the tube to oscillate.

This method of pulsing the reflector circuit is illustrated in Fig. 12–4. The pulse voltage that is superimposed on the reflector voltage is indicated in the lower part of the illustration. The reflector voltage and the amplitude of the pulse voltage are adjusted so that the tube does not oscillate during the "off" period and operates at the point of maximum output during the "on" period. The pulse output is shown at the right of the typical output characteristic. If the sides of the pulse are very steep and the pulse is flat-topped, the frequency will remain constant with this type of modulation.

12.10. Pulse Circuits.—A circuit for obtaining the pulse voltage and applying it to the reflector circuit is shown in Fig. 12–5. A multivibrator

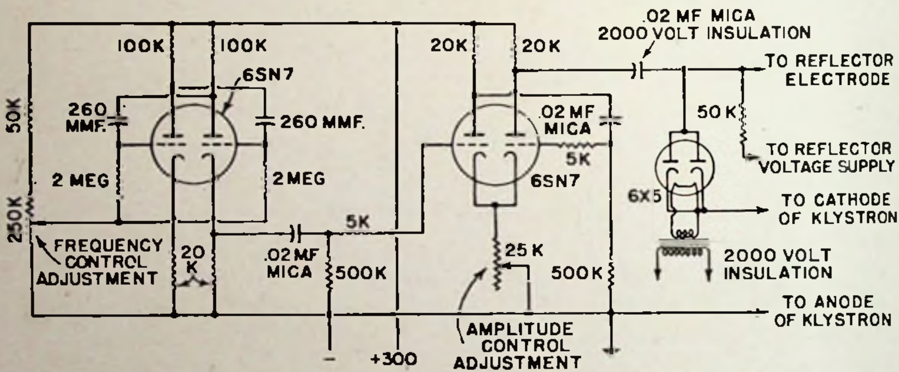


FIG. 12–5.—Circuit diagram for obtaining square wave modulation of the reflector voltage.

circuit generates the pulses. A clipper tube is used to ensure a steep pulse wave form with a flat top. Controls to change the frequency and the amplitude of the square wave are included.

12.11. Frequency Modulation.—A frequency-modulated output is useful for many tests and is easily obtained with a reflex klystron by varying the reflector voltage. Saw-tooth modulation is particularly useful for oscillators to be used in spectrum analyzers and similar applications because there is no difficulty due to the pattern on the return sweep failing to coincide with the pattern when the sweep is in the opposite direction.

The effect of saw-tooth modulation on the output of a reflex klystron is illustrated in Fig. 12–6, which is similar to Fig. 12–4 for pulse operation, but is designed to utilize the frequency-modulation characteristic rather than to suppress it. Note that some amplitude modulation is present.

12.12. Two-resonator Klystron Oscillators.—In many respects, a two-resonator klystron oscillator is similar to a reflex oscillator. There are voltage modes, but only the beam voltage can be varied to choose the proper operating point. The tuning of the two resonators must be correct when the voltage is varied, or oscillation will not be observed. Similarly, the voltage must be of the correct value when changing the tuning or the tube will not oscillate. Satisfying these two conditions simultaneously presents a very difficult problem if the proper equipment is not available.

12.13. Pretuning with a Signal Source.—A source of radio-frequency power in the desired tuning range simplifies the tuning of a two-resonator

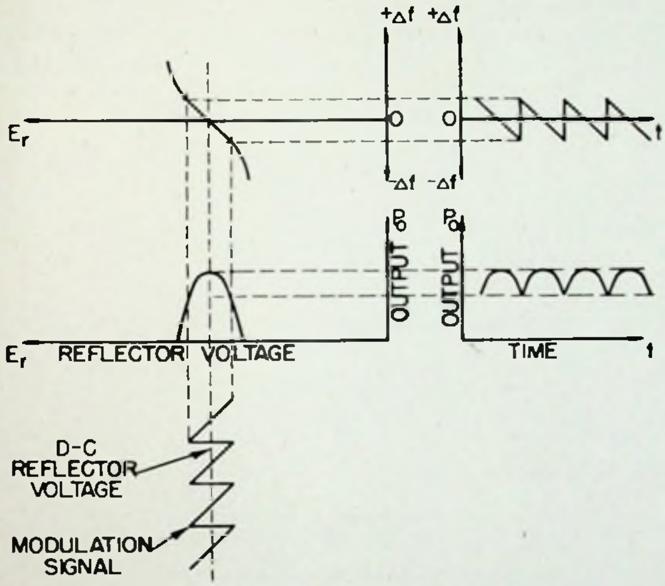


FIG. 12-6.—Frequency-modulated output of a reflex klystron with saw-tooth modulation of the reflector voltage.

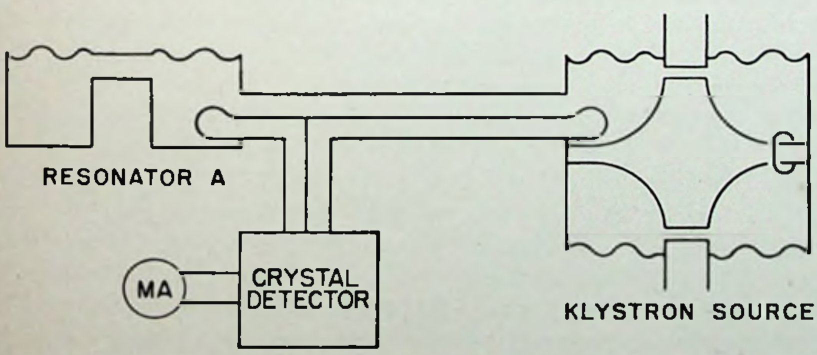


FIG. 12-7.—Pretuning a klystron resonator with a signal source.

oscillator considerably. Figure 12-7 shows a typical setup for pretuning the resonators. One of the klystron resonators is connected with a transmission line to another klystron oscillator used as a signal source. A "tee" in the coaxial line is connected to a crystal detector. The tee may be located at any point in the line, but the location near the resonator being tuned (resonator A) is recommended. The current in the crystal detector usually varies as shown in Fig. 12–8 as the resonator is tuned through the oscillator frequency. The shape of the curve indicating resonance depends on the position of the tee and the length of the line. Both resonators should be tuned to the same frequency; the klystron will oscillate when proper beam voltage and current are applied to the tube. Slight retuning may be necessary in order to obtain maximum output from the tube.

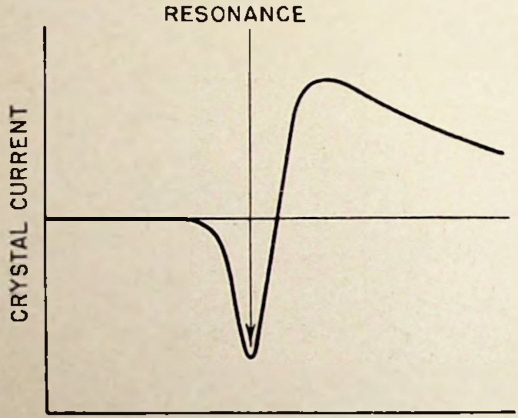


FIG. 12–8.—Reactance-type indication of resonance with the signal source.

If two coaxial terminals are available on a single resonator, the crystal may be connected to one coaxial terminal and the signal source to the other. In this case resonance will be indicated by a crystal current maximum when tuning through the oscillator frequency.

A two-resonator klystron with an internal feedback line and a single coaxial terminal in one resonator may be pretuned quite easily. The resonator with the coaxial terminal is first tuned, using a crystal detector in the tee connection. The first resonator should be tuned for the current minimum in Fig. 12–7. Then the second resonator is tuned and the crystal current will again vary, somewhat less sharply than indicated in Fig. 12–7, as the second resonator is tuned through the oscillator frequency. After the resonators have been pretuned in this manner, the klystron will oscillate when the correct beam voltage and current are applied. Some slight retuning adjustments may be required to obtain the best output from the tube.

12.14. Tuning without a Signal Source.—Finding the proper adjustment of tuning and acceleration voltage for a two-resonator klystron oscillator requires special technique if the tube is being tuned for the first time and no other source of r-f power at the desired frequency is available. The procedure is not at all difficult, and experience with the tubes will allow many of the following instructions to be ignored. These instructions apply primarily to klystrons similar to the Type 3K30/410R which do not have the tuning mechanism permanently attached to the tube.

Some means of varying the beam voltage automatically during the tuning adjustment is desirable. Either self-rectified alternating current

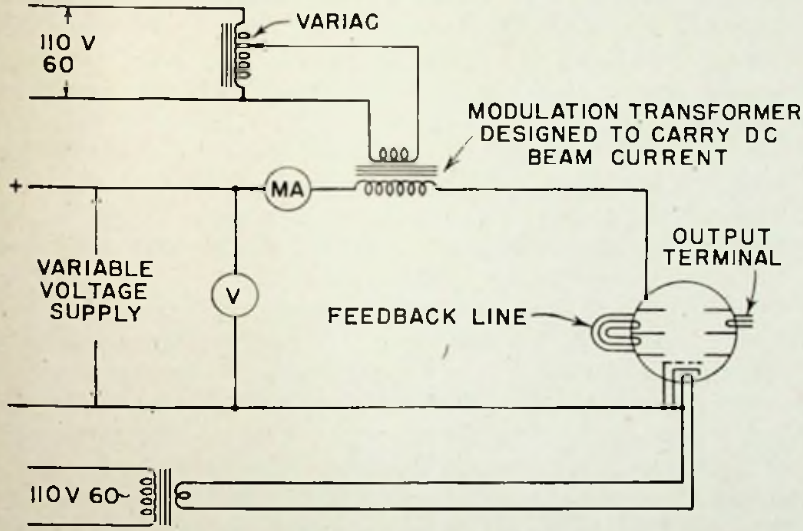


FIG. 12-9.—Circuit diagram for modulating the beam voltage of a two-resonator klystron oscillator during tuning adjustments.

may be used by applying a 60-cycle voltage directly to the klystron, or a combination of a-c and d-c voltage may be used, as shown in Fig. 12-9. The latter method is similar to the circuit used for examining the characteristics of a reflex klystron dynamically and may also be used for that purpose. The procedure for adjusting a two-resonator oscillator is as follows. *Do not consider these suggestions as inflexible rules.*

1. Determine the spacing of some reference points for each resonator on the klystron or tuner when the klystron grids are touching. This adjustment may be determined by turning the rough adjustment screws until the struts are loose and the diaphragms are collapsed completely. If the diaphragms have been "work-hardened" by use, it may be necessary to apply slight pressure in addition to the springs that normally collapse the resonator diaphragms. Do not force the tube, or damage to the grids may occur.
2. After the reference spacings have been determined for each resonator, increase both distances approximately 0.030 in. for a klystron operating with a wavelength of 10 cm. This adjustment will make the

resonator gaps approximately correct and equal. For shorter wavelength tubes decrease the gap spacing proportionately. Move each rough adjustment screw equally so that the tuning rings remain approximately parallel.

3. Choose some operating voltage and current from past experience or from data supplied with the tube. Do not exceed the maximum voltage rating of the tube. Apply about 110 volts rms of 60-cycle modulation.
4. Vary the tuning of one resonator, either the input or the output resonator. Variations as great as 0.010 in. may be necessary if the original settings were not made accurately. Do not move the diaphragms too far.
5. Vary the acceleration voltage in 200-volt steps and repeat the tuning procedure if oscillation cannot be detected.
6. Turn off the beam voltage modulation after oscillation is found. Readjust the beam voltage until oscillation is at maximum strength.
7. Retune the resonators for maximum output.
8. Crystal detectors connected directly to the kystron should not be used as oscillation detectors when the beam voltage is greater than 600 volts, as damage to the crystal detector is possible if excessive power is applied to it. The crystal-detector current is not a good indication of overloading, since excessive input may reduce the current or even cause it to reverse in direction. Use a neon lamp across the output terminal as an indicator for high-power tubes, or use one of the suggestions in Sec. 14.20 for monitoring the power.
9. When changing the frequency to the desired range, vary both resonators simultaneously, or vary first one resonator and then the other a small amount, so that the tube continues to oscillate at all times. Readjust the beam voltage when necessary to improve the output, if the frequency change is considerable.

12.15. Families of Voltage Modes.—When the input and the output resonators are tuned to the same frequency, as with a separate signal source, the output is not so great as the maximum output that can be obtained by detuning one of the resonators. This effect, introduced because the two circuits are coupled by the feedback line, was discussed in Chap. 8. If the two resonators are overcoupled, there will be two separate families of voltage modes. This behavior is typical of oscillators such as the Type 412 or 710 klystrons. The voltage modes correspond to transit times in Eq. (8–16) which make n an integer. One family of modes corresponds to the plus sign in this equation, and the second family is associated with the minus sign.

If the two resonators are tuned to the same frequency, one family will operate at a lower frequency and the other modes will occur at a higher frequency (see Fig. 8–10). When the input resonator is tuned to a higher frequency than the output resonator, the output from the first family of modes will be increased and the frequency of oscillation will approach the resonant frequency of the output resonator. Detuning the input resonator in the opposite direction will emphasize the other family of modes. Oscillation corresponding to one family may not be detected between two voltages corresponding to the other family when the relative tuning has been changed to maximize the power output. Or a higher voltage mode may fail to give as much output as a lower voltage mode until the relative tuning of the resonators is readjusted.

Two modes of different families merge into a single mode if the beam current is large enough. This effect may be observed in the oscillator section of a Type 2K34 oscillator-buffer klystron. The modes also merge into one when the coupling is reduced. Figures 8–9 and 9–11 illustrate this characteristic, which is typical for the Type 3K30/410R klystron oscillator.

12.16. Range of Tuning without Changing Voltage.—Two-resonator klystron oscillators require a different beam voltage if the tuning is changed over a considerable range. This behavior is similar to that of a reflex klystron described in Sec. 12.2 and will not be discussed in detail.

12.17. Effect of the Load.—A two-resonator klystron oscillator differs from a reflex oscillator in its dependence upon the output load because the input resonator is a separate resonator and the bunching voltage can be controlled by detuning the input resonator. In a reflex oscillator, the bunching voltage in the single resonator can be reduced to prevent overbunching only by loading the resonator. For this reason, modes corresponding to a large value of N require heavier loading than modes with fewer cycles during the bunching time. This situation does not exist in a two-resonator oscillator because optimum bunching is obtained by detuning the input resonator. Maximum output is obtained when the load corresponds to an equivalent load resistance R_L equal to the cavity shunt resistance R_s (see Secs. 8.15 and 4.11). The optimum load is the same for all modes.

Although the discussion of detuning in the preceding paragraph mentions tuning only the input resonator, it is obvious that either resonator may be tuned in order to maximize the output. It is the relative tuning of the two resonators that is important.

A Rieke diagram for a klystron oscillator (Fig. 8–14) shows the effect of the load on both the output and the frequency of oscillation. If the load on the klystron corresponds to a large standing wave ratio, small changes in frequency may shift the phase of the load into an unstable region. "Dead spots" may occur; *i.e.*, there may be frequencies at which the tube will refuse to oscillate. A similar effect may be introduced if a resonant load such as a wavemeter is coupled to the tube. As the wavemeter is tuned through resonance, the load may correspond to a point in a region of nonoscillation, and the indication of the wavemeter may have two peaks with oscillation suppressed at the setting when the wavemeter is tuned to the frequency of the tube.

The effect of the load on the frequency of oscillation is known as "pulling." The "pulling figure" of a tube is defined as the difference in megacycles per second between the maximum and the minimum frequencies of oscillation reached when the load impedance varies through 180 deg, while the absolute magnitude of the VSWR is constant and equal to 1.5.

12.18. Dynamic Measurement of Characteristics.—A circuit similar to Fig. 12–9 may be used in the manner described in Sec. 12.3 in order to observe the characteristics of a two-resonator oscillator with a cathode-ray oscilloscope. Or, if half the sweep range of the oscilloscope is sacrificed, the d-c power supply is unnecessary and a transformer may be used to supply power that is self-rectified by the klystron. This arrangement is not suitable for operating a tube, since the oscillator will be swept continuously through many voltage modes, but the circuit is ideal for observing the different modes simultaneously and determining the effect of tuning on the output of a klystron oscillator. The data for the curves in Figs. 8–11 and 8–12 were obtained with such a circuit.

12.19. Length of the Feedback Line.—This dynamic method of observing the output of a tube simplifies the problem of determining the correct length for the feedback line of a two-resonator klystron oscillator. The simple theory discussed in Sec. 8.14 implies that a change in beam voltage could be compensated by a change in the length of the feedback line. This result would be true if the line were matched but, when standing waves are present, certain line lengths may prevent the tube from oscillating at any voltage.

A trombone or variable length of line in the feedback circuit can be used to show the effect of the line length on the output of an oscillator. The correct beam voltage for any adjustment of the feedback line is automatically provided by sweeping the beam voltage over a wide range at a 60-cycle rate. Any change in the power output appears immediately on the screen of the cathode-ray oscilloscope. The phase of the reflections in the line will vary as the length is changed, and the shape of the mode will vary. It may appear that the relative tuning of the two resonators is changing. This effect may be avoided by retuning one resonator in order to maximize the output each time the length of the feedback line is changed.

As the line length is changed, the output will remain satisfactory over a large variation of length, indicating that the length of the line is not critical if the line is nearly correct. Varying the length of the line will shift the mode slightly to a different voltage. Eventually the output will fall off, and no adjustment of the tuning will give the maximum output that can be obtained with the correct length of line. The tube may stop oscillating for some adjustment of the line length. When the line length is increased further, the tube will oscillate again. The position of the mode will be the same as before when the line length has been changed one-half wavelength.

It would be possible to use an impedance transformer in the feedback line to obtain output at any chosen voltage. However, the two additional adjustments that would be required would complicate the tuning of the oscillator considerably and are not recommended. The simplest solution is to provide a number of lines of slightly different length and use the one that is best for the frequency range to be covered. Four different lengths of line differing by one-eighth of a wavelength would be adequate for the range of most two-resonator oscillators. Remember that a wavelength is a shorter distance in a solid dielectric line than in air.

12.20. Adjustment of Amplifiers.—Klystron amplifiers are simpler to adjust than two-resonator oscillators and involve several differences in technique. There are no voltage modes as in oscillators and any voltage may be used on an amplifier, although variations in output resembling voltage modes appear owing to overbunching if the excitation is constant and the voltage is reduced (see Fig. 4–5). The obvious remedies for overbunching are reducing the driving power, detuning the buncher resonator, or increasing the acceleration voltage. When the emission is space-charge limited, the beam current follows the $\frac{3}{2}$-power law and the output increases with acceleration voltage after the voltage for optimum bunching is exceeded. This case is illustrated in Fig. 4–5.

12.21. Tuning Procedure.—If desired, a klystron amplifier may be pretuned before the beam voltage is applied. Each resonator is tuned separately with a signal source as described in Sec. 12.12. Then the beam voltage is applied to the tube, and the drive-power source is coupled to the input resonator. A slight readjustment of the tuning of the resonators may be required after the tube has reached its operating temperature.

A klystron amplifier may also be tuned with the beam voltage applied. If the excitation is reduced to ensure that overbunching does not occur, the tuning of a klystron amplifier is identical to the procedure used at the lower frequencies with conventional vacuum-tube amplifiers. The input resonator is tuned until the excitation is a maximum as indicated by a crystal detector coupled to the input resonator. If the output resonator is tuned near the correct frequency, it is possible to use a detector loosely coupled to the output resonator as an indicator for tuning the input resonator. When only one coaxial terminal is available in the input resonator, resonance may be determined by the reaction on the source of the excitation by connecting a detector to a "tee" in the input line, as indicated in Fig. 12–7.

After the input resonator has been tuned to approximately the correct frequency, the output circuit is tuned until a maximum occurs in the power output. Readjustment of the input resonator may increase the output slightly. Then the drive power may be increased until a maximum in the output indicates that optimum bunching has been obtained.

12.22. Tuning Cascade Amplifiers.—Cascade amplifiers may also be tuned as described in the previous section, tuning the first two resonators as a single-stage amplifier; then the third resonator can be tuned. The precaution regarding overbunching is more important with this type of tube because the drive power required is so much less.

If the second resonator of a cascade amplifier is not provided with an output line, the correct tuning adjustment may be obtained by tuning the first and third resonators as a single-stage amplifier; then the second resonator is tuned until the output at the third resonator is a maximum. The output will vary as indicated in Fig. 9–5 when the tuning of the second resonator is varied. It is important to avoid having the second cavity tuned to the frequency that cancels the bunching when making the preliminary adjustment of the first and third resonators. The best procedure is to tune the second resonator to a high frequency by making the gap spacing greater than that of the other two resonators during the preliminary adjustments. This procedure ensures that the output at the third resonator is always greater than it would be if the second resonator were not present.

12.23. Adjustment for Optimum Bunching.—It is easy to tune an amplifier with reduced drive power and then increase the excitation until optimum bunching occurs and the output is a maximum. An excess of drive power is desirable, and some means of controlling the amount of drive power is necessary. A variable attenuator with a low insertion loss is ideal for this purpose.

When an attenuator is not available, overbunching may be used to determine the correct adjustment of a klystron amplifier. Determination of the adjustment for optimum bunching is certain only when excess driving power is available. This is most easily obtained by reducing the acceleration voltage of the amplifier. Overbunching will be indicated by a double peak in the output as the input resonator is tuned through resonance with the source of drive power. These two output peaks occur when the input resonator is detuned sufficiently from the driver so that the radio-frequency voltage at the input gap is correct for optimum bunching, and the output peaks are found on either side of resonance with the source of drive power. At resonance, the radio-frequency voltage at the input gap is too great, the output is reduced, and a minimum in the power output occurs. This behavior is illustrated in Fig. 4–6 and described in Sec. 4.8.

After overbunching has been observed, the acceleration voltage of the amplifier should be increased until optimum bunching occurs when the input circuit is tuned to resonance. Overbunching may be avoided by detuning the input resonator until the output is a maximum, but this procedure is not recommended. Reducing the drive power and tuning the input to resonance give better stability. The latter adjustment allows operation at a point where a small frequency change causes only a small change in the input voltage, and a small change in input voltage affects the output only slightly. This adjustment corresponds to operation on the flat portion of both the input resonance curve and the Bessel function bunching curve (Fig. 4–6, Sec. 4.8).

12.24. Adjustment of Frequency Multipliers.—The tuning procedure for a klystron frequency multiplier is quite similar to the adjustment of a two-resonator amplifier. However, multiplier excitation is more critical. Reference to Figs. 5–2 and 5–3 shows that the bunching curve for a multiplication factor of ten is almost zero until optimum bunching is approached. The basis for this characteristic may be readily understood because the Bessel function of the nth order is approximately proportional to x^n for small values of x. An apparent increase in the Q of the input resonator is one result of this multiplier bunching characteristic. This effect was discussed in Sec. 5.5.

Amplifier-multiplier klystrons have characteristics similar to those of cascade amplifier klystrons and are tuned in a similar manner. The efficiency can be improved by overdriving these tubes and detuning the second resonator to a frequency higher than the input frequency. These tubes combine the excitation or bunching characteristics of a multiplier with the tuning characteristics and high gain of a cascade amplifier.

CHAPTER 13

KLYSTRON POWER SUPPLIES

13.1. Special Requirements.—Several of the electrical characteristics of klystrons, such as the existence of voltage modes and the frequency dependence upon voltage, and some of the mechanical features of the tube design impose conditions upon the power supplies for these tubes such as are not encountered in the design of power supplies for more conventional tube types. These conditions are not difficult to meet, but they do require special care in the design of the power supplies.

As an example of a simple difference in the design of klystron power supplies, the beam current of these tubes is much less than the usual maximum rating of commercially available transformers. It may not be economical to use special transformers for a small, low-voltage klystron power supply, but a considerable saving in size, weight, and cost may be gained if special transformers are built for a larger klystron power supply. Other considerations, which apply in general to the design of klystron power supplies, will be discussed in the following sections; then a few typical power supplies for several types of applications will be described.

13.2. Ground Connection on Power Supplies.—The resonant cavities used in klystron tubes form the anode of the electron gun and, since the radio-frequency output line is usually directly connected to the cavity, it is desirable to ground the positive terminal of the power supply so that the output terminal may be connected directly to a grounded load that can be handled with safety. Exceptions to this rule are possible, but these exceptions emphasize the importance of the precaution. For example, a Type 723A reflex oscillator, which is frequently used as the local oscillator in microwave receivers, has an output line which is part of the shell of the tube and is therefore at anode potential. However, this type of tube is often plugged into a wave guide with the output line of the tube insulated from the wave guide. This arrangement is used in order to operate the tube with the cathode grounded so that a single power supply may be used for the local oscillator and the other vacuum tubes in the receiver. It has the disadvantage that the shell of the tube is not grounded and must be protected from accidental contact by a grounded shield surrounding the tube, and the tube must be tuned with an insulated tool.

When klystrons are operated at higher voltages, the positive terminal of the beam-voltage power supply is usually grounded as a safety precaution. The desirability of grounding the negative lead becomes less important when a separate power supply is required for the klystron. However, there are some cases when the operation of a high-power klystron with the cathode at ground potential is advisable. For example, the operation of a reflex klystron transmitter tube with the cathode grounded will simplify the use of reflector voltage modulation. The choice between grounding the positive or the negative terminal of the power supply will depend upon the relative difficulty of insulating the radio-frequency output line or insulating the modulation circuit in the reflector lead.

The power transformer should be insulated for a higher voltage when the positive terminal is grounded instead of the negative terminal. This statement refers to a power supply using a full-wave thermionic rectifier. Grounding the negative terminal is equivalent to grounding the center tap of the power transformer; therefore, the maximum voltage from the secondary winding to ground is equal to the voltage from the center tap to one side of the winding. If the positive lead is grounded, first one and then the other side of the secondary winding is grounded as the rectifier becomes conducting, and the opposite side has twice the center tap voltage appearing between it and the ground. This factor becomes quite important in high-voltage supplies, and the insulation of the power transformer should be designed to withstand this extra voltage.

13.3. Frequency Stability Requirements.—Probably the greatest difference between klystron power supplies and the supplies for lower frequency vacuum tubes is the degree of voltage stability required. Conventional low-frequency tubes are relatively insensitive to small changes in voltage. For example, a ripple voltage of one-tenth of 1 per cent would introduce a small amount of amplitude modulation and practically no frequency deviation in a low-frequency triode oscillator. However, with a klystron operated at 500 volts, one-tenth of 1 per cent ripple corresponds to a ½ volt variation in the power supply voltage. Since the electronic tuning rate for a klystron oscillator may be 100 kilocycles per volt, or even greater, this amount of ripple would introduce an amount of frequency modulation that would be undesirable in some applications.

A frequency deviation of several kilocycles can be tolerated in many applications but would not be satisfactory for a narrow-band (60-kilocycle) frequency-modulation communication system. In the latter case the power-supply ripple must be reduced to a few millivolts. It is possible to build filter circuits to suppress the ripple adequately, but it is more convenient to use an electronic voltage regulator which permits the use of a less effective filter and also regulates the output voltage.

13.4. Electronic Regulators.—All electronic voltage regulators may be considered variations or refinements of three basic regulator circuits. The circuit used most frequently is the degenerative type. All the circuits to be described in this chapter are of this type. The degenerative type of regulator consists of

1. A voltage divider across the output,
2. A battery or some other source of constant voltage, such as a gaseous discharge tube,
3. A d-c amplifier,
4. A control tube or gate tube in series with the output lead from the power supply.

When the voltage across the divider exceeds the value of the comparison source, the d-c amplifier increases the negative grid voltage on the control tube in order to reduce the output voltage. Such a circuit may be considered as a negative feedback circuit with a high gain and a large feedback ratio.

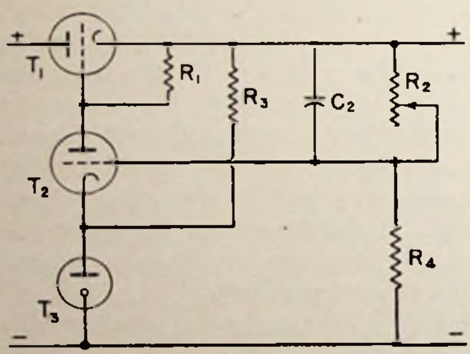


FIG. 13–1.—Basic circuit diagram of a degenerative-type voltage regulator.

The basic circuit of the degenerative type of regulator is shown in Fig. 13–1. Actual values of the circuit constants are not shown because this circuit will be used only for a discussion of general principles and a comparison with other simple regulator circuits. This circuit has some advantages, principally simplicity, and a few disadvantages. The voltage across R_4 will be substantially constant and equal to the voltage of the VR tube T_3. The resistor R_3 is required to furnish the correct operating current for the VR tube. The output voltage will be approximately equal to the ratio of $(R_2 + R_4)/R_4$ times the voltage across the VR tube T_3; therefore the value of R_2 will control the output voltage. The minimum voltage that can be obtained is the sum of the voltage across the VR tube T_3 and the voltage required for the amplifier tube T_2. The maximum voltage that can be obtained is limited by the maximum voltage rating of the amplifier tube T_2.

Only a fraction of the d-c voltage variation is applied to the grid of the amplifier tube, depending on the ratio of R_2 and R_4. However, the addition of condenser C_2 applies all of the ripple voltage to the grid of T_2, so that this circuit is more effective in suppressing ripple than it is in maintaining the output voltage constant.

13.5. Improving the Regulation.—The obvious method of improving the regulation is to increase the gain of the feedback amplifier. For this reason pentodes are normally used instead of triodes. When pentode amplifiers are used, it is possible to apply some of the voltage variations to the screen electrode and improve the operation of the regulator. This method is used in some of the regulators described in the sections that follow.

Other methods of increasing the gain of the feedback amplifier use multistage d-c amplifiers in the feedback circuit. A regulator with a two-stage amplifier is described in Sec. 13.16.

13.6. Protecting the Gate Tube.—All of the voltage variation that would appear at the output of an unregulated power supply must be absorbed by the series control tube, which is commonly known as the "gate tube." Since the total current drawn by the regulator itself and by the load is also passed by the gate tube, this tube must have a fairly high voltage rating and be capable of passing a large current. If one tube does not have a sufficiently high current rating, a number of tubes may be used in parallel. Also, some protection must be provided so that the gate tube will not be overloaded accidentally.

If the power supply is to operate at a fixed output voltage, the voltage across the gate tube may become somewhat greater than the product of the output voltage and the percentage variation of the line voltage. For example, a 20 per cent change in line voltage corresponds to a 200-volt change at 1,000 volts. If the load current is decreased, an additional voltage will occur at the gate tube, and the total voltage can exceed the amount due to the change in the line voltage. Therefore, a well-designed regulator should provide for a gate tube with a plate dissipation equal to the product of the maximum current and about 30 per cent of the maximum voltage expected from the supply.

Either the voltage rating of the gate tube must be increased, or some means of reducing the voltage input to the regulator (such as a Variac in the primary circuit of the power transformer) must be provided if the output voltage of the regulated supply is to be varied. The latter method is used in the high voltage supply described in Sec. 13.16. When the input voltage is not varied as the output voltage is reduced, the voltage drop appears across the gate tube, and a tube with a high voltage rating must be used. If the input voltage is reduced simultaneously, it is possible to use a gate tube which is only required to control the fluctuations caused by variation of the line voltage.

A precaution must be observed when the input voltage is independently variable. It is possible to overload the regulator by continuing to increase the input voltage beyond the proper point. The gate tube may fail to control the output voltage when it is overloaded, and damage to the circuit and tubes will occur. Therefore it is desirable to have a voltmeter across the gate tube to indicate the proper setting of the input voltage. As an added precaution, a voltage overload relay may be used to open the primary circuit if the rating of the gate tube is exceeded.

13.7. Heater Delay.—One difficulty, which is often overlooked in designing a voltage regulator, occurs because the amplifier tubes have indirectly heated cathodes, while the rectifier and gate tubes are usually filament types. As a result, the full rectified voltage builds up before the amplifier tube passes any current. The gate tube has zero grid voltage until the amplifier tube begins to draw current; therefore, the full voltage from the rectifier appears at the output of the regulator for a few seconds until the heater of the amplifier tube reaches its operating temperature. Blown fuses or more disastrous damage may be the result. The cure for this situation is obvious after the existence of the difficulty is recognized. A separate switch for the heater and filament circuits, or a time-delay relay for the high voltage, would solve the problem.

13.8. Elimination of Transients.—The use of several tubes in parallel, as suggested in the previous section for increasing the current rating of a regulator, may introduce difficulties in some cases caused by parasitic oscillations at some audio frequency. The oscillations usually depend upon some obscure arrangement of the wiring which causes feedback. They may be eliminated by adding an *RC* filter to the circuit after the source of the trouble has been located. No general rule can be suggested to solve this problem. In one case it was found that the removal of one particular tube in the parallel combination eliminated the difficulty.

Transients may also be introduced in the output because the regulator is inserted in the grounded lead. The "commutating pulse" from the rectifier tubes may cause a large, highly damped, high-frequency oscillation to appear in the output twice each cycle. The peak value and the appearance of this transient will depend upon the distributed capacity of the transformer windings and may also depend upon the arrangement of the wiring. It will also be affected by the high-frequency response of the feedback amplifier. Since this transient is caused by shock excitation, a small condenser between the plates of the rectifier tubes will reduce the effect to a satisfactory degree. A capacity of 0.005 μf is usually sufficient. This capacity is not always necessary and may be omitted if the ripple is satisfactorily small without it. An inductance in the grounded lead of the filter is another satisfactory method of eliminating the commutation transient.

13.9. Bias Voltage Supplies.—Some klystron types require additional power supplies for beam control electrodes or reflector electrodes. Although all tubes do not use these voltages, they are frequently included in klystron power supplies in order to use a single supply for a variety of klystron types. Supplies for the reflector voltage will be discussed in a separate section. Bias voltage for the control grid or focusing electrode may be obtained from a bleeder resistor across the beam voltage supply if a positive voltage is required, or a self-biasing cathode resistor may be used to obtain a negative bias voltage. However, bias voltages are usually obtained from separate low-voltage power supplies which have been insulated to withstand the maximum beam voltage plus twice the bias voltage (see Sec. 13.2). This means that specially insulated power transformers must be used.

Regulation of the bias voltage may be obtained with a very simple circuit, using a series resistor and a voltage regulator tube as shown in Fig. 13–2. In order to obtain a satisfactorily low ripple voltage from such a supply, it is important to have the cathode of the klystron connected directly to the power transformer, not through the filter, in order to reduce the pickup of transients. When either polarity of bias voltage is provided

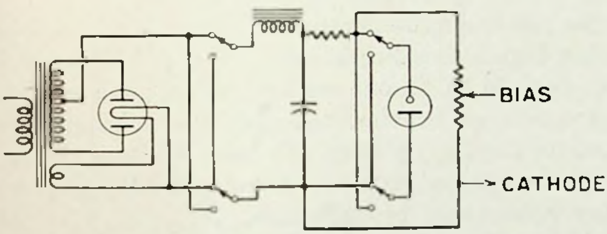


FIG. 13–2.—Circuit diagram for a bias voltage supply for a klystron.

by a reversing switch, a multipole, two-position switch should be used as indicated in Fig. 13–2 in order to avoid inserting the filter in the lead to the cathode. As pointed out in Sec. 13.8, this precaution is not always necessary, and occasionally the accidental arrangement of the wiring may allow the use of a simple two-pole, two-position switch as a reversing switch without introducing serious ripple.

13.10. Cable Pickup.—It is not sufficient to eliminate the ripple from the power supply if the leads from the supply to the klystron are not properly shielded. An improperly shielded cable may introduce enough hum voltage to modulate the output of the klystron appreciably. The use of a braided shield surrounding the cable to the klystron, with a ground connection at only one end, is not only desirable as a shield but a worth-while safety precaution. However, such a shield will not eliminate hum pickup completely if a high impedance is inserted in any lead at the power-supply end of the cable. The same effect may be noticed with an open-circuited cable connected to a cathode-ray oscilloscope. The pickup will be very large until the cable is terminated with a low impedance.

13.11. Power Supplies for Reflex Klystrons.—The voltage required by the reflector electrode makes a power supply for reflex oscillators somewhat more complicated than the power supply for other types of klystrons, but this slight additional requirement is more than offset by the advantages of this type of oscillator. The reflector electrode normally draws no current, although a few milliamperes may be collected when the reflector is operated near the cathode potential. The power requirements are therefore quite small, and it is a simple problem to provide a well-regulated variable voltage for the reflector electrode.

Since the positive terminal of the reflector supply is connected to the cathode, the warning regarding insulation of the transformer, discussed in the last paragraph of Sec. 13.2, applies to power supplies for reflex tubes. If the cathode of the reflex klystron is grounded, the transformer for the reflector supply should be insulated for twice the maximum reflector voltage. The insulation must withstand the maximum beam voltage plus twice the reflector voltage if the anode of the reflex klystron is grounded.

Although the power requirements are small, it is not satisfactory to use a supply with a high internal resistance. The possibility of hum pickup, mentioned in Sec. 13.10, is one reason for avoiding a high impedance circuit. If a condenser is connected from the reflector terminal to the cathode to reduce the hum pickup, the time constant of the circuit may become large if the internal resistance is high, and the response to changes in the reflector voltage may be quite slow.

13.12. Secondary Emission.—Another difficulty appears if the impedance of the reflector voltage supply is high and the tube is operated with the reflector electrode near the cathode potential. Electrons that have been speeded up may strike the reflector with sufficient energy to cause secondary emission. If the reflector surface is a good secondary emitter, the number of secondary electrons may be greater than the number of primaries and the current may flow from the reflector to the anode.

The disadvantage of this secondary-emission behavior is the possibility that the circuit may become unstable. If the reflector accidentally becomes more positive than the cathode, owing to some transient in the power supply, the secondary current may be large enough to maintain the reflector at a positive potential. The energy in the primary electrons which would then be collected by the reflector electrode may be sufficient to overheat the electrode and cause permanent damage to the tube. For this reason, it is important to avoid accidental open circuits in the reflector lead of a high-voltage reflex klystron, since the heat generated by the electron bombardment may be sufficient to melt the reflector electrode.

If it is necessary to introduce a high resistance in the reflector lead, in order to apply a modulation voltage, it is possible for the reflector electrode accidentally to become positive, and the voltage drop in the resistor due to the secondary electron current will not allow the reflector electrode to return to its normal potential. Protection for the klystron can be provided by adding a diode clipper tube from the cathode to the reflector electrode. The clipper tube will not only limit the voltage drop across the resistor by providing a source of electrons; it will also prevent damage to the klystron if the reflector lead is accidentally opened. Of course, such a circuit cannot safeguard the tube if the circuit is opened beyond the point where the diode clipper is connected.

Another simple circuit for limiting the voltage that can appear across a high resistance in series with reflector lead uses a small neon tube shunted across the resistance. This device has the advantage of simplicity and does not require a source of heater power. However, it does not give protection against an open circuit and cannot be used when large modulation voltages appear across the resistance.

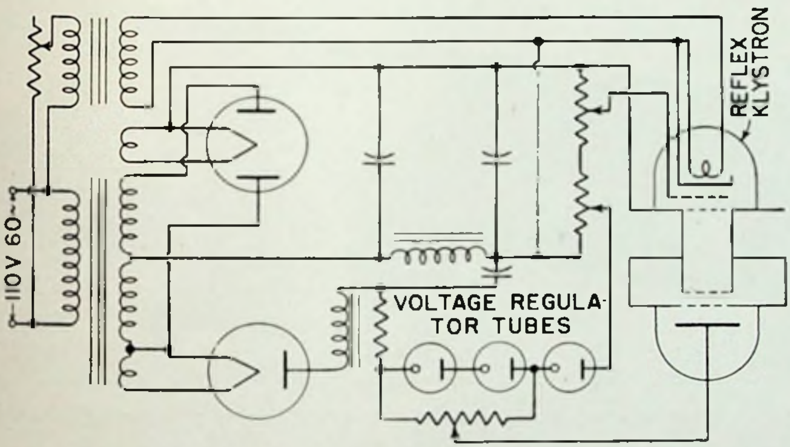


FIG. 13-3.—Voltage compensating circuit for a reflex klystron.

13.13. Voltage Compensating Circuit.—It is possible to maintain the proper voltages on a reflex klystron without regulating the beam voltage if a constant voltage is available for the reflector circuit. This constant voltage is not connected between the cathode and the reflector, however. The phase of the feedback in a reflex oscillator depends only upon the transit time of the electrons in the reflector field. Neglecting the field-free space in a klystron grid having an appreciable thickness, assuming a uniform electric field in the reflector space, and neglecting space-charge effects, the transit time of an electron will be independent of its velocity if the change of reflector voltage is half as great as the change of beam voltage. Reference to Fig. 6–3 shows that this relation is true if the reflector voltage is small. A circuit that utilizes this principle is given in Fig. 13–3.

Actual tubes will deviate from these simple assumptions. It is also obvious from inspection of Fig. 6–3 that the relation is incorrect when the reflector voltage becomes large. The correct adjustment of the circuit constants can be determined experimentally or calculated from measurements of the reflector voltage change necessary to maintain the output at a maximum as the beam voltage is varied.

Test of the circuit in Fig. 13–3 has shown that the compensation is satisfactory for many purposes when the line voltage changes ±10 per cent. However, the circuit does not compensate for frequency changes caused by thermal expansion of the cavity due to the changed power input to the tube when the beam voltage changes. Therefore the use of a regulated beam voltage is recommended for applications where the fre-

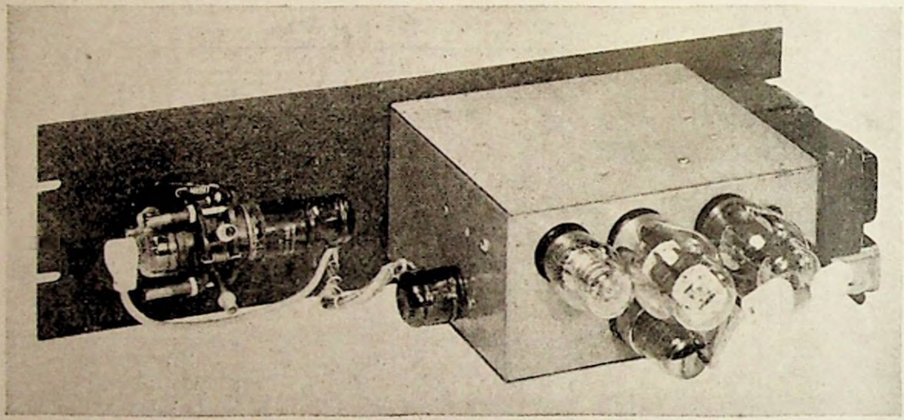


FIG. 13–4.—Type M2 power supply for reflex klystrons. (*Courtesy of Browning Laboratories.*)

quency stability is a critical factor, but this circuit is satisfactory for applications that require constant output and are insensitive to frequency changes.

13.14. Typical Supplies for Reflex Tubes.—It should be realized that the design factors discussed in the previous sections may be varied in innumerable ways to fit particular power-supply requirements. Relatively simple power supplies may be used when the allowable ripple is large. More complicated circuits become necessary when the stability requirements are tightened and when higher powered tubes are used.

A simple power supply which can be used with low-voltage tubes is shown in Fig. 13–4, and the circuit diagram is given in Fig. 13–5. This unit is the Type M2 power supply manufactured by the Browning Laboratories. The source of the regulated reflector voltage is an interesting feature of this power supply. The beam current for the klystron is carried by the voltage regulator tube, and the cathode connection for the klystron is not the negative terminal of the power supply. The voltage across the regulator tube acts as both the comparison voltage for the feedback amplifier and the source of reflector voltage.

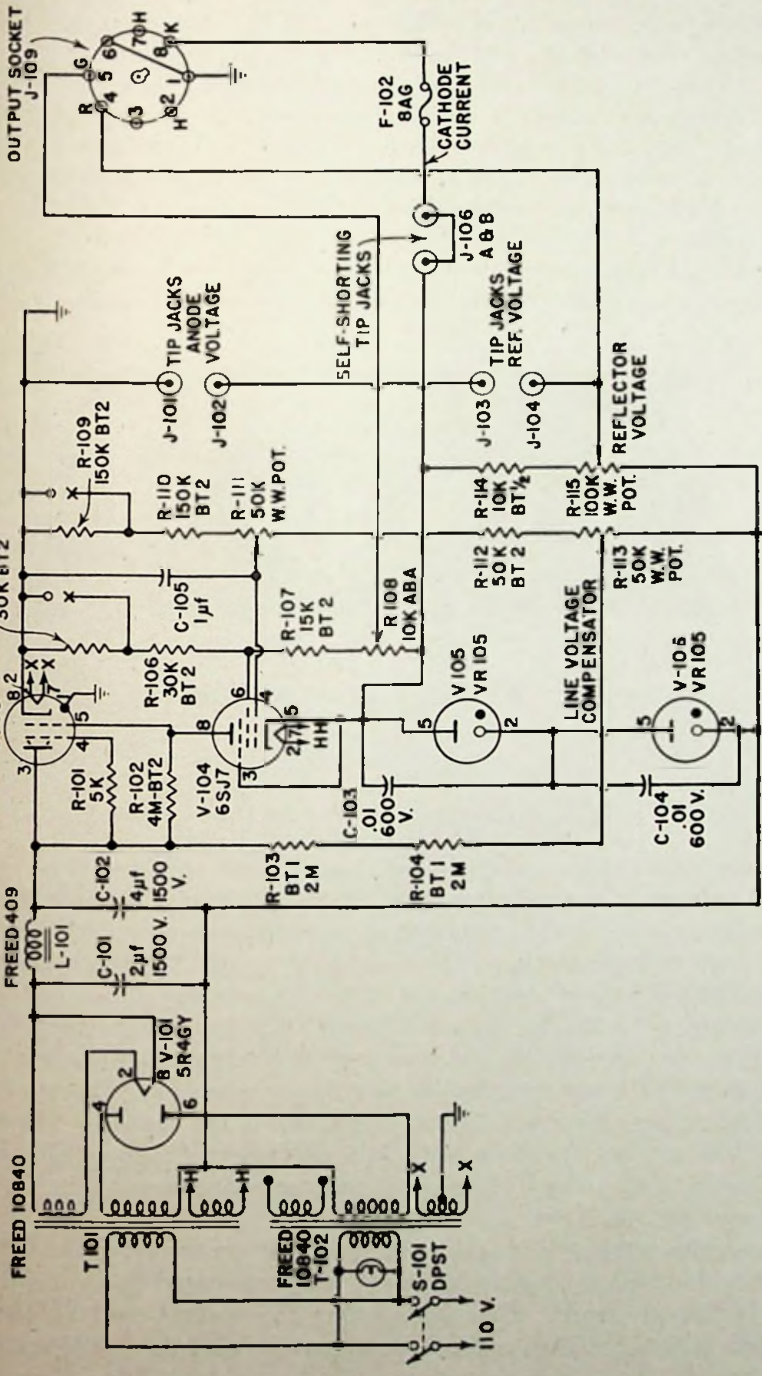


FIG. 13-5.—Circuit diagram of Browning Type M2 power supply.

Another power supply, the Sperry Mk. SX121, has been designed for higher operating voltages and currents and the components are therefore much larger. A photograph of the SX121 power supply is shown in Fig. 13–6. This power supply is designed for medium power reflex klystrons which are frequently used as bench oscillators. A modulator unit including a klystron and tuner, a blower, and an impedance transformer can be added, and the combination is known as the Mk. SX12 klystron signal source.

The circuit diagram for the SX121 power supply is given in Fig. 13–7. A selector switch S101 permits the choice of three different line voltages.

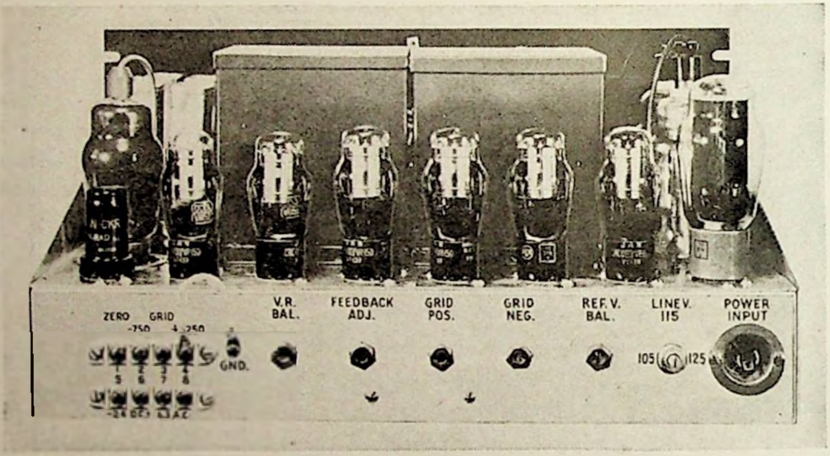


Fig. 13–6.—Sperry Microline Type SX121 power supply.

The gate tube V103 is a high-voltage beam power tube which allows the output voltage to be controlled from 750 to 1,250 volts without changing the input voltage to the regulator. The regulator amplifier V104 also has a high voltage rating. The voltage reference for the regulator is furnished by V105. Potentiometer R119 controls the output voltage of the unit and is located on the front panel. The other adjustments, R112 and R114, are used to change the range of the panel adjustment R119 and to balance the regulator for the best hum suppression. These controls are seldom changed and are located at the rear of the chassis. The screen grid of the amplifier pentode V104 is connected to the input side of the gate tube in order to apply some of the voltage variations to the amplifier to improve the regulation.

A separate rectifier and VR tube regulator are used for the reflector supply and furnish the higher reflector voltages required by some medium-power reflex klystrons. Bias voltage for the control grid or focusing electrode is obtained from potentiometers across VR tubes V105 and V106.

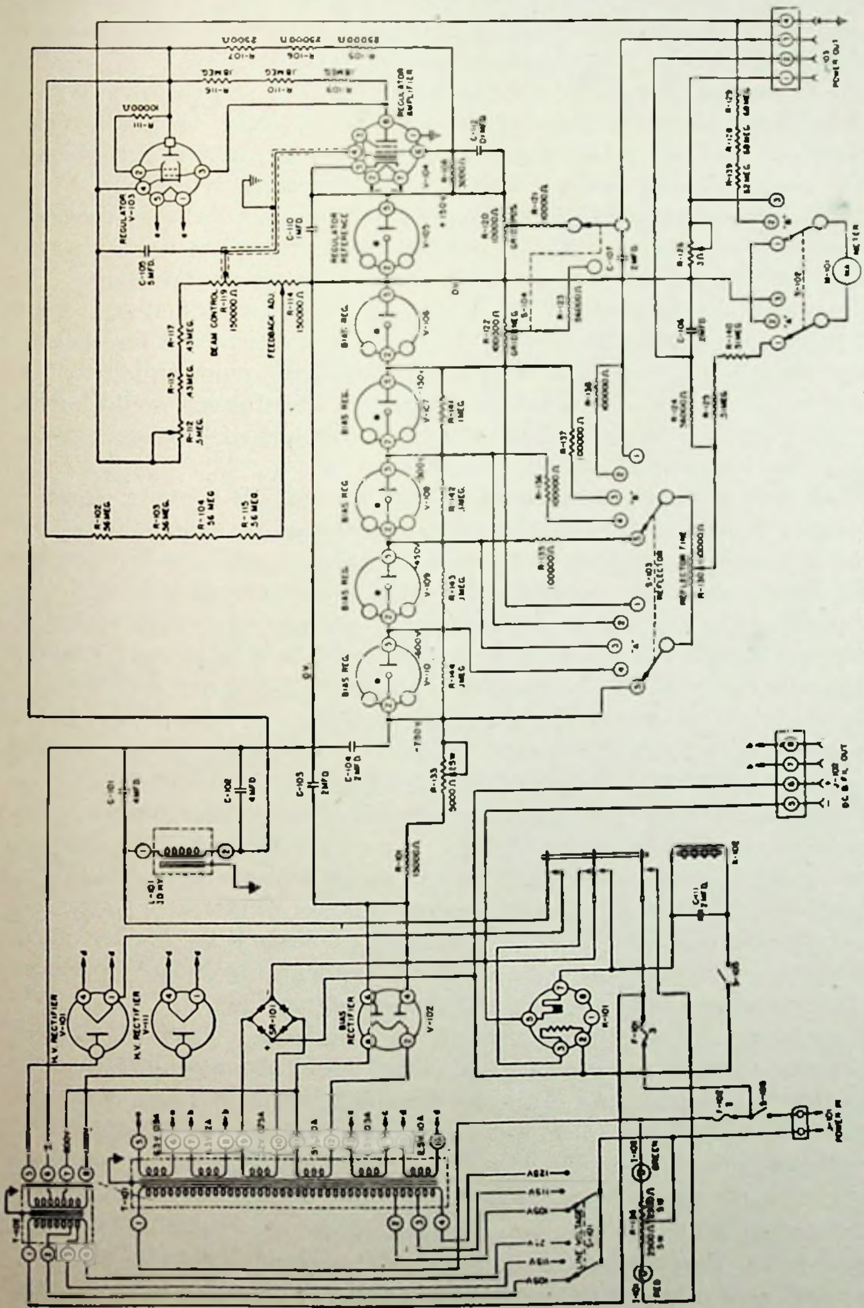


FIG. 13-7.—Circuit diagram for Type SX121 power supply.

Positive bias voltage is obtained when the negative bias control is turned beyond the zero position, actuating a switch which shifts output terminal 3 from potentiometer R122 to potentiometer R120.

The bridge rectifier SR101 furnishes a 28-volt d-c source which is used for pilot lights and to actuate a time-delay relay. The d-c power is also available at terminals 5 and 6 for connection to the d-c motor for the blower in the modulator unit of the Mk. SX12 klystron signal source.

13.15. Radio-frequency Power Supplies.—A number of the requirements of the reflector voltage supply suggest the possibility of rectifying the output of a radio-frequency oscillator to obtain the reflector voltage. The design would not be identical to the supplies used for cathode-ray tubes, but there is enough similarity to make the idea practical. A low-voltage, high-current supply operated at ground potential would supply the d-c power to the oscillator, and a radio-frequency transformer would furnish the necessary insulation so that the low-current, high-voltage reflector supply could operate at cathode potential.

13.16. Power Supply for a Large Voltage Range.—If a power supply is required to cover a large range of voltage, for example, from 250 to 3,000 volts, it is impractical to use an input of more than 3,000 volts and decrease the output voltage by increasing the drop across the gate tube. Very few applications will require such an extreme range of voltage variation, but power supplies of this type are quite useful in a laboratory where a single power supply may be used with numerous types of tubes, or when complete tests of a tube at all voltages are desired. In order to save power and use smaller tubes, the input voltage of such power supplies is controlled manually with a Variac or similar adjustable transformer.

Figure 13–8 shows the circuit diagram for such a power supply. Two gate tubes are used in parallel in order to pass the required beam current. A voltmeter is shown across the gate tubes to indicate the voltage drop across them. This meter can be used to indicate whether the gate tubes are overloaded and also shows when the input voltage is too low for proper regulation. Automatic protection against accidental overload is provided by an overvoltage relay which opens the primary circuit.

A two-stage d-c amplifier is used in this regulator. Two VR105 tubes are used as the voltage reference, and one of these tubes stabilizes the voltage across the second stage of the d-c amplifier. Since the gain of the two stages is high and most of the voltage deviation of the output appears on the grid of the amplifier tube when the voltage control is set for minimum output, the percentage regulation will be quite good. When the voltage control resistance has a high value, giving maximum output voltage, only a small part of the deviation of the output voltage is applied to the amplifier. The regulation is less complete, but the output voltage is also greater and the percentage regulation is the same. All of the ripple voltage appears

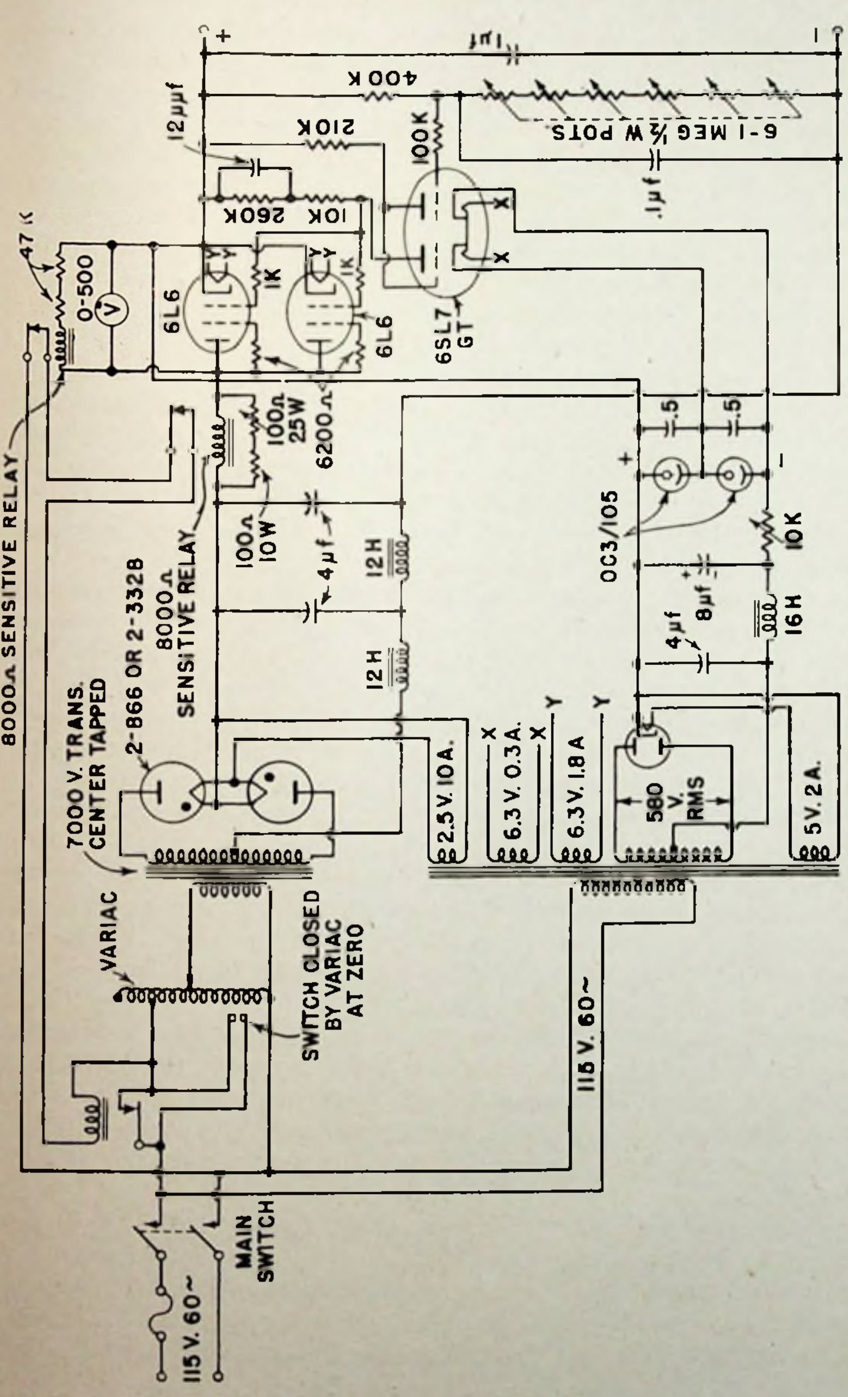


FIG. 13-8.—Circuit diagram for a high-voltage regulated power supply.

at the input of the amplifier because the control resistance is by-passed by a condenser; therefore, the ripple component will be very small.

It is possible to use the main power source as the supply for the amplifier and its VR tubes, but such a circuit would be wasteful of power when operating at high voltages because considerable power would be dissipated in a dropping resistor or additional VR tubes. Therefore a separate rectifier has been provided for the amplifier circuit. This supply also furnishes the voltage reference. The simple VR tube and series-resistor type of regulator may not be satisfactory if large variations of line voltage occur, but may be replaced by an electronic regulator if better control of the output voltage is required.

13.17. Heater Power.—A unipotential cathode is required for velocity-modulation tubes because the beam velocity would be modulated if different parts of the cathode were at different potentials. Indirectly heated cathodes are always used for this reason. Alternating current is quite satisfactory for the heater in most applications. Since the cathode is usually operated below the ground potential, the heater transformer must be insulated to withstand the maximum beam voltage of the tube.

In most klystron applications, the heater and the cathode terminals are connected together at the tube socket. When a direct connection cannot be made, the voltage between the heater and the cathode should be limited to a small value. The connection between the heater and the cathode should preferably be made directly at the socket of the klystron, although it is satisfactory to make it at the power supply, provided separate leads are used for the heater and the cathode so that the a-c voltage drop in the heater lead does not appear as a ripple voltage in the cathode lead.

Since it is difficult to make a heater coil completely noninductive, there may be some variation of the beam current produced by the residual magnetic field. The amount of ripple from this source is small, but for applications that require an extremely low noise level it may be necessary to provide a rectifier and filter for the heater supply, or use batteries.

13.18. Bombarded Cathodes.—Indirectly heated cathodes are usually oxide-coated surfaces because the temperature required for space-charge limited emission from such surfaces is low. However, the oxide-coated surface has serious limitations in high-power, high-voltage tubes. The surface may be easily "poisoned" by impurities that reduce the emission, and bombardment by positive ions may destroy the coating in high-voltage tubes.

A unipotential cathode which is not subject to these difficulties may be obtained by heating a pure metallic surface by electron bombardment. A typical bombarded cathode structure is illustrated in Fig. 13–9. It consists of two separate cathodes and accelerating systems. A directly heated tungsten filament is the cathode for the bombarder system. The electrons emitted from the tungsten filament are accelerated by a bombarder voltage supply. The anode of the bombarder system is a metallic button which is heated by the electron bombardment and serves as the cathode or electron source for the main electron beam of the klystron. The electrons that are emitted from this bombarded cathode button are accelerated along the axis of the tube by the beam voltage supply. The cathode button is made of a metal that is a satisfactory thermionic emitter

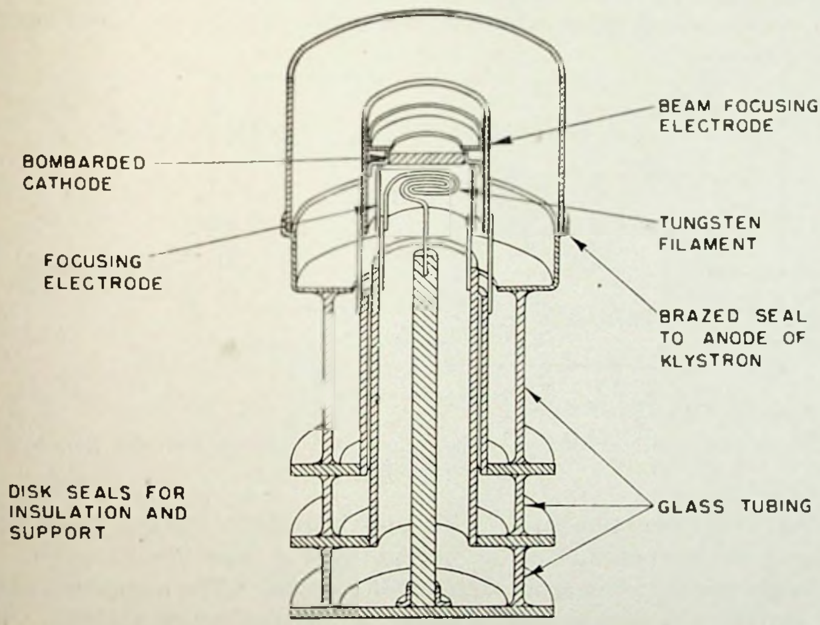


FIG. 13-9.—Construction of a bombarded cathode for a high-voltage klystron.

at a temperature safely below its melting point. Since there is no oxide surface to become contaminated or destroyed, the life of a bombarded cathode can be quite satisfactory.

13.19. Circuits for Bombarded Cathodes.—There are two acceptable ways to operate the bombarder portion of this electron-gun structure. In one case the tungsten filament is operated at a temperature low enough so that the electron current available is limited by the operating temperature of the filament. This type of operation is called "temperature limited." The other method is to operate the tungsten filament at a temperature high enough so that the electron current to the cathode button is determined by the bombarder voltage and is limited by the electron space charge that surrounds the tungsten filament. This type of operation is called "space-charge limited." In both cases the operating temperature of the cathode button is made high enough so that the electron emission from it is space-charge limited.

It is sometimes desirable to operate the tungsten filament at as low a temperature as possible in order to reduce the rate of evaporation of the tungsten and increase the life of the filament. Operation in the region of temperature-limited emission will increase the life of the filament, but it may introduce instability which can cause failure of the tube unless the heater supply is very carefully regulated. Consider a case when the bombarded cathode is at some particular temperature and the tungsten

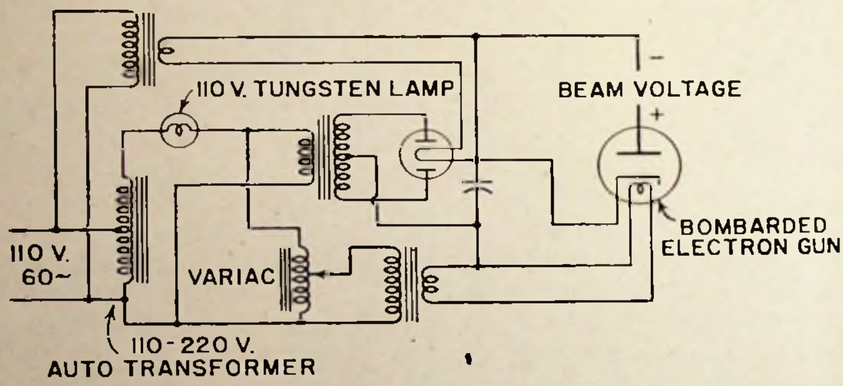


FIG. 13-10.—Protective circuit for temperature-limited emission from the filament of a bombarded cathode.

filament is temperature-limited. A small increase in the line voltage will increase the temperature of the filament and increase the emission. At the same time the bombarder voltage will increase. The increased voltage and current will raise the temperature of the bombarded cathode. The increased cathode temperature will not affect the emission from the cathode button if it is limited by space charge, but some heat will be radiated and cause an additional heating of the tungsten filament. The increased filament emission will cause the cathode to become hotter, and the process can continue until the filament burns out or a portion of the cathode button is melted.

Protection against the thermal instability of a temperature-limited tungsten filament in a bombarded cathode structure may be provided by a simple negative feedback circuit as shown in Fig. 13–10. An incandescent tungsten lamp is inserted in series with the primary of the transformer for the bombarder supply. Any increase in the filament current or the bombarder current will increase the drop in the lamp and reduce the voltage on the filament. The feedback is quite effective because a small change in filament voltage produces a large change in the bombarder current when the filament is temperature-limited. The nonlinear resistance characteristic of the tungsten lamp increases this effect and the circuit becomes quite stable.

The positive terminal of the bombarder supply is also the cathode connection for the klystron; therefore, the insulation for the filament transformer must be equal to the sum of the beam voltage on the klystron and the bombarder voltage. The insulation requirement for the transformer for the bombarder supply is somewhat greater because the center tap of the bombarder transformer is connected to the filament; therefore, it must withstand the beam voltage plus twice the bombarder voltage.

CHAPTER 14

MICROWAVE MEASUREMENT TECHNIQUES

14.1. Introduction.—The problems of making measurements at microwave frequencies are often intimately associated with the problems of design and operation of klystrons. For this reason, it was thought advisable to include in this book a chapter dealing with some of the problems encountered in measurement work and the techniques that have been developed. It is quite beyond the scope of this chapter to give a full and comprehensive treatment of the subject. Some of the salient considerations will be pointed out, however, and comparisons will be made between the techniques of microwaves and the corresponding techniques that are used at lower frequencies.

14.2. The Problems of Microwaves.—One of the most important things to keep in mind about microwave equipment is that nearly all components are electrically "long," and that a piece of equipment that is physically small may nevertheless have dimensions that are an appreciable fraction of a wavelength. It is uncommon to have microwave components that may be regarded as lumped-constant circuit elements over any but a very narrow band of frequencies, although it is often customary and very convenient to explain the behavior of a structure in terms of an equivalent lumped-constant network. But lumped-constant circuit theory has lost much of its usefulness at these short wavelengths.

14.3. Transmission Lines and Microwaves.—The components of a microwave system are generally connected together by electrically long transmission lines, which are in most cases integral parts of the components themselves. For these reasons, transmission-line theory is often very useful in describing the behavior of microwave components, and the characteristics of various components are frequently expressed relative to the characteristics of the interconnecting lines.

The limitations of microwave instruments and components are in many instances the limitations of the transmission lines that are of necessity an integral part of their construction. A microwave transmission line must be shielded to avoid loss of energy by radiation. The three types of line that are most commonly used are flexible coaxial cable, air-dielectric coaxial line, and hollow-pipe wave guides. Flexible cables may be used over a wide band of frequencies but are limited in their application because of the higher attenuation that results from the addition of dielectric losses to conductor losses. Air-dielectric coaxial lines have lower attenuation, but the band width is usually restricted because of the necessity of supporting the center conductor at intervals. Each of these supports, whether it is a dielectric bead or a quarter-wave stub line, is a possible discontinuity in the transmission system which may interfere with the flow of energy along the line. By careful design the effect of the support may be minimized over a band of frequencies, but this band may not be large.

The size of a coaxial line must also be restricted so that only the principal or dominant mode may carry energy down the line. If the arithmetic mean circumference of inner and outer conductors exceeds a wavelength, it is possible for so-called "higher order" or "wave-guide" modes to carry energy down the line, and these may seriously affect the line's characteristics. For frequencies higher than 10,000 megacycles, coaxial lines are excessively small.

14.4. Wave Guides.—Hollow wave guides, which are metallic tubular sections, are widely used for the transmission of radio-frequency power at frequencies above 2,500 megacycles. Their attenuation is low and the absence of a center conductor greatly simplifies the problems of construction. Wave guides are not generally used for frequencies below 2,500 megacycles, as their size becomes excessive. A wave guide of given dimensions has a cutoff frequency, below which energy will not propagate through the guide. If the cross section of the guide is rectangular, the cutoff frequency corresponds to a wavelength that is twice the larger, interior dimension of the guide. The dimensions of wave guides are therefore quite large for frequencies below 2,500 megacycles, and coaxial lines are a more practical form of transmission line for most applications below this frequency.

If the larger dimension of the rectangular wave guide exceeds one wavelength, an additional mode of propagation is above cutoff, and interference between the modes can occur. As a result, a wave guide is a transmission line which may only be used over a rather narrow band of frequencies. The lowest frequency that can be propagated is determined by the cutoff wavelength of the guide. It is undesirable to operate too close to the cutoff frequency, as the attenuation becomes large. Frequencies greater than twice the cutoff frequency should not be used, because of the possibility of multiple modes of transmission.

Other forms of wave guide may be used, but they have similar limitations regarding cutoff frequencies and multiple-mode propagation. Guides of circular cross section have even less band width than do rectangular guides. In addition, it is difficult to maintain the plane of polarization in a guide of circular cross section. Guides of rectangular cross section are therefore almost universally used.

IMPEDANCE MEASUREMENT

14.5. Standing Waves on Transmission Lines.—The impedance of a microwave component or structure is nearly always referred to the transmission line that is connected to, or forms an integral part of, the component. Impedance relationships on transmission lines have been discussed in a number of publications,[1–3] and we shall review here only the fundamental principles which are involved in the measurement techniques. These apply rigorously only to lines whose attenuation is zero, but the attenuation per wavelength of most microwave transmission lines is low enough for this approximation to be satisfactory.

Any transmission line has associated with it a quantity known as its "characteristic impedance," indicated by the symbol Z_0. This is the impedance which will be measured at the input terminals of an infinitely long transmission line and which will be encountered by an electromagnetic wave that is traveling along the line. If at the end of a transmission line the traveling wave encounters a load impedance Z_L that is different from Z_0, part of the traveling wave will in general be absorbed in this load impedance, but part of it will be reflected back down the transmission line toward the source of energy. The electric field associated with the incident wave will be represented by the symbol E_1 and the field associated with the reflected wave will be designated E_2.

The strength of the electric field as measured at any point along the transmission line that is carrying energy to the load impedance will be the vector sum of the contributions from each of the traveling waves. The maximum field will be measured at the points where the two contributions add in phase, and is given by the sum of the magnitudes of the two fields.

$$E_{\max} = |E_1| + |E_2| \tag{14-1}$$

The points of maximum electric field strength will be spaced along the transmission line at intervals of a half wavelength as measured on the line. For a uniform, low-loss, air-dielectric coaxial line, a half wavelength on the line will be very nearly equal to a half wavelength in free space. If the conductors of the coaxial line are separated with a solid dielectric, the wavelength on the line will be reduced by the square root of the dielectric constant, and the voltage maxima will be spaced correspondingly closer together. The wavelength as measured in a uniform, air-dielectric wave

[1] J. C. Slater, *Microwave Transmission*, McGraw-Hill Book Company, Inc., New York, 1942.

[2] F. E. Terman, *Radio Engineers' Handbook*, McGraw-Hill Book Company, Inc., New York, 1943.

[3] S. Ramo and J. R. Whinnery, *Fields and Waves in Modern Radio*, John Wiley & Sons, Inc., New York, 1944.

guide is greater than the wavelength in free space, and the spacing between voltage maxima is correspondingly greater.

Midway between the maxima, the electric field components of the traveling waves will be of such phase as to tend to cancel each other, and a minimum value of field strength will be measured whose magnitude is given by the difference between the magnitudes of the two fields.

$$E_{min} = |E_1| - |E_2| \tag{14-2}$$

14.6. Standing Wave Ratio.—It will be seen that when the electric field strength is measured as a function of distance along the line, a standing wave pattern will be observed. The ratio of the peak or maximum field strength E_{max} to the minimum field strength E_{min} is known as the "standing wave ratio," frequently abbreviated SWR. In some cases this ratio may be designated VSWR. The latter term denotes "voltage standing wave ratio," and is used to differentiate between the voltage standing wave ratio and the ratio of the readings from a square-law detector, which if used directly gives the standing wave ratio in terms of a power ratio. The definition of the voltage standing wave ratio may be written

$$\text{VSWR} = \frac{E_{max}}{E_{min}} = \frac{|E_1| + |E_2|}{|E_1| - |E_2|} \tag{14-3}$$

If the load impedance at the end of the transmission line is a short circuit, an open circuit, or a pure reactance, and is unable to absorb any power from the incident wave of electromagnetic energy, there will be complete reflection of this incident wave at the load, and $|E_1|$ will be equal to $|E_2|$. When the two traveling waves are in opposing phase on the input transmission line, there will be complete cancellation because the two waves are equal in amplitude, and E_{min} will equal zero. The standing wave ratio on the input line will therefore be infinite. If the load impedance is equal to the characteristic impedance of the transmission line, the incident wave will be completely absorbed by the load impedance and there will be no reflected wave. Therefore $E_{max} = E_{min} = |E_1|$, and the standing wave ratio will be unity.

14.7. Reflection Coefficient.—It will be convenient to express the standing wave ratio in terms of the ratio of the field strength of the reflected wave to that of the incident wave. This ratio E_2/E_1 is known as the "reflection coefficient" and is designated by the symbol ρ. Equation (14–3) may be rewritten

$$\text{VSWR} = \frac{1 + |E_2/E_1|}{1 - |E_2/E_1|} = \frac{1 + |\rho|}{1 - |\rho|} \tag{14-4}$$

Complete reflection corresponds to an infinite standing wave ratio but is represented by a unit value of the reflection coefficient. It is often more convenient to plot the magnitude of the reflection coefficient, which does not become infinite. Since the VSWR is the quantity that is convenient to measure, a relation giving the magnitude of the reflection coefficient in terms of the standing wave ratio is quite useful. This relation may be obtained by rearranging Eq. (14-4).

$$|\rho| = \frac{\text{VSWR} - 1}{\text{VSWR} + 1} \tag{14-5}$$

The value of the reflection coefficient and therefore the value of the standing wave ratio are determined by the ratio of the load impedance Z_L to the characteristic impedance Z_0 of the transmission line. It is a measure

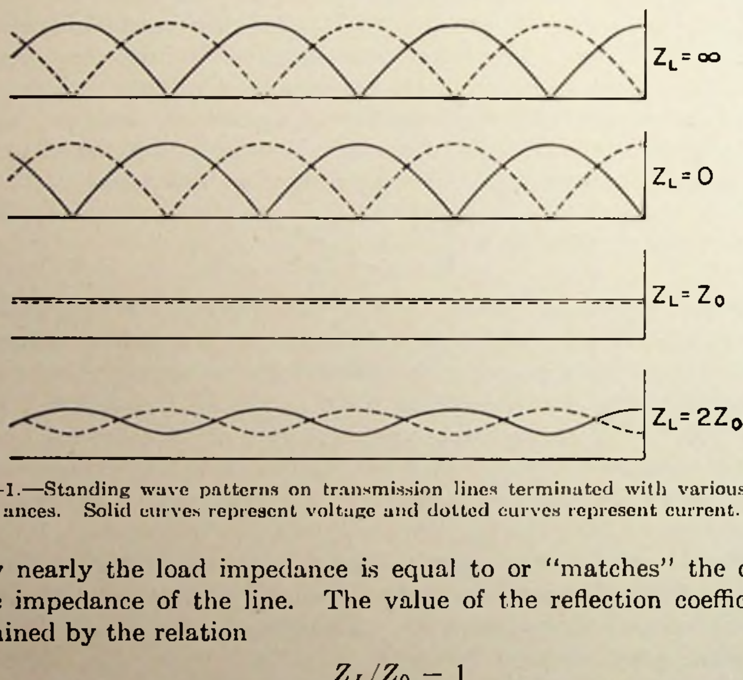


FIG. 14–1.—Standing wave patterns on transmission lines terminated with various impedances. Solid curves represent voltage and dotted curves represent current.

of how nearly the load impedance is equal to or "matches" the characteristic impedance of the line. The value of the reflection coefficient is determined by the relation

$$\rho = \frac{Z_L/Z_0 - 1}{Z_L/Z_0 + 1} \tag{14-6}$$

Equation (14-6) involves complex quantities, since the load impedance may have a reactive as well as a resistive component. The reflection coefficient will have a phase angle as well as a magnitude. The phase angle is determined by the complex ratio Z_L/Z_0, but it is usually measured in terms of the position of the standing waves along the transmission line.

Typical standing wave patterns are shown in Fig. 14–1 for an open-circuited line ($Z_L = \infty$), for a shorted line ($Z_L = 0$), for a matched termination ($Z_L = Z_0$), and for a load impedance that is twice Z_0.

14.8. Use of Standing Waves to Measure Impedance.—The standing wave ratio and position of the standing waves on a lossless transmission line are established by the ratio Z_L/Z_0, and the standing wave pattern may therefore be used to determine this ratio. If the absolute magnitude of the quantity Z_0 is known, the absolute magnitude of Z_L may be established. The characteristic impedance of coaxial lines may be calculated from the radii of the conducting surfaces of the line.

$$Z_0 = 138 \log_{10} \frac{r_{\text{outer}}}{r_{\text{inner}}} \tag{14-7}$$

where r_{outer} is the inside radius of the outer conductor and r_{inner} is the

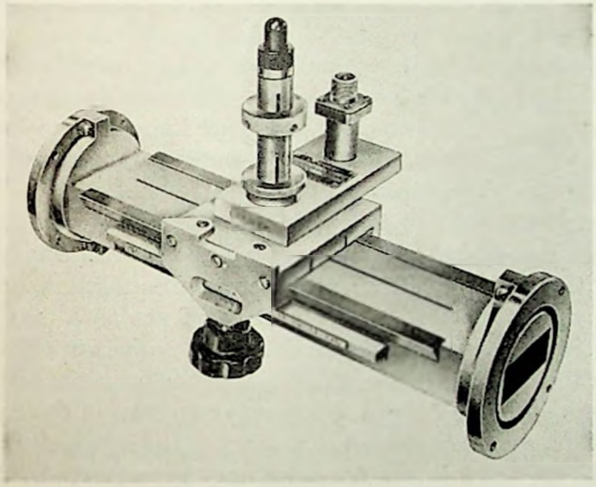


Fig. 14-2.—Standing wave detector using a slotted wave guide.

radius of the inner conductor. There is no similar unambiguous definition for wave guides. Nevertheless, in the great majority of instances one is interested in the ratio Z_L/Z_0 rather than in the absolute magnitude of Z_L, and any ambiguity in the definition of characteristic impedance is seldom troublesome.

The microwave impedance meter is a device to measure the standing waves along a transmission line that is carrying energy to the unknown impedance and is known as a "standing wave detector." The most common form uses a pickup device which is capable of indicating the relative field strength at any one point on the line and which may be moved along the line to measure the variation of field strength at different positions along the line. A photograph of a typical standing wave detector is shown in Fig. 14-2.

In a coaxial line or wave guide, the electromagnetic fields are completely enclosed within a tubular outer conductor. This outer conductor is slotted so that measurements can be made of the fields inside. A pickup probe is inserted through the slot parallel to the electric lines of force, and projects a short distance inside the outer conductor.[1] This probe extracts from the line an amount of energy proportional to the square of the electric field strength at the point of insertion, but the amount of energy extracted is in general small enough not to disturb appreciably the fields within the line. The energy picked up by the probe is fed to some sort of detector, such as a crystal rectifier or hot-wire detector. As the probe is moved along the slot, the detector indication will provide a measure of the variation of field strength as a function of position along the line.

The standing waves that are measured in this manner may be used to calculate the load impedance at the end of the transmission line. Frequently there is no need for this calculation, as a knowledge of the standing wave pattern is in itself sufficient information for many purposes. For this reason the slotted line and traveling probe are known as a "standing wave detector," since it is information about standing waves rather than impedances that is directly obtained by the instrument.

Many advantages are gained by using a standing wave detector constructed on the same type of transmission line that leads to the unknown impedance. If an adaptor is used to connect the unknown impedance to the standing wave detector, the adaptor must act as a matching device between the two types of transmission line. Changes in the shape of the conductors or the dielectric between them must be properly designed; otherwise an adaptor will affect the standing wave pattern in the input line and cause a resulting error in the measurement. If the discontinuity of the adaptor is known, the measured results may be corrected to account for it, but the measurement procedure will be greatly complicated thereby.

14.9. Graphical Aids to Impedance Calculations.—Analytical methods by which impedance relationships on transmission lines may be determined from standing wave patterns have been outlined in various publications.[2,3] These calculations are frequently tedious and time-consuming. A great deal of time and effort may be saved by the application of graphical methods of calculations; at the same time it is often possible to obtain a clearer picture of the variation of impedance with some parameter such as

[1] A coupling loop which links magnetic lines of force within the line may also be used as a pickup device, and the energy extracted will be proportional to the square of the magnetic field strength. The probe is more satisfactory for most uses because of its greater simplicity and smaller dimensions.

[2] J. C. Slater, *Microwave Transmission*, McGraw-Hill Book Company, Inc., New York, 1942.

[3] S. A. Schelkunoff, *Electromagnetic Waves*, D. Van Nostrand Company, Inc., New York, 1943.

changing frequency. The graphical calculations are greatly aided by the use of various forms of transmission-line charts, also known as "circle diagrams" and "impedance diagrams." These charts are directly applicable only to lines with no attenuation, although losses may in some cases be taken into account by additional manipulations.

One form of transmission-line chart, often known as a "rectangular impedance chart," is shown in Fig. 14–3. The rectangular coordinate background of this chart is an impedance plane, and any point on this plane represents a particular value of impedance. The real component

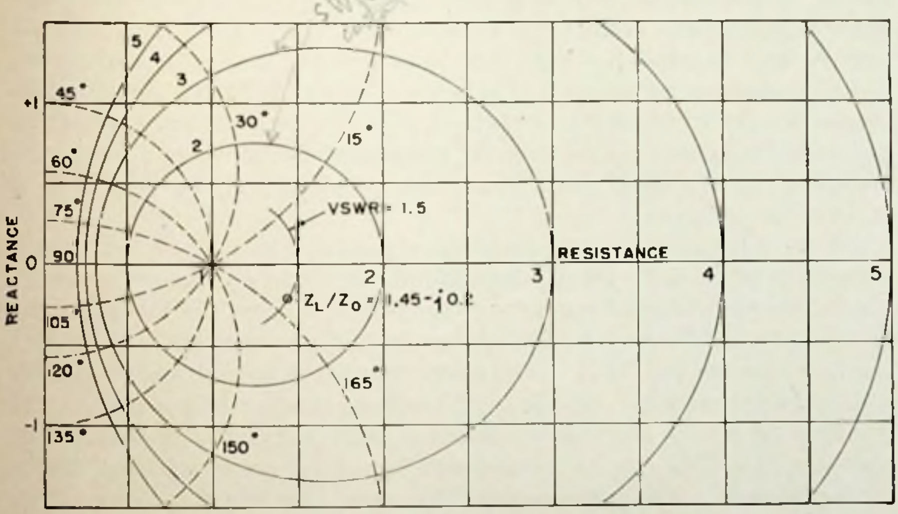


FIG. 14–3.—Impedance chart using rectangular coordinates.

of any given impedance is measured along the horizontal axis of this coordinate system, and the reactive component is measured along the vertical axis. All values of impedance are normalized with respect to Z_0, the characteristic impedance of the input transmission line along which the standing waves are measured.

An impedance equal to the characteristic impedance of the transmission line is represented on this plane by the point 1,0. Each of the family of eccentric circles that enclose this point is a locus of impedances which, when used to terminate the transmission line, will produce equal standing wave ratios. The standing wave ratio that will be produced by these impedances is indicated by the number with which each circle is labeled.

The second family of circles, orthogonal to the first and passing through the point 1,0, provides a measure of distance along the transmission line. Each of these circles is labeled with a value in degrees, which indicates the electrical length of the transmission line between the impedances through which the circle passes and the nearest point of maximum electrical field strength on the input transmission line.

As an example, the normalized impedance $Z_L/Z_0 = 1.45 - j0.2$ is plotted on Fig. 14–3. This impedance is seen to set up a standing wave ratio of 1.5 on the input transmission line, and the nearest electric field maximum on the line will be located at a distance corresponding to 170 electrical degrees from the load impedance.

If the equations that express the impedance relationships on a transmission line are rewritten in terms of admittances, they will have identical forms. It is therefore possible to use a chart such as Fig. 14–3 to represent either impedance or admittance relationships on a transmission line. It must be kept in mind, however, that an impedance maximum corresponds to an admittance minimum. The second family of circles therefore indicates the electrical length of transmission line between the load admittance and the nearest minimum in the electrical field rather than the nearest maximum, on the input line. Otherwise problems may be dealt with in an identical manner.

The rectangular impedance coordinates used in Fig. 14–3 are not always convenient because the plane is semiinfinite in extent and points are somewhat crowded in a small region near the origin. These disadvantages are overcome by a different form of impedance chart proposed by P. H. Smith [1] and illustrated in Fig. 14–4. The entire complex plane is transformed into a circle of unit radius. The lines of constant resistance become a family of circles tangent to the unit circle at the right side of the diagram. The reactance lines become an orthogonal family of circles tangent to the horizontal axis. The phase-angle lines are now equally spaced lines radiating from the center of the unit circle and the curves of constant standing wave ratio are concentric circles.

Large values of standing wave ratio are closely spaced near the outer circle. This crowding may be avoided by plotting the reflection coefficient instead of the standing wave ratio. Curves corresponding to a constant magnitude of the reflection coefficient are concentric circles with radii equal to the magnitude of the reflection coefficient. The phase angle of the reflection coefficient is given by the radial lines. This simplification allows plotting the position of the standing wave or phase angle and the reflection coefficient on ordinary polar coordinate paper. Then the point may be transferred to an impedance circle diagram with the same unit radius.

14.10. Rieke Diagrams.—These transmission-line charts or impedance charts are frequently used to help analyze the behavior and characteristics of microwave oscillators. The power output and frequency of the oscil-

[1] P. H. Smith, *Transmission Line Calculator, Electronics*, January, 1939; January, 1944; and March, 1945.

lator are measured as functions of changing load impedance, and contours of constant power and constant frequency that are characteristic of the oscillator may be plotted on the impedance charts. Charts of this kind are generally known as "Rieke diagrams" (see Sec. 4.15). Although the name was originally applied to polar diagrams for magnetron oscillators, the term has been adopted for similar diagrams for other microwave oscillators. Examples of Rieke diagrams plotted on rectangular impedance paper are shown in Figs. 4-12, 7-11, 7-12, and 8-14.

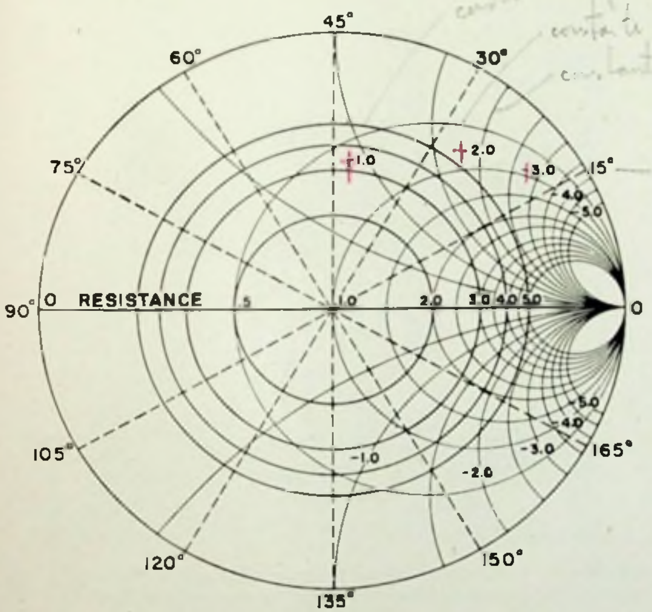


FIG. 14-4.—Circular impedance chart.

14.11. Impedance Transformers.—The variation of impedance along a transmission line when standing waves are present can be used as a means of matching impedances. Microwave impedance transformers are based on this principle.

One of the simplest impedance transformers is a variable length line, or trombone. Changing the length of the line is equivalent to traveling around a circle of constant standing wave ratio. As an example, if the standing wave ratio is 1.5, changing the length of the line by a quarter wavelength varies the impedance between 0.67 and 1.5. It should be noted that such a transformer also changes the reactive component of the impedance; in many cases this changing reactive component causes no trouble, although it will affect the frequency of an oscillator. The principal advantages of a trombone as an impedance transformer are its simplicity and the fact that it requires only a single adjustment. Its disadvantages are the possibility of introducing serious losses from poor sliding contacts and the fact that a flexible line is usually required. Trombones are often convenient in experimental setups but are seldom recommended for permanent installations.

Shorted sections of transmission line whose lengths are adjustable may be used to provide a variable reactance. A single, adjustable stub line

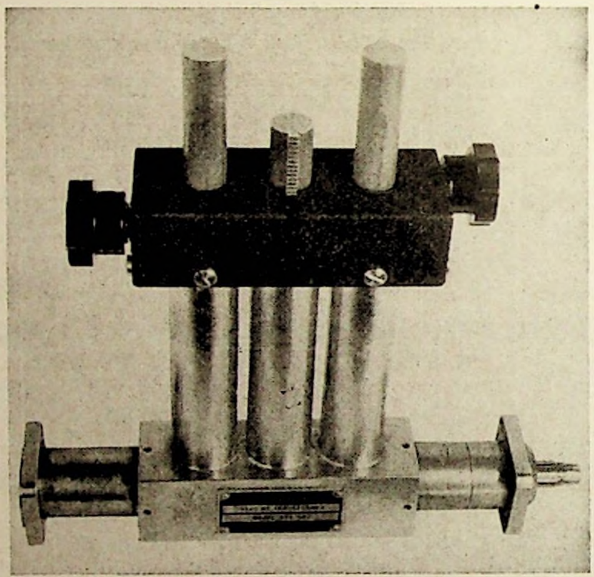


FIG. 14-5.—Impedance transformer using three variable stub lines with the two outside stubs varied simultaneously by the same control.

in shunt with the main transmission line makes a satisfactory impedance transformer if its position along the main line can be varied. One method of effectively varying the position of the stub line is to use a trombone in the main transmission line. In most cases the use of a trombone to vary the position of the stub line is not advisable because both the input and output of the impedance transformer must be rigid. Two or more variable length stubs may be placed at fixed positions along the line to overcome this difficulty. One rather satisfactory design shown in Fig. 14-5 uses three adjustable stubs, spaced apart a quarter wavelength, with the first and third stubs ganged together. If the losses in the stubs are negligible, this device will match any two impedances provided that one is not a pure reactance.

Variable impedance transformers in wave guide may be built using stub lengths of guide connected to the main guide, but it is often simpler mechanically to use a metal probe or rod that extends into the wave guide parallel to the electric lines of force. This probe is equivalent to a shunting reactance, whose magnitude may be varied by changing the probe penetration. A single probe may be used whose position is variable, or multiple probes may be installed that are fixed in position, but of adjustable penetration.

To provide a fixed impedance transformation in coaxial lines, it is customary to use a shorted stub line in shunt with the main line as shown

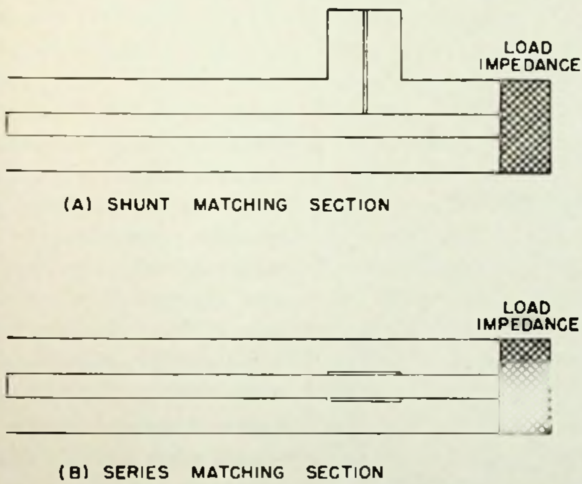


FIG. 14–6.—Matching sections in coaxial lines.

in Fig. 14–6*A*, whose length and location are chosen to accomplish the desired transformation. It is also common practice to put a short section of line of a different impedance in series with the main line. This type of matching section is illustrated in Fig. 14–6*B*. The length, position, and impedance of this matching section may be chosen to accomplish the desired results. Quarter-wave transformer sections are most widely used and are constructed by placing a metal sleeve over the center conductor or inside the outer conductor to change the line impedance.

Fixed impedance transformations in wave-guide transmission systems are accomplished with the use of thin metal diaphragms that partly block the guide, or with tuning screws or probes that extend part way across the guide. Both structures are equivalent to reactances that shunt the transmission line. A diaphragm inserted in the side of a wave guide as shown in Fig. 14–7*A* introduces an inductive shunt reactance, and a similar structure at the top or bottom of the wave guide introduces a capacitive shunt reactance.

14.12. Double-slug Transformer.—Sliding contacts between conductors are frequently unsatisfactory at microwave frequencies because they may introduce losses that are appreciable and erratic. It is therefore often difficult to determine whether trombones or stub transformers that employ sliding contacts are operating satisfactorily. Designs that avoid sliding contacts are preferred for this reason.

One type of transformer that avoids the difficulties of sliding contacts is the "double-slug" transformer. A schematic diagram of a coaxial double-slug transformer is shown in Fig. 14–8. Two dielectric beads or metal slugs, preferably a quarter wavelength long, are mounted with the transmission line and change the impedance of the line. If the two quarter-wave slugs are together, they represent a half-wave section of transmission line of different characteristic impedance, which does not affect the standing waves on the input line. But if the slugs are separated, they will introduce a reflection whose magnitude depends upon the spacing between the slugs and which is a maximum when this spacing is a quarter wavelength.

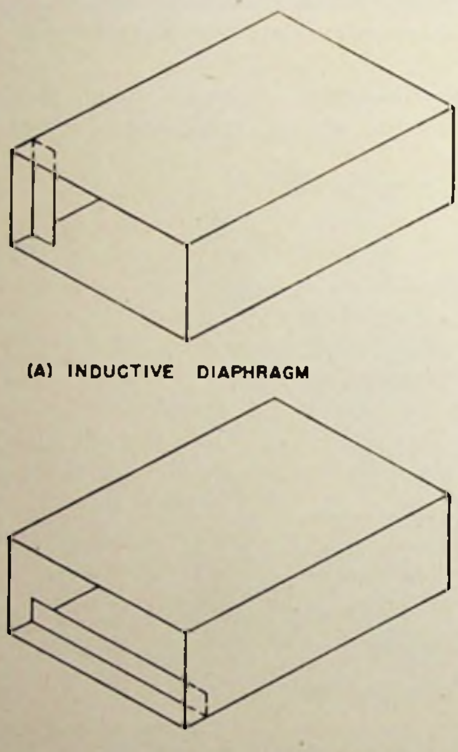


Fig. 14–7.—Wave-guide diaphragms.

Two controls are provided for the movable slugs. The first changes their position on the line, while holding constant the spacing between them. This changes the phase of their over-all reflection without altering its magnitude. The second control changes the spacing between the slugs by moving each an equal amount in opposite directions. This changes the magnitude of the over-all reflection without appreciably varying its phase. It is therefore possible to control independently the magnitude and phase of the standing wave ratio intro-

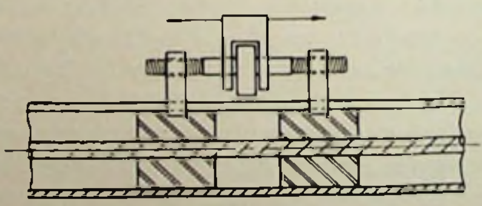


Fig. 14–8.—Double-slug impedance transformer.

duced by the transformer, which may be used to cancel any standing wave already present on the line. Tuning adjustments are rapidly convergent, and much experimental work is simplified, such as measuring Rieke diagrams of oscillators (see Secs. 14.10 and 14.34).

14.13. Band-width Considerations.—A practical consideration that frequently arises in connection with the use of variable or fixed impedance transformers is the "long line effect." This effect results from the use of electrically long transmission lines to supply energy to mismatched load impedances. Even though the load impedance at the far end of the line may not vary with changing frequency, the electrical length of the transmission line does change. Correspondingly, the input impedance of the mismatched line may be a rapidly changing function of frequency. The frequency sensitivity of the input impedance becomes greater as the electrical length of the transmission line is increased.

An oscillator such as a klystron that is required to deliver power into a transmission line under these conditions may be unstable because the load presented to the klystron can change considerably with small changes in frequency. The frequency of oscillation may jump between two different frequencies, and there may be "forbidden" frequencies to which the oscillator cannot be tuned. An amplifier will not become unstable when long line effects are present because its frequency is determined by an independent source, but the frequency-modulation characteristic of an amplifier will have considerable amplitude distortion. The frequency-modulation characteristic of an oscillator may be even more seriously distorted and have discontinuities in the frequency characteristic.

For these reasons, it is desirable to have a transmission-line system of minimum length. If a long transmission system is unavoidable, it is of primary importance that the sections of the system that are resonant, *i.e.*, on which standing waves exist, are held to a minimum length. If some discontinuity or source of reflections in the system is unavoidable, the impedance transformer that is required to cancel the undesired reflection should be inserted as near as possible to the source of the reflection. If the klystron requires a large standing wave ratio to obtain maximum output, the impedance transformer introducing the standing wave should be located as close as possible to the klystron.

Even if care is taken to minimize resonant lengths of line in this manner, the results may still be unsatisfactory. Appreciable lengths of transmission line may be an integral part of the impedance transformer that is inserted, or of the component that is producing the undesired reflection. For this reason, it is usually considered good practice to minimize reflections by constructing a fixed matching device as an integral part of each microwave component. In this manner, the length of line between the source of the reflection and the associated matching transformer may be held to a minimum, and the frequency sensitivity of the impedance of the component will be minimized.

A variable impedance transformer, designed as a separate component to be inserted in a transmission line, will usually be more sensitive to frequency changes than a fixed transformer. There may be an appreciable length of line in the transformer itself, and the connecting lines that are required will add to the difficulty of obtaining a good design. As a result, the band width over which a variable transformer presents a well-matched impedance is usually smaller than the band width obtainable with a fixed transformer.

14.14. Improving the Band Width of a Component.—Although proper care is taken in the design of a microwave component and it is matched at a single frequency, the frequency sensitivity of its impedance may be greater than desired. The band width over which the component presents a well-matched impedance may usually be improved by the addition of some sort of microwave equalizer which is made an integral part of the component. One of the simplest forms of such a network is a parallel resonant circuit shunting the transmission line. If the constants of the resonant circuit are properly chosen and the circuit is correctly located along the line, the susceptance-frequency characteristic of the resonant circuit may be used to cancel the susceptance-frequency characteristic of the component. With coaxial transmission lines, the parallel resonant circuit usually takes the form of a shorted stub coaxial line, whose length, position, and characteristic impedance are chosen to give the desired compensating action. With rectangular wave guides, a resonant window of the type shown in Fig. 14–9 may be used. This window is a combination of the inductive and capacitive windows shown in Fig. 14–7. The height and width of the opening and the position of the window are the design factors that must be properly chosen in this case.

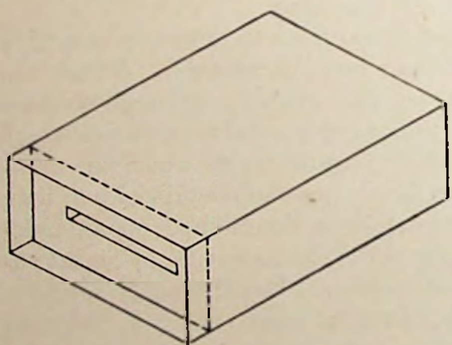


FIG. 14–9.—Resonant window in a wave guide.

POWER MEASUREMENT

14.15. The Need for Microwave Power Measurements.—At low frequencies, the power level in a circuit may be measured by finding the voltage across, or the current through, a known impedance. But at higher frequencies it is increasingly difficult to construct accurate voltage or current meters that do not disturb the circuits being tested. Furthermore, at microwave frequencies where components are connected by transmission lines of appreciable electrical length, voltage and current as well as impedance are rapidly varying functions of position unless the transmission line is matched. The power level in a line does not vary along the length of the line if losses in the line may be neglected, and is not changed if the impedance of the line varies. Power is therefore a quantity with more meaning at microwaves than voltage or current, and power is measured instead of these quantities.

14.16. Low-frequency Techniques That May Be Used with Microwaves.—Certain techniques of voltage and current measurement that are used at lower frequencies are still of value when applied to microwaves. Crystal rectifiers may be used to give an indication of power level but must be calibrated against some standard of measurement. Crystals also suffer from instability and temperature sensitivity and, unless carefully used, they have little value except to give qualitative indications. Thermocouples are also used with microwaves, but suffer from a necessity for calibration, and have not found wide application.

14.17. Calorimeters for Microwave Power Measurement.—Calorimeters may be used as a standard of power measurement at microwave frequencies. The power to be measured is converted into heat energy that is used to heat a liquid. The amount of power may then be measured by determining the flow rate and temperature rise of the liquid—usually water. This technique of measurement, although useful for high powers, suffers from a number of limitations. The minimum power that can be accurately measured is a few watts. Furthermore, the required equipment is bulky, and a considerable time is required to perform a measurement. In addition, all of the power being measured is dissipated in the liquid; it is not practical to monitor the power flow into a more useful load.

14.18. Bolometric Methods of Measurement.—Many of the handicaps associated with calorimeter measurements are overcome with bolometric methods of measurement. A simplified circuit diagram of a bolometric wattmeter is shown in Fig. 14–10. The heart of the instrument is a power-sensitive resistance element, or bolometer. The resistance of this element varies with its temperature, which in turn is a function of the power being absorbed in the element. Used as a power-measuring instrument, the bolometer is first heated by the direct current in the bridge circuit until its resistance reaches a value that brings the bridge to balance. It is then heated further by microwave energy, and the resulting change in resistance is used to measure the microwave power.

Bolometers may be used as absolute standards for power measurement if direct current and microwave current cause equivalent changes in resistance. In well-designed units, this assumption is justified. Several types of bolometer are used at microwave frequencies. Short lengths of fine wire are frequently used. Some early instruments employed 5- or 10-ma Littelfuses as bolometer elements. The sensitive resistance element is the fuse wire, which is a short length of Wollaston wire (0.00006 in. diameter platinum). More recently, these wires have been incorporated in a cartridge that has been designed specifically for microwave power

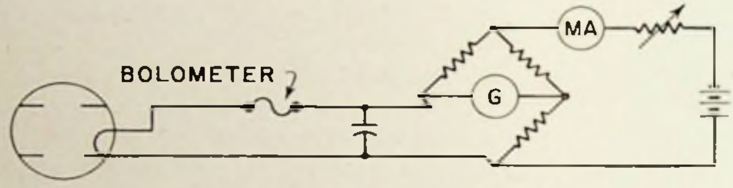


Fig. 14-10.—Bridge circuit for a bolometer-type wattmeter.

measurements. An illustration showing the construction of a hot-wire unit is shown in Fig. 14-11. These elements are known as "barretters" and are capable of measuring power in the range from 10^{-5} to 10^{-2} watt. Elements for measuring higher powers are built with heavier wires of tungsten or carbon, and the power-handling capacity may be increased to over 10 watts by suspending a 12 mm long, 0.0005 in. diameter tungsten wire in an atmosphere of hydrogen.

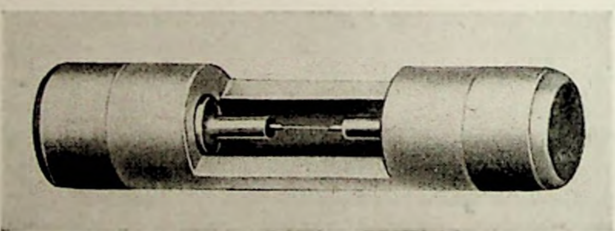


Fig. 14-11.—Cutaway view of a barretter type of bolometer element.

Thermistors have been very widely used as bolometer elements. These are fine beads of temperature-sensitive oxides suspended between two fine lead-in wires. The most sensitive thermistors are several times more sensitive than the Wollaston wire units. They can withstand much greater overloads but are more sensitive to changes of ambient temperature than the Wollaston wire elements. A thermistor has a greater heat capacity than a wire element; therefore, its thermal lag is greater. As a result, thermistors are particularly adaptable to the measurement of pulse power.

14.19. Wattmeter Bridge Circuits.—Two different methods are used to measure power with bolometers. In one method, the galvanometer reading of an unbalanced bridge is used as an indication of the amount of microwave power. The bridge is balanced when no microwave power is present, and the addition of microwave power changes the resistance of the bolometer and unbalances the bridge. An unbalanced-bridge type of wattmeter may be calibrated by adding a known amount of low-frequency power to the bolometer and recording the galvanometer reading. This type of bridge has the advantage of being direct reading, but the microwave impedance of the bolometer unit changes as the microwave power input varies the resistance of the bolometer.

In the second method of measurement, a balanced bridge is used and the microwave impedance of the bolometer is maintained at a constant value. The bridge is rebalanced by reducing the direct current in the bridge after the microwave power has been added. The amount of microwave power is then said to equal the amount of d-c power that it was necessary to subtract from the bolometer in order to regain the balanced condition.

A variety of bridge circuits have been developed for use with these bolometer elements. In practical forms of the balanced and unbalanced bridges mentioned above, a great deal of attention has been paid to temperature compensation to minimize the effects of ambient temperature changes. More recent circuit developments include self-balancing bridges. The bridge circuit is incorporated in the feedback circuit of an oscillator, and the strength of oscillation automatically adjusts itself to maintain the bridge in a condition near balance. The voltage applied by the oscillator to the bridge may be calibrated to indicate the microwave power absorbed in the bolometer element.

14.20. Extending the Range of a Wattmeter.—If the power to be measured is greater than the capability of the bolometer element, it is possible to extend the range of the wattmeter by using a calibrated attenuator between the power source and the wattmeter. Attenuators may be divided into three main types: resistive attenuators, reactive attenuators, and power-monitoring devices. Resistive attenuators reduce the power level by absorbing the power in a lossy material. A length of lossy transmission line is an example of this type of attenuator. Reactive attenuators depend upon loose coupling to an electric or magnetic field to reduce the power level. A wave guide that is too small to allow propagation of the frequency under consideration is frequently used as a reactive attenuator. Power monitors which sample a known amount of the total power are the third type of attenuator. These three types may be further subdivided into fixed attenuators and variable attenuators.

14.21. Fixed Attenuators.—It is possible to obtain several decibels of attenuation in a small space by filling a coaxial line or wave guide with a lossy dielectric. A length of lossy cable is often used as a simple, although somewhat crude, attenuator. An alternative construction uses a resistive material for the center conductor of a coaxial line; for example, a metallized glass rod may be used.

Wave-guide fixed attenuators are often made by inserting a resistive film into the wave guide. This resistance may be either a metal film on glass, or a carbon film on bakelite. The lossy material is usually a tapered section, as shown in Fig. 14–12, in order to prevent reflections. A groove has been cut in the walls of the wave guide to hold the attenuator card in place. This design also allows the lossy material to be removed and reinserted easily.

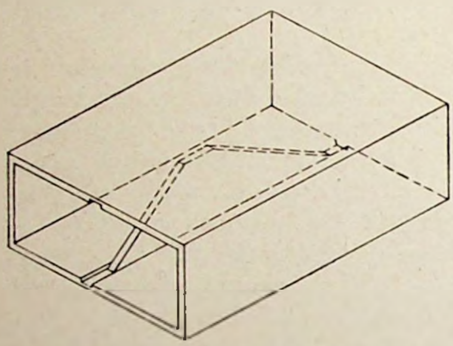


FIG. 14–12.—Fixed attenuator in a wave guide.

Reflections should always be avoided in attenuators because the calibration of the attenuator is not reliable if power is reflected when the attenuator is inserted in the transmission line. The use of tapered sections illustrated in Fig. 14–12 is an excellent method of avoiding reflections. An undercut center conductor may be used in a coaxial attenuator employing lossy dielectric in order to maintain the same characteristic impedance as the line with air dielectric, or a matching section may be used in the line to cancel the reflections from the attenuating section.

14.22. Variable Attenuators.—Variable attenuators are as useful at microwave frequencies as at ordinary radio and audio frequencies. They are built in all degrees of complexity and accuracy, but nearly all types operate on one of two basic principles.

Resistive attenuators depend for their attenuation upon resistive films, such as are used in the fixed attenuator described in the previous section. The attenuation is varied, however, by moving the lossy film or resistance strip in and out of regions of strong field in the transmission line. In a rectangular wave guide, the strip may be inserted parallel to the electric field through a slot in the broad face of the guide, and the attenuation will vary with changing penetration into the guide. Or the strip may be completely enclosed within the guide, and the attenuation increased by moving the strip toward the region of maximum field strength in the center of the guide.

A simple, uncalibrated attenuator of the first type is shown in Fig. 14–13. It is as useful at microwave frequencies as a simple potentiometer or volume control at lower frequencies, and it is often used to isolate an oscillator from the effects of a changing load impedance. If sufficient care is put into the design and construction, an attenuator of this type may be calibrated and used as a measuring instrument to reduce power levels by a known amount. These attenuators have found wide use in signal generators. One of the principal advantages is that the strip may be completely removed from the guide or moved to a region of negligible field strength within the guide. The insertion loss is therefore extremely low.

The greatest precision in a microwave attenuator is attained by using the reactive attenuation in a wave guide below cutoff, *i.e.*, a wave guide that is too small to allow transmission of energy. The rate of attenuation per unit length of a wave guide below cutoff may be accurately calculated from theoretical considerations, if troubles from leakage and the simultaneous presence of more than one mode of propagation are overcome.

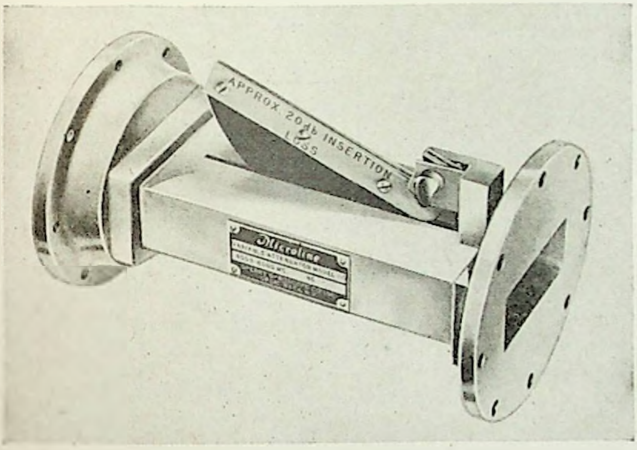


FIG. 14–13.—Variable "flap" attenuator in a wave guide.

When these troubles are eliminated, the attenuation expressed in decibels is a linear function of the length of the wave guide below cutoff. If the operating frequency is very much lower than the cutoff frequency of the wave guide, the rate of attenuation will be essentially independent of the frequency. Further discussion of the relations giving the attenuation of wave guides below cutoff may be found in the references listed below.[1,2]

A schematic illustration of a wave guide below cutoff type of attenuator using the dominant transverse electric field mode of propagation is shown in Fig. 14–14. Shunt resistances are shown across the input and output of the attenuator. Some loss must be introduced if the input and output impedances are to match the impedance of the connecting lines, since

[1] T. Moreno, *Microwave Transmission Design Data*, Sperry Gyroscope Co., Inc., Great Neck, N. Y., 1944.

[2] S. Ramo and J R. Whinnery, *Fields and Waves in Modern Radio*, John Wiley & Sons, Inc., New York, 1944.

the impedances of the input and output couplings are essentially pure reactances. Although a fixed attenuator or lossy cable may be used to match the input and output impedances, the insertion loss will be less if simple disks of resistance film are used. The calibration of the attenuator will be nonlinear if the spacing between the input and the output coupling loops becomes too small. This nonlinear region is usually avoided by limiting the minimum spacing adjustment. As a result of this restriction on the minimum spacing and the loss required to match the impedance to the line, the insertion loss of an attenuator using a wave guide below cutoff is usually in excess of 20 db.

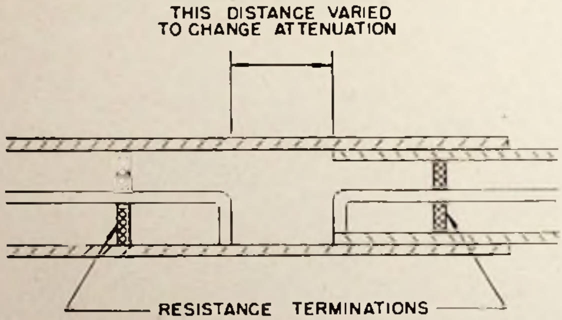


FIG. 14-14.—Variable attenuator utilizing a wave guide below cutoff.

14.23. Power Monitoring Devices.—Loss-type attenuators cannot be used when it is necessary to measure the power flowing into a useful load; they are also impractical when the power level is so high that dissipating the heat produced in a lossy attenuator becomes a serious problem. In these cases a power monitor that measures a known fraction of the total power is invaluable.

The simplest monitoring devices are small loops or probes which are coupled to the field in a transmission line. The probe may be calibrated as an attenuator, but such a crude monitoring device is seldom used because its indication can fluctuate widely if the standing wave ratio or the phase of the reflections in the transmission line varies. Transmission-line "tees" and similar devices are unsuitable for accurate power measurements for the same reason.

14.24. Directional Couplers.—Directional couplers are widely used as power-monitoring devices and have enough other applications to warrant special consideration. In a directional coupler, an auxiliary transmission line is coupled to a main transmission line in a manner such that a single traveling wave on the main transmission line, for example, the incident wave, will induce a traveling wave in one direction on the auxiliary line. The reflected wave, traveling in the opposite direction on the main line, will induce a traveling wave on the auxiliary line in a direction opposite to the wave induced by the incident wave on the main line.

The simplest of the many forms that a directional coupler may take is shown in Fig. 14–15. Wave guides are chosen for illustration, but the principles may be applied equally well to coaxial lines. The auxiliary guide runs parallel to the main guide, and two holes that are spaced one-quarter of a wavelength apart are cut through the wall common to the two sections of wave guide. Since the wavelength in the guide is greater

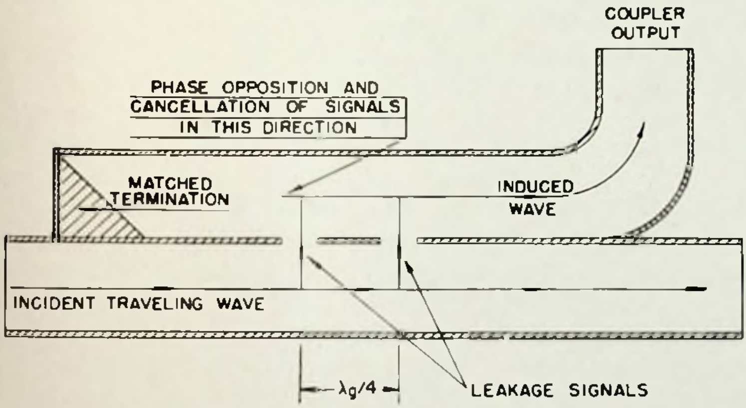


FIG. 14-15—Wave-guide directional coupler.

than the wavelength in free space, the symbol $\lambda_g/4$ is used to denote this dimension. Leakage energy from the incident wave in the main guide will pass through each hole and would travel in both directions in the auxiliary guide if only one hole were present. The coupling through the second hole will be retarded 90 deg with respect to the coupling through the first hole because of the spacing between them. The component of radiation through the second hole which is traveling toward the first hole is delayed an additional 90 deg in the auxiliary guide, and therefore cancels the component of radiation from the first hole in the same direction. In the forward direction in the auxiliary guide, the radiation from the two holes will be in phase and hence will add to give a traveling wave in that direction in the auxiliary guide.

If there is a reflected wave traveling the main guide in a direction opposite to the incident wave, there will be an induced wave in the auxiliary guide which will travel toward and be absorbed by the termination. If no power is reflected by the termination and the holes between the two guides are properly spaced, there will be no energy from a reflected wave in the main guide which will appear at the output of the directional coupler.

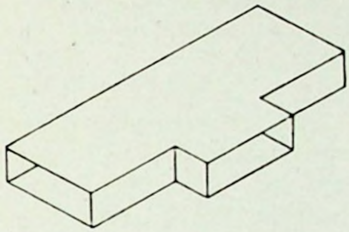
The relative strength of the induced wave in the auxiliary guide and the incident wave in the main guide will depend upon the amount of coupling through the holes, and this in turn depends upon the size of the holes. It is customary in a directional coupler to have the induced wave 20 db or more weaker than the wave in the main line; the relative strength of these waves is known as the "attenuation" of the coupler. The other coupler characteristic that is of importance is its "directivity." This is the ratio of the field strength of the induced wave in the preferred direction to the field strength of the induced wave in the other direction, for a single traveling wave in the main line. This is also expressed in decibels. In an ideal coupler, the directivity is infinite; in practical couplers, the directivity is usually between 15 and 40 db.

It is frequent practice to terminate one end of the auxiliary guide in its characteristic impedance, as shown in Fig. 14–15, while a power-measuring device is attached to the other end. The power that will be measured will be proportional to the strength of an incident traveling wave in the main guide, because the signal coupled into the auxiliary guide by the reflected wave will be absorbed by the termination in the other end of the auxiliary guide. The coupler may therefore be used as a fixed attenuator, if the main guide is also terminated in its characteristic impedance, and the amount of attenuation will depend only upon the size of the holes between the two guides.

If the power flowing down the main guide is dissipated in some useful load, such as an antenna, the power measured at the coupler output will monitor the power delivered to the useful load. There will be an error equal to the amount of power reflected from the load impedance, but this is usually small in a well-designed transmission system. In any event, the error will be much less than would be expected if the monitor were coupled to the main guide at a single point. With the latter technique the monitor indication will be proportional to the field strength at the point of coupling, and therefore very sensitive to the presence of standing waves in the guide.

14.25. Other Applications of Directional Couplers.—In most radar systems and in certain types of communication systems, it is customary to connect both transmitter and receiver through a common transmission line to a single antenna. If a directional coupler is fastened to this line, it may be used to monitor the transmitter power, as mentioned in the previous paragraph. Also, a signal generator may be connected to the coupler output in place of the wattmeter, and measurements made of receiver sensitivity. The signal fed from the coupler into the main line will be directed toward the receiver, and none will be radiated from the antenna. It is not necessary to break the transmission line connecting transmitter and receiver to the antenna for either of these measurements.

A coupler may also be attached to a transmission line in the reverse direction, and its output will then be an indication of the reflected wave in the transmission line. The termination shown at one end of the auxiliary guide in Fig. 14–15 may be replaced by a detector to measure the reflected wave. A similar detector at the output of the directional coupler will measure the relative magnitude of the incident wave, and the device can be used to measure the standing wave ratio automatically. It is usually preferable to use two separate directional couplers for this purpose, since any reflection from one of the detectors would be absorbed in the termination and not affect the reading of the other detector in a separate directional coupler. No information regarding the phase of the reflections can be obtained with this method.



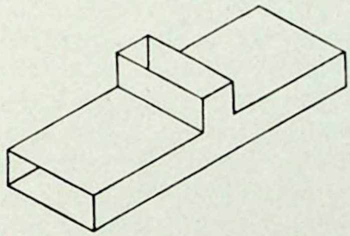


FIG. 14–16.—Wave-guide junctions.

14.26. Wave-guide Junctions.—A directional coupler is a special type of wave-guide junction which introduces attenuation and directivity. Two other common types of wave-guide junctions are illustrated in Fig. 14–16. Both of these junctions are "tee" connections. Figure 14–16*A* shows a tee that allows the input to one wave guide to divide between the other two wave guides. This type of junction is called a "shunt" tee and is the wave-guide counterpart of a tee in a coaxial line. The junction in Fig. 14–16*B* is a series tee; *i.e.*, the impedance presented at the junction by the branch arm acts in series with the main guide.

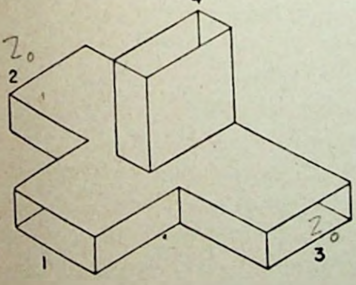


FIG. 14–17.—Wave-guide hybrid junction.

Another very useful device is illustrated in Fig. 14–17. This unit is one of the wave-guide forms of a hybrid junction and consists of a series and a shunt junction at the same point. This structure possesses the following characteristics. If arms 2 and 3 are terminated with their characteristic impedance, a signal applied to arm 1 will divide equally between 2 and 3 and no signal will appear at arm 4. If arm 2 is terminated by its characteristic impedance but a reflection occurs at the termination of arm 3, the signal at a m 4 will be proportional to the magnitude of the reflection coefficient in arm 3.

Hybrid junctions are used in antenna systems for duplex transmission and reception with a single antenna, as automatic standing wave detectors, in microwave frequency discriminators, and for many other applications. Other forms of hybrid junctions are possible, including combinations of wave guides and coaxial lines.

FREQUENCY AND WAVELENGTH MEASUREMENTS

14.27. Low-frequency Techniques That May Be Extended to Microwaves.—Well-known techniques of generating signals of precisely known frequency are readily extended to microwaves. The output of a crystal-controlled oscillator may be passed through a chain of frequency multipliers to reach as high a frequency as desired. The frequency at the output of the multiplier chain will be known to the same percentage accuracy as the frequency of the master oscillator, neglecting any phase modulation that may be introduced by the multiplier chain. It is difficult to use triodes at frequencies much higher than 1,000 megacycles, but other techniques may be used for frequency multiplication in this region.

Frequency-multiplier klystrons are very useful in the final stages of a multiplier chain that extends to the microwaves. Because of the rich harmonic content of the bunched-beam current in a klystron, appreciable output may be obtained if the output resonator of a klystron is tuned to a frequency that is the tenth or even the twentieth harmonic of the frequency to which the input resonator is tuned. For example, if the input resonator is tuned to 300 megacycles, the output resonator of the klystron may be tuned to the ninth, tenth, eleventh, etc., harmonics of this frequency. Appreciable signal output (several milliwatts or more) will be obtainable from the output resonator at the corresponding microwave frequencies of 2,700, 3,000, and 3,300 megacycles, which are spaced apart 300 megacycles.

The utility of the klystron multiplier may be further increased if the klystron is phase-modulated at a lower frequency by a signal in series with the beam voltage. For example, if the klystron is phase-modulated with a 30-megacycle signal obtained from one of the earlier stages in the multiplier chain, the final resonator may be tuned to give output at the additional side-band frequencies. These are spaced every 30 megacycles, rather than 300 megacycles. This greatly increases the usefulness and flexibility of the frequency standard.

Klystron frequency multipliers have the advantage of developing signals of appreciable power at microwave frequencies, but possess the disadvantage that the output resonator must be tuned each time a new frequency is to be obtained. For many purposes it is preferable to use as a final multiplier stage an untuned multiplier, which develops much weaker signals but a large number of harmonics simultaneously. Crystal rectifiers are ideal for this purpose because of their nonlinear characteristics, and with a sensitive receiver it is possible to detect harmonics as high as the fiftieth. Spectrum analyzers, discussed in Sec. 14.32, have sufficient sensitivity for this purpose and are widely used to compare precisely known frequencies generated in this manner with frequencies that are to be measured. In this fashion it is possible to measure to a very high degree of accuracy the frequency of an unknown signal.

14.28. Microwave Wavemeters.—Where less accurate measurements are to be made of the signal frequency, precision wavemeters and search wavemeters are used. But to be useful at microwave frequencies, wavemeters must have much greater relative accuracy than the corresponding lower frequency instruments. Lower frequency search wavemeters that use a calibrated resonant circuit for reference are seldom accurate to better than 1 per cent, and an instrument with a 2 or 3 per cent error is still very satisfactory. However, a microwave search wavemeter must usually measure wavelength or frequency with an accuracy better than 1 per cent, since 1 per cent of 3,000 megacycles is equal to an error of 30 megacycles. Precision wavemeters must be accurate within 0.01 per cent or even better.

To possess this degree of accuracy and to be able to separate two signals that are close together, it is desirable to have a resonant circuit with high Q. Fortunately, the cavity resonators that are most suitable for microwaves are almost universally characterized by very high Q, and these cavities lend themselves to the construction of wavemeters that have the desired characteristics.

14.29. Coaxial Line Wavemeters.—A resonant coaxial line is a simple resonator which satisfies the requirements for a search-type wavemeter. A shorted coaxial line is used in the same manner that Lecher wires are used at lower frequencies. The self-shielding coaxial construction is required to eliminate radiation losses at the microwave frequencies. A section of coaxial line that is shorted at both ends will be resonant when its length is an integral number of half wavelengths. In an air-dielectric, low-loss coaxial line, the wavelength is very nearly equal to the free-space wavelength at the same frequency.

Calibration of a shorted line wavemeter is not required because the wavelength can be measured directly from the dimensions of the resonant circuit. Any perturbations that are introduced by the loop or loops that couple to the cavity may be compensated for by finding the change in length between two resonances instead of the length required for resonance. This will also overcome any ambiguity that may arise from not knowing the number of half wavelengths within the coaxial cavity at resonance. A closed resonator of this type usually has an unloaded Q of a few thousand, which is sufficiently high for many purposes.

The physical dimensions of the wavemeter may be reduced by using a quarter-wave coaxial cavity that is shorted at one end and open at the other. A wavemeter of this type is illustrated in Fig. 14–18. Resonance will not occur when the length of the center conductor is exactly a quarter wavelength because of the effective capacity that results from the fringing fields at the open end of the center conductor. For this reason calibration is required, but the relationship between the length of the center conductor and the resonant wavelength is very nearly linear.

In Fig. 14–18 it will be noted that there is no metal-to-metal contact between the movable center conductor and the end of the cavity. Spring fingers may be used here, if desired, but if they are not carefully built the contact may be erratic, and the cavity Q lowered by losses in the sliding

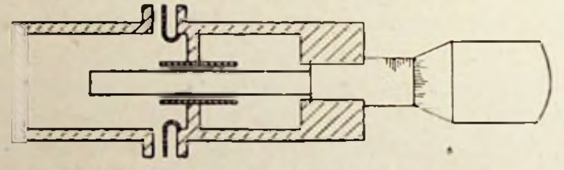


FIG. 14–18.—Coaxial line wavemeter.

contacts. A wave trap is shown instead, which offers an equivalent electrical short circuit at its input, without any physical contact being made. The micrometer screw that is shown provides a precision mechanical drive for the plunger that forms the center conductor, and gives an indication of the setting of the plunger which may be calibrated against wavelength and frequency.

14.30. Precision Wavemeters.—The Q that may be obtained from coaxial lines is not sufficiently high for precision wavemeters, and circular cylindrical cavities are used when greater precision is required. Resonance will be found when the length of the cylinder is an integral number of half wavelengths for some mode of transmission in a circular wave guide.

The chief problem that is encountered in the design of these wavemeters is the possibility of multiple resonances for a single input frequency, resulting from the multiple modes of transmission in a circular wave guide, with a resulting ambiguity in the indication. This may be avoided if the mode that is chosen to indicate resonance is the dominant, or $TE_{1,1}$ mode in circular wave guide, shown in Fig. 14–19. This mode has the longest cutoff wavelength, and cavities may be designed which will be resonant in this mode over a band of frequencies but which are too small to be resonant in any other mode. The Q that is obtainable with this mode is appreciably higher than with coaxial lines, and a working or loaded Q in the order of 10,000 in the frequency range from 3,000 to 10,000 megacycles can be obtained. The coupling circuit must be carefully designed;

otherwise it will introduce an equivalent ellipticity that leads to mode-splitting. This will cause two closely spaced resonances, rather than a single resonant peak.

Resonant cavities that operate with the $TE_{0,1}$ mode, which is the circular electric mode in a wave guide shown in Fig. 14–19, have an exceptionally high Q, and in many ways are ideal for high-precision wavemeters. An unloaded Q above 50,000 is possible at a frequency of 3,000 megacycles. Ellipticity will cause no mode splitting. No currents flow from the cavity end plates to the cylinder walls, and a good contact or wave trap is not required.

A cavity that is large enough to be resonant in the $TE_{0,1}$ mode may also be resonant in a number of other wave-guide modes. Some of these can

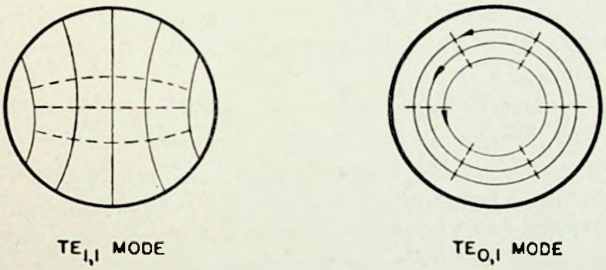


FIG. 14–19.—Field patterns in cylindrical cavity resonators.

be avoided by choosing the proper means of coupling to the resonator. To avoid multiple indications completely, it is necessary to use some sort of mode damper which will suppress all resonances except those in the desired mode. Various schemes for damping out the undesired modes have been found successful.

To utilize fully the accuracy possible with these precision cavity wavemeters, they must be constructed with mechanical precision. Appreciable errors may also be introduced by dimensional changes resulting from changes in temperature. These errors may be reduced by the proper choice of materials. Changes of humidity can also introduce errors greater than one part in ten thousand. The most accurate wavemeters must therefore be either hermetically sealed or equipped with dehydrators.

Sealed, temperature-compensated cavities are often used as fixed secondary standards. Although their accuracy is far less than may be obtained by the use of crystal-controlled sources, they can be used as calibration standards for less accurate wavemeters.

14.31. Methods of Using Cavity Wavemeters.—Resonant cavities are used in two ways to indicate wavelength. With transmission-type wavemeters, the signal to be measured is coupled through the cavity to a detector. The detector output is zero except when the cavity is in resonance. Reaction-type wavemeters are coupled to a transmission line that connects to a detector. The detector output is a maximum when the cavity is off resonance. When the cavity is tuned to resonance, part of the energy is absorbed in the cavity, and there is a corresponding dip in the detector output. This characteristic indication with a reaction-type wavemeter is illustrated in Figs. 12–3 and 12–8.

Transmission-type wavemeters are preferred for most accurate measurements, because it is possible to couple both input and output loosely to the cavity and thereby minimize pulling effects. Reaction-type wavemeters must be coupled more heavily to the transmission line to give an appreciable dip in the detector output, but the detector gives a continuous indication of oscillation. The wavemeter should be isolated from the oscillator by an attenuator or by loose coupling. Otherwise the wavemeter may react upon the oscillator while being tuned and change the oscillator frequency.

Cavities are coupled to transmission lines by several methods. The transmission line, if coaxial, may terminate in a loop that extends into the cavity and links some of the magnetic field. The center conductor of the coaxial line may also terminate in a probe that couples to the electric field in the cavity. When the transmission line is a wave guide, it is customary to open a hole in the wall between the wave guide and the cavity. There will be coupling through the hole if the magnetic field at the surface of the wave guide has a component parallel to the magnetic field at the surface within the cavity.

14.32. Spectrum Analyzers.—Microwave spectrum analyzers are used in connection with frequency measurement but have a wide variety of other applications. A spectrum analyzer is similar to a panoramic receiver. The input signal is fed to a crystal mixer where it beats with a local oscillator signal. The difference frequency is fed through a narrow-band intermediate-frequency amplifier to the vertical deflection plates of an oscilloscope. A block diagram is shown in Fig. 14–20. The frequency of the local oscillator of the receiver is swept by a low-frequency saw-tooth or sine-wave voltage, which is also applied to the horizontal deflection plates of the oscilloscope. The pattern on the oscilloscope will then appear as a horizontal line, with a vertical deflection or "pip" marking the point at which the frequencies of the local oscillator and input signal differ by an amount equal to the frequency of the intermediate-frequency amplifier. The width of the pip will depend upon the frequency range over which the local oscillator is swept and upon the band width of the intermediate-frequency amplifier, while the height of the pip is determined by the strength of the input signal. If there are two signals of different frequency being received, two pips will appear upon the screen of the oscilloscope.

The spacing between these two pips is determined by the frequency difference of the two signals.

The input signal is usually fed to the mixer through a variable microwave attenuator, which adjusts the signal strength to a convenient level. Frequently multiple input channels are provided, each equipped with a variable attenuator, in order that the relative strength of two unequal signals may be compared.

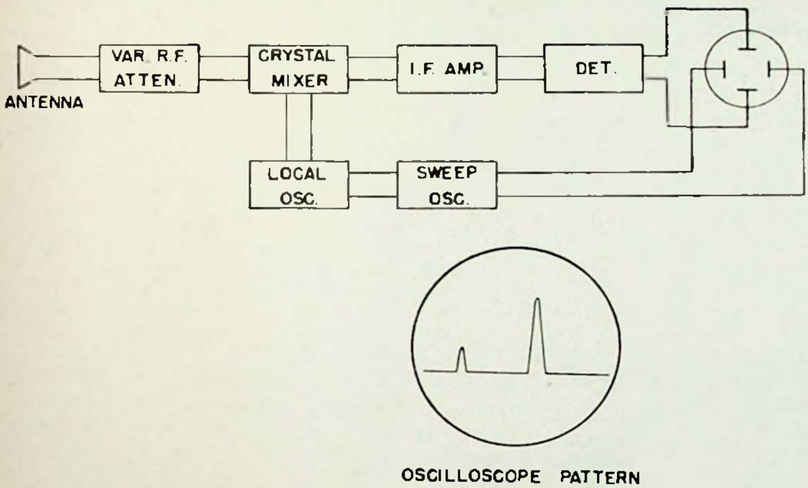


FIG. 14–20.—Block diagram of a spectrum analyzer and typical oscilloscope pattern indicating two signals with different frequencies.

The local oscillator of a microwave spectrum analyzer is most commonly a reflex klystron, whose frequency is easily swept over a range of a number of megacycles by a sweep voltage applied to the reflector electrode.

A spectrum analyzer may be used to compare the frequencies of two signals or to measure their relative strength. It may be used in conjunction with a traveling probe to measure very high standing wave ratios. While the probe is moved, the variable attenuator is adjusted to maintain constant input to the mixer. Then the standing wave ratio can be calculated from the attenuator readings.

The analyzer is very useful in investigating the characteristics of pulsed oscillators, such as magnetrons. The frequency spectrum of the oscillator output is often an indication of whether the oscillator is functioning properly. The analyzer may be used to measure the Q of a cavity, or to calibrate wavemeters in conjunction with a crystal-controlled source or standard cavities. There are many additional applications.

14.33. Signal Generators and Test Sets.—Signal generators and test sets are as useful at microwave frequencies as at lower frequencies, and the same fundamental techniques are used in their design. Signal generators that have been designed for the purpose of testing radar sets often incorporate, in addition, equipment with which the transmitter power of the radar set can be measured.

The block diagram of a typical microwave signal generator of this type is shown in Fig. 14–21. The oscillator is a source of continuous wave power in the order of milliwatts; lighthouse-type triodes or local oscillator-type reflex klystrons are most commonly used. The oscillator output is fed through a wave-guide variable attenuator which adjusts the initial

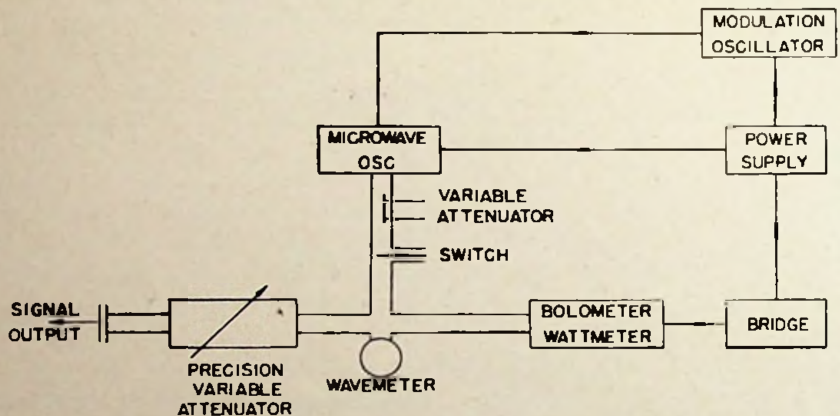


FIG. 14–21.—Block diagram of a microwave signal generator.

power level, and past a wave-guide switch to a tee connection. The power divides here, half flowing to a bolometer-type wattmeter which is used to monitor the power level. The other half passes through a precision variable attenuator to the output connection. A cavity-type wavemeter is coupled to the wave guide at the tee connection. This is a reaction wavemeter which indicates signal frequency by causing a dip in the monitor reading when tuned to resonance.

The oscillator may be modulated in any manner desired, and for radar testing it is customary to include in the signal generator provisions for pulsing or frequency-modulating the oscillator. Provisions for external modulation are also included.

It is possible to measure transmitter power with this test set by closing the wave-guide switch and connecting the output fitting of the test set to a directional coupler in series with the transmitter power. The power to be measured is fed back through the precision attenuator into the power monitor. The power is measured by finding what setting of the attenuator is required to give a standard indication on the power monitor.

SOME TYPICAL MICROWAVE EXPERIMENTS

14.34. Measuring the Rieke Diagram of an Oscillator.—The experiments that are outlined in this and the following section are not presented as the only method, or even the preferred method, by which the desired results may be found. They are merely typical experiments involving applications of the various measurement equipments that are involved.

To plot the Rieke diagram of an oscillator, such as a klystron, it is necessary to measure the effect of a changing load impedance upon the power output and frequency. A typical experimental arrangement that might be used is shown in Fig. 14–22. The oscillator is connected through

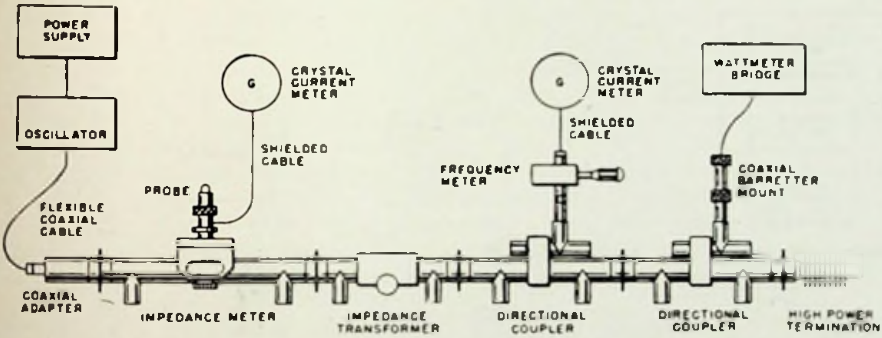


FIG. 14–22.—Block diagram of apparatus for obtaining the Rieke diagram of a klystron oscillator.

an adaptor to a rigid coaxial transmission line. The power output is fed through a standing wave detector, impedance transformer, and directional coupler in that order, and the signal is finally dissipated in a matched dummy load. The output of the directional coupler is fed past a reaction wavemeter, which is coupled to the auxiliary transmission line, and into a bolometer wattmeter.

The bolometer wattmeter may be used to measure the power output of the oscillator as the load impedance is varied, if the attenuation of the coupler is known. A wavemeter is used to determine the frequency. The wavemeter is attached to the output of the directional coupler so that any pulling effects from the wavemeter will be isolated from the oscillator by the attenuation of the coupler. The impedance transformer is changed to vary the load impedance presented to the oscillator, and the standing wave detector measures the load impedance. If the impedance transformer is calibrated, the standing wave detector may be omitted. A double-slug transformer is recommended for this application, because independent controls are provided for the amplitude and phase of the standing waves introduced into the matched line by the transformer.

14.35. Measuring the Reflection Introduced by an Antenna.—A reflex klystron may conveniently be used in this experiment as a signal source, if power is available from a well-regulated power supply. The klystron is one hundred per cent amplitude-modulated or keyed by a square wave of audio frequency applied to the reflector electrode. A square wave is used rather than a sine wave to avoid frequency modulation of the klystron. The experimental setup is illustrated in Fig. 14–23.

The klystron output is fed through a variable wave-guide attenuator and into one end of a series tee section (see Sec. 14.23 and Fig. 14–16*B*). A transmission-type wavemeter is coupled to the branch arm of the tee. The signal that continues past the branch arm passes through a standing wave detector and into the antenna that is to be tested.

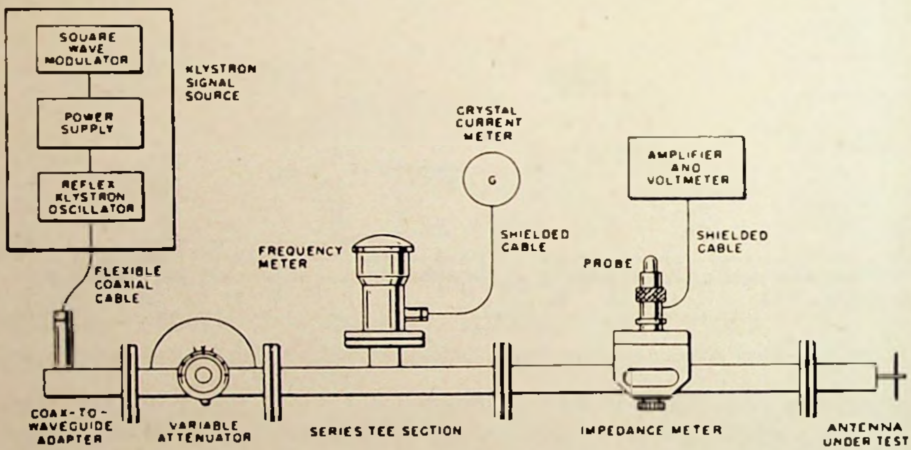


FIG. 14–23.—Block diagram of apparatus for measuring the impedance of an antenna.

The signal that is picked up by the traveling probe is fed to a detector, which may be a crystal or a bolometer element. The output of this detector is an audio signal whose fundamental frequency is the modulation frequency applied to the klystron. This audio signal is amplified and then measured with a voltmeter, and the voltmeter indication is a measure of the field strength in the wave guide at the location of the traveling probe. Measurement of the standing wave ratio is simplified when an audio amplifier and a voltmeter are used to replace a galvanometer because the response of the system is much more rapid, and the position and amplitude of the standing waves can be determined more quickly.

The reflection introduced at some other frequency is obtained by repeating the measurement with the klystron tuned to the new frequency. The wavemeter is used to measure the new oscillating frequency of the klystron. The variable attenuator reduces the oscillator power to a convenient level and also isolates the oscillator from the wavemeter and the other measuring equipment.

APPENDIX A

GLOSSARY OF TERMS AND SYMBOLS

acceleration voltage is the voltage between the cathode and the anode. This voltage determines the average velocity of the electrons in the beam. It is also referred to as the *beam voltage.*

Applegate diagram is a space-time diagram used to illustrate electron bunching.

barretter is a type of bolometer with a positive temperature coefficient.

beam-coupling coefficient is a measure of the interaction between a resonator and the electron beam when the gap transit time is not negligible.

beam current usually refers to the total cathode current. The actual current in the electron beam is smaller because the focusing of the electron beam is never perfect.

beam voltage (see *Acceleration voltage*).

bolometer is a device that changes its electrical resistance when its temperature is changed. This characteristic is used as a means of measuring radio-frequency power by observing the change in resistance when a bolometer is heated by the application of radio-frequency power.

buffer refers to an amplifier that isolates the oscillator from any changes (load, tuning, or beam-power input) in the output stage.

buncher is an abbreviation for *buncher resonator.* This term has been used for the resonator of a klystron that produces the variations of velocity in the electron beam. *Input resonator* is now a preferred term.

bunching is the process of converting a d–c electron beam into a beam with a radio-frequency current component as a result of motion along a drift space or reflection space after velocity modulation has been introduced. *Optimum bunching* is the bunching condition required for maximum output in a klystron amplifier. *Underbunching* and *overbunching* refer to the conditions existing when the input gap voltage is less or greater than the value required for optimum bunching.

bunching parameter combines a number of factors that affect the electron bunching into a single term which determines the theoretical operation of a klystron [see Eqs. (3–15) to (3–21)].

capacity loading refers to the tuning effect of a lumped capacity at some point on a transmission line, usually the open-circuited end. The *capacity section* of a transmission line or resonator is that part which may be considered as a lumped capacity. The grids of a klystron resonator are the capacity section of the resonator.

cascade amplifier is a multistage klystron amplifier which uses three or more resonators in a single tube utilizing the same electron beam.

cascade bunching refers to the superposition of velocity modulation from more than one resonator.

catcher is an abbreviation for *catcher resonator.* This term has been used for the resonator that is excited by the bunched electron beam. *Output resonator* is now a preferred term.

cavity resonator is the term for a metallic cavity that is resonant to some radio frequency or wavelength determined by the dimensions of the cavity and the electromagnetic field within the cavity.

coaxial terminals are the concentric output or input lines that are coupled to a klystron resonator (see Fig. 1–1).

coaxial wavemeter is a concentric resonant transmission line used for determining wavelength.

control grid is the electrode in the electron gun that controls the cathode emission or beam current of a klystron.

debunching is a space-charge effect which tends to destroy the electron bunching.

directional coupler is a device that samples a small fraction of the power traveling in one direction but is not sensitive to the reflected wave on a transmission-line system.

drift space is the region between two resonators in a klystron. The modulation of the electron velocities causes the electron beam to form into bunches while traversing this drift space.

electronic tuning refers to the frequency change produced by a change of voltage applied to an electrode in the tube. The term is not applied to *thermal tuning* but is reserved for an electronic effect.

hybrid junction is a special form of *wave-guide junction* which possesses the properties of the hybrid coils used in audio-communication circuits.

impedance transformer is a device for changing the input impedance of a transmission-line system by introducing reflections at the proper point in the system.

input gap is the region of the input resonator where the electric field is concentrated and where the velocity modulation is superimposed upon the electron beam (see *Resonator grids*).

input gap voltage is the radio-frequency voltage across the *input gap*. In most discussions, the term implies the peak value of this oscillating voltage.

klystron grids (see *Resonator grids*).

klystron tubes are electron beam tubes in which a periodic variation of the velocity of the electrons, without appreciable simultaneous variation of the beam current, is converted into radio-frequency energy by utilizing a transit time to produce electron bunching.

mode refers to one of several methods of exciting a resonant system. The term has also been used to describe the existence of a series of different input voltages which cause a klystron to oscillate.

optimum bunching (see *Bunching*).

output gap is the region where the bunched electron beam transfers its radio-frequency energy to the output resonator.

output gap voltage is the radio-frequency voltage across the output gap. In most cases, the term implies the peak value of the oscillating voltage.

overbunching (see *Bunching*).

pulse amplitude modulation uses pulses separated by equal intervals, and information is transmitted by controlling the amplitude of these pulses.

pulse-length modulation varies the duration of the pulses in order to transmit information.

pulse time modulation controls the time of occurrence of pulses of equal amplitude in order to transmit information.

Q is 2π times the ratio of the energy stored in a resonant system to the energy lost per cycle.

reaction-type wavemeter indicates resonance by its effect on the reflections in an oscillating system.

reentrant describes the shape of an object with one or more sections directed inward. The resonator in Fig. 2–4 is an example.

reflection bunching occurs when a velocity-modulated beam traverses a reflecting field (see *Bunching*).

reflection coefficient is the complex ratio of the field strength of the reflected wave on a transmission line to the field strength of the incident wave.

reflector refers to the electrode in a reflex klystron which reverses the direction of the electron beam and returns the electrons through the resonator grids a second time.

reflector voltage is the voltage between the reflector electrode and the cathode.

reflex klystron is a single-resonator klystron oscillator depending primarily on electron bunching in a reflecting field for its operation.

resonator (see *Cavity resonator*).

resonator grids are the part of a klystron resonator through which the electron beam passes and form a *resonator gap*. In some cases the gap may consist of an aperture without a grid structure.

resonator wavemeter may be any resonant circuit used to determine wavelength. The term may be restricted occasionally to describe a *cavity resonator wavemeter*.

Rieke diagram is a plot of constant-power output and constant-frequency contour lines on a chart representing the impedance or admittance of the load.

shape factor is a proportionality factor in the equations defining the different parameters describing cavity resonators. The shape factor is different for each parameter and is determined by the ratio of two integrals which include the effect of the boundary conditions and the field within the resonator. As an example, the statement that Q is proportional to the ratio of the volume to the surface area is an approximation of the more exact relation

$$Q = \frac{\int B^2\, dV}{\lambda \int B^2\, dS}\, 2\sqrt{2}\, \frac{\lambda}{d}$$

where B is the magnetic flux density, V and S represent the volume and surface area, λ is the wavelength, and d is the skin depth.

shunt resistance is a characteristic constant of a resonant circuit which may be compared to the effective resistance of a parallel LC circuit at resonance.

spectrum analyzer is a panoramic type of receiver.

standing wave ratio is the ratio of the magnitude of the maximum electric field strength on a transmission line to the magnitude of the minimum field strength.

starting current is the minimum beam current required to start oscillation in a klystron.

synchrodyne is used to refer to a system producing a second frequency which is always separated a fixed amount from a control frequency.

thermal tuning utilizes controlled thermal expansion to change the tuning of a resonator

thermistor is a type of bolometer with a negative temperature coefficient.

transreduction factor is the ratio of the small-signal transadmittance to the large-signal transadmittance, and determines the strength of oscillation.

underbunching (see *Bunching*).

velocity modulation refers to the modification of the velocity of the electrons in a beam. The term usually implies that the variations of electron velocity are utilized to convert d-c beam energy into radio-frequency energy.

velocity sorting produces a radio-frequency component of current by the selection or rejection of electrons according to the magnitude of their velocity, in contrast to electron bunching which groups electrons with different velocities.

voltage modes refers to the existence of regions in the beam voltage or reflector voltage which allow a klystron to oscillate, separated by regions of nonoscillation, in contrast to the continuous oscillation of low-frequency tubes as the voltage is varied.

voltage standing wave ratio (see *Standing wave ratio*).

wave guide is a type of transmission line with only one conductor and the radio-frequency energy is confined within this conducting tube.

wave-guide junction refers to any system of waveguides with more than two connections for input and output.

LIST OF SYMBOLS

a	Radius of a cylindrical cavity resonator or a coaxial line
a	Deceleration caused by retarding force F due to the reflecting field
A	Area of the resonator gap
A_n	Amplitude of the nth side band
b	Length of a cavity resonator
B_L	Shunt susceptance of the load
B_s	Shunt susceptance of a cavity resonator
C	Equivalent capacitance of the resonator gap
d	Spacing of the resonator gap
e	Charge of an electron
E	Instantaneous value of the radio-frequency voltage across the resonator gap
E_g	Control grid voltage
E_m	Peak value of the modulation voltage
E_r	Beam voltage or acceleration voltage
E_0	Reflector voltage (voltage between the cathode and the reflector electrode)
E_1	Peak value of the radio-frequency voltage across the input resonator gap
E_1	Electric field strength of the incident wave on a transmission line
E_2	Peak value of the radio-frequency voltage across the second resonator gap
E_2	Electric field strength of the reflected wave on a transmission line
E_3	Peak value of the radio-frequency voltage across the third gap
E_{max}	Magnitude of the maximum electric field strength on a transmission line
E_{min}	Magnitude of the minimum electric field strength on a transmission line
eff.	Efficiency (ratio of radio-frequency output power to beam power input)
f	Frequency of oscillation
f_0	Resonant frequency of a cavity
f_1	Input frequency
f_2	Output frequency
f_m	Modulation frequency
F	Retarding force due to a reflecting field
$\frac{\Delta f}{f_m}$	Modulation index
g_m	Transconductance
g_{ms}	Small-signal transconductance
G_L	Load conductance
G_s	Shunt conductance of a cavity resonator
i_p	Plate current of a triode
i_2	Fundamental component of the radio-frequency current in the bunched electron beam
I	Current
I_0	Average beam current (value of direct current)
I_{start}	Minimum value of beam current required to maintain oscillation
I_2	Instantaneous value of the bunched-beam current
J_n	Bessel function of the first kind and nth order
J_n'	Derivative of J_n
k	Root of the Bessel function J_n
k'	Root of the derivative of J_n
l	Length
L	Equivalent inductance of a cavity resonator

m	Mass of an electron
M	Multiplication ratio of a frequency multiplier
n	Represents an integer in many different equations
N	Transit time in the drift space, expressed as the number of oscillation cycles
N'	Effective bunching time, which may differ from the actual transit time
N_{ff}	Transit time, expressed as the number of oscillation cycles, through a field-free drift space
N_r	Transit time, expressed as the number of oscillation cycles, in a reflecting field
N_t	Total transit time through a combination of a field-free drift space and a reflecting field
N_1	Transit time in the drift space, expressed as the number of oscillation cycles at the input frequency f_1
N_2	Transit time in the drift space, expressed as the number of oscillation cycles at the output frequency f_2
p	Number of nodes in the electric field along the length of a cavity resonator
p'	Number of nodes in the magnetic field along the length of a cavity resonator
P_L	Power delivered to the load
P_2	Power delivered to the resonator and load
Q	Unloaded Q of a cavity resonator
Q_L	Loaded Q of a cavity resonator
Q_1	Unloaded Q of the first resonator
Q_2	Loaded Q of the second resonator
$\frac{2Q_L\Delta f}{f}$	Frequency deviation from the resonant frequency of the cavity
$2Q_L\left(\frac{2\Delta F}{f}\right)_0$	Band width between zero-power output points
$2Q_L\left(\frac{2\Delta f}{f}\right)_{½}$	Band width between half-power points (frequently called "electronic tuning band width")
r	Radius
R_L	Equivalent load resistance
R_s	Shunt resistance of a cavity resonator
R_{sL}	Loaded shunt resistance of a cavity resonator
s	Distance measured from a resonator gap
s_0	Distance between resonators or between the resonator and a reflector electrode
SWR	Standing wave ratio
t	Time
t_1	Departure time when an electron leaves the first resonator gap
t_2	Arrival time when an electron reaches the second resonator gap
t_3	Arrival time when an electron reaches the third gap
T	Transit time in the drift space or reflecting field
T_0	Average transit time of an electron with average velocity v_0
T_{12}	Transit time between first and second resonators
T_{23}	Transit time between second and third resonators
TE	Transverse electric field mode
TM	Transverse magnetic field mode
v	Velocity of an electron
v_0	Average velocity of an electron (corresponds to E_0)
VSWR	Voltage standing wave ratio
W	Energy transferred from the electron beam to the cavity resonator

x	Bunching parameter equal to $\beta\pi N \frac{E_1}{E_0}$ $= \beta \pi \alpha \xi$
$\frac{x}{2J_1(x)}$	Transreduction factor (magnitude of the ratio of the small-signal trans-admittance to the large-signal transadmittance)
$xJ_1(x)$	Universal efficiency parameter for reflex klystrons
X	Reactance
X_L	Reactive component of the load impedance
X_s	Reactive component of the shunt impedance of a detuned resonator
y_{m1}	Small-signal transadmittance
Y_s	Total admittance of the cavity resonator
Y_2	Equivalent admittance due to the bunched beam
Z_0	Characteristic impedance of a transmission line
β	Beam coupling coefficient equal to $\frac{\sin \delta/2}{\delta/2}$
δ	Transit angle across the resonator gap
λ	Wavelength in centimeters
ϕ	Phase angle between the bunched-beam current and the input gap voltage
ϕ_L	Phase angle introduced by the length of the feedback line
ϕ_2	Phase angle between the bunched-beam current and the output gap voltage
ρ	Reflection coefficient
τ	Transit angle in radians corresponding to the drift transit time T
τ_0	Average value of the transit angle
τ_2	Phase angle at the second gap referred to the phase at the input gap
ω	$2\pi f$
ω_0	2π times the resonant frequency of the cavity
ω_1	2π times the input frequency of an amplifier or frequency multiplier

APPENDIX B

KLYSTRON DESIGN CHARTS

Two types of charts are included in this Appendix. The first nine charts give the necessary data for calculating the resonant wavelength of a reentrant cavity resonator from its physical dimensions; the last four are graphs of various Bessel function relationships which are important in electron bunching theory. These charts have been plotted on cross-section paper so that values may be read accurately from the curves.

The resonator charts are plotted in terms of dimensionless ratios so that any desired range of wavelength may be used. The ratio of the length of the resonator Z_0 to the radius of the inner conductor ρ_1 is the horizontal coordinate. The ratio of the gap spacing δ to the radius of the inner conductor is the vertical coordinate. Families of curves for two different ratios of the outer radius ρ_2 to the radius of the inner conductor ρ_1 are given on each chart. Curves for the highest ratio of ρ_2/ρ_1 are repeated on the next chart in order to facilitate interpolation.

Each line on the chart corresponds to a constant value of $k\rho_1$, *i.e.*, a constant value of λ, the resonant wavelength. The straight line in the upper left-hand corner of the chart corresponds to the limiting case when the gap spacing is equal to the length of the resonator and gives the resonant wavelength of the mode that is independent of length and is determined only by the radius of the cylindrical cavity.

A typical example of the use of these resonance charts may be obtained by assuming some practical values for the dimensions and calculating the resonant wavelength. Consider a resonator with the following dimensions:

$$\begin{aligned}
\rho_1 &= 0.5 \text{ cm} \\
\rho_2 &= 1.5 \text{ cm} \\
Z_0 &= 1.0 \text{ cm} \\
\delta &= 0.1 \text{ cm} \\
\rho_2/\rho_1 &= 3.0 \\
\delta/\rho_1 &= 0.2 \\
Z_0/\rho_1 &= 2.0
\end{aligned}$$

The solid curves on Chart VII correspond to ρ_2/ρ_1 equal to 3.0, and the intersection of the vertical line for Z_0/ρ_1 equal to 2.0 and δ/ρ_1 equal to 0.2 occurs between the curves for $k\rho_1$ equal to 0.3 and 0.4. Interpolation between the two curves gives a value of $k\rho_1$ equal to 0.325, which corresponds to a resonant wavelength of 9.6 cm.

Charts X to XIII are similar to Figs. 3–7, 7–5, 7–7, and 8–6 in the text. Reference to the sections in the text that discuss these figures will explain the use of these charts.

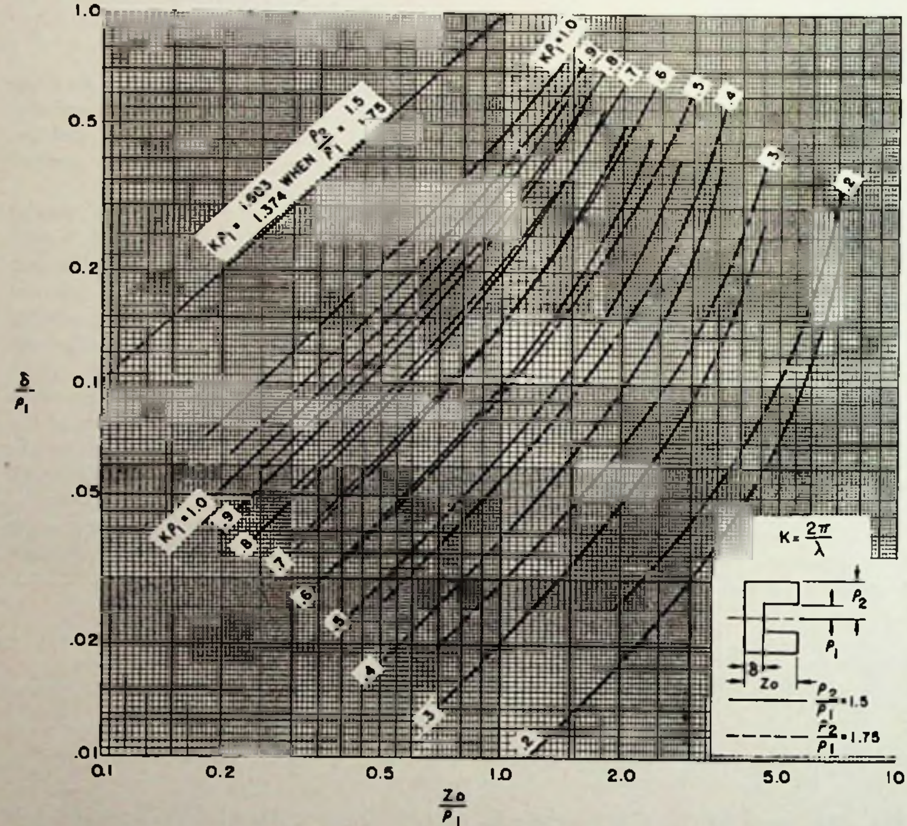


CHART I.—Resonator design curves for radius ratios of 1.50 and 1.75.

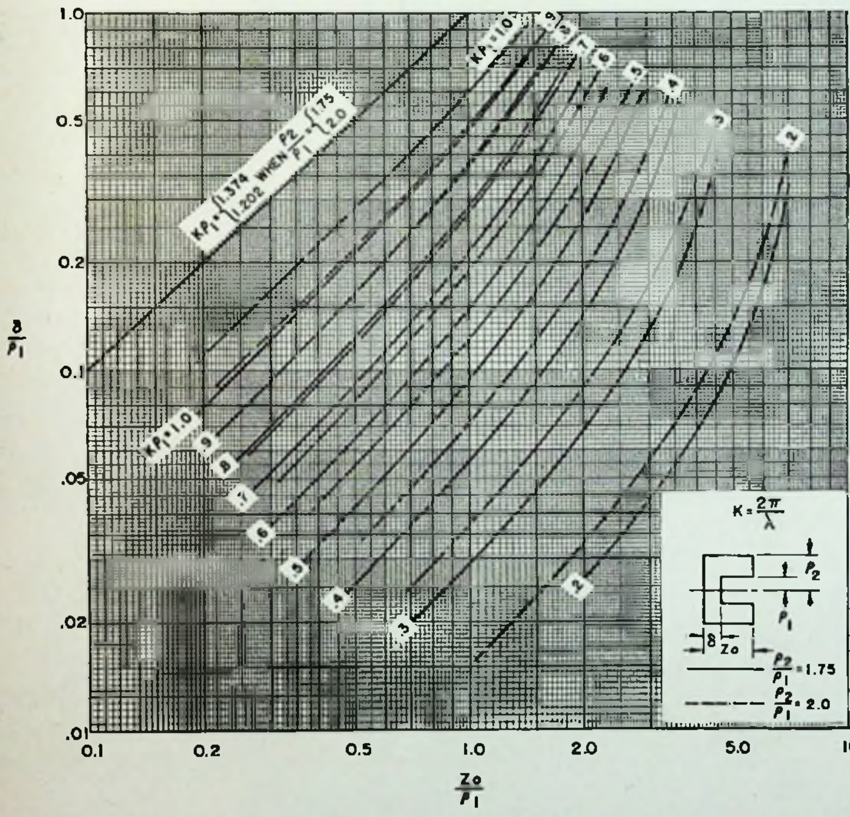


CHART II.—Resonator design curves for radius ratios of 1.75 and 2.00.

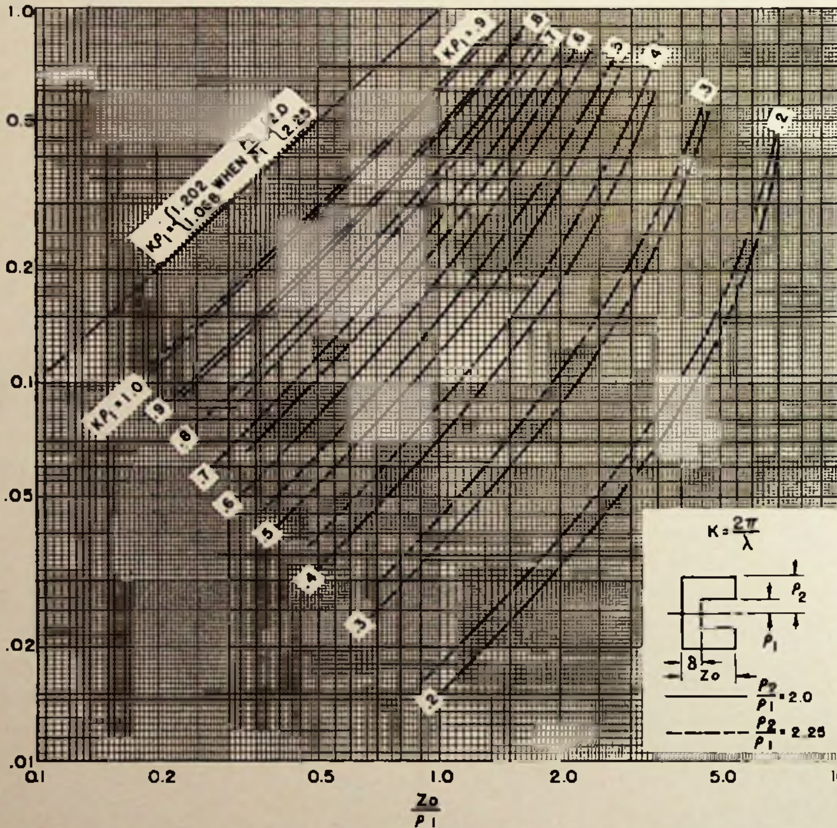


CHART III.—Resonator design curves for radius ratios of 2.00 and 2.25.

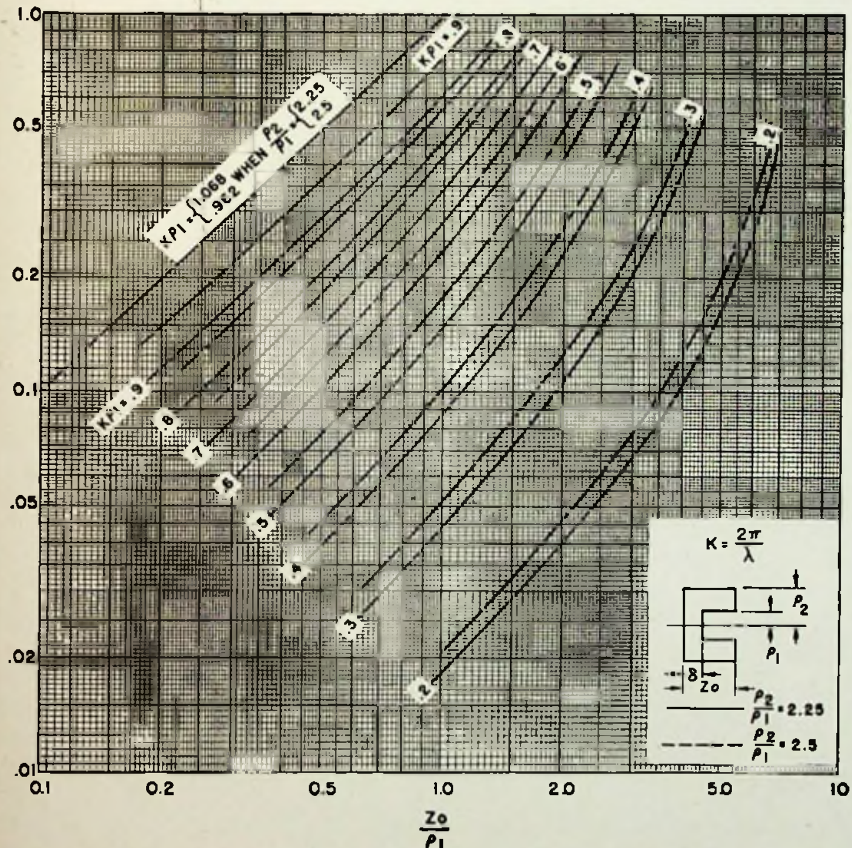


CHART IV.—Resonator design curves for radius ratios of 2.25 and 2.50.

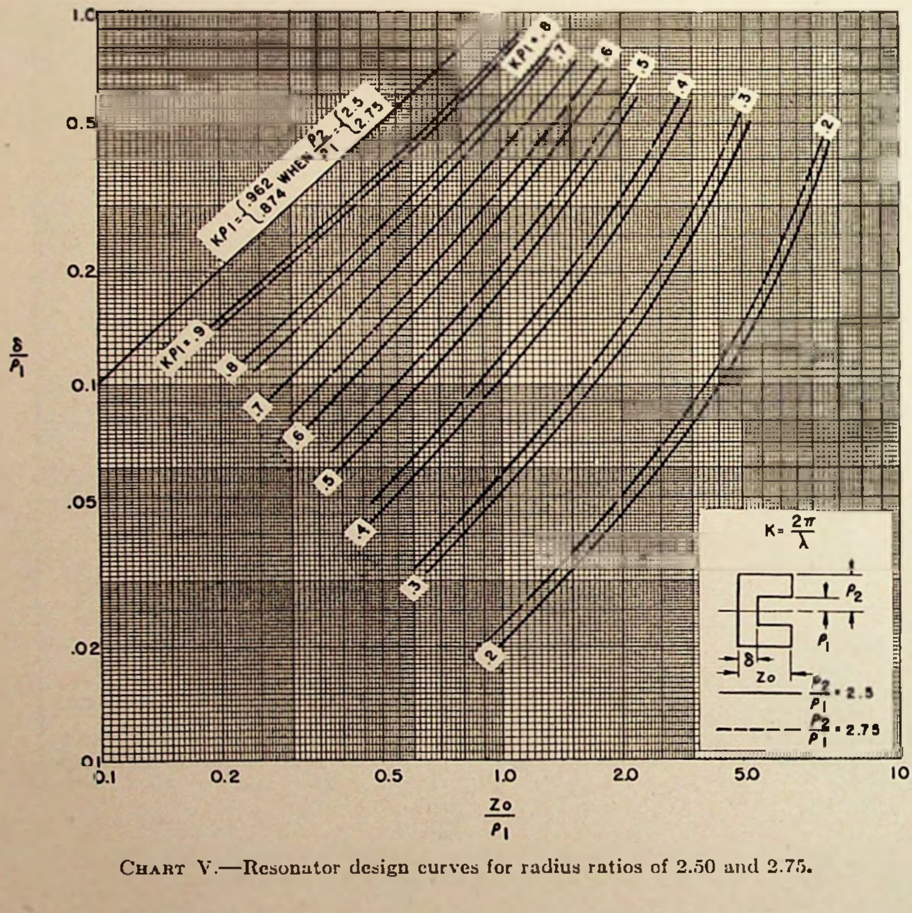


CHART V.—Resonator design curves for radius ratios of 2.50 and 2.75.

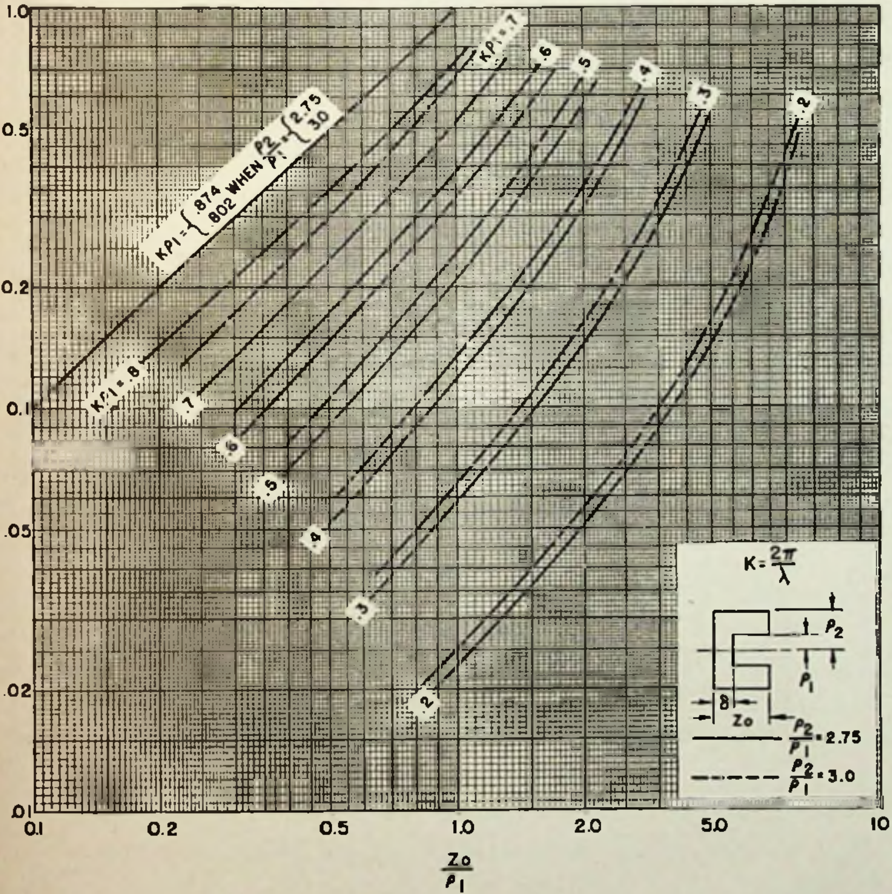


CHART VI.—Resonator design curves for radius ratios of 2.75 and 3.00.

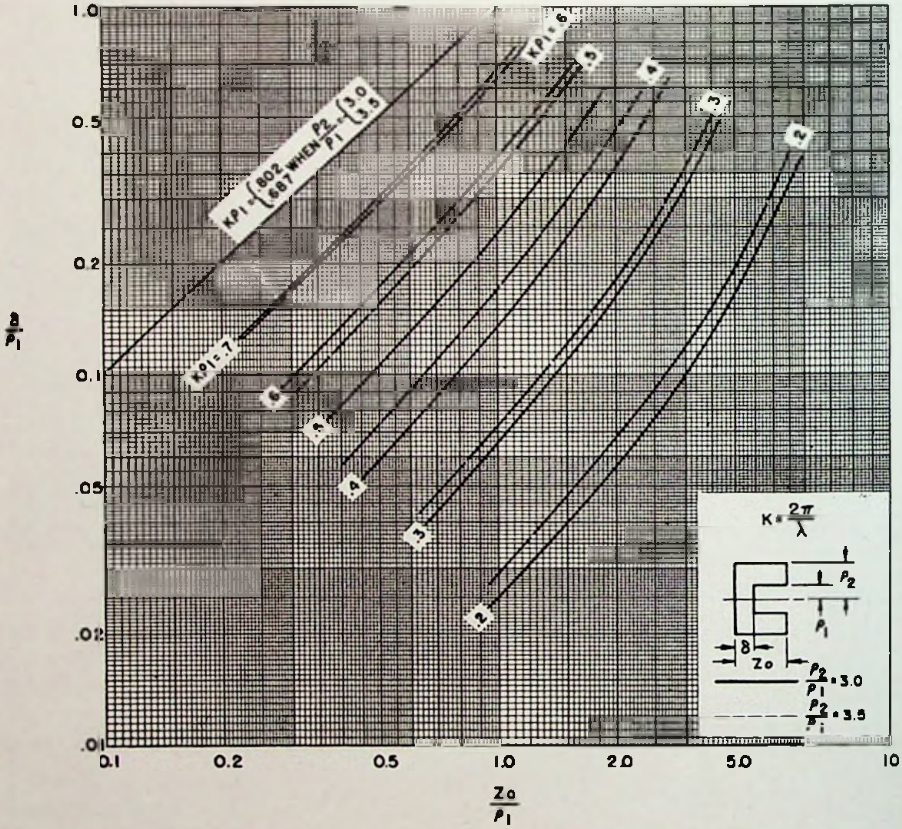


CHART VII.—Resonator design curves for radius ratios of 3.00 and 3.50.

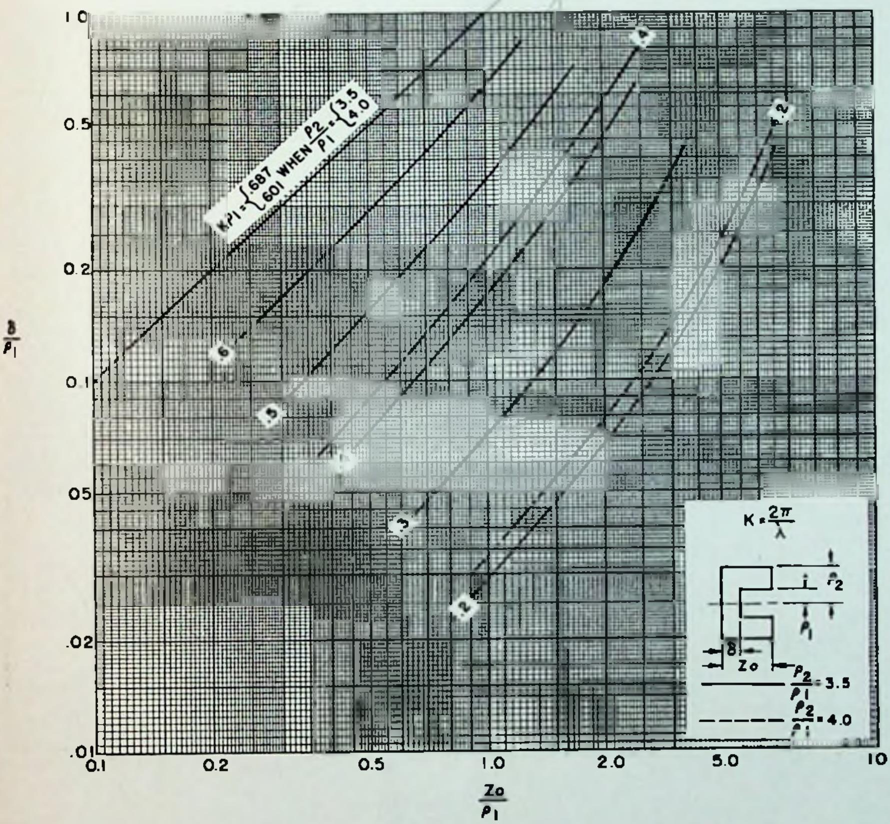


CHART VIII.—Resonator design curves for radius ratios of 3.50 and 4.00.

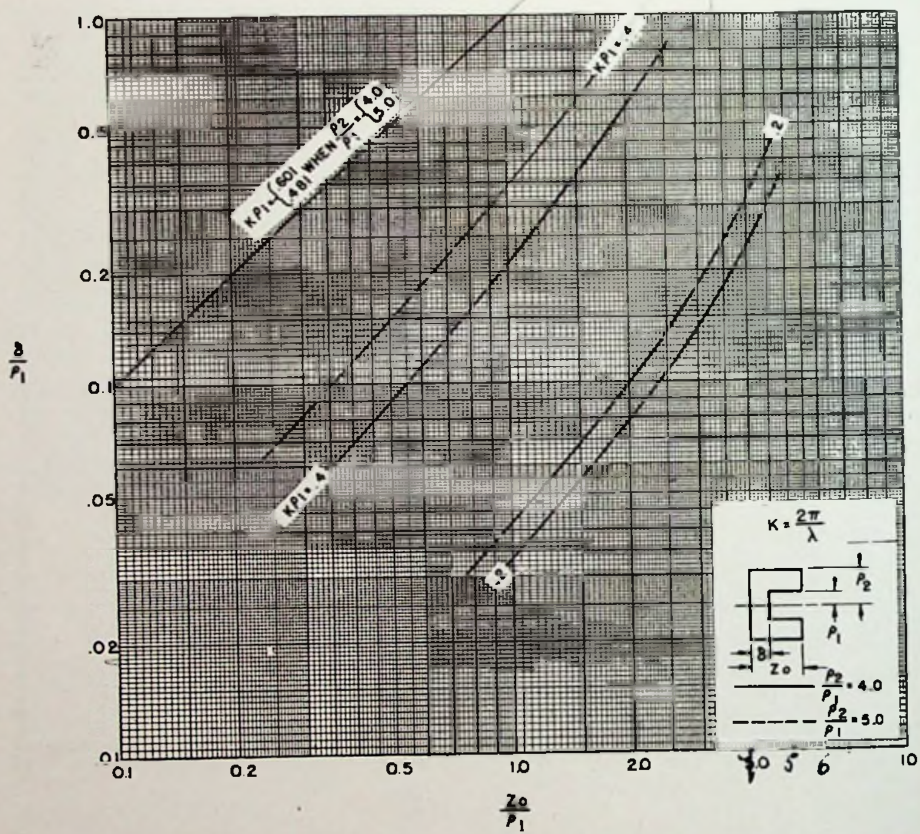


CHART IX.—Resonator design curves for radius ratios of 4.00 and 5.00.

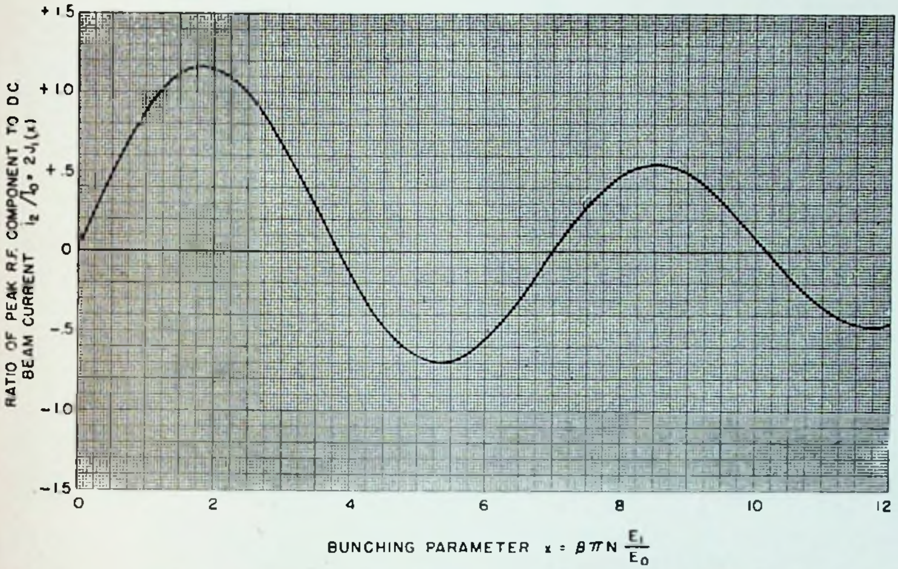


Chart X.—Ratio of the fundamental component of the bunched-beam current to the direct current in the electron beam as a function of the bunching parameter (see Figs. 3-7, 6-8, and 7-4).

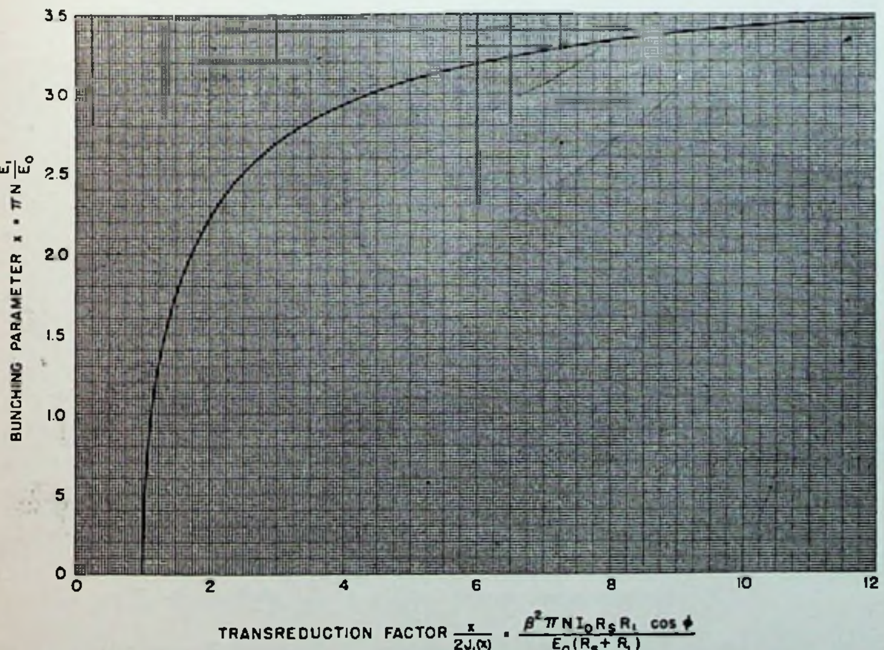


Chart XI.—Bunching parameter as a function of the transreduction factor (see Fig. 7-5).

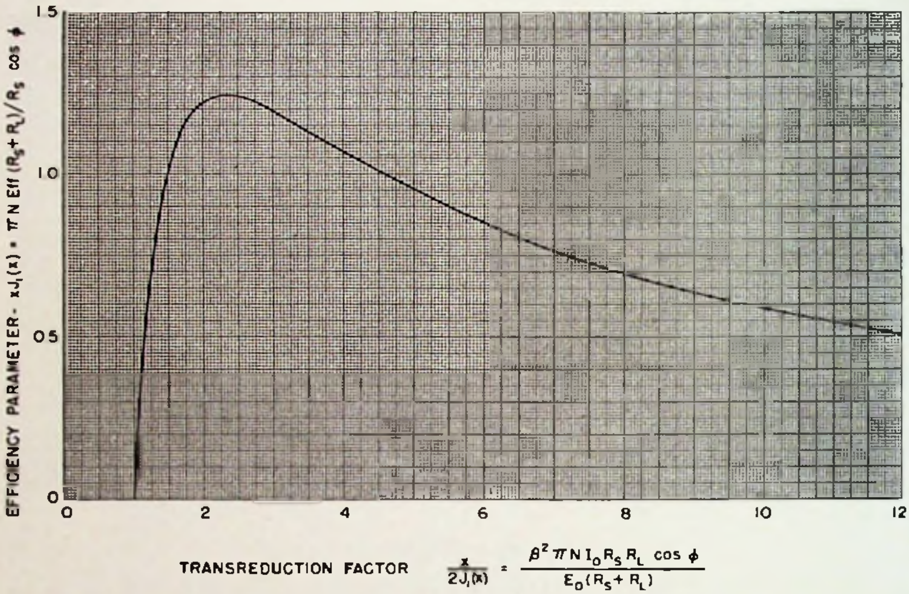


CHART XII.—Universal curve for the efficiency of a reflex klystron oscillator (see Fig. 7–7).

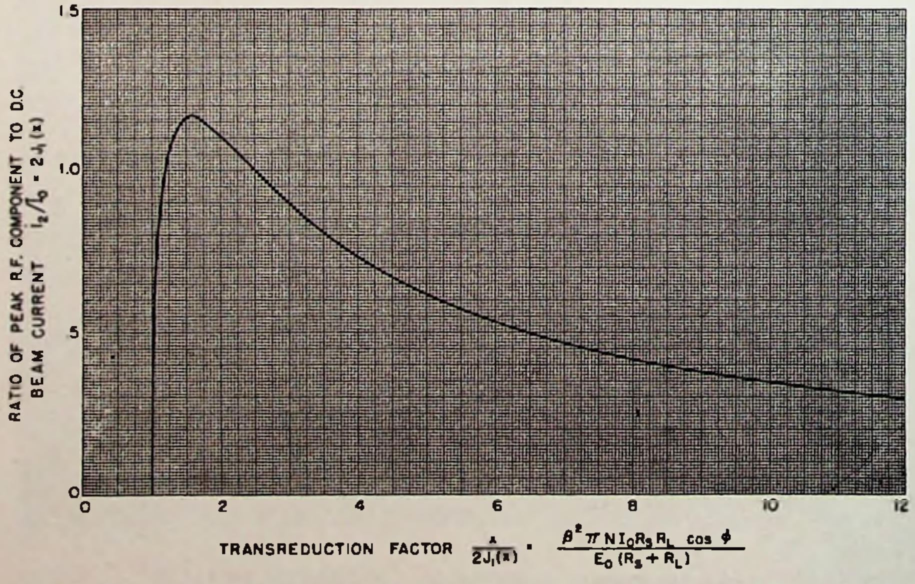


CHART XIII.—Ratio of the fundamental component of the bunched-beam current to the direct current in the electron beam as a function of the transreduction factor (see Fig. 8–6).

APPENDIX C

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INDEX