CHAPTER 7.2

MICROWAVE GENERATORS AND AMPLIFIERS

Robert S. Symons, Donald H. Preist, M. B. Shrader, Pamela L. Walchli, George K. Farney, Howard R. Jory

KLYSTRONS

Robert S. Symons

Introduction

For high frequencies, linear-beam tubes overcome the transit-time limitations of grid-controlled tubes by accelerating the electron stream to high velocity before it is modulated. Modulation is accomplished by varying the velocity, with consequent drifting of electrons into bunches to produce RF space current. The RF circuit for coupling signals to and from the electron beam are generally integral parts of the tube. Two basic types are important today, klystrons and traveling-wave tubes. Different versions of each are used as oscillators and amplifiers.

In a klystron, the RF circuits are resonant cavities which act as transformers to couple the high-impedance beam to low-impedance transmission lines. The frequency response is limited by the impedance-bandwidth product of the cavities but can be increased by stagger tuning and by multiple-resonance filter-type cavities.

Reflex Klystrons

In the reflex klystron a single resonator is used to modulate the beam and extract RF energy from it, making the tube simple and easy to tune. The beam passes through the cavity and is reflected by a negatively charged electrode to pass through again in the reverse direction. With proper phasing determined by applied voltages, oscillating modes occur for n + three-quarters-cycle transit time between passes through the cavity. The frequency can be modulated by varying voltage on the reflector (which draws no current). Reflex klystrons have been used as test signal sources, receiver local oscillators, pump sources for parametric amplifiers, and low-power transmitters for FM line-of-sight relays. Reflex-tube frequencies cover the entire microwave range from 1 to 140 GHz. In new applications, they have largely been replaced by solid-state devices.

Two-Cavity Klystron Oscillators

In all klystrons except the reflex, the beam goes through each cavity in succession, and so there is no feedback. The tube is a buffered amplifier, with each stage isolated from those upstream. Electromagnetic feedback may be provided to make an oscillator.

The specialized two-cavity oscillator has a coupling iris in the wall between the cavities. This tube is more efficient and more powerful than the reflex klystron. It can be frequency-modulated by varying the cathode voltage around the center of the oscillating mode but requires more modulator power than a reflex klystron.

Two-cavity oscillators have been used where moderate power and stable frequency for low side band noise are needed. Examples are the transmitter source in Doppler navigators, pumps and parametric amplifiers, and master oscillators for cw Doppler radar illuminators.

Again in new applications, most requirements that were met by two-cavity oscillators are now met by solid-state devices unless extremely low noise requirements exist.

Extended-Interaction Oscillators

At millimeter wave frequencies the losses in klystron cavities make it hard to build up the impedance necessary to oscillate with the very small low-current beams required.

If a series of cavities are coupled together and interact sequentially with the beam in the proper phase, the total interaction impedance increases directly with the number of cavities. The circuits of extended-interaction oscillators resemble those of traveling-wave tubes. Since they operate with a complete standing wave (at the cutoff of the traveling-wave passband), the tubes can be classed as klystrons. Various names are used for tubes of this type. The Laddertron uses a ladder-shaped periodic circuit and a flat-ribbon electron beam. Multicavity klystron oscillators use coupled cavities and cylindrical beams. Communication and Power Industries (formerly Varion) makes a number of extended-interaction klystron oscillators with power output ranging from 1 kW at 15 GHz to 1 W at 300 GHz. Most of these oscillators operate with a beam voltage near 10 kV. They provide very low noise characteristics in the order of −120 dB below carrier power in a 1-Hz bandwidth close to the carrier when used with quiet power supplies.

Two-Cavity Amplifiers

In the simplest klystron amplifier, the driving signal is coupled through a transmission line to input cavity. The cavity voltage produces velocity modulation of the beam. After a single drift space, the resultant density modulation induces current in the output resonator, from which power is extracted through another transmission line. As with nearly all klystron amplifiers, for efficiency, the Q of the output cavity is adjusted so the RF voltage almost stops electrons at the center of the bunch. The beam is usually focused electrostatically.

The gain of a two-cavity klystron is about 10 dB. Use is limited because more gain is desired in high-power tubes and solid-state amplifiers are available at low powers.

Multicavity Klystron Amplifiers

Downstream from the input cavity, cascaded intermediate cavities are inserted between the input and output cavities. They usually have no external coupling and are driven by the RF beam current and in turn remodulate the beam velocity. Figure 7.2.1 shows a three-cavity klystron schematically.

Each cavity tuned to the signal frequency adds about 20 dB of gain to the 10 dB of a two-cavity klystron. Net gain of up to about 60 dB is practical. If higher gain is attempted, one must do something to suppress fast secondary electrons traveling in the reverse direction through the tube and creating a feedback path, and one must be careful that one does not create a tube with a high noise figure by exciting a high-power beam with a very low-power drive signal. In klystrons with five or more cavities, they are usually added to increase bandwidth or efficiency rather than gain and consequently are stagger tuned. The characteristics that make the klystron desirable in the various applications in which they are used include fairly linear amplification of the RF input signal and separation of the beam formation, RF interaction and collection functions in different parts of the electron tube. The separation, in turn, allows the designer of

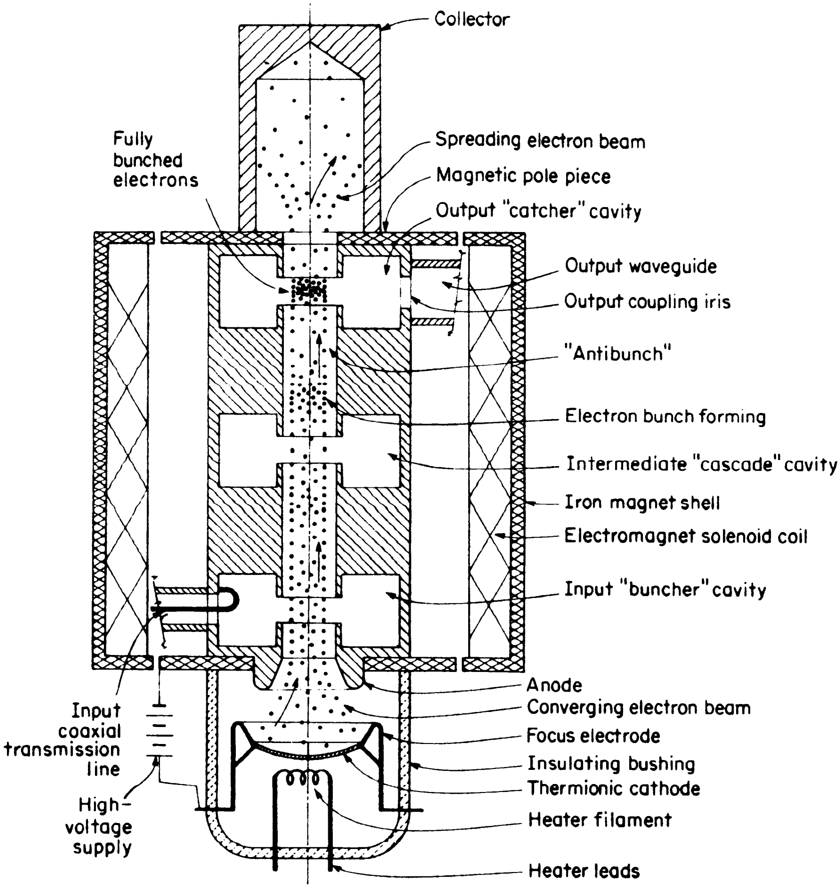

FIGURE 7.2.1 Cross section of cascade klystron amplifier. (*Varian Associates*)

the tube to optimize it for life, gain, bandwidth, low noise, efficiency or power output. Multicavity klystrons provide the economic solution to many requirements. Air cooling is often used for tubes with output power below 5 kW, and boiling water or forced liquid cooling is used for higher power. Magnetic focusing is used to control the beam in most power amplifier klystrons. Permanent magnets are used if size and required field strength permit, and solenoids or other electromagnets are used for larger and higher-power klystrons.

CW multicavity klystron amplifiers are employed in industrial-heating applications, electron-storage rings and superconducting electron linear accelerators used in high-energy nuclear physics experiments, satellite-ground-station tropospheric-scatter and space-communications transmitters, and UHF television broadcast transmitters. Pulsed multicavity klystrons are used in scientific and medical electron linear accelerators and in proton linear accelerators. They are also used in many radars in which pulse compression, Doppler processings, or spectral purity is an important requirement. A discussion of klystron designs optimized for these various applications follows:

CW klystrons for industrial heating and those used to make up the synchrotron losses of electrons in the storage rings used in nuclear physics experiments must have very large output power and high efficiency. A number of klystrons ranging in operating frequency from 200 to 2450 MHz with continuous output power up to one-half or 1 MW have been manufactured. At 200 MHz such a klystron is about 6 m in length, but the length scales inversely as frequency and directly as the square root of beam voltage. In such tubes, efficiency is achieved by using enough cavities so that, with the desired drive power, the fundamental component of current in the beam reaches the maximum that can be achieved with synchronously tuned cavities prior to reaching the output cavity. At this distance down the electron beam, the second harmonic content of the current may be used to excite a cavity tuned near the second harmonic in order to sweep electrons located between the electron bunches into the bunches. Alternatively, the beam is allowed to drift for a fairly long distance while the electrons between bunches continue to drift toward the bunches and the bunches deteriorate to some extent. In either approach, following the second-harmonic cavity or the long drift length, usually two cavities tuned to frequencies somewhat above the fundamental frequency (inductively tuned cavities) provide RF electric fields that push most electrons back toward the centers of the bunches and raise the fundamental component of current to a very high level. Using such techniques, together with high beam voltages (V) and low beam currents (I), efficiencies (η) reaching 70 to 80 percent have been achieved. An empirical relationship that fits the experimental data is $\eta = [90 - 20 \times 10^6(I/V^{3/2})]$ for $(I/V^{3/2}) \leq 3 \times 10^{-6}$.

Klystrons for communications applications require bandwidth, and so it is customary to use somewhat lower beam voltages and higher beam currents. Bandwidth scales directly as the dc beam conductance $[I/V]$ so it increases slowly as the power output of the tube is increased by raising beam voltage (because $[I/V^{3/2}]$, the electron beam "perveance," is a constant of the electron gun design. Bandwidth will increase more rapidly as the perveance, itself, is raised. As discussed above, a perveance increase will reduce efficiency, so there is a trade-off. Nevertheless, the efficiency increasing techniques discussed above remain effective, and bandwidth can be increased further by stagger tuning the input cavity and low-level intermediate cavities. Many tropospheric scatter communications tubes have been designed to operate at frequencies from UHF through 5 GHz at power levels up to several tens of kilowatts, and satellite communications klystrons with power outputs from 800 W to 15 kW have been designed to operate from 5 to 30 GHz. For deep-space communication even higher powers are used ranging from 500 kW at 2450 MHz to 200 kW at X-band. Figure 7.2.2 shows the gain characteristics of a VA-884D, a 14-kW cw broadband klystron designed for ground-to-satellite communication and tunable over the 5.925 to 6.425 GHz band. The klystron is a fairly good linear amplifier from zero signal up to 2 to 3 dB below saturated output, but the curvature of the gain characteristic does introduce some third-order intermodulation products, $(2f_2 - f_1)$ and $(2f_1 - f_2)$. Figure 7.2.3 shows the necessary reduction in the average power of a two-equal-carrier signal relative to the klystron saturation power as a function of the desired third-order intermodulation product level relative to the carrier powers. Another type of distortion in klystrons is am-to-pm conversion. It is caused by the increase in transit angle at the output cavity gap when the tube is operating very efficiently near maximum power and electrons in the bunches are nearly

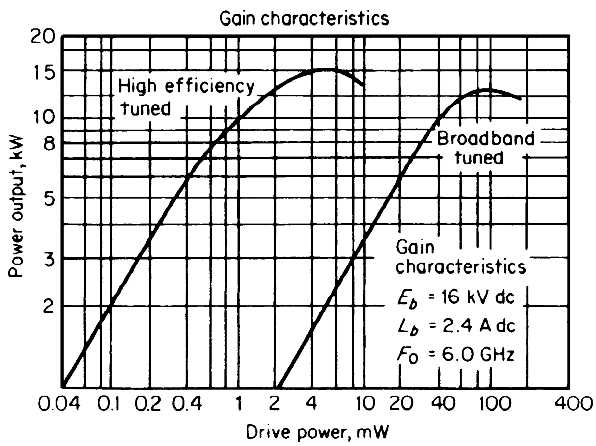

FIGURE 7.2.2 Typical gain, output power and drive power characteristics.

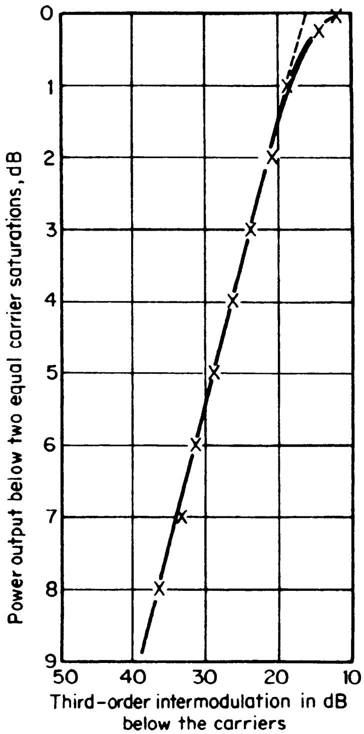

FIGURE 7.2.3 Third-order intermodulation distortion under two equal carrier conditions.

stopped. The variation in phase amounts to about 20°. Lowering the output cavity Q slightly below that value which gives maximum efficiency will substantially reduce the am-to-pm conversion. Variation of the beam voltage also causes phase modulation, so good filtering of the beam voltage supply is important. The percentage change in phase through a klystron is equal to a maximum of one-half the percentage change in beam voltage because of the square root relationship between beam voltage and electron velocity. At very high beam voltages this is somewhat less because of relativistic effects.

Figure 7.2.4 shows bandwidth curves for the klystron with various levels of drive power. Figure 7.2.5 shows the trade-off between gain and bandwidth as the cavity resonant frequencies are stagger tuned, and Fig. 7.2.6 shows fundamental and harmonic power outputs as a function of drive power. The operating bandwidth of many communications klystrons may be located anywhere within an entire communications band by means of tuners in the cavities. These sometimes take the form of a capacitive paddle which may be moved relative to the cavity gap, or sometimes the form of a moveable copper cavity wall with tungsten spring contacts which can vary the volume or "inductance" of the cavity. In either case, motion is transmitted through the vacuum wall of the klystron by means of a metal bellows. "Channel tuners," which move all cavities simultaneously to settings appropriate to one of a fixed number of broadband channels, are common on tubes designed for satellite-communication ground stations. A typical power supply schematic for a cw amplifier is shown in Fig. 7.2.7. Protective devices often required for klystron amplifiers include sensors to monitor cooling air or water flow, collector overtemperature, cathode-heating-time delay, cathode overcurrent, body overcurrent, and output waveguide arcing (photodetector and PIN diode switch).

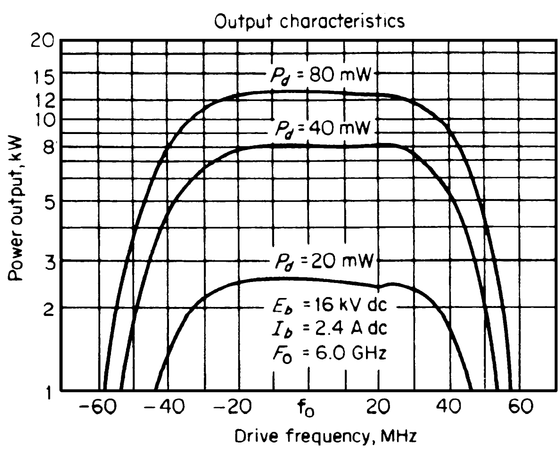

FIGURE 7.2.4 Gain and output vs. frequency characteristics under saturated and unsaturated RF drive conditions.

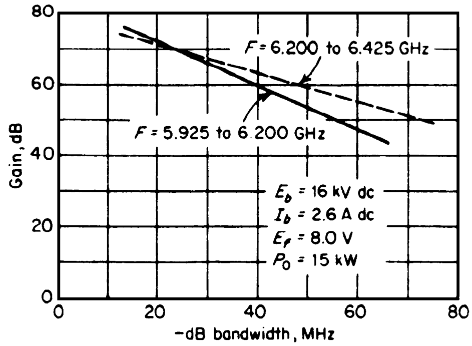

FIGURE 7.2.5 Type VA-884 series klystron gain-bandwidth characteristics.

UHF television broadcast klystrons are like other communications klystrons except that the designs are optimized to amplify the NTSC signal efficiently. In the NTSC signal, the lower sideband is vestigial and the carrier is suppressed. Typical peak output power of television klystrons is 30 to 60 kW. Operating beam voltages range from 20 to 30 kV. The required bandwidth is 6 MHz. Thus, four or five cavity stagger-tuned klystrons are used to provide gain bandwidth and efficiency. In "internal-cavity" klystrons, the cavity wall is the vacuum wall and tuning is accomplished with a capacitive paddle near the interaction gap of the cavity. The paddle is actuated through a bellows. With this type of construction, tubes with three different cavity sizes are required to cover the 470 to 806 MHz UHF TV band. "External-cavity" klystrons have a ceramic vacuum wall inside each cavity, and tuning over the entire 470 to 806 MHz band is accomplished by moving two cavity walls at each end of an air-filled waveguide-like cavity. The synchronizing pulses that are transmitted at the end of each line scan require full output power from the transmitter about 10 percent of the time. White level is 10 percent of the peak RF voltage or 1 percent of the peak power, and black level is about 78 percent of peak voltage or 60 percent of peak power. Thus, for an average gray raster, the average output power is about one-third the peak-of-sync power. Before 1974, when electric power was cheap, it was usual to operate klystrons at constant input power, and the output cavity was heavily loaded to a low Q to improve linearity and reduce am-to-pm conversion at the expense of efficiency. Thus the average efficiency was about 11 or 12 percent. After 1974, as the cost of power rose, it became attractive to operate klystrons at two different beam current levels, a lower current during the visual scan period and a higher current during the synchronizing pulses. To vary the current, a control electrode was incorporated in the electron gun. The Q of the output cavity was also increased. The increased amplitude and phase distortion in the klystron is compensated by predistortion in the exciter at the expense of partially regenerating the lower sideband which is then again eliminated by a high-power filter between the final amplifier and the

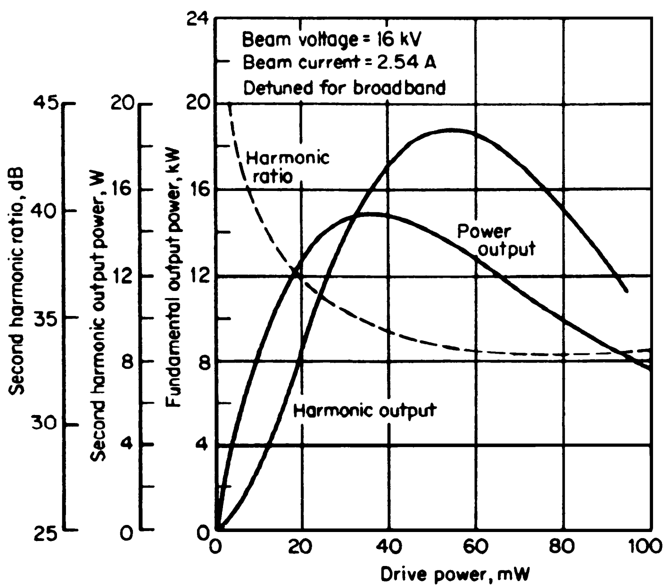

FIGURE 7.2.6 Harmonic output of a typical klystron.

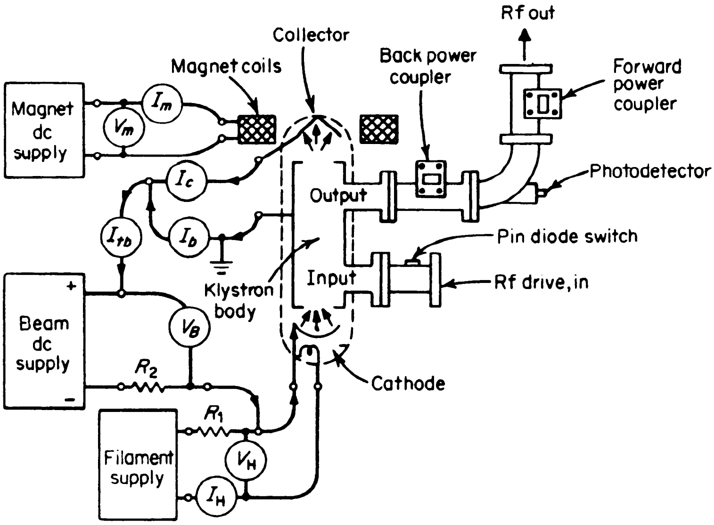

FIGURE 7.2.7 Circuits for a cw klystron amplifier.

antenna. In this way, average efficiency is increased to about 18 percent. More recently, UHF TV klystron efficiency has been increased further by means of multiple electron collector electrodes at reduced potentials, progressively closer to cathode potential. In such "multistage depressed collectors" (MSDC) klystrons, electrons tend to be collected at the lowest possible potential, and the power input goes up and down with drive level and output power in much the same way as it does in a triode or tetrode class-B amplifier. In this way efficiency is further increased by a factor of 2.5 to 3. Figure 7.2.8 shows a simplified schematic diagram of a circuit in which an MSDC klystron might be used.

Klystrons for radar and accelerator applications operate at higher voltages and currents than the cw klystrons described above. Because the electrons are traveling faster, the drift lengths between cavities and the cavity gaps can be longer. Because the gaps can be longer the drift-tube radius can also be larger without excessively lengthening the transit angle for electrons on the axis over that for electrons near the drift-tube wall. For this reason, pulse klystrons can be big enough to handle large average powers. Radar and accelerator klystrons operate at beam potentials and beam currents up to several hundred kilovolts and several hundred amperes. The Stanford Linear Accelerator Center (SLAC) has developed and manufactured many fairly conventional klystrons with peak output power at 2856 MHz ranging from 20 to 60 MW and pulse lengths of several microseconds. A 12 MW peak, 20 kW average power, 115 ms pulse length, 805 MHz klystron is used in the ion injector at the Fermi Laboratory (Fig. 7.2.9). The tube operates at 180 kV and 155 A. Long-pulse (several hundreds of microseconds) klystrons with frequencies between UHF and S-band, peak power of several megawatts and average power of several hundred kilowatts have been used to detect ballistic missiles. The Federal Aviation Agency uses many 2-to-5-MW klystrons in air-route surveillance radars (ARSR) operating in the 1250 to 1350 MHz band, and even more 1-to-2-MW klystrons in airport surveillance radars (ASR) operating in the 2700 to 2900 MHz band. An identical klystron is used in the National Oceanic and Atmospheric Administration's NEXRAD Doppler weather radar which gives superior tornado and clear-air-turbulence warnings. The coherence and high-peak-power, short-pulse performance of klystrons make them ideal for detecting close-in low-radar-cross-section moving targets such as weather disturbances. It is interesting to note that the ASR/NEXRAD klystron will operate over the voltage range from 60 to 80 kV and produce output from 500 kW to 2 MW with efficiency ranging from 45 to 52 percent over the full range of voltages. This is not atypical; most klystrons will operate from one-half to about twice their design power by varying beam voltage and allowing the current to follow. It is also interesting to note that some of the 5-to-7 MW S-band klystrons used to power 16-to-20 MeV cancer-therapy linear accelerators are made using many of the same parts used in ASR/NEXRAD radar klystrons. The ASR/NEXRAD klystron is air

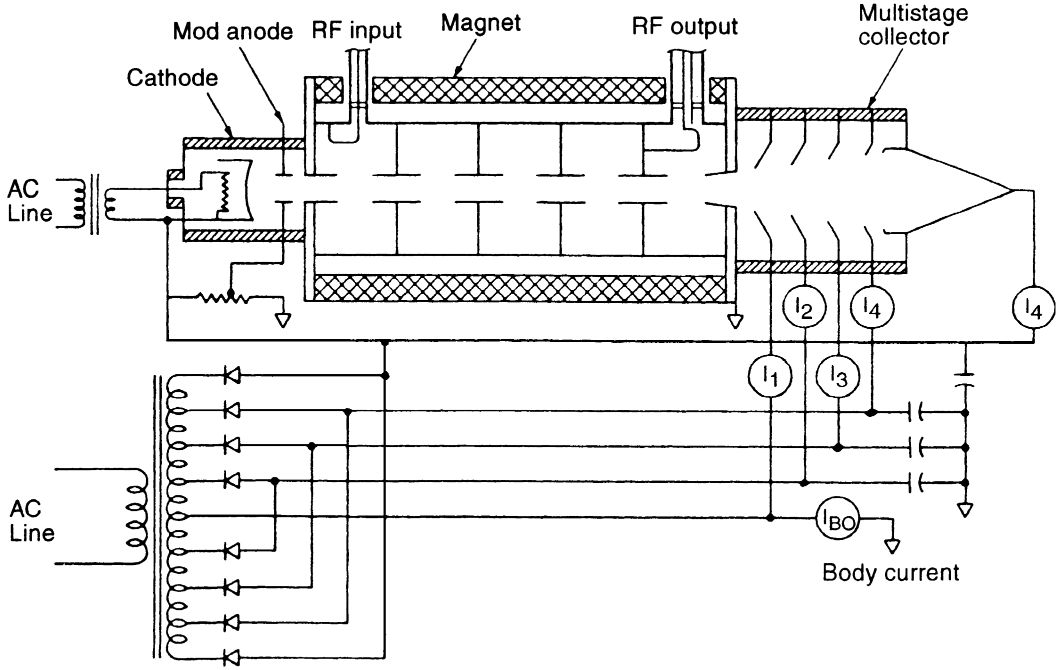

FIGURE 7.2.8 Simplified schematic for high-efficiency, depressed-collector-klystron UHF-TV final amplifier.

cooled; other radar and accelerator klystrons are liquid cooled. Nearly all are solenoid focused. Because the military, for good reason, places a premium on large instantaneous bandwidth, very few tunable klystrons remain in military service. By raising perveance, by full exploitation of stagger tuning, and by using single-gap, double-tuned output circuits, it is possible to achieve about 4 percent bandwidth at the 1-MW level and 10 percent at the 10-MW level in otherwise conventional klystrons. More modern broadband klystrons usually use some form of extended interaction which will be discussed in the next section.

All klystrons are usually operated with the cavities at ground potential. Short-pulse klystrons are usually "cathode pulsed." That is, the cathode is connected to the negative going terminal of the secondary of a pulse transformer. The primary is connected by means of a switch (solid-state or gaseous electron tube) to a charged artificial transmission line (pulse-forming network). Longer pulse lengths are usually produced by incorporating a control electrode such as a floating (or "modulating") anode in the klystron. When such an anode is used, the cathode of the klystron is connected to the negative end of a power supply incorporating a large energy storage capacitor, and electronic switches are connected between the cathode and the modulating anode (the "off switch") and between the modulating anode and ground (the "on switch"). Sometimes control grids are used in front of the cathodes of pulsed klystrons, but they are fragile and are easily damaged by arcs. When dc power supplies are used either with modulating anodes or grids, it is common to use rapid discharge or "crow bar" circuitry across the power supply to protect the tube if an arc is sensed. These usually operate in less than 1 μs and are used in conjunction with an inductor which limits the rate of rise of the fault current.

Extended Interaction Klystrons

Extended interaction klystrons (EIK) use resonant circuits in which a weak electric field interacts with the electron beam over some distance or over several gaps which are so phased that the energy gain or loss of an

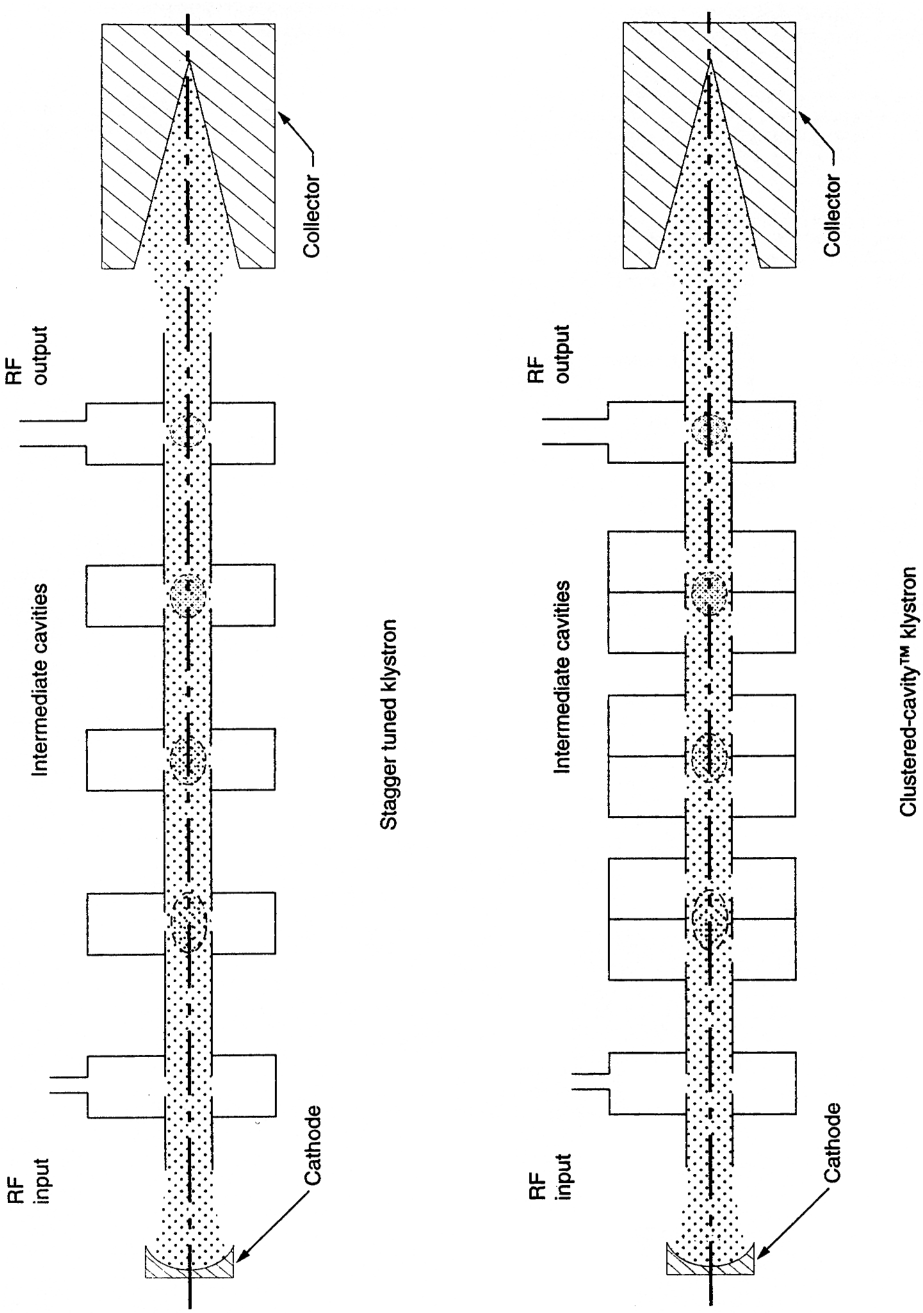

FIGURE 7.2.9 Conventional and clustered—cavity™ klystrons.

individual electron is cumulative. The advantage of such resonant circuits is that the integral of E^2 over the volume or the stored energy of the circuit is always less than it would be if the same voltage were developed at a single gap. This fact offers several possibilities:

Extended interaction output circuits (EIOC) can develop a greater impedance over a larger bandwidth, or in the highest power accelerator klystrons, the fields in the output circuit can be reduced to prevent RF arcing. Because stagger tuning of the cavities of conventional klystrons can achieve bandwidths which exceed those of single-gap output circuits, EIOCs have sometimes been used on such tubes, which have then been referred to as EIKs. At SLAC double-, triple-, and quadruple-coupled cavity circuits are being used in 150 MW, 3 GHz klystrons and 50 MW, 14 GHz klystrons. The circuits usually have been constant impedance filter circuits which support either a standing wave or a traveling wave. In standing wave resonators, the beam interacts with the forward traveling component of the wave which has been designed to have a phase velocity very nearly the same as the average beam velocity. Most traveling wave circuits are really much like traveling-wave tube output circuits in which the fields grow exponentially toward the output end. Such a circuit is used in the Twystron[R], so called because it uses a klystron buncher and a cloverleaf TWT output circuit. A more sophisticated approach to EIOC design makes use of a tapered impedance output circuit in which the filter circuits which couple the output gaps have image impedances which are progressively reduced in an inverse arithmetic taper (i.e., 1, 1/2, 1/3, . . .). Thus, when used with a current saturated electron beam in which the RF current can be made fairly constant over some distance, the current on the circuit builds up linearly (i.e., 1, 2, 3, . . .) and the voltage at all gaps is the same. This results in lower stored energy and reduces arcing for the same output power.

Extended-interaction buncher circuits have been used to a more limited extent. The first extended interaction klystron built by Wessel-Berg at Stanford about 1960, used three short-circuited helix sections as cavities and gave excellent bandwidth. Several millimeter wavelength klystrons, with power output ranging up to 2 kW at 94 GHz, have been built using ladder circuits to minimize circuit losses. Various manufacturers have proposed klystrons using pairs of cavities coupled together to achieve greater bandwidth. In the mid-1980s during computer simulations of such tubes, it was found that the coupling between the cavities of the pair was largely irrelevant because cavities, so closely spaced that there is little gain between them, are already coupled in the proper phase by the exciting current bunch so they cooperate in forming a new bunch farther downstream. This kind of klystron was named a clustered-cavity™ klystron, and in a sense, it is a special case of extended interaction which is somewhat easier to design because one does not have to deal with coupling between cavities. Figure 7.2.9 shows a schematic comparison of a conventional klystron and a clustered-cavity klystron of the same power and gain. The clustered-cavity klystron will have twice the bandwidth if the paired cavities are artificially loaded to one-half the Q values of the single cavities they replace. A 13 percent bandwidth, 3 MW, 3 GHz clustered-cavity klystron is being manufactured. Computer simulations indicated that 30 percent bandwidths may be available from megawatt klystrons of this type.

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"Twystron" is a registered trademark of Varian Associates Inc.

"Clustered-Cavity" is a registered trademark of Litton Systems, Inc.

Portions of the material on klystrons were adapted from material supplied by Richard B. Nelson.

INDUCTIVE OUTPUT TUBES (KLYSTRODES)

Donald H. Preist, M. B. Shrader

Tube and Cavity Configurations

The name *Klystrode* was coined to indicate that the output cavity, load system, and collector are basically as in a klystron. The input cavity and the input part of the tube resemble a triode. In all the UHF/TV amplifiers so far built the output cavity system is the external cavity type using a cylindrical ceramic window concentric with the electron beam. Many klystrons in service use this arrangement and a coupled secondary cavity connected to the output load, to maximize bandwidth (see Fig. 7.2.10). The input cavity is also external in that the major part of the circuitry is outside the vacuum envelope. Two types are in service.

The older type (Fig. 7.2.10) uses a small amount of controlled positive feedback and gives 1 to 3 dB more power gain than the second type (Fig. 7.2.11), but is hard to adjust. The second type is much easier to adjust and gives adequate gain, typically 22 dB. This is the favored approach in new amplifiers. In the very high power amplifiers built for "Big Science" applications the output cavity is in vacuum and the output window is in the coaxial transmission line to the load (Fig. 7.2.11). The input cavity is external as in the TV amplifiers. Both types have been used successfully, but the second type is preferred.

The most significant difference between the IOT/Klystrode and the klystron is the input cavity/tube system. The grid and cathode are at high negative dc potential with respect to ground. The correct RF voltage must appear between grid and cathode, even though the source of the RF drive power must be at dc ground potential. Also the grid to anode RF impedance must be minimized to prevent feedback; if positive this will tend to cause self-oscillation, and if negative it will tend to reduce power gain.

Transmitter Configurations for UHF/TV

Most of the earlier high-power installations used a pair of 60 kW tubes giving 120 kW P. S. visual and a third giving 13 kW CW aural. Later, transmitters featuring combined aural and visual amplification in the same tube were put into service. The exacting NTSC specifications could be met only by a high degree of amplitude and

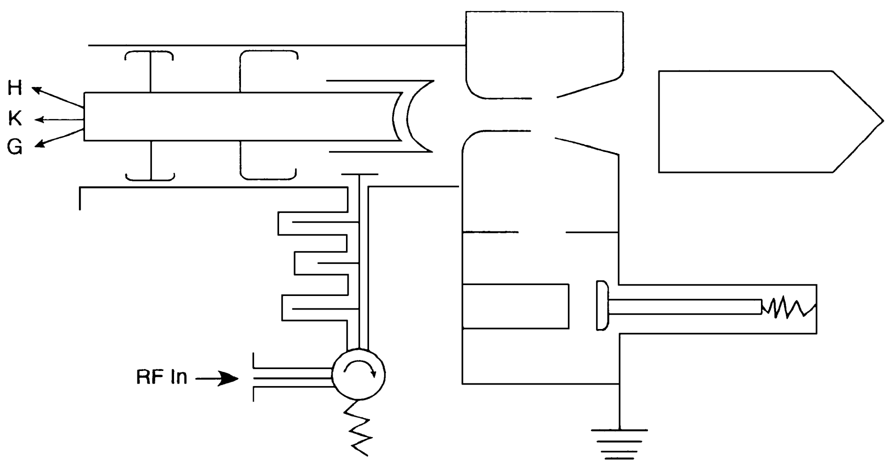

FIGURE 7.2.10 Input cavity provides controllable positive feedback to enhance gain. It is stabilized by stub tuner/Ferrite circulator combination. Resonant cavities iris coupled. (Loop has been used.) Preferred for 450 MHz and higher. Magnetic field between anode and collector not shown.

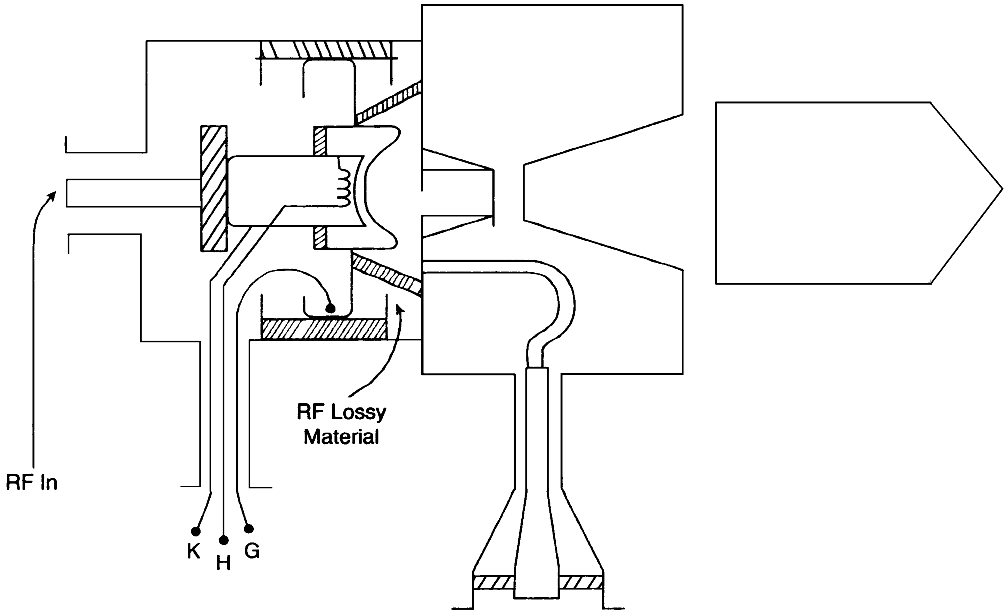

FIGURE 7.2.11 Grid-anode part of input cavity is heavily damped by lossy material for stabilization. Bringing out H, K, G connections radially makes cathode extension available for coupling RF drive through HV dc insulation. Output cavity in vacuum shows loop coupling to load. Fine-tuning “paddle” not shown. Magnetic field not shown.

phase linearity. This was achieved by a combination of careful tube design and a sophisticated precorrection module which allowed the residual nonlinearity of the IOT/Klystrode and the solid-state driver amplifier to be reduced to a very low level.

The IOT itself is more linear than the klystron at black level in TV service, and it resembles a tetrode at low level (Fig. 7.2.12).

Properties of the IOT/Klystrode as an Electron Device

First, it must be emphasized that the tube, unlike the klystron and the TWT is not a velocity-modulated tube. It uses a density-modulated electron beam as in a triode or tetrode. If the electron transit time through the tube is small in terms of the operating frequency, e.g., 1 rad, the efficiency will be high and may approach the theoretical limit of 78.5 percent. The “Big Science” tubes have shown 70 to 74 percent at output powers of several hundred kW. However, as the frequency is raised and the transit time or angle increases, the efficiency will eventually fall, but it will exceed the triode efficiency because the IOT anode is at high dc potential at the time of maximum current flow.

The tube (Fig. 7.2.10) is a five-electrode device with a spherical cathode and grid and an apertured anode at high dc potential. A reentrant cavity is placed between the anode and the next electrode, the tailpipe. Insulated from this is the final electrode, the collector. The electron beam is guided through the cavity by a magnetic field.

Because of the action of the negatively biased grid and the superimposed RF voltage the beam current flows for about half the RF cycle and is zero for the rest of the cycle. It passes through the aperture in the anode and the first part of the cavity without interception, at constant velocity. Next, it passes through the gap in the cavity where the RF field is decelerating, then through the second field-free region in the tailpipe with minimal interception due to the magnetic field, and finally enters the collector which is usually at ground dc potential. The two field-free regions are field-free because they are cylindrical metal pipes long enough to behave as waveguides beyond cut-off at the operating frequency.

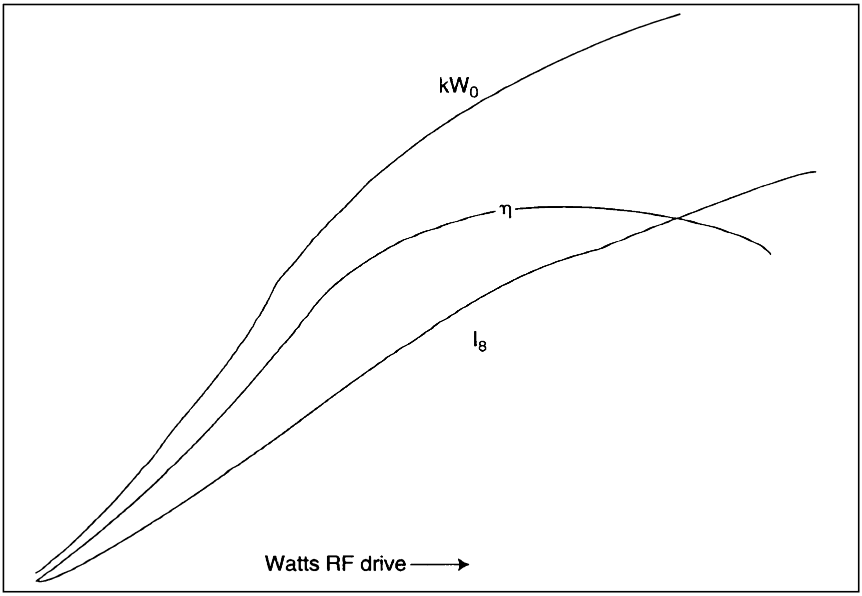

FIGURE 7.2.12 Power output vs. RF drive shows fair linearity. Efficiency stays relatively flat over 2:1 change in drive. Beam current is zero for zero drive.

Reference 2 contains a detailed analysis of the electron field and wave interactions. From the user's viewpoint the outstanding features are: (a) a complete absence of interaction between the output load impedance and the grid-cathode conditions, except when the load impedance approaches infinity; (b) the absence of a "dc blocking capacitor" carrying RF current as in a tetrode; (c) the output cavity system is at dc ground potential; (d) the power dissipated on the tailpipe is small; (e) the power output is not limited by the collector as this can be made indefinitely large and easy to cool, in principle. In practice there may be constructional problems with a very large collector on a super power tube. It should be noted, though, that the collector has to dissipate only the dc beam power minus the RF output power. In a conventional klystron the entire beam power has to be dissipated when the RF drive goes to zero.

Other significant performance characteristics can be seen from Fig. 7.2.12. Because the beam current falls to zero when the RF drive falls to zero, and varies monotonically with drive level, the device resembles the class B linear amplifier well-known to tetrode users. It is this feature that provides high average efficiency with an amplitude-modulated signal such as in TV, compared to a klystron. This is even more important with systems having a high peak-to-average signal ratio such as some digital HDTV systems.

In high power CW or long pulse (10 ms) service the IOT/Klystrode has performed well at 425 MHz (Ref. 5) and at 267 MHz (Ref. 4) as an RF power source for driving proton accelerators (Fig. 7.2.13). Here a set of characteristics become important which do not apply in TV service. Among these are: (1) absence of need for a high level pulser because the RF drive can be pulsed; (2) at VHF the physical size is very much smaller than a Klystron; (3) a high-power ferrite circulator may not be needed between the tube and the accelerator, and (4) the efficiency is higher than that of a klystron or a tetrode (70 to 74 percent measured).

Basis for Output Improvements

There are several reasons why IOT/Klystrodes can produce hundreds of kW compared to the 100 watts of Haeff's developmental tube, described in 1939 (Ref. 1).

1. The evolution of techniques for designing and fabricating electron guns, beam-focusing systems, cooling systems, cavities, and output windows over decades of microwave tube development,

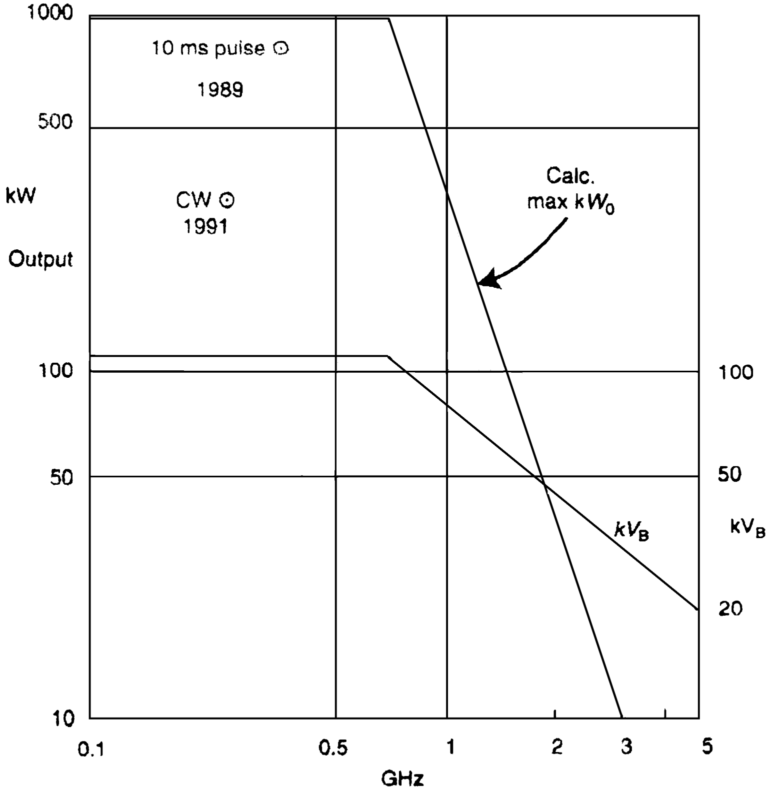

FIGURE 7.2.13 Assumptions made in calculating maximum kW are (*a*) constant dc perveance 0.3 microperv, (*b*) maximum usable gun anode voltage 113 kV up to 750 MHz 20 kV at 3 GHz, (*c*) tube dimensions vary inversely with *f* above 750 MHz, (*d*) efficiency 70 percent.

2. The availability of pyrolytic graphite, an excellent material for making grids,

3. The availability of high purity alumina for output windows and coatings for multipactor suppression, and interelectrode insulators, and

4. The invention of the tungsten matrix impregnated cathode.

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TRAVELING-WAVE TUBES

Pamela L. Walchli

Introduction

The traveling-wave tube (TWT) is a linear-beam device which amplifies microwave signals to high-power levels over broad bandwidths. It was invented by Rudolf Kompfner in the latter part of World War II and developed into a viable device by J. R. Pierce and L. M. Field at Bell Telephone Laboratories in 1945. Today, the TWT finds diverse application in communications, radar guidance, and electronic countermeasure systems.

Basic structure. All TWTs comprise four basic elements: (1) an electron gun, (2) an RF interaction circuit, (3) an electron-beam magnetic focusing system, and (4) a collector to dissipate the spent beam power.

The major difference between the various types of TWTs lies in the RF interaction structure. A schematic representation of a typical TWT is shown in Fig. 7.2.14. At the left is the electron gun, which forms the beam; at the center are the RF interaction structure (in this case, a helix) and the magnetic beam-focusing system; on the right is the collector that absorbs the spent beam power.

Theory of operation. The purpose of the interaction structure is to slow the RF signal so that it travels at the same speed as the electron beam. Electrons enter the structure during both positive and negative phases of an RF cycle. Those entering during a positive phase are accelerated; those entering during a negative phase are decelerated. The electrons that experience a velocity increase catch up with the electrons that have been slowed down, forming electron bunches. These bunches produce an alternating current superimposed on the dc beam current. The alternating current induces growth of the RF circuit wave, which, in turn, forms tighter electron bunches and thus a larger component of alternating current.

Growth of the wave on the circuit occurs because the velocity at which the beam is traveling forces the electron bunches to enter a decelerating phase of the RF field. In the decelerating field, the electrons are slowed, transferring their energy to the RF wave. This cycle is limited by one or more severs and ultimately by the extraction of the RF power through the output connector. The sever absorbs the RF power that has been built up on the circuit but does not affect the ac component of current in the beam.

The modulated beam drifts through the sever region, and induces a new RF wave in the next circuit section, where the interaction process begins again. The purpose of the sever is to absorb reflected power, which travels in a backward direction on the circuit. The reflected power arises from an imperfect match between the RF circuit and the output connector. Without the sever, regenerative oscillations would be induced.

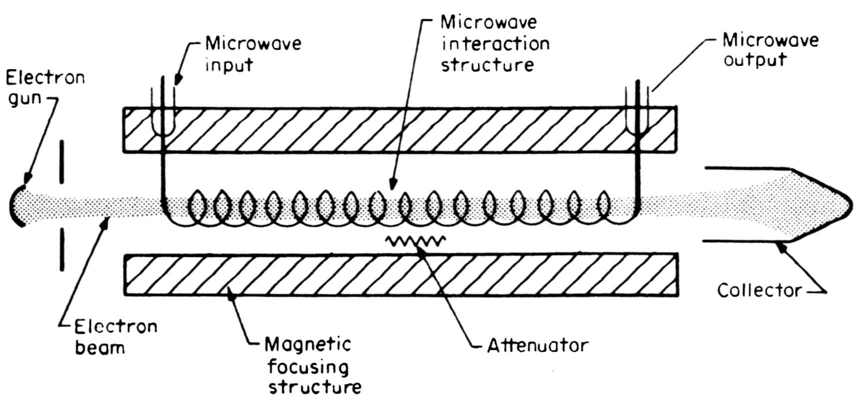

FIGURE 7.2.14 Basic elements of a typical TWT.

At any given frequency, a certain level of drive power will cause the maximum degree of bunching and thus the greatest amount of output power. This condition is known as *saturation.* For small drive signals, typical TWTs have 40 to 70 dB of gain.

The Electron Gun

The electron gun forms a high-current-density pencil beam of electrons, part of whose energy is converted to RF power through interaction with the wave traveling along the RF circuit. In the typical electron gun, electrons are emitted from a spherical cathode and converged to the required beam size by focusing electrodes. The final converged beam size is maintained through the interaction structure by either permanent magnet or electromagnet focusing.

On-off switching modulation of the electron beam is accomplished by applying a pulse to one of four electrodes: (1) cathode, (2) anode, (3) focus electrode, or (4) grid.

Cathode and anode pulsing. In the first method, the cathode is pulsed negatively with respect to the grounded anode, requiring both the full beam voltage and current to be switched. Alternatively, anode modulation involves switching the full beam voltage between cathode potential and ground, but the current switched is just that intercepted on the anode, usually only a few percent of the full beam current.

Focus-electrode pulsing. The focus electrode, which normally operates at or near cathode potential, can be biased negatively with respect to the cathode to turn the beam off. The voltage swing required is usually one-third or less of the full cathode voltage. Since the focus electrode draws no current, reduction in power requirements is significant.

Grid pulsing. Switching power requirements are minimized with grid modulation. A grid structure, to which the modulating voltage is applied, is placed directly in front of the cathode surface. The amount of voltage swing needed is typically only one-twentieth or less of the full beam voltage.

Some common grid structures in use are shown in Fig. 7.2.15. The grid in Fig. 7.2.15*a* is a simple, single intercepting grid. To turn the beam on, a voltage positive with respect to the cathode is applied to the grid, drawing current from the full cathode surface. The grid webs intercept the current drawn from the cathode area directly behind the grid. This interception limits the duty cycle at which the tube can operate.

Current drawn by the grid can be minimized by schemes such as those shown in Figs. 7.2.15*b* and 7.2.15*c*. The structure in Fig. 7.2.15*b* is composed of two grids, the one nearest the cathode surface operating at cathode potential and the outer one operating at the modulating voltage. The inner grid, identical in pattern to the outer grid, prevents emission from the cathode surface directly behind it, effectively eliminating intercepted current on the control grid. This inner grid is referred to as a *shadow grid.* The shadow grid may be attached directly to the cathode surface, as in Fig. 7.2.15*c*. This kind of structure has the trade name Unigrid.

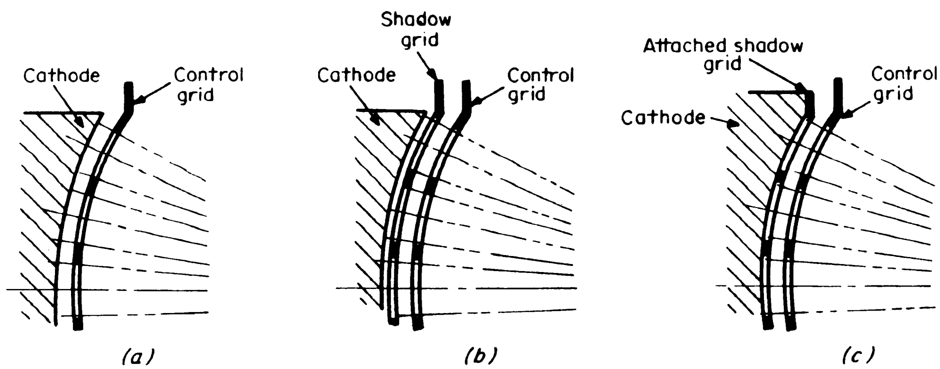

FIGURE 7.2.15 Grid structures used in TWTs: (*a*) simple intercepting grid; (*b*) double grid, with shadow grid operated at cathode potential; (*c*) unigrid type, with shadow grid attached directly to cathode.

A simple grid has application in smaller guns, where high amplification factors are not required. The grid operates at cathode potential and therefore intercepts no current. To turn off the beam, a voltage negative with respect to the cathode is applied to the grid. Amplification factors around 10 are typical for this kind of electron gun.

Magnetic Beam Focusing

Without a focusing system the electron beam would spread due to the mutually repulsive forces on like-charged particles, causing the electrons to strike the RF circuit. A magnetic beam-focusing system is the most widely used and is usually implemented in one of three ways: (1) electromagnetic, (2) permanent magnet, or (3) periodic permanent magnet.

Electromagnet Focusing. Electromagnet focusing is used primarily on very high power coupled-cavity TWTs. Tight beam focusing is required in these tubes because significant interception on the RF circuit is intolerable at the power levels in question. Disadvantages of this kind of focusing are size, weight, and consumption of power, but all can be reduced somewhat by wrapping the windings of the solenoid directly on the tube body. Solenoidal focusing is illustrated in Fig. 7.2.16*a*.

Permanent-Magnet Focusing. Permanent-magnet focusing is possible where the interaction structure is short, e.g., in low-gain or millimeter-wave tubes. It can be used in place of solenoidal focusing in these kinds of tubes. This focusing system is shown in Fig. 7.2.16*b*.

Periodic-Permanent-Magnet Focusing. Periodic-permanent-magnet focusing is used on almost all helix TWTs and most coupled-cavity TWTs. A periodic-permanent-magnet (PPM) structure is shown in

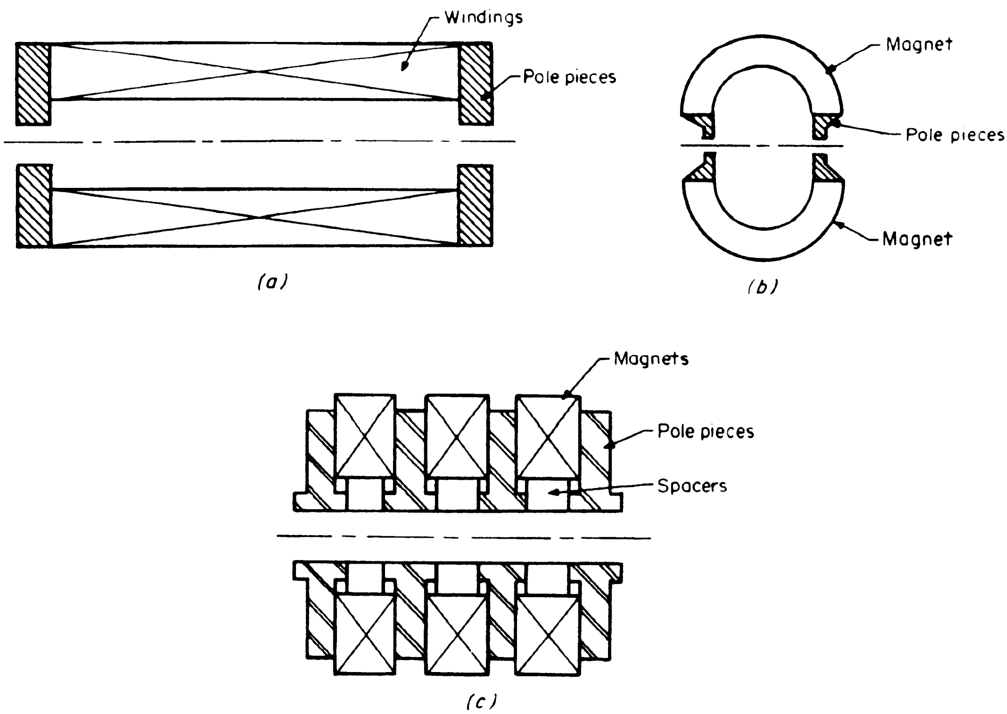

FIGURE 7.2.16 Magnetic focusing arrangements: (*a*) solenoidal type; (*b*) permanent-magnet type; (*c*) periodic permanent-magnet structure.

Fig. 7.2.16*c*. The magnets are arranged with alternate axial polarity in successive cells. In helix TWTs the pole pieces (with nonmagnetic spacers) may form the tube's vacuum envelope, or the pole pieces and spacers may be slipped over a stainless-steel tube that serves the same purpose. In coupled-cavity TWTs the cavity walls themselves are the pole pieces.

This kind of focusing provides a major reduction of tube size and weight, along with the elimination of the magnet power supply. The drawback of this scheme is that the electron beam ripples with a periodicity of the length of one magnet cell. This increases beam interception on the RF circuit and thus generally limits the use of PPM focusing to lower average-power TWTs.

The Interaction Circuit

The fundamental principle of operation of a TWT is that an electron beam moving at approximately the same velocity as an RF wave traveling along a circuit gives up energy to the RF wave. Since the RF wave travels at the speed of light, a method must be found to slow the forward progress of the wave to roughly the same velocity as that of the electron beam. The beam speed in a TWT is typically between 10 and 50 percent of the velocity of light, corresponding to cathode voltages of 4 to 120 kV. The two structures that accomplish the slowing of the RF wave are the helical and coupled-cavity circuits.

Helix circuits. The helix (Fig. 7.2.17) is supported inside the vacuum envelope by three or more ceramic support rods, which also conduct heat away from the helix. A helix interaction structure is used where bandwidths of an octave or more are required, since over this range the velocity of the signal carried by the helix is almost constant with frequency. For greater than octave-bandwidth operation, the variation of velocity with frequency can be modified by the introduction of metal loading segments near the helix, causing the phase velocities of a wider range of frequencies to be more nearly in synchronism with the beam velocity.

The helix provides satisfactory performance over the range of frequencies from 500 MHz to over 40 GHz. However, the typical helix circuit is limited in average power-handling capability to a few hundred watts. Peak power levels above several kilowatts cannot, in general, be achieved because of circuit RF instabilities. Higher peak power levels can be obtained by eliminating these oscillations with a special type of helix circuit consisting of the superposition of a helix wound in a right-hand sense on a helix wound in a left-hand sense. Two practical implementations of this configuration are the ring-loop and ring-bar circuits (Fig. 7.2.18). Peak powers of hundreds of kilowatts are attainable, but average power capability is no better than that of the conventional helix

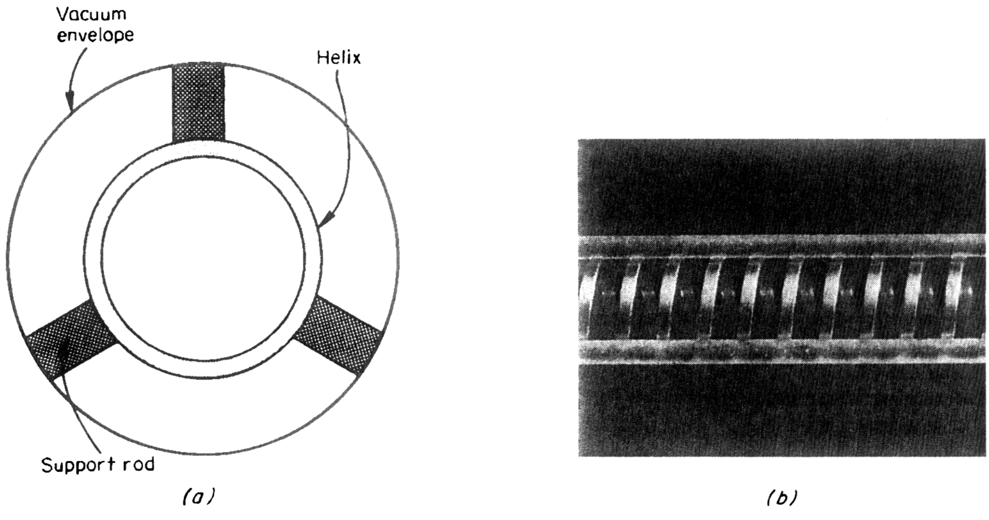

FIGURE 7.2.17 Helix circuit: (*a*) end view; (*b*) side view.

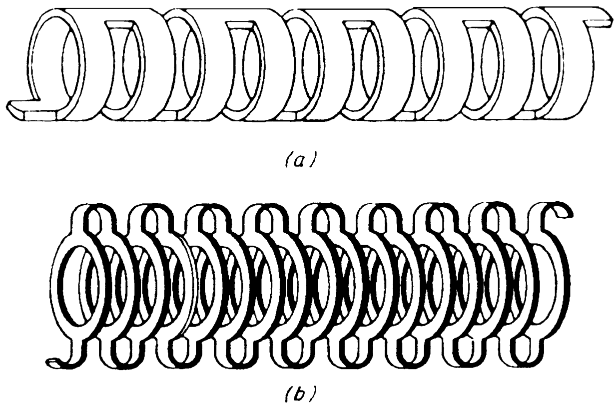

FIGURE 7.2.18 Structures composed of two helixes superimposed in opposite sense of rotation: (*a*) ring-bar circuit; (*b*) ring-loop circuit.

circuits, since the structures are supported in a like manner. Because the ring-bar and ring-loop circuits are dispersive, the maximum bandwidth of a tube using them is typically only one-third octave.

Coupled-Cavity Circuits. Because of its superior ability to dissipate heat, the coupled-cavity structure is capable of both high peak and average power over moderate bandwidths. Coupled-cavity tubes find applications from 2 GHz up to nearly 100 GHz. Bandwidths of 10 percent are typical, although tubes with 40 percent bandwidth have been developed.

The coupled-cavity circuit consists of resonant cavities coupled through slots cut in the cavity end walls, resembling a folded waveguide. This arrangement results in a bandpass filter network that is highly dispersive, limiting the tube bandwidth. The two most common kinds of coupling schemes are illustrated in Fig. 7.2.19. The structure in Fig. 7.2.19*a* is a forward fundamental circuit, also called a cloverleaf circuit from the shape of its cavities. It is used primarily on extremely high peak power coupled-cavity tubes or in the output section of a hybrid klystron TWT, known as a Twystron. Typical performance of a tube with a forward fundamental circuit is 3 MW peak and 5 kW average at S band. Figure 7.2.19*b* illustrates the more commonly used coupled-cavity structure, the single-slot space harmonic circuit. A peak power of 50 kW and an average power of 5 kW at X band are typical for space harmonic TWTs.

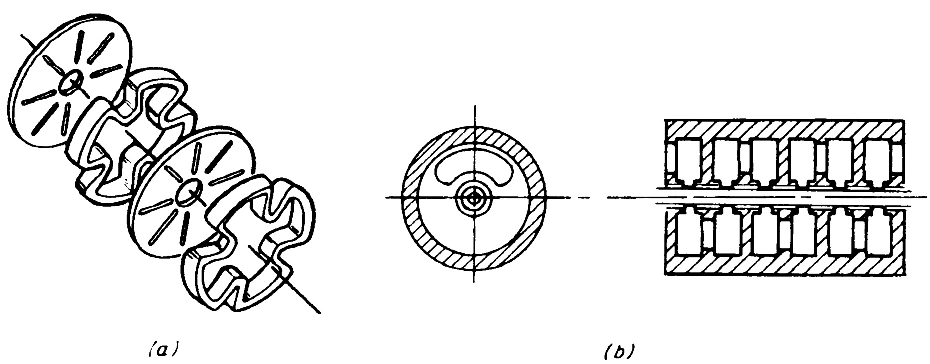

FIGURE 7.2.19 Coupled-cavity circuits: (*a*) forward fundamental circuit ("cloverleaf"); (*b*) single-slot space harmonic circuit.

The Collector

The function of the collector is to collect the electron beam after it has passed through the interaction structure and dissipate the remaining beam energy. During interaction electrons give up various amounts of energy, and some actually gain energy. Typically, the slowest electrons lose no more than 50 percent of their original energy; the fastest gain at most 20 percent with the remainder distributed between these extremes. If a TWT had an interaction efficiency of 20 percent, the average electron would possess 80 percent of its original energy.

Single-Stage Collectors. The overall efficiency of the TWT can be increased by operating the collector at a voltage lower than the full beam voltage, a practice known as *collector depression.* This introduces a potential difference between the interaction structure and the collector through which the electrons pass. The amount by which a single-stage collector can be depressed is limited by the remaining energy of the slowest electrons; i.e., the potential drop can be no greater than the amount of energy of the slowest electrons or they will be turned around and reenter the interaction structure, causing oscillations.

Multistage Collectors. Efficiency can be increased still more by introducing multiple depressed-collector stages. This method provides for the collection of the slowest electrons on one stage, while allowing those with more energy to be collected on other stages depressed still further. Figures 7.2.20*a* and *b* represent the configuration of power supplies (less the heater supply) to operate a gridded TWT with a single-stage and a multistage depressed collector, respectively. Calculations of the overall efficiency of such TWTs are shown in the following table, assuming a beam power of 5 kW (10 kV, 0.5 A) and an interaction efficiency of 15 percent:

	Voltage, kV	Current, A	Power, W
	Single-stage collector		
Helix supply	10	0.025	250
Collector supply	5	0.475	2,375

$$\text{Overall efficiency} = \frac{750\text{ W}}{250\text{ W} + 2375\text{ W}} = 29\%$$

	Voltage, kV	Current, A	Power, W
	Multistage collector		
Helix supply	10	0.025	250
Collector stage, 1	5	0.23	1,150
2	2.5	0.15	375
3	1	0.085	85
4	0	0.01	0

$$\text{Overall efficiency} = \frac{750\text{ W}}{250\text{ W} + 1150\text{ W} + 375\text{ W} + 85\text{ W}} = 40\%$$

Helix TWTs are cooled conductively by mounting the tube in a metal baseplate, which is in turn attached to an air- or liquid-cooled heat sink. Coupled-cavity tubes below 1 kW average power are cooled convectively by drawing air over the entire tube length. Higher-power coupled-cavity tubes are cooled by circulating liquid over the tube body and collector.

Microwave Power Module

In keeping with the general trend in electronic components toward minimization of system size, the microwave power module (MPM) is a complete microwave power amplifier contained within a volume of less than 20 in^3. This system consists of a micro-miniature helix TWT driven by an MMIC preamplifier, and a

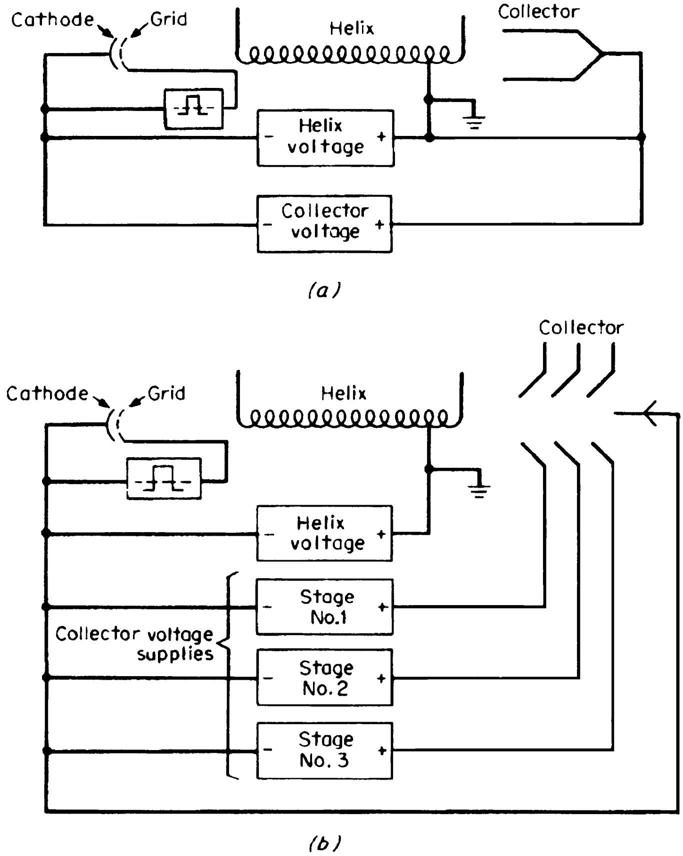

FIGURE 7.2.20 Power supplies for TWTs: (*a*) single collector TWT; (*b*) multistage depressed collector TWT.

high-density electronic power conditioner that powers both the TWT and the MMIC amplifier. Applications include electronic countermeasures, radar and communications transmitters. These modules may also be combined for use in shared aperture phased arrays.

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CROSSED-FIELD TUBES

George K. Farney

Crossed-Field Interaction Mechanism

A crossed-field microwave tube is a device that converts dc electric power into microwave power using an electronic energy-conversion process similar to that used in a magnetron oscillator. These devices differ from beam tubes in that they are potential-energy converters rather than kinetic-energy converters. The term *crossed field* is derived from the orthogonality of the dc electric field supplied by the source of dc electric power and the magnetic field required for beam focusing in the interaction region. Typically, the magnetic field is supplied by a permanent-magnet structure. These tubes are sometimes called *M tubes*.

The electronic interaction is illustrated schematically in Fig. 7.2.21. Electrons moving to the right in the figure experience electric field deflection forces ($f_e = -\epsilon E$) toward the electrically positive anode, while the magnetic deflection forces ($f_m = -\epsilon v \times B$) resulting from the motion of the negatively charged electron in the orthogonal magnetic field cause deflection toward the negative electrode. This electrode is also called the *sole*.

The forces are balanced when an electron is traveling in a parallel direction between the electrodes with a velocity numerically equal to the ratio of the dc electric field to the magnetic field ($v_e = E/B$). Any alteration of the electron velocity leads to an unbalanced condition. Reduction of the electron forward motion causes the magnetic deflection force to become less, and the electron trajectory is deviated toward the positive electrode. Conversely, an increase of velocity causes a greater magnetic deflection force, which causes trajectory deviation toward the negative electrode.

Electronic interaction with a traveling wave occurs when the positive electrode is an RF-guiding slow-wave circuit whose phase velocity for the traveling wave is numerically equal to the ratio of the dc electric field to the magnetic field ($v_p = E/B$). Under these conditions synchronous interaction occurs between the RF fields on the slow-wave circuit and the stream of electrons traveling in the interaction region.

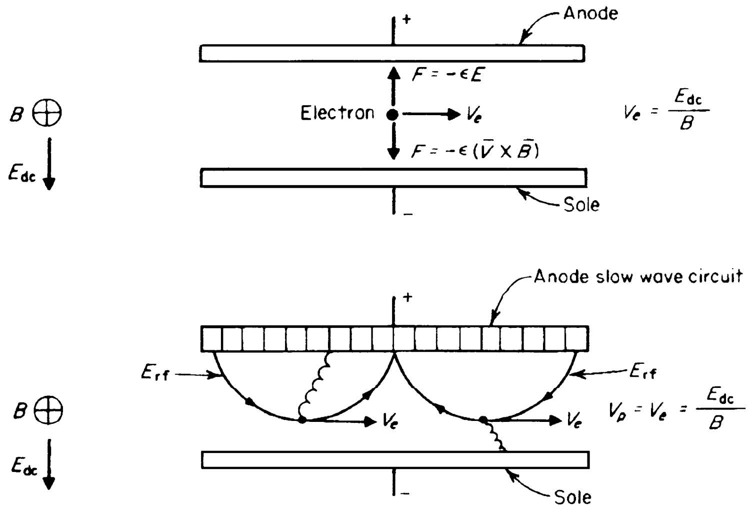

FIGURE 7.2.21 Forces exerted on a moving electron in a crossed-field environment.

Two general kinds of motion result, as illustrated schematically in Fig. 7.2.21, where a moving frame of reference is shown traveling from left to right at a velocity equal to the phase velocity of the circuit wave, so that the instantaneous RF fields are seen as stationary. The electronic motion resulting from interaction with the tangential components of the additional RF electric fields depends on the location of the electron relative to the phase of the RF fields of the slow-wave circuit. Those located so that their forward motion is retarded by the RF electric field are slowed, and the energy they lose is transferred to the RF wave on the circuit. These slower-moving electrons are subsequently accelerated toward the anode by the dc electric field, and their velocity is increased to the synchronous condition. The energy-exchange cycle can then be repeated. Electrons moving in this phase of the RF field pattern transfer energy to the RF wave on the circuit while maintaining nearly constant kinetic energy. The energy transfer results from the loss of potential energy of the electrons as they move to the anode.

Electrons located in the alternate phase of the RF field pattern are accelerated by the RF field and move away from the anode. The intensity of the slow wave decreases exponentially with distance away from the slow-wave circuit so that the magnitude of this interaction decreases. The result is the transfer of dc electric power to microwave power on the slow-wave circuit, with the phase-sorted electrons in the crossed-field interaction region providing the necessary coupling mechanism. Electron current thus flows to the anode only in the region of suitably phased RF electric fields.

The components of the RF field which are perpendicular to the forward motion of the electrons exert forces which phase-lock the electron near the center of the pattern. These regions are called *spokes* because of the similarity, in a magnetron oscillator, to the spokes in a rotating wheel. The phase locking of the sorted space charge relative to the traveling RF wave on the slow-wave circuit reduces the effect of power-supply variations on the electron trajectories. The details of the electron trajectories are extremely complex and have been calculated only approximately, using sophisticated computer techniques.

It is an important fundamental of crossed-field interaction that very high electronic conversion efficiency can be obtained because the kinetic energy of the electrons lost as heat upon ultimate impact with the slow-wave circuit can be designed to be a small fraction of the total potential energy transferred from the power supply. The ideal electronic conversion efficiency is given by $\eta = 1 - V_0/V$, where η is efficiency, V_0 is the synchronous voltage, and V is the cathode-to-anode voltage. Large ratios of V/V_0 lead to high efficiencies.

The crossed-field magnetron oscillator achieved prominence as a source of microwave power for radar applications during World War II. Since that time many kinds of crossed-field devices were investigated. One type of device obtains current from a thermionic cathode, electron gun located external to the crossed-field interaction space similar to that used in electron beam tubes. A second type uses electron current supplied by thermionic and/or secondary emission from a negative electrode facing the slow-wave circuit. This is similar to a cathode in a magnetron. These are called emitting sole tubes. Both types are illustrated schematically in Fig. 7.2.22.

Slow-Wave Circuits for Crossed-Field Tubes

Electron current in crossed-field interaction moves toward the slow-wave circuit rather than through the circuit as in beam tubes. This leads to the use of open circuits that present an RF waveguiding surface to the electron stream. Maximum energy conversion efficiency is usually obtained when the current is intercepted on the slow-wave circuit; so the structures must withstand the thermal stress associated with electron bombardment. Electronic interaction can occur using either forward-wave or backward-wave traveling-wave circuits, as well as with circuits supporting a standing wave. Examples of circuits suitable for use in forward-wave interaction are various *meander lines, helix-derived structures, bar* and *vane structures*, which are capacitively loaded by ground planes, and *capacitively strapped-bar circuits.*

A helix-coupled vane circuit and a ceramic-mounted meander line are shown in Fig. 7.2.23*a*. The most common backward-wave circuits are derivatives of the interdigital line and strapped-bar and vane circuits. Examples of a choke-supported interdigital line and a strapped-bar circuit are shown in Fig. 7.2.23*b*. Traveling-wave circuits are used mostly in amplifiers. Standing-wave circuits are resonant and used typically in magnetron oscillators. The most commonly used standing-wave circuits are composed of arrays of quarter-wave resonators that may or may not be strapped for improved oscillating-mode stability.

Variations of these circuits include *hole-and-slot resonators* and *rising-sun* anodes. Examples are shown in Fig. 7.2.23*c*. Cooling of vane structures for high average power is obtained by heat conduction along the vanes to the back wall of the anode to a heat sink, which may be liquid- or forced-air-cooled. Bar structures are cooled by passage of liquid coolant through the tubular bars of the slow-wave circuit.

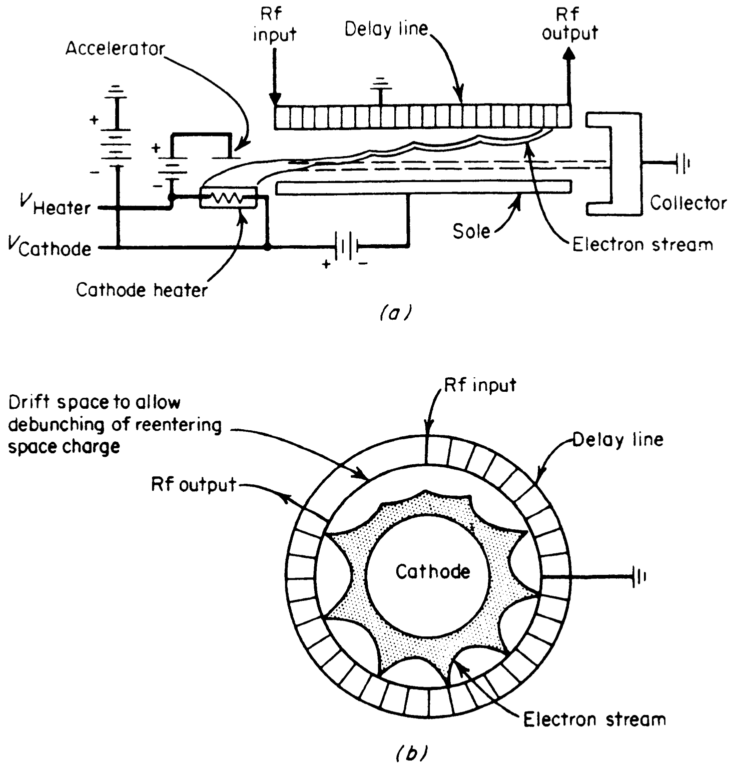

FIGURE 7.2.22 (*a*) Linear injected beam and (*b*) reentrant emitting-sole crossed-field amplifier.

During the time of intense research and development for crossed-field tubes, there were a large number of concepts under investigation. It was common to display a family tree of these to illustrate their similarities and differences. They were separated into two major groupings of injected beam and emitting sole devices. There were subgroups in each dependent on whether the device was an oscillator or amplifier, whether it used reentrant or nonreentrant electron streams and whether it used forward wave, backward wave or standing wave interactions with a slow-wave circuit. Many of these are no longer of general interest since some were not fully developed and others were made obsolete by newer technologies and/or by changes in performance requirements.

Magnetron oscillators are single-port devices. Both the slow-wave circuit and the electron stream are reentrant; i.e., the circular geometry is always used. Traveling-wave crossed-field oscillators are single-port devices but use a nonreentrant electron stream. They use either the linear or circular format.

Injected-beam and emitting-sole amplifiers are two-port devices with RF input and output ports. They are fabricated in both linear and circular format. Linear tubes must use a nonreentrant electron stream. Some circular-format amplifiers use a reentrant electron stream and some do not. Both forward-wave and backward-wave amplifiers have been developed.

Crossed-Field Oscillators

Conventional Magnetrons. The conventional magnetron is an emitting-sole, circular-format, reentrant-stream device with electronic interaction between the circulating current and a π-mode, RF standing wave on the slow-wave circuit. Oscillation builds up from noise contained initially in thermionic-emission current from

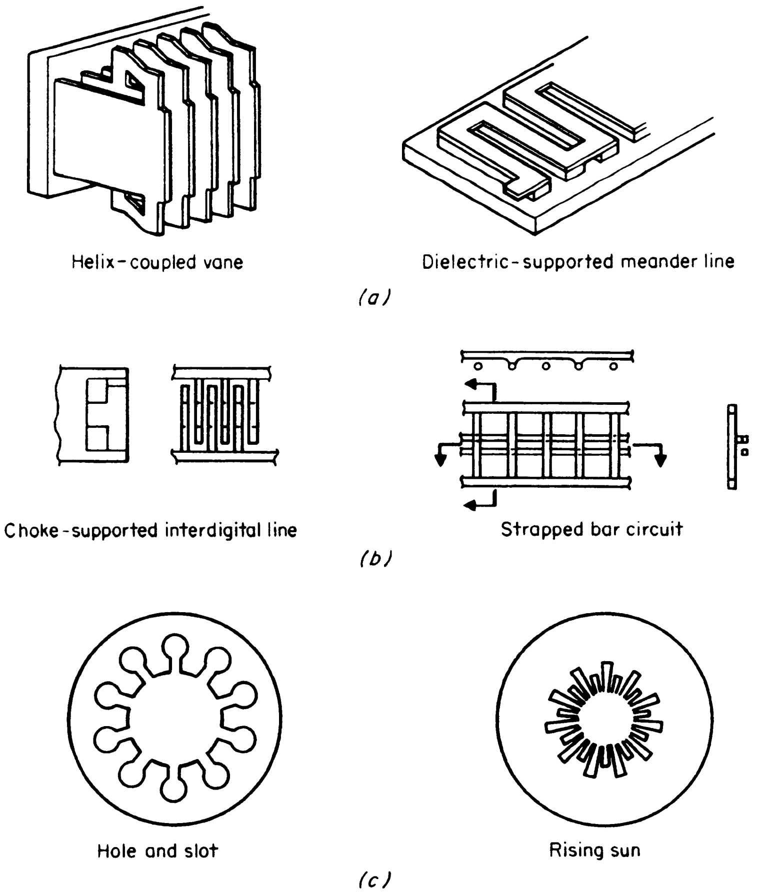

FIGURE 7.2.23 Slow-wave circuits for crossed-field tubes.

a heated cathode. During operation interaction current is obtained from a circulating hub of space charge supplied primarily by secondary electron emission from the cathode surface. This is illustrated in Fig. 7.2.24. Large peak currents are obtainable, permitting the generation of high peak power at lower voltages than are used for beam tubes of comparable peak power.

Pulsed magnetrons have been developed covering frequency ranges from a few hundred megahertz to 100 GHz. Peak power from a few kilowatts to several megawatts has been obtained with typical overall efficiencies of 30 to 40 percent, depending on the power level and frequency range. Continuous-wave magnetrons have also been developed with power levels of a few hundred watts, in tunable tubes, at an efficiency of 30 percent. As much as 25 kW cw has been obtained for a 915-MHz fixed-frequency magnetron at efficiency greater than 70 percent.

Pulsed magnetrons are used primarily in radar applications as sources of high peak power. Low-power pulsed magnetrons find applications as beacons. Magnetrons operate electrically as a simple diode, and pulsed modulation is obtained by applying a negative rectangular voltage pulse to the cathode with the anode at ground potential. Voltage values are less critical than for beam tubes, and line-type modulators are often used to supply pulse electric power. Tunable cw magnetrons are used in electronic countermeasure applications. Fixed-frequency magnetrons are used as microwave heating sources.

Mechanical tuning of conventional magnetrons is accomplished by moving capacitive tuners, near the anode straps or capacitive regions of the quarter-wave resonators, or by inserting symmetrical arrays of

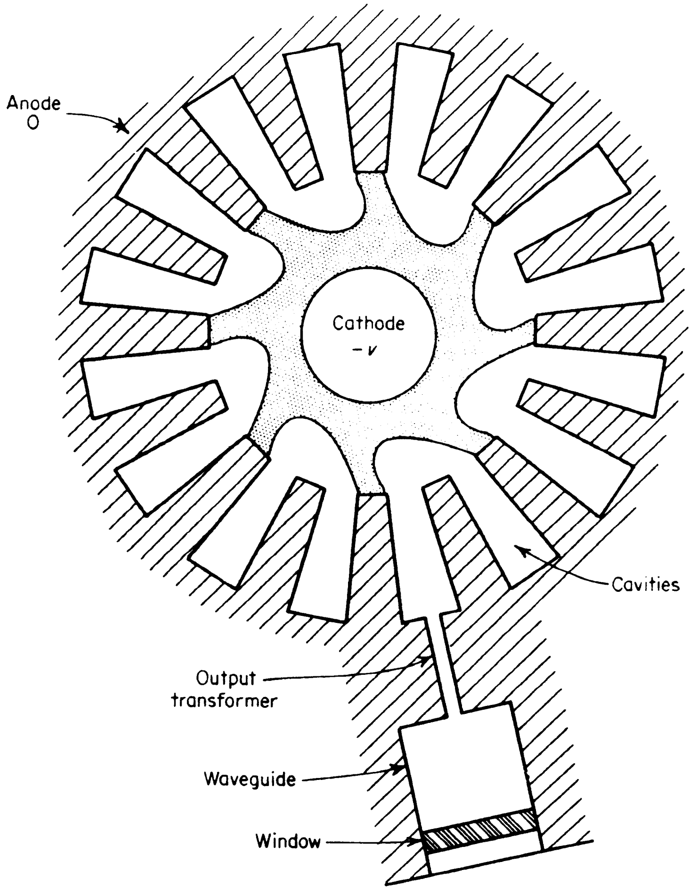

FIGURE 7.2.24 Conventional magnetron structure.

plungers into the inductive portions. Tuner motion is produced by a mechanical connection through flexible bellows in the vacuum wall. Tuning ranges of 10 to 12 percent bandwidth are obtained for pulsed tubes and as much as 20 percent for cw tubes.

Coaxial Magnetrons

The frequency stability of conventional magnetrons is affected by variations in the microwave load impedance (frequency pulling) and by cathode current fluctuations (frequency pushing). When the mode control becomes marginal, the tube may occasionally fail to produce a pulse. The coaxial magnetron minimizes these effects by using the anode geometry shown in Fig. 7.2.25. Alternative cavities are slotted to provide coupling to a surrounding coaxial cavity. π-mode operation of the vane structure provides in-phase currents at the coupling slots which excite the TE_{011} circular electric coaxial mode. The unique RF field pattern of the circular electric mode permits effective damping of all other cavity modes with little effect on the TE_{011} mode, and oscillation in other cavity modes is thereby prevented. Additional resistive damping is used adjacent to the slots but removed from the vanes to prevent oscillation in unwanted modes associated with RF energy stored in the vanes and slots that does not couple to the coaxial cavity.

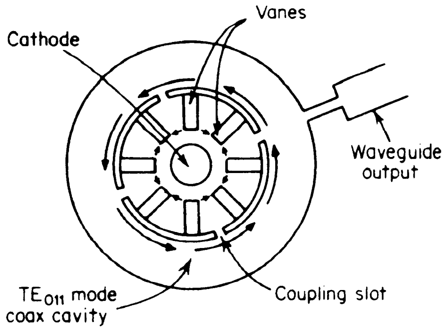

FIGURE 7.2.25 Coaxial magnetron coupling.

The oscillation frequency is controlled by the combined vane system and resonant cavity. Sufficient energy is stored in the TE_{011} cavity to provide a marked frequency-stabilizing effect on the oscillation frequency. Hence the coaxial magnetron is much less subject to frequency pushing and pulling than conventional magnetrons, and it exhibits fewer missed pulses. Tunable versions of this tube type are tuned by a movable end plate in the coaxial cavity similar to a tunable coaxial wavemeter. This is illustrated in Fig. 7.2.26. The larger resonant volume for energy storage leads to a slower buildup time for oscillation than in conventional magnetrons. This causes greater statistical variation in the starting time for oscillation (leading-edge jitter).

These factors are compared in Table 7.2.1 for the SFD-349 coaxial magnetron, which was designed as an improved retrofit for the 7008 conventional magnetron. The operating efficiency of the SFD-349 was deliberately degraded to meet retrofit requirements. Typically, coaxial magnetrons operate with an efficiency of 40 to 50 percent or higher.

Spurious Noise. The circulating space charge in the hub of both conventional and coaxial magnetrons contains wide-band noise-frequency components that can couple to the output. In conventional magnetrons this spurious noise can couple directly to the output waveguide. Spurious noise power measured in a 1-MHz bandwidth is typically greater than 40 to 50 dB below the carrier. The coaxial cavity in the coaxial magnetron provides some isolation between the spurious noise coupled to the vanes and the output waveguide. The spurious-noise power from coaxial magnetrons is typically 10 to 20 dB lower than conventional magnetrons of comparable peak power level.

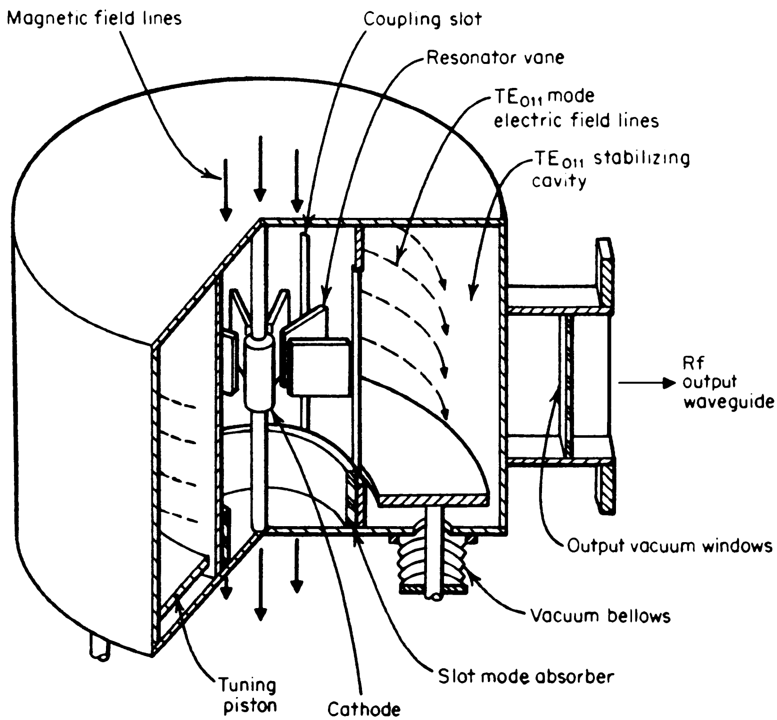

FIGURE 7.2.26 Schematic of coaxial magnetron.

TABLE 7.2.1 Comparison of the 7008 Magnetron and the SFD-349 Coaxial Magnetron

	7008	SFD-349
Efficiency, %	38	38
Leading-edge jitter, rms, ns	1.2	1.5
Pushing factor, kHz/A,	500	100
specified Typical	200	50
Pulling factor (VSWR 1.5), MHz	15	5
Spectra side lobes, dB	8–9	12–13
Missing pulses, %	1	0.01
Pulse-frequency jitter, rms, kHz	60	5
Life, h, specified	500	1,250
Typical	700–800	3,000–3,500

Frequency-Agile Magnetrons

To improve radar-signal detection and electronic countermeasures, rapid frequency-changing signal sources have been developed. Frequency-agile conventional magnetrons are available with rapidly rotating capacitive tuners (*spin-tuned magnetrons*) or hydraulic-driven, *mechanically tuned* tubes. The operational advantages of the coaxial magnetron are preserved in frequency-agile dither-tuned and gyro-tuned coaxial magnetrons.

Dither-tuned magnetrons use a mechanically tuned coaxial magnetron with an integral motor and *resolver* to provide high-speed, narrow-band frequency-agile operation. Mechanical linkage between the rotating motor and the tuning plunger provides approximately sinusoidal tuning of the magnetron frequency. Mechanical limitations imposed by acceleration forces determine the attainable tuning range and tuning rates. A voltage output from the resolver is made proportional to the magnetron frequency and is used to adjust the receiver local oscillator to track the rapidly tuned frequency of the magnetron. X-band tubes, 200 kW, with narrow-band dither-tuned frequency ranges of 30 to 50 MHz, are tuned at rates of 200 Hz. Wider-band frequency excursions of 250 to 500 MHz are dithered at rates of 25 to 40 Hz. Some of these tubes are equipped with servo motors for tuning. Frequency can be set electronically to provide rapid changes of fixed-frequency operation or can be dither-tuned with various shapes of frequency-tuning curves. These tubes are called *Accutune magnetrons.*

Gyro-tuned coaxial magnetrons use several rotating dielectric ceramic paddles in the stabilizing coaxial cavity, which cause frequency variation as they are rotated in a plane normal to the RF electric field of the TE_{011} mode. The anode vane system of the tube is surrounded by a ceramic cylinder bonded to the ends of the coaxial cavity to form the vacuum wall for the electronic interaction region. The stabilizing cavity, containing the tuning, is outside of this vacuum wall and is pressurized with sulfur hexafluoride, to inhibit arcing or corona caused by high RF fields. The tuner drive motor and frequency readout generator are also located within the pressurized section of the magnetron. The symmetry and inherently low rotational mass of the dielectric paddles result in a mechanism in which tuning speed and RF frequency excursion are essentially independent. It is therefore possible to attain higher tuning rates and relatively broader frequency excursion simultaneously than is achieved currently with dither tuning. K_u band, 60 kW peak power, gyro-tuned magnetrons obtain frequency excursions of 300 MHz at 200-Hz tuning rates.

Crossed-Field Amplifiers (CFAs)

Injected-Beam Types. Extensive development efforts have been devoted to nonreentrant forward-wave amplifiers (*TPOM*) for both cw and pulsed application. Continuous-wave amplifiers with a few hundred watts output have been developed with gain of 20 to 30 dB and with efficiency of 20 to 35 percent. Proper control of the electron stream at large values of gain requires the beam to be physically close to the anode at synchronism. Operation at low values of V/V_0 (3 to 6), together with increased insertion loss for tubes with greater circuit length for greater gain, leads to lower efficiency values. Like TWTs, the attainable bandwidth is dependent on the electron-beam optics and the dispersiveness of the slow-wave circuit. Half-octave bandwidth has been obtained at constant-voltage settings, and full-octave bandwidth has been obtained with adjustment of the anode-to-sole voltage. Stable operation at gain in excess of 20 dB requires circuit severs or distributed attenuation as in TWTs. Useful gain in excess of 30 dB is difficult because of excessive noise buildup in the electron stream.

Narrow-band (10 percent), pulsed, high-peak-power (5 MW) injected-beam amplifiers, which operate with efficiencies greater than 50 percent, were developed in France for use in radar application. These tubes have gain of 11 to 15 dB. The lower gain values (greater input signal level) permit the use of less critical beam optics, shorter slow-wave circuits, and greater ratios of V/V_0.

Injected beam CFAs were developed in the United States for electronic countermeasure applications, but the tubes were not widely deployed.

Emitting-Sole Crossed-Field Amplifiers

Electron current for emitting-sole crossed-field amplifiers is obtained from the sole electrode in the interaction space by electron-beam back bombardment, as in the magnetron oscillator. Unlike the magnetron, these amplifiers do not require thermionic emission to initiate current flow. Current flow can be started in an emitting-sole CFA by the admission of an RF signal to the input of the slow-wave circuit when the proper magnetic field and anode-cathode voltage are present in the interaction region. Amplifiers with RF-induced current flow are called *cold-cathode* amplifiers, regardless of cathode temperature, provided there is no thermionic emission from the cathode. In the absence of an RF input signal, these amplifiers remain quiescent even with full operational voltage applied. The details of the starting mechanism of RF-induced current flow are not fully understood, but the phenomenon is reliable in a properly designed amplifier.

Radio-frequency-induced current flow permits several modulation techniques for pulsed emitting-sole CFAs. These include cathode-pulsed CFAs, dc-operated CFAs with combination of dc voltage and a pulsed turn-off voltage, and dc-operated CFAs with only dc voltages applied. Cathode-pulsed amplifiers, as do magnetron oscillators, use pulse modulators to supply the required dc electric power during amplification. Direct-current-operated amplifiers obtain electric input power for amplification from a dc power supply. Electron current flow is initiated by an RF input signal and is terminated at the end of the RF input signal either by a voltage pulse or a dc bias voltage applied to a quench electrode.

Cold-cathode starting for cathode-pulsed amplifiers is assured by correct temporal alignment so that the RF input pulse bridges the cathode voltage pulse. The RF signal is present on the slow-wave circuit as the applied voltage pulse increases to synchronous value. Current flow is initiated, and amplification occurs during the voltage pulse and ceases upon removal.

Both forward-and backward-wave cathode-pulsed CFAs are available. They use the circular-formal reentrant-stream geometry illustrated in Fig. 7.2.27. A reentrant electron stream is advantageous because electrons that have delivered only part of their available potential energy to the circuit wave as they leave the interaction region can reenter for further participation, thereby leading to higher electronic conversion efficiency. (Overall amplifier efficiencies of 45 to 50 percent or more are common.) The leaving phase-sorted electrons contain RF modulation. Some reentrant-stream amplifiers use a circuit geometry with the RF input and output ports spatially separated by a sufficiently long circuit-free region (called the *drift space*) for internal space-charge forces to cause dispersal of the electron spokes. This removes the RF modulation from the reentrant stream while preserving the reentrant-stream efficiency advantage. Other reentrant backward-wave amplifiers (*Amplitrons*) use the modulated electron stream to create a regenerative amplifer. A minimal drift space is used, so that the electron spokes reenter the interaction region before the modulation is dispersed. By suitable design of the shorter drift-space dimensions, the modulated stream reenters with positive phase to enhance the interaction.

Greater electronic conversion efficiency (amplifier efficiency in excess of 70 percent has been obtained) can be obtained at the expense of lower gain-bandwidth product than can be obtained with amplifiers which remove the reentrant modulation. The use of regenerative amplification is not feasible with a forward-wave amplifier because, at a fixed operating voltage, the simultaneous regenerative gain at frequencies other than the drive signal could lead to unwanted auxiliary oscillations or selective peaks in spurious-noise output power. This is avoided with a backward-wave amplifier because the dispersive slow-wave circuits require different voltages to obtain adequate amplification of separated frequencies.

Spurious Noise. Dispersal by space-charge forces of the nonphase-locked electrons in a drifting spoke can lead to a rapid buildup of broadband spurious-noise components in the reentering electron stream. This is

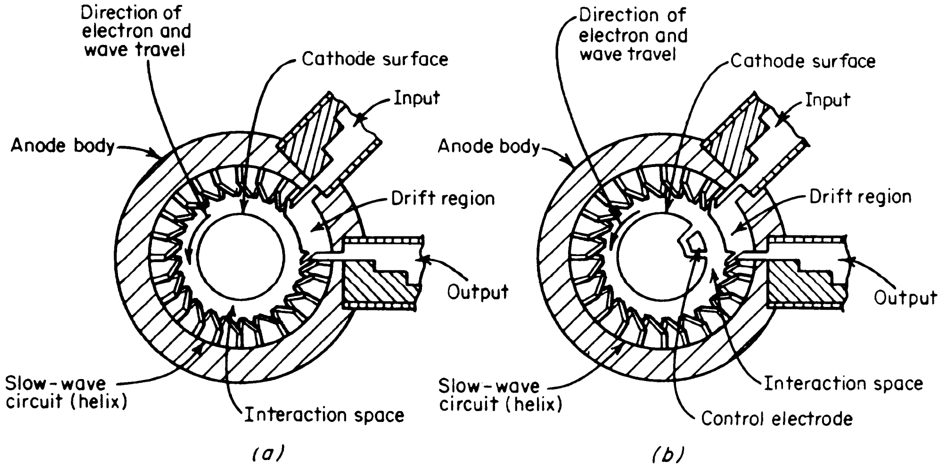

FIGURE 7.2.27 Diagrams of reentrant-stream crossed-field amplifiers: (*a*) cathode-pulsed; (*b*) with control electrode turn-off.

prevented from becoming severe by use of a sufficiently large RF input signal to lock out noise-signal growth. CFA power output as a function of RF drive signal is illustrated in Fig. 7.2.28. Reduction of the RF input signal leads to a reduced amplifier output signal at higher values of gain but is accompanied by an increased relative amount of broadband noise-power output. (Reentrant emitting-sole CFAs with terminated input have been used for high-efficiency broadband noise generators.) Adequate lockout of the noise power is obtained at reduced signal gain when the amplifier is driven well into saturation. The rapid growth of noise at small drive signals precludes the use of distributed attenuation and of circuit severs for emitting-sole amplifiers.

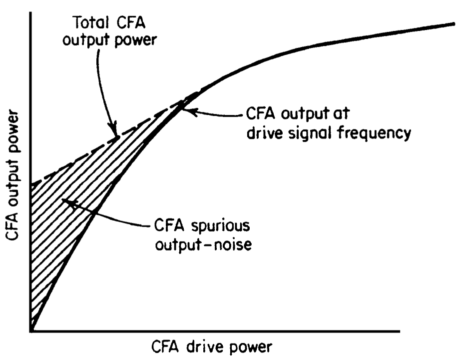

FIGURE 7.2.28 Emitting-sole crossed-field power output.

Relatively short slow-wave circuits are used with the minimum attainable insertion loss between the RF input and output connections. These amplifiers are called *transparent tubes*. Reflected signals from a load mismatch travel backward through the amplifier to the RF input, where they can be reflected, possibly leading to oscillation. Judicious use of ferrite isolators and circulators at the RF input and/or output minimizes this effect. The overall stable gain of emitting-sole CFAs is limited to 20 dB or less (typically 13 to 15 dB) because of the requirement for a sufficiently large RF input signal for adequate lockout of spurious-noise power and the need to limit gain to avoid oscillations caused by multiple reflected signals. Transparency is often used to advantage in radar systems employing the final amplifier in the transmitter chain in a nonoperating feedthrough mode to provide coarse programming of the output power level.

Bandwidth Characteristics. Cathode-pulsed forward-wave amplifiers offer 10 to 15 percent instantaneous bandwidth at a fixed value of pulsed voltage. Backward-wave amplifiers provide only 1 to 2 percent instantaneous bandwidth under comparable conditions but can accommodate 10 percent bandwidth by adjustment of cathode voltage. The static impedance of both tube types varies as a function of frequency. The constant voltage versus frequency characteristics of forward-wave amplifiers is readily accommodated by a hard-tube cathode modulator, providing nearly constant power output across the frequency band of the amplifier. Constant power output versus frequency for a variable-voltage backward-wave amplifier is nearly achieved with a constant-current modulator. For restricted bandwidth (4 to 6 percent) this condition is approximated by using a line-type modulator.

Modulation Requirements. A simplification in modulator requirements is obtained with a broadband, dc-operated, RF-triggered CFA. The termination of the RF input signal after RF turn-on leaves uncontrolled circulating space charge that can generate a cw spurious output signals of magnitude as large as half the amplified signal output. To avoid this, a *control electrode* (also called quench electrode) isolated from the cathode is mounted as part of the cathode structure and is located in the drift space. This location minimizes interference with amplifier performance (Fig. 7.2.27*b*). This electrode is pulsed positive with respect to the cathode, coincident in time with removal of the RF input signal. The circulating space charge in the interaction space is collected upon the control electrode, and the cathode current flow is terminated. Modulator requirements are simplified because the pulse voltage required for the turn-off electrode is typically one-quarter to two-thirds of the anode-cathode voltage and the peak collected current is less than one-half of the peak cathode-current flow during amplification. The duration of the collection time for the circulating current is approximately one transit time for electron flow around the interaction region. Typically, this is a small fraction of the time duration of the amplified signal. Consequently, the modulator energy required for the control electrode per amplifier pulse is much less than that required from the modulator for full-cathode-pulsed amplifiers.

Secondary Emission Cathodes. A variety of materials are used for secondary-electron-emission cathodes in RF-triggered amplifiers. The selection is based on the amplifier drive signal level, peak power output, and the intended operating voltage for the tube. Materials used include pure metals, such as aluminum, beryllium, and platinum, as well as a variety of composite materials, such as dispenser cathodes and cermets. Dispenser cathodes and pure-platinum cathodes are suitable for drive signal levels in excess of 10 kW. Amplifiers with drive signals from a few hundred watts to 10 kW are better accommodated with metals supporting oxide surface layers such as aluminum and beryllium. Oxide layers are susceptible to erosion under electron-beam bombardment, and some CFAs employ a low-level background pressure of pure oxygen supplied from a suitable reservoir to rejuvenate and extend the active cathode life.

Frequency range. Emitting-sole amplifiers have been developed at frequency ranges extending from VHF to K_u band, with experimental models at lower and higher frequencies. Examples of peak power levels available include 100 kW at K_u band, 1 MW at X band, and 3 MW at S band. Average power levels vary from 200 W at K_u band to several kilowatts at lower frequencies. Laboratory models have demonstrated as much as 400 kW of cw power at S band.

Noise Power. Noise-power output from reentrant stream emitting role CFA's with a space charge dispersing drift space measured in a 1-MHz bandwidth far from the carrier frequency is typically greater than 35 dB below the carrier power level. Noise levels greater than 45 dB below the carrier are not uncommon. Many radar applications require very low noise power between spectral lines of a pulsed signal. Much effort has been expended in recent years for further reduction in noise power. Signal-to-noise ratios (S/N) close to the carrier frequency are now exceeding 50 dB_c/MHz. In some experimental tubes, S/N ratios of greater than 65 dB_c/MHz have been measured, but this performance is not now obtainable across a full operating band by production tubes. Phase locking of the space charge by the drive signal also minimizes phase variation due to voltage change and drive signal variation. Saturated amplifiers with 12 to 15 dB gain have output phase variations of 3 to 8° for a 1 percent change in anode/cathode voltage. A comparable phase change occurs with a 1 dB variation in drive-signal level.

Applications. The primary use for emitting-sole CFAs is for transmitter tubes in coherent radars. CFAs have been used in pulse compression radars and pulse-coded and phased-coded radars, as well as phased-array radars. Lightweight low-voltage cathode-pulsed amplifiers are attractive for airborne applications. High-power dc-operated CFAs are used in ground-based radars.

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CYCLOTRON RESONANCE TUBES (GYROTRONS)

Howard R. Jory

Introduction

Gyrotrons or cyclotron resonance masers are a class of microwave generators that make use of the cyclotron resonance condition to couple energy from an electron beam to a high-frequency electromagnetic (em) field. This type of coupling allows the beam-wave interaction region to be large compared to a wavelength. As a result, gyrotrons can produce orders of magnitude higher power at a given frequency compared to other microwave devices without exceeding power density limits.

The basic equation for cyclotron resonance coupling to an em field stationary in space is given by

$$\omega = \frac{neB}{\gamma m_0}$$

where ω = operating frequency
B = applied dc magnetic field
e/m_0 = charge to mass ratio for the electron
γ = relativistic mass factor
n = an integer

The strongest coupling occurs for the fundamental resonance where $n = 1$. Harmonic interactions where n has larger values are generally progressively weaker unless some special boundary geometry is used to shape the em fields. Second harmonic coupling can be very effective and harmonic interactions as high as $n = 12$ have been demonstrated.

The coupling equation illustrates one of the limitations of gyrotrons in that the required magnetic field is proportional to the frequency of operation. Operation at 30 GHz with fundamental resonance requires a magnetic field of about 1T, which is near the limit of practical magnets based on room temperature copper technology. Higher-frequency gyrotrons generally employ superconducting magnets that require a supply of liquid nitrogen and liquid helium, but negligible electrical power.

In its simplest form, the cyclotron resonance interaction requires only electrons making circular orbits in a plane perpendicular to a dc magnetic field, and a time-varying electric field also in a direction perpendicular to the magnetic field. In a practical embodiment, the electrons will also have a component of axial velocity and the electric field will have both transverse variations in amplitude and possibly phase. The cyclotron resonance interaction is quite flexible. It can be realized with a single cavity (gyrotron), multiple cavities (gyroklystron), or traveling waves (gyro TWT). It can use simple cylindrical cavities using any TE (transverse electric) modes or quasi-optical cavities formed by mirror reflectors.

Gyrotron Oscillators

Most gyrotron oscillators have been built with the configuration shown in Fig. 7.2.29. The cavity where the interaction with the electron beam takes place is cylindrical with some tapers or steps in diameter to control the axial variation of electric field amplitude. The diameter can range in size from 1 to 20 wavelengths. The axial length is typically in the range of 5 to 10 wavelengths.

In the simplest case, the microwave output from the cavity propagates axially through the beam collector and out through a window at the top of the gyrotron. Since all of the structures are large compared to a wavelength, many waveguide modes could propagate. However, it is generally important to have all of the output power contained in a single waveguide mode. Therefore, care must be taken in designing the tapers in diameter such that mode conversion does not occur. Generally it is possible to achieve an output where 90 to 95 percent of the power is in a single mode.

The beam collector area must be large enough to avoid excessive beam impact density. A typical design value for long pulse or CW operation is 1 kW/cm^2. Hence a collector for a 1-MW beam should have a beam impact area of 1000 cm^2 or more. Collector magnet coils are often used to help distribute the beam in the collector.

Desirable properties for the output window are low microwave loss, high strength, and high thermal conductivity. Typical materials used are alumina, sapphire, berrylia, and boron nitride. Pulsed gyrotrons generally use single disc windows cooled only at the edges of the disc. CW gyrotrons normally use two discs with a modest axial space between the discs through which a dielectric cooling fluid flows. In this way one face of each window disc is cooled, and the window can transmit much higher CW power without overheating.

The type of electron gun that has been most successful for gyrotrons is the magnetron injection gun. The desired electron beam is one where electrons have helical motion. In the cavity region the electrons should have a transverse to axial velocity ratio of the order of 2. Then about 80 percent of the beam kinetic energy is transverse motion, which can couple in the cyclotron resonance interaction. The axial velocity in this case serves only to determine the transit time of the electrons in the cavity and to keep electron space charge fields to a reasonable value. The magnetron injection gun, combined with magnetic compression of the beam between the cathode and the cavity, effectively provides a good beam for interaction.

Figure 7.2.30 shows a simulation of the electron motion in the gun as well as through the gyrotron to the collector. The cathode (emitting portion) of the gun is a section of a cone. The cathode and its associated support structures operate below ground potential at the full beam voltage desired (typically 20 to 80 kV). The simulation shows the case where the conical anode of the gun is at ground potential along with the body and collector. This provides a large radial component of electric field at the cathode which results in a component of transverse velocity for the electrons. The simulation shows a projection of the electron motion on the *R-Z* plane. The true motion of the electron is helical. Figure 7.2.29 shows a version of the gun which has an intermediate electrode operated at a potential between cathode and ground. This electrode can be used to control the electric field at the cathode and, therefore, control the ratio of transverse to axial velocity in the final beam at the cavity.

The gyrotron will oscillate when the main magnetic field is adjusted so that the beam cyclotron resonance frequency and a resonant frequency of the cavity are nearly equal and when the electron transverse-to-axial velocity ratio is high enough. Since a cavity that is large compared to a wavelength will have many resonances spaced perhaps a few percent apart in frequency, it is often possible to step tune from one frequency to the next by changing the main magnetic field value. This has been demonstrated at MIT with a pulsed gyrotron covering a 2 to 1 frequency range with steps about 5 percent apart in frequency. High power CW gyrotrons, however, are generally limited to single frequency operation because of bandwidth limitations imposed by double disc windows and collector tapers. As the main magnetic field is varied by a few percent about a normal operating value the oscillator frequency will change slightly and the output power will vary. The frequency variation is related to the Q of the cavity and typically has a value of 0.1 percent. The output power variation can be of the order of 10 to 1.

Output power can also be varied over a range of the order of 10 to 1 by changing the magnetic field at the cathode or by changing the gun anode voltage, if the design includes an intermediate anode. Beam voltage can be used to change power output over a range of about 2 to 1. Beam current can be used to change power output by the order of 10 to 1 also. In the magnetron injection gun beam current is controlled by changing the cathode heating power. The cathode is operated in the mode where emitted current is limited by temperature.

Peak operating efficiencies for gyrotron oscillators are in the range of 30 to 50 percent. When output power is varied by changing beam current the variation of efficiency will be minimized. When one of the other means discussed above is used, the efficiency will be reduced proportionally to the power output.

The high-power levels of gyrotrons require the use of oversize waveguides at the output. Various modes have been used such as TE_{01} or TE_{02} or whispering gallery modes such as $TE_{12,1}$ or $TE_{15,2}$ or $TE_{22,2}$. Waveguide diameters, large compared to a wavelength, are required to avoid breakdown as well as excessive loss. Typical sizes are 1 to 4 in. in diameter. Systems have been built, which transmit power for 50 to 100 m length with losses as low as 5 percent.

As gyrotron frequencies and power levels increase, it becomes particularly advantageous to separate the output waveguide from the beam collector. Figure 7.2.31 shows a technique to accomplish this, which has been used in Russia for about 10 years, and is now also being used in other countries. From the electron gun up through the output taper there are no essential changes. Above that point, the wall of the waveguide is cut in an appropriate way to cause the microwave power to radiate in the direction shown by the arrows. Perturbations in the wall of the waveguide just before the cut can be used to better control the radiation pattern. The radiated power is handled by a number of mirrors to form a Gaussion-like beam which passes through a window on the side of the gyrotron. The electron beam is confined magnetically to pass through the output coupler region and then is allowed to expand to hit a conveniently large collector structure. The output microwave beam can be transmitted further using quasi-optical techniques or by injection into an HE_{11} type of low loss waveguide.

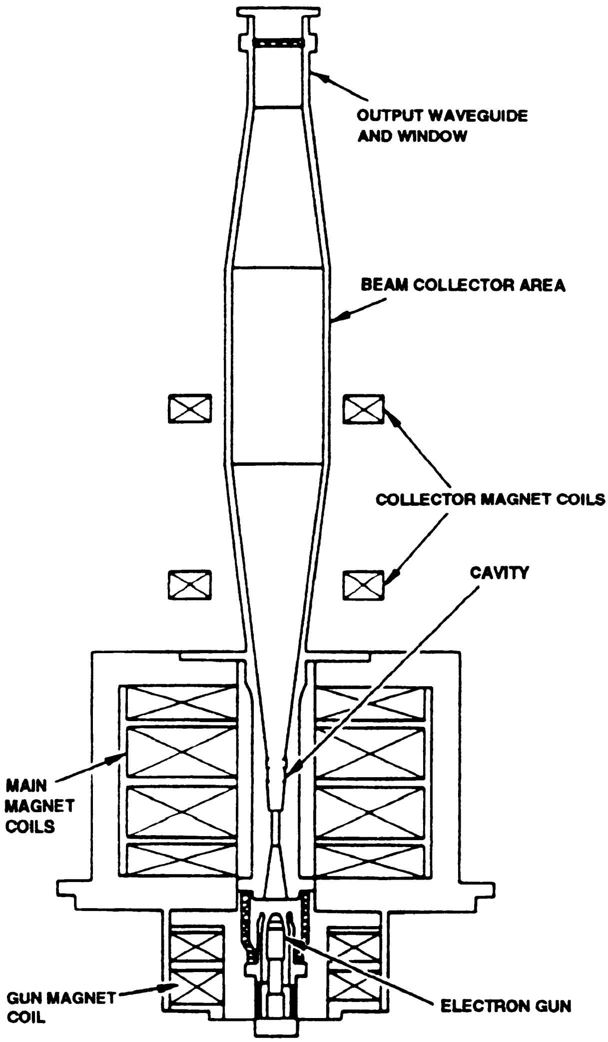

FIGURE 7.2.29 CW gyrotron with an axisymmetric RF output.

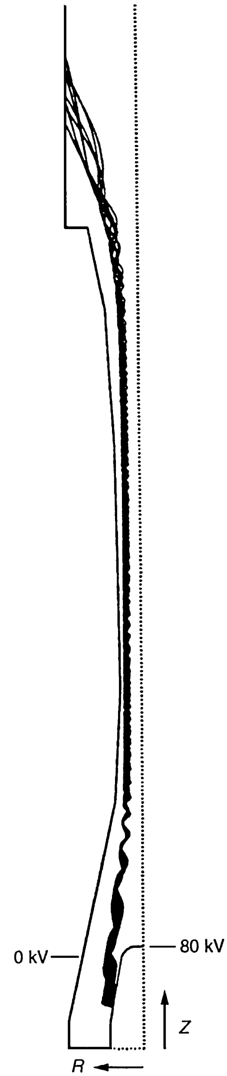

FIGURE 7.2.30 Electron trajectories in gyrotron.

Coaxial interaction cavities have produced good results recently as a means of achieving mode separation in very large cavities for higher power output. Coaxial modes such as the $TE_{28,16}$ have been used.

Gyrotrons for Fusion Applications

The gyrotrons used in connection with magnetic fusion have been mainly in the frequency range from 8 to 140 GHz. They have usually been operated under pulsed conditions with pulse lengths of 50 ms to 5 s and duty factors less than 1 percent. The measure of importance in this application is the output power in joules per pulse. Figure 7.2.32 shows the capability of various gyrotrons as a function of frequency. For each data point the first number gives the power output in MW and the second number the pulse length in seconds. Also given

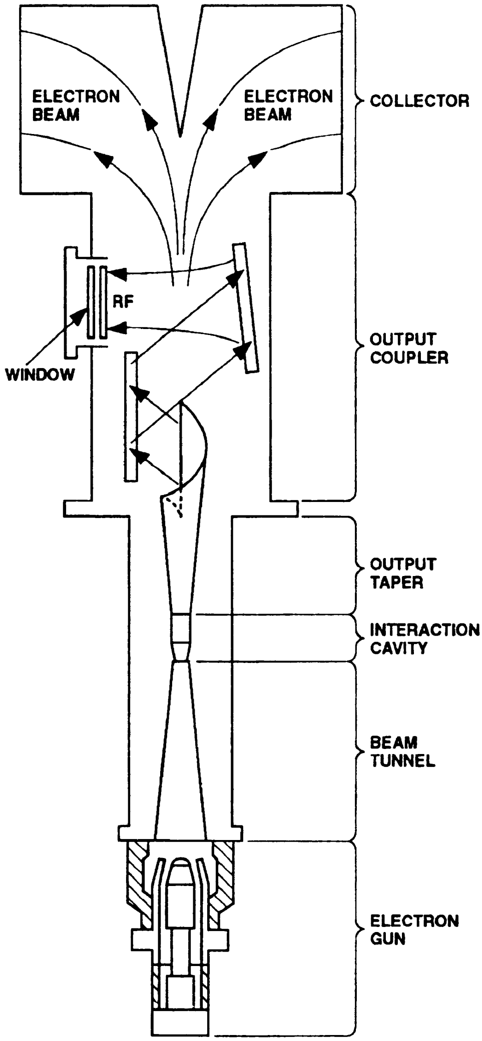

FIGURE 7.2.31 Gyrotron with coupler to separate RF output.

are the year of achievement and group involved. The dotted curves are for constant product of energy *x* frequency squared, which is a rough measure of technical difficulty.

Development programs are currently in progress in several countries to produce 1 MW at 110 GHz with pulse lengths of several seconds or CW. There is also development work for 1 MW at 170 GHz which corresponds to the current plan for the International Thermonuclear Experimental Reactor (ITER) project.

Other Gyrotrons

Gyrotrons have been built for operation at frequencies as high as 850 GHz with outputs at watt levels. Near 300 GHz, CW power of tens of watts has been produced and pulsed power of hundreds of kW. Some of these have been developed for plasma diagnostics or spectroscopy applications.

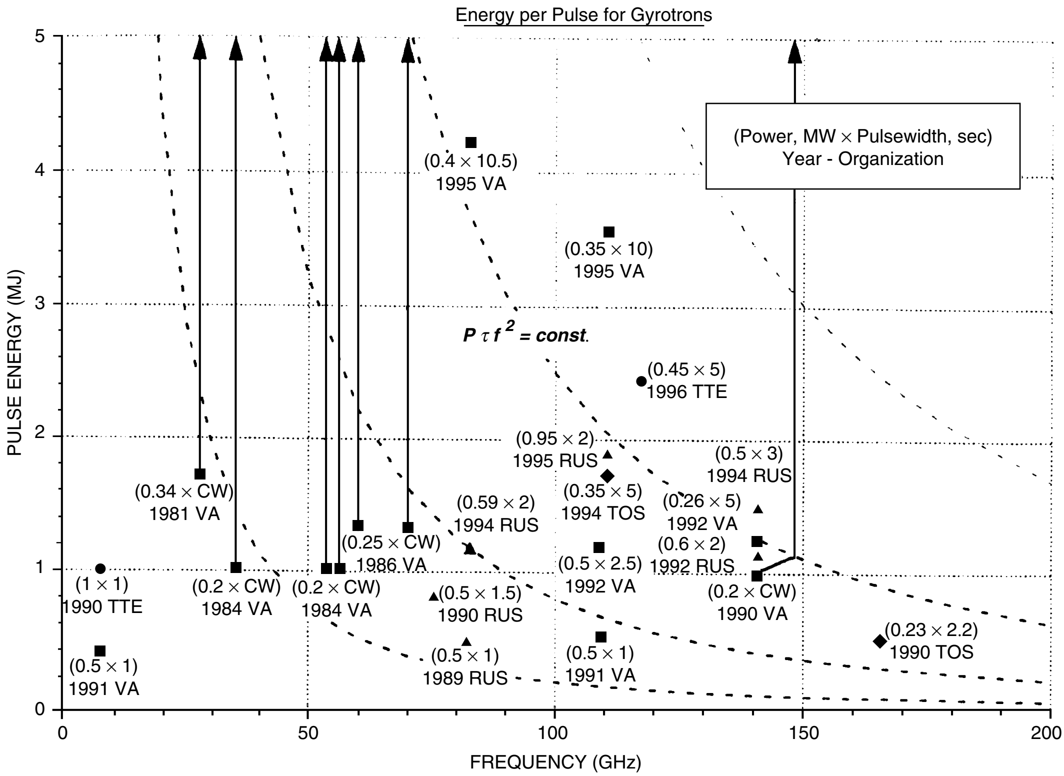

FIGURE 7.2.32 Energy per pulse for gyrotrons "VA = Varian/CPI, Rus = Russia, TTE = Thomson, TOS = Toshiba (data points for CW gyrotrons shown for 5-s pulse).

Gyrotrons using quasi-optical (confocal) resonators have been built near 100 GHz. Pulsed power outputs of hundreds of kW have been achieved but with efficiencies lower than the conventional cavity gyrotrons. Typical values are 10 to 15 percent.

Short-pulse, high-peak-power gyrotrons have been built based on cold-cathode flash x-ray technology. Outputs of hundreds of MW have been achieved at frequencies of 10 to 35 GHz. Lower power levels have been demonstrated up to 100 GHz. Typical pulse lengths are less than 100 ns. Beam voltages up to 1 MV and currents up to 10 kA have been employed. Typical efficiencies are 5 to 10 percent. The cold-cathode technology limits these devices to short pulse length and relatively short life compared to hot cathode devices. Repetitive pulsing has been demonstrated.

Gyroklystrons

The gyrotron can be made into an amplifier by using two or more cavities along the axis of the beam. The cavities are shorter to avoid self-oscillation. As in the conventional linear beam klystron, the RF input to the first cavity modulates the beam. But in this case it is an angular velocity modulation. As the beam drifts, the angular velocity modulation converts to bunching in angle. Intermediate cavities intensify the modulation and the output cavity extracts angular kinetic energy from the beam.

The most impressive results with gyroklystrons have been produced at the University of Maryland where 27 MW output at 9.85 GHz with μs pulse length was achieved. Beam voltage was 440 kV and efficiency 32 percent. This device was also modified to extract power with a second harmonic interaction in the output cavity. The results were 32 MW at 19.7 GHz with 29 percent efficiency.

A gyroklystron built by Varian demonstrated pulsed output of 75 kW at 28 GHz with 41 dB gain and 9 percent efficiency. Gyroklystrons have been built in Russia at XBand and other frequencies. For example, 60 kW pulsed output at 94 GHz with 34 percent efficiency and 40 dB gain is reported.

Major issues with gyroklystrons are stability and efficiency. The beam tunnels between cavities can be nonpropagating for the normal cavity modes but will usually propagate for lower-order modes. Mode conversion and reflection back through the beam tunnel can easily result in spurious oscillation. This problem is generally controlled by careful cavity design and the inclusion of absorbing material in the beam tunnels.

Bandwidth of gyroklystron amplifiers is in the range of 0.1 to 1 percent because of the high cavity Q's required. Stagger tuning of cavities might allow small increases.

Gyro Traveling-Wave Tube

The gyro TWT is similar to the gyroklystron except that the resonant cavities are replaced by traveling-wave circuits. Stability in this case is even more difficult. In addition to spurious oscillations caused by reflections at the ends of each circuit, there are potential instabilities at the cutoff frequency of the traveling-wave circuit and with beam interactions with the backward circuit wave.

Nevertheless, reasonable results have been demonstrated for gyro TWTs. An experiment in Taiwan produced pulsed output of 25 kW at 35 GHz with 23 percent efficiency and gain of about 20 dB. Varian pulsed gyro TWTs have operated at 95 GHz with 30 kW output 16 dB gain, 8 percent efficiency and 2 percent bandwidth and at 5 GHz, 120 kW output, 16 dB gain, 26 percent efficiency, and 6 percent bandwidth.

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