PRODUCT INFORMATION

SAFETY PRECAUTIONS

WARNING: WITHOUT PROPER AND ADEQUATE PRECAUTIONS, THE OPERATION, HANDLING, OR SHIPMENT OF MICROWAVE AND HIGH-VOLTAGE ELECTRONIC TUBES CAN BE HAZARDOUS TO PERSONNEL AND PROPERTY. READ THE FOLLOWING INFORMATION. TAKE ALL REQUIRED PRECAUTIONS.

GENERAL

This information is provided to alert the purchaser of high-voltage tubes and microwave tubes to the potential hazards which may be created by improper operation, handling or shipment of these devices. All persons responsible for the operation, handling and shipping of these tubes should familiarize themselves with the potential hazards, and suitable safety precautions should be established and followed for the protection of personnel and equipment.

Do not operate high-voltage and microwave tubes except in accordance with adequate understanding of the potential hazards and with proper equipment-operating instructions and safety precautions.

Questions regarding proper and safe use of such tubes should be addressed to:

General Electric Company
Microwave Tube Operation
Building 269 – Application Engineering
1 River Road
Schenectady, New York 12305

Several of the potential hazards are defined and regulated by state or federal governmental agencies and bureaus. Since the documentation and specifications of such agencies and bureaus are frequently revised, it is not feasible to make full or precise reference to their content in this publication. If current governmental information is desired or if there are questions, the appropriate agency should be consulted.

HIGH VOLTAGE

The voltages used to operate microwave and high-voltage electronic tubes can cause death or serious injury due to electric shock and burns. Depending on the device and equipment designs, and considering the possibility of malfunctions in either, part or all of the exterior tube surfaces may be at, or may quickly reach, dangerous voltages. Equipment design and laboratory testing must take this into account by following design and operating precautions so that contact with, and proximity to, high-voltage circuits is not possible under operating conditions. High-voltage circuits should be enclosed in protective housings, and interlock circuits should be provided so that primary power is removed, and high-voltage terminals and capacitors are quickly grounded, whenever the enclosure is open. It is always dangerous and unsafe practice to defeat or avoid the proper safety devices and safety procedures (as bypassing an interlock circuit) while operating or testing the equipment.

GROUNDS

Many microwave tubes are operated in a grounded electrode mode in which the envelope and output cables are operated at ground potential. Care must be taken to be certain that the tube envelope is properly grounded before the operating voltages are applied. The grounding should never be done through the output cables since a break in the cable will then result in the tube envelope being raised to high voltage.

X-RADIATION

X-radiation is produced by the impact of high energy electrons on electron tube surfaces. Such high-energy electrons are produced when accelerated by the applied electrode voltages. Depending on the construction of the electron tube and the materials involved, X-radiation may be produced at voltages as low as 5 kilovolts. The production of highly penetrating X-radiation and energy increases to relatively more dangerous proportions as the electrode voltages and currents are increased. All electron tubes operating in high-voltage ranges constitute potential hazards, and applications of such tubes should be carefully reviewed before operation.

When X-radiation shielding is required, it should be provided with proper interlocks to prevent accidental exposure of personnel to X-radiation. Where hazards are high, periodic X-radiation level surveys should be made. Furhter, when continuous operation is in effect, personnel-monitoring devices should be worn by the personnel and controlled access to the area implemented.

Most high-voltage and microwave electronic devices are not designed, nor intended, to be fully self-shielded to X-radiation under all possible conditions of their application and use. External radiation shielding will usually be necessary. This shielding should be designed by the equipment manufacturer as a part of the user's equipment to protect the user against possible personal injury. It is the responsibility of the manufacturer of the equipment using such tubes to provide any and all enclosures required, and to provide the instructions and maintenance procedures for the proper use of the equipment.

Generally, the spatial distribution of X–radiation from power tubes is complex and changes from tube to tube. The same tube does not radiate the same 360^{o} around. Also, the surrounding metallic construction will tend to prevent, distort, or further filter the passage of X–radiation to regions external to the tube. Of major concern are the areas in which materials used in tube construction present the least attenuation of X–radiation.

The search for possible X–radiation is not to be confined to those directions in which emission may be expected; unintended emissions in high power tubes have sometimes caused X–radiation in unexpected directions. A thorough search in all directions around the tube is necessary to ensure that the regions of emissions is correctly determined.

Tubes presenting X–radiation hazards or other possible hazards will have radiation precaution labels or tags affixed to the device at the time of shipment. These should not be removed at any time. If these labels or tags are removed by the user, they should be prominently displayed in close visual proximity to the device.

MICROWAVE RADIATION

The radio-frequency output power of many electron tubes may exceed those power densities considered safe for human exposure. The design, operating instructions, and maintenance procedures of equipment utilizing such tubes must ensure that the radio-frequency energy is properly restricted to and contained in the circuits, transmission lines, waveguides, or cavity resonators and that these are frequently monitored to ensure that the radiation of radio-frequency energy from joints or connectors is below the hazardous limit. Antenna systems should also be frequently monitored for stray or indirect radiation. Operating and service personnel should be advised of exposure hazards and arrangements made to prevent accidental exposure.

MERCURY

Some devices contain mercury as a necessary constituent to their operation. Under certain circumstances, the presence of free mercury may generate air contamination or other pollution that is considered toxic. Disposal of tubes or handling of damaged tubes must be done with adequate precaution given to this possible hazard. If disposal presents questions, these questions should be directed in writing to the General Electric Company, Microwave Tube Operation, at the address shown on the front side of this sheet.

Air shipment regulations allow air transportation of devices containing mercury only under special packing and marking requirements. The current requirements should be obtained directly from the airline.

The packing containers of devices containing mercury will be marked accordingly when they are shipped from the tube manufacturer.

IMPLOSION

Most electronic tubes and devices operate with their internal volumes under high vacuum, and many gas-filled tubes also have their internal volumes considerably below atmospheric pressure. In the event that the envelope of some of these tubes is punctured or broken, the inrush of air can be violent under certain conditions. Tubes with large glass envelopes should be handled and stored with particular care, and implosion-proof shields should be installed in operating equipments. Particular care should also be given to shielding of the eyes and face.

MAGNETIC FORCES

The attractive force between magnetic and feromagnetic objects increases rapidly as separation between the objects is decreased and the objects will be accelerated toward one another, meeting with considerable impact. When handling or working near large permanent magnets, care must be taken to prevent injury which could result from this hazard.

Air shipment regulations allow air transportation of devices containing magnetized materials only under special packing and marking requirements. The current requirements should be obtained directly from the airline.

PRECAUTIONS TO BE OBSERVED IN TESTING HIGH-FREQUENCY PLANAR TUBES

Introduction - Testing of close-spaced, high-performance, high-frequency planar tubes presents difficulties that may be overlooked and may account for misleading results or damage to the tubes being tested. Many commercially avilable tube checkers are not satisfactory for checking these tubes, and an effort should be made to determine if the checkers meet the requirements listed below before they are used.

Short and Leakage Tests - When grid-to-cathode leakage and shorts are checked, the maximum voltage applied between grid and cathode should be 100 volts, which with the grid negative with respect to the cathode. Some checkers use a neon bulb in series with an a-c source and a capacitor to check for shorts and leakage, and apply peak-to-peak voltages as high as 250 volts between grid and cathode. This type of circuit can indicate shorts and leakage when it should not, and its use may permanently damage the tube being tested.

Test Conditions - In order to obtain values of plate current and transconductance comparable to those listed on the tube data sheets as "Initial Characteristics Limits", it is necessary that the tubes be tested under the conditions given on these sheets. This includes using the indicated values of heater voltage, plate voltage, and grid voltage.

Oscillation - When high-Gm tubes are tested, radio-frequency tank circuits are often formed by the leads external to the tube, and oscillation often results. This oscillation will give misleading results and is usually manifest by variations in plate current as leads external to the tube are moved or a hand is brought near the tube. This oscillation can usually be stopped with chokes and bypass capacitors at the test socket.

Cooling - It is important that the envelope temperature rating is not exceeded during testing. If testing is prolonged, some means of cooling may be required. This may be accomplished by means of a heat sink or with forced air.

Sockets for Testing - Sockets suitable for use in fabricating adapters, and complete adapters for some tube types, may be obtained from several socket manufacturers. The following manufacturers may be contacted for information on sockets and adapters:

Community Engineering Corporation
State College, Pennsylvania

Instruments for Industry, Inc.
101 New South Road
Hicksville, New York

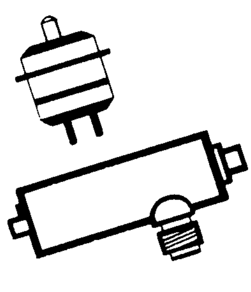
Jettron Products, Inc.
56 Route 10
Hanover, New Jersey

In Case of Difficulty - If your results in testing planar tubes are unsatisfactory, contact your General Electric Sales Representative, giving details of your test.

Prepared and distributed by Technical Data Unit, Receiving Tube Engineering, Owensboro, Kentucky, on the basis of information supplied by Mr. S. E. Peach of Application Engineering.

Performance and application of Microwave Gridded Vacuum Tubes and Microwave Circuit Modules

to meet the challenge of the new generation.



GENERAL ELECTRIC

Notes on the performance and application of the Planar Triode:

MODERN ELECTRONIC SYSTEM NEEDS

Modern electronics has seen significant changes in recent years. For example, radar has become more and more complex.

- Elaborate schemes using pulse compression and doppler returns are operational.
- Very large radars have been designed using electronically steerable arrays capable of tracking many targets at the same time.
- New, higher-performance navigational aids are being designed.

New telemetry systems are using complex coding and operating at higher frequencies. These and other electronic systems have placed more stringent requirements on active electronic components.

More CW power at higher frequencies is being used; therefore overall efficiencies are very important. Operation at frequencies up to 10 gigahertz is common with many designs using frequencies to K-Band, 16 gigahertz and higher. Frequency agility and phase fidelity are requiring wider instantaneous bandwidths. Advanced doppler radars are limited in target-recognition ability by the noise sidebands on the transmitted frequency and by the sidebands present on the receiver local oscillators. The usefulness of adaptive filters is also limited by the generation of undesirable modulation distortion products. Extreme linearity and wide dynamic range radar receivers are very difficult to design.

Pulse radars, radar beacons and other pulsed electronic systems require larger power outputs from smaller and more efficient power sources. C and X Bands are being used more and more — and gridded tube transmitters are being designed to their maximum capabilities. Magnetrons, klystrons, BWO's and C.F.A. are often too large and expensive. Solid-state sources cannot provide the minimum acceptable power outputs. In addition to more power, some systems must meet other difficult requirements. A good example is the pulsed radar altimeter. The transmitted pulse must be as short as possible with very fast rise-times. For better accuracy and range, high duties must be used. Any jitter present on the transmitted pulse destroys the radar accuracy. The radar altimeter as well as almost all other pulsed systems cannot tolerate FM or AM distortion during the pulse period. Pulsed phased array radars require the maximum state-of-the-art gain bandwidth products to become practical, and steering accuracy is limited by the phase distortion introduced during amplification of the transmitted or received signals.

Many other needs are present in addition to the electrical requirements already mentioned. There is a constant desire to reduce system size and weights. The mechanical features of the active components must be compatible with the circuit techniques used. Component packaging is important. Many missile or airborne systems meet extreme variations of temperature, tolerate high levels of shock and vibration, and in some cases, tolerance to nuclear radiation is required. Certain new electronic systems must operate instantly or within a few seconds after voltages are applied. The equipment must not only tolerate wide ranges of ambient temperature but must also operate in a stable fashion at the same time. Long life and extreme reliability are essential in almost all systems, either because of their complexity, vital function or to provide economical operation.

All circuit designers are familiar with most, if not all, of these requirements. **The component designer must also appreciate the circuit designer's needs.** The discussion presented here relates recent efforts by one gridded vacuum tube manufacturer to meet all, or as many as possible, of the requirements mentioned. It is recognized that all microwave functions cannot be performed by the gridded tube, but recent improvements are resulting in tube usage in many functions previously relegated to more expensive and complex modulated electron beam devices. Solid-state equipments have found usage only at the lower power and frequency levels.

NECESSARY GRIDDED TUBE IMPROVEMENTS

The vacuum tube of yesterday must be significantly improved if it is to be competitive for today's needs. To satisfy these needs, certain tube characteristics and geometries are essential. To reduce transit-time loading and phase delay, closer element-to-element spacings must be used. Smaller cathode areas must be used since transit-time effects are proportional to the active emission areas. Smaller element-to-element capacitances are mandatory to resonate tube and circuit at higher frequencies. Series inductance inside the vacuum enclosure can have serious effects on tube operation. More efficient conduction of internally generated heat away from the tube itself is necessary if smaller size and weight and more power output is to be obtained. All insulating portions of the tube must be of low loss materials at temperatures much higher than ambient.

Immobile internal structures are essential. The most difficult components in this respect are the heater-cathode structure and the grid. For efficient use in strip-line, cavity and/or waveguide circuit, the external surfaces of the tube must be suitable for proper RF connection. Another mechanical requirement is the maintenance of uniform element-to-element spacing over wide temperature ranges. Use of very low coefficient of expansion materials is essential if complex compensating circuitry is to be avoided.

For microwave use, many additional improved electrical characteristics are essential. High levels of transconductance are required to provide acceptable levels of gain-bandwidth product, and in narrowband circuits the pulsed start or rise-times are highly dependent upon tube transconductance. In large signal devices, such as class C operation, high transconductance, well into the positive grid region, is essential. Not only must the tube cathode supply very high currents from small areas, but long life at the same time must be assumed. The goal is always to obtain more and more emission from cooler cathodes. Many recent successes with high current density cathode materials cannot be applied to the microwave gridded tube. Even moderate cathode sublimation can seriously affect the grid performance. Lower cathode temperatures also improve tube life and reduce the required heater power. To obtain lower heater power at the same cathode temperature, efficient heat transfer from the heater to the cathode is important. Closer mechanical contact must be used and more mechanically rugged designs result. Closer bonding of the heater to the cathode also helps fulfill the system need for fast warm-up. Another electro-mechanical characteristic not often appreciated is the elimination of serious RF discontinuities due to complex seals, radical changes in tube dimensions and other internal features that produce parasitic capacitance or inductance. In other words, it is essential that a minimum amount of any circuit must be inside the vacuum enclosure.

This is a partial summary of the required electrical and mechanical features of a gridded tube designed to work into the higher microwave frequencies. The following describes a new family of tubes using manufacturing techniques proven most successful in improving the present tube state-of-the-art.

A NEW FAMILY OF GRIDDED TUBES

One of the most important design features of a high performance tube is a low loss, high temperature and vacuum tight metal-to-ceramic seal. One of the biggest obstacles in this respect has been in obtaining a seal between the ceramic material and a metal of equal or similar coefficient of expansion. This has been done using a special ceramic designed to duplicate the coefficient of expansion of titanium metal. Titanium is also used to provide the efficient tube gettering action necessary to reduce the effect of gas on

ENLARGED VIEW OF A NICKEL-TITANIUM SEAL

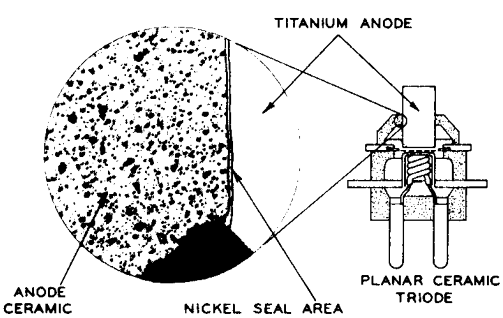


FIG. 1

tube life and cathode poisoning. **Figure 1** is an enlargement of a titanium-to-ceramic seal using nickel as a eutectic. This seal is made at over 1000°C and some of the most successful life tests made on a tube design using this sealing technique were made at 400°C ambients. This tolerance to high temperature is more important in obtaining higher levels of plate dissipation since 400°C ambients are seldom required.

Using this basic sealing technique, very accurate control of tube dimensions can be maintained. **Figure 2** demon-

SECTIONED CERAMIC TRIODE

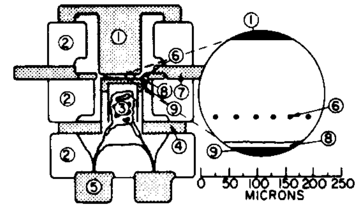


FIG. 2

strates this feature, as well as other features to be described later. This is an artist's sketch of a cutaway section of the 7077 triode — a tube marketed using nickel-titanium metal-to-ceramic seals. In this tube type, all seals are parallel and planar. The ceramics are diamond-lapped to close tolerances. The active cathode area is about 0.05 square centimeters for about 10 ma per volt transconductance at about 7 ma of plate current. This small area results in low capacitance and high electrical performance at higher frequencies. The heater used in the 7077 uses radiation to heat the cathode, since little or no direct contact is made to the cathode itself.

CUTAWAY SKETCH OF THE Y-1124 TRIODE

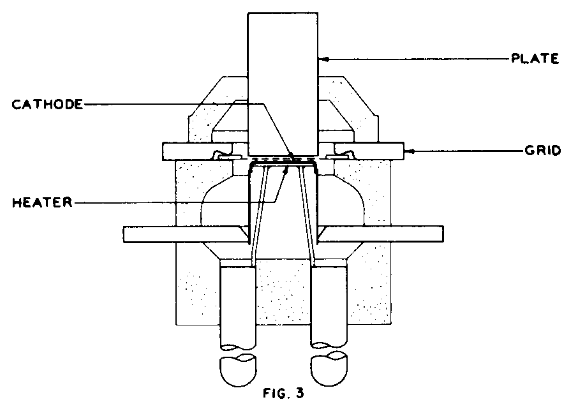


FIG. 3

Figure 3 is a simplified sketch of a more recent design, developmental type Y-1124. Note the reduced volume of ceramic used. The Y-1124 also uses a combination of planar and coaxial seals. The coaxial anode seal provides a lower capacitance and more efficient RF design. This basic configuration has been used commercially to 9.6 gigahertz as a radar beacon local oscillator. **Figure 3** also shows the basic features of a new bonded heater-cathode structure. The heater is bonded inside a flat insulating material attached to the back side of the oxide coated cathode. The cathode is heated by conduction rather than by radiation. Several advantages result from this bonded heater. One of the most important is a drastic reduction in warm-up time required for the plate current to reach 90% of the steady-state value after heater power is applied. **Figure 4** shows this. The reduced mass curve is the basic structure

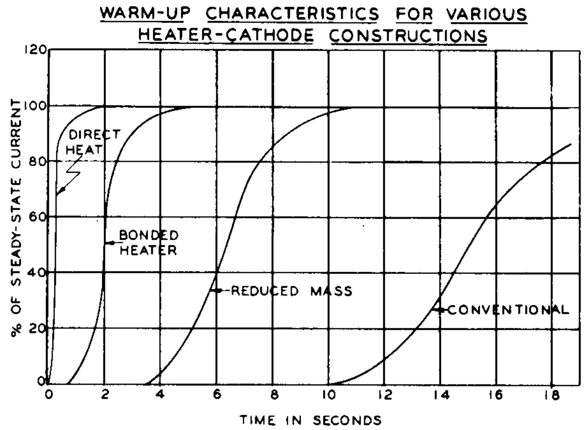


FIG. 4

shown in the 7077, **Figure 2,** using low cathode mass and extremely thin support wall material. This structure was not mechanically secure, and the regular 7077 dimensions must be used. The regular 7077 is shown on the right. The fourth curve shows the warm-up characteristics of a developmental directly heated cathode. This design is not presently offered for sale but is adaptable to the planar structure.

A second advantage of the bonded heater design is the very significant reduction of tube microphonics. (Microphonics are electrical signal outputs generated by internal element movements when the tube is shocked or vibrated).

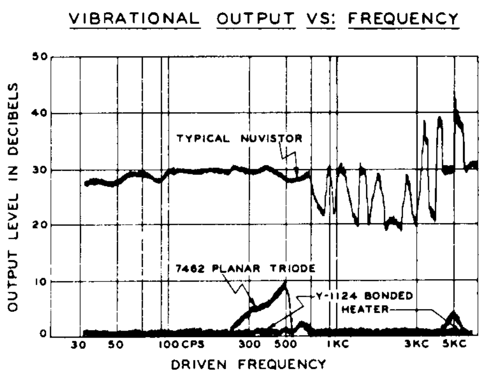


FIG. 5

Figure 5 is a reproduction of the microphonic outputs as a function of frequency as the tube under test is vibrated at a 10g level from 30 hertz to 5 kilohertz. Note the bonded heater construction is for all practical purposes microphonic free. Microphonic levels can be used to predict microwave performance where such undesirable results, such as pulse jitter, pulse bounce, and FM and AM distortion can limit tube usefulness.

The extra performance available when the tube is operated at high cathode current densities is useful only if acceptable life can be obtained. The use of titanium as the major metallic portion of the tube, use of high temperature bake-out and the use of very good vacuums available from bell-jar exhaust systems result in extremely low levels of gas within the tube. The ability of any gas to cause cathode poisoning and short life increases when the tube is operated at high current densities. The proof of very low gas levels in the new tube family is shown in **Figure 6.** This data shows that

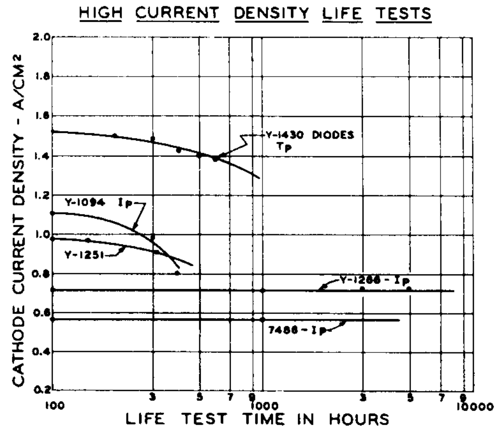


FIG. 6

excellent life is obtained at current densities greater than 1 ampere per sq cm of active cathode area. This level of operation can be compared to the highest level of current density used in a TV set, less than 100 ma per sq cm. Similar good results have been obtained on pulsed rated types. **Figure 7** is a plot of pulse power output as a

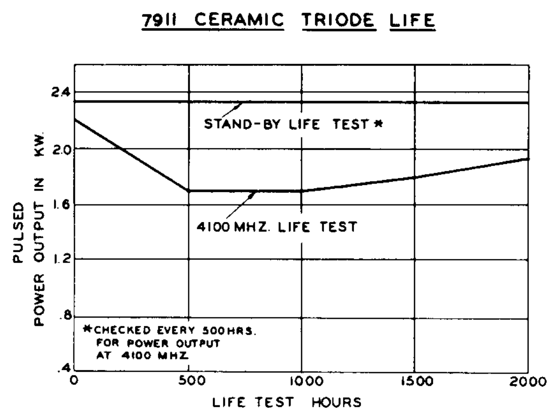


FIG. 7

function of tube life. This data was taken on the 7911 triode operating at about 9 amperes per sq cm during the plate pulsed period. The peak RF currents would be near 30 amperes per sq cm. The top curve is a plot of power output vs. time with the tubes actually being life tested with only the heater voltage applied. In some applications, this type of life test is more difficult than actual operating life. This standby life is the average of 12 tubes. The second curve is a plot of power output vs. time with the tubes operating in a 4100 mhz life test cavity. This is the average result on 16 tubes.

Significant improvements in the electrical, mechanical and thermal characteristics of the grids used for the new family were necessary. Two basic grid fabrication techniques are used. **Figure 8** is a sketch of the two constructions used. The sketch on the upper left is a mechanism for obtaining a very high degree of grid wire tensioning. The materials and mechanical features are arranged so that as the grid cools after the exhaust bake-out, the difference between the coefficient of expansion of the tungsten grid frame and the

PLANAR TUBE GRID AND ANODE CONSTRUCTION

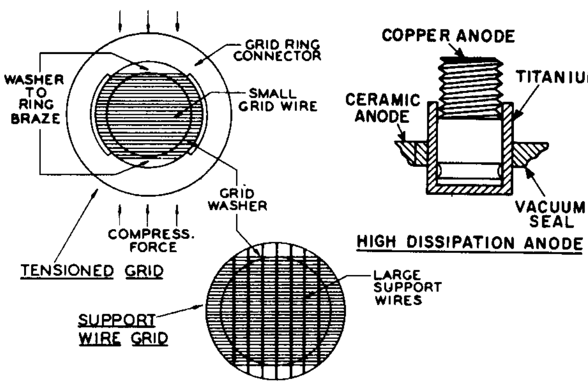


FIG. 8

titanium distorts the grid frame in a direction that tensions the small lateral wires on the grid frame washer. This construction has resulted in significant reductions in the level of microphonics measured on tubes using more conventional grid-making techniques. The sketch at the bottom-center uses a grid ring similar to the other sketch on the left, but in this case, heavy support wires are wound at right angles to the small grid wires. All physical connections between both wire sizes and the grid ring are brazed together in a high temperature furnace. This grid receives its rigid characteristics from the rigidity of the large support wires. These large wires also greatly improve the thermal properties of the grid. The higher powered tube types use the support-wire grid for this reason. Further work is being done at this time to provide even better grids to provide even higher performing tubes.

Most recent efforts to obtain the optimum grid has resulted in an etched-frame support grid structure. This mechanical configuration is shown in the photo, **Figure 8A.** The large vertical bars are electro-etched and are typically 20 to 40 mils wide. The smaller horizontal bars are chemically-etched and are typically 2 to 4 mils wide. The actual high performance portions of the grid are the small wires running diagonally. These wires are typically .3 to 1.0 mils in diameter depending upon the triode performance desired. After final assembly, high temperature brazing bonds each portion of the grid to its adjacent component. Each of the small wires are inspected under a microscope to assure proper brazing between each wire and its etched frame support. Grids can be constructed with various combinations of these techniques. The large vertical bars can be nested into slots in the cathode to provide close spacing between the small grid wires and the active portion of the cathode. Grids using only the chemically-etched frame can be used with the frame facing the tube anode and more efficient use of the available cathode area is possible. This structure results in grids capable of conducting larger amounts of heat, covering larger active cathode areas and extra high performance. Triodes using these techniques have been built with cathode areas of over two square centimeters, transconductance approaching one mho, and extra high dissipation capabilities.

Most of the heat generated in the tube must be dissipated by the anode and its heat-sink. Unfortunately, titanium is not an excellent conductor of heat, and other than solid titanium anodes must be used on higher power rated types. The sketch of a cutaway view of a combination copper and titanium anode is shown on the left in **Figure 8.** Heat dissipation capabilities sufficient to prevent tube failure at maximum cathode current capabilities has been obtained using this bi-metal anode design.

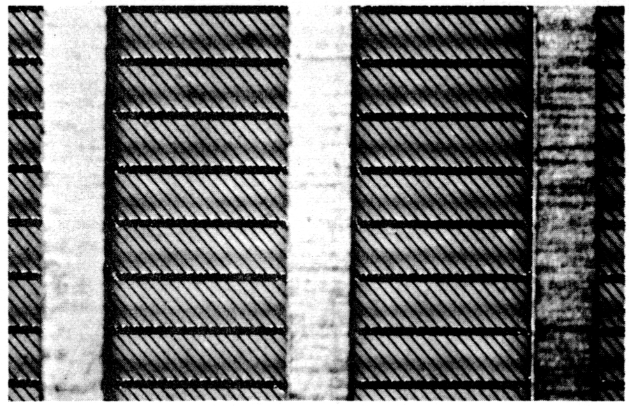


FIG. 8 A

The basic construction techniques used on the new tube family permits the modular construction of several possible external configurations with identical internal features. **Figure 9** shows this feature. The types 7077, GE 14501 and Y-1266 have similar internal construction. All three types have the same grid connector size and configuration. The 7077 uses button-type heater connections with a recessed cathode terminal. This provides an external outline more suited to clip-type sockets useful at the lower microwave frequencies, and the GE 14501 is the small tube adapted for use in coaxial cavity circuits. The Y-1266 is similar to the GE 14501 except for the anode. The larger Y-1266 anode can dissipate more heat since more contact and heat-sinking areas are provided.

VARIOUS GEOMETRIES SHOWING MODULAR CONSTRUCTION

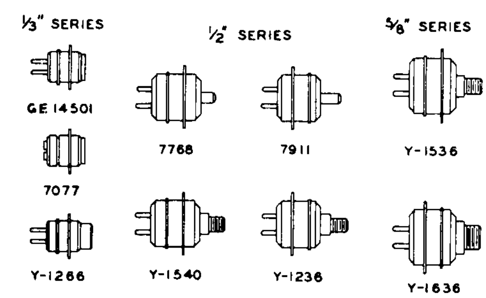


FIG. 9

In the half-inch series (approximate diameter of the ceramics) the 7768 can be compared directly with the Y-1540. The 7911 can be compared with the Y-1236. Each one is identical to its counterpart, except for the anode. The 7768 and 7911 are rated for about 6 watts of anode dissipation. The Y-1540 and Y-1236 are rated at 30 watts and use the anode design shown in **Figure 8.** The largest series (five-eighths of an inch ceramic diameter) is shown at the right. The first two developmental types are the Y-1536 and Y-1636. The Y-1536 has 0.6 sq cm of cathode surface and is designed for grounded grid amplifier use. The Y-1636 has 0.8 sq cm of cathode area and is the largest of the new family. This type has an enlarged copper-titanium anode that has dissipated 100 watts. The Y-1636 is designed for grounded cathode use in a re-entrant cavity oscillator.

Significant data has been taken to demonstrate the power output versus frequency capabilities of the new ceramic tube family.

PLANAR TRIODE MICROWAVE PERFORMANCE
(C W OPERATION)

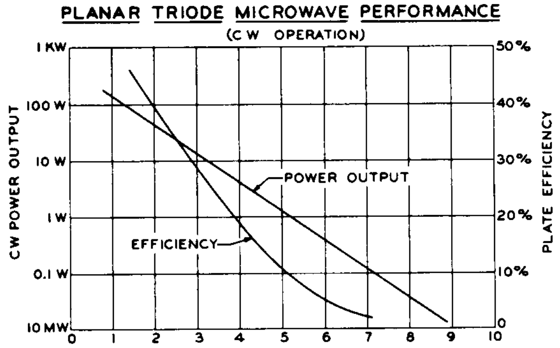


FIG. 10

Figure 10 is a plot of CW output versus frequencies. The approximate plate efficiency is also shown. This curve was constructed from a variety of measured results on a variety of tube types. At lower frequencies, the larger tubes are recommended. At higher frequency the smaller tubes were evaluated to determine the power outputs available

PLANAR TRIODE MICROWAVE PERFORMANCE
(PULSED OPERATION)

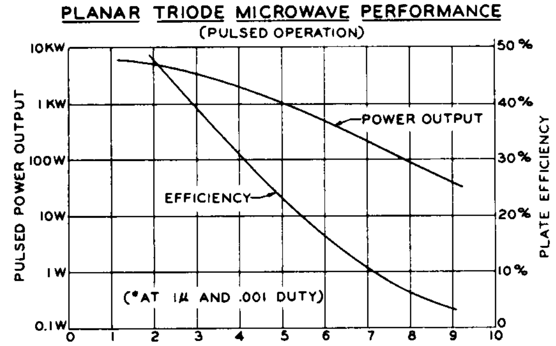


FIG. 11

within tube maximum rating. The data shown in **Figure 11** shows the plate pulsed capability of the pulse rated types. This data was obtained in a similar fashion.

Using the grid techniques shown in **Figure 8,** very high levels of transconductance are obtained. For example, the type 7768 is specified at about 50 ma per volt and this is obtained with about 25 ma of plate current. The 7768 has demonstrated very high levels of small signal gain-bandwidth products. The 7768 has been evaluated in a triple-tuned pulsed circuit at 1.3 gigahertz. A gain of about 14

BROAD-BAND PULSED AMPLIFIER

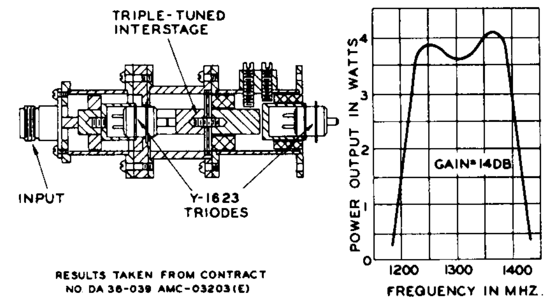


FIG. 12

db was measured with a three db bandwidth of near 165 megahertz. This calculates to about 4800 megahertz gain-bandwidth. **Figure 12** also shows the circuit arrangement used for the triple-tuned 1.3 gigahertz amplifier. The response obtained is also shown.

CERAMIC TUBE TOLERANCES TO ADVERSE ENVIRONMENTS*

1. NO DAMAGE TO 10^{19} NEUTRONS FAST
2. NO DAMAGE TO 10^{11} ERGS PER GRAM CARBON
3. NO RADIATION RATE EFFECTS NOTED
4. SURVIVES 20,000 G'S CONSTANT ACCELERATION-CENTRIFUGE TESTS
5. SURVIVES 20 G'S FOR 10'S OF HOURS AT MOST CRITICAL FREQUENCY
6. OPERATE IN A $1G^2$ PER CYCLE PER SECOND AT 50-2000 CPS.
7. EXCELLENT LIFE AT 400°C AMBIENT
8. SURVIVES AT LEAST 20,000 G'S IN "GUN-SHOT" TESTS
9. SURVIVAL AT 3000 G'S FOR 3 TO 5 MILLISEC.

*ON SELECTED TYPES

FIG. 13

New military electronics systems must tolerate a large variety of adverse environments. **Figure 13** is a brief resume of the conditions which the metal ceramic triode has survived. The most severe of these required the combination of the bonded heater-cathode structure, extra strong ceramics, mechanically rugged grids and new sealing techniques available only in the new planar ceramic tube family.

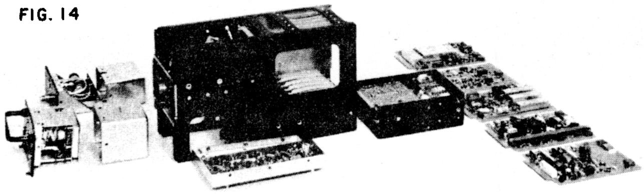
NEW AND IMPROVED EQUIPMENTS MADE PRACTICAL

It is, however, fair to state that many new equipment concepts have been made practical as a result of the extra performance available from the new tube designs. There are several examples of this.

Distance measuring transponders are being used today aboard military, commercial and private aircraft. These units send out coded pulses which interrogate a special transmitter located at a known location. The roundtrip time of the interrogating and reply pulse are used to determine the line of sight distance from the aircraft to the ground station. These equipments are used in the military TACAN and VORTAC systems. In commercial and private usage, they are referred to as DME (distance measuring equipments). To identify the large number of ground stations, many different frequencies must be used. The most used spectrum is from 1125 megahertz to 1250 megahertz with one megahertz channel separation. Previous designs use four stages of single-tuned RF amplifiers that must be mechanically tuned and tracked across the assigned spectrum. This equipment was large, expensive and heavy. The newer version of TACAN-DME will use four stages of double-tuned RF amplifiers broadbanded to cover the complete spectrum of 125 megahertz. One designer reports a 10 to 1 reduction in both size and weight using the new ceramic tubes described here. There are at least five companies in the United States with this new system concept in design or prototype production. Almost without exception all stages for all five companies are using the new ceramic tube family.

Radar altimeters for aircraft use have been in service for years. However, for modern aircraft, more accurate instruments are needed. More accuracy requires higher transmitting frequencies and shorter rise-times and durations for pulsed systems. The small planar tube has met these needs. Pulse durations of a fraction of one microsecond are easily obtained and pulse powers up to over one kilowatt are practical for long-life transmitters. **Figure 14** is a photo of the APN-171 pulsed radar altimeter. Pulsed powers of over 150 watts are available for pulse durations of less than 100 nanoseconds. The transmitted frequency is approximately 4300 megahertz. This unit also uses a small planar ceramic tube as the local oscillator for the receiver portion of the altimeter.

FIG. 14



The higher transconductance triode types are being used in other broadband applications. The 7768 test results shown in **Figure 12** relate the performance in the pre-amplifier stages of a phased array module. Triodes were evaluated in these tests because of their low phase distortion and delay. The complete module which is not shown here was being developed to compete with the TWT. Other broadband amplifications include ECM amplifiers and broadband Doppler radar amplifier chains. The triode offers small size, high efficiency and an economical solution to the problem of obtaining wideband operation and high power outputs.

In the United States, there is a program to up-date the present aircraft handling facilities at large, metropolitan airports. There is a similar program to provide better identification for military aircraft. These programs have been combined under the **AIMS** Program. The hoped-for mass employment of identification beacons on all aircraft of all sizes demands a low cost, small size and high performance beacon transmitter. One offering by General Electric uses two of the new family for a master oscillator-power amplifier arrangement. This is done to provide the required frequency stability. **Figure 15** is a photo of this unit. The Y-1537 triode is designed specifically for this application requiring long life and good reliability. These equipments are often referred to as ATC, air traffic control, and/or IFF, identification friend or foe, transponders.

Most radars used for aircraft and missile tracking use radar beacons to augment the radar returns. These beacons must operate at the radar frequency. Several designs have been manufactured using the new ceramic planar triodes. The local oscillator for the beacon receiver uses the smaller triodes up to about 10 gigahertz. Some designs operating at lower frequencies, up to 6 gigahertz, also use the triode in a pulsed oscillator transmitter. Triodes are being used here because of their small size, low cost and simple power supply requirements. The frequency stability of the triode is important in these applications. Triodes have also been shown to produce less sideband noise when compared to the reflex klystron, magnetron and varactor multiplier. This desirable feature is very important in low noise receivers and in Doppler radar transmitters.

Another high frequency use for the small planar triode is in hand-held radar applications. Many of the performance features mentioned for the radar beacons apply here with the extra requirement for low power consumption. The triode is being evaluated for use as both a local oscillator and a pulsed transmitter.

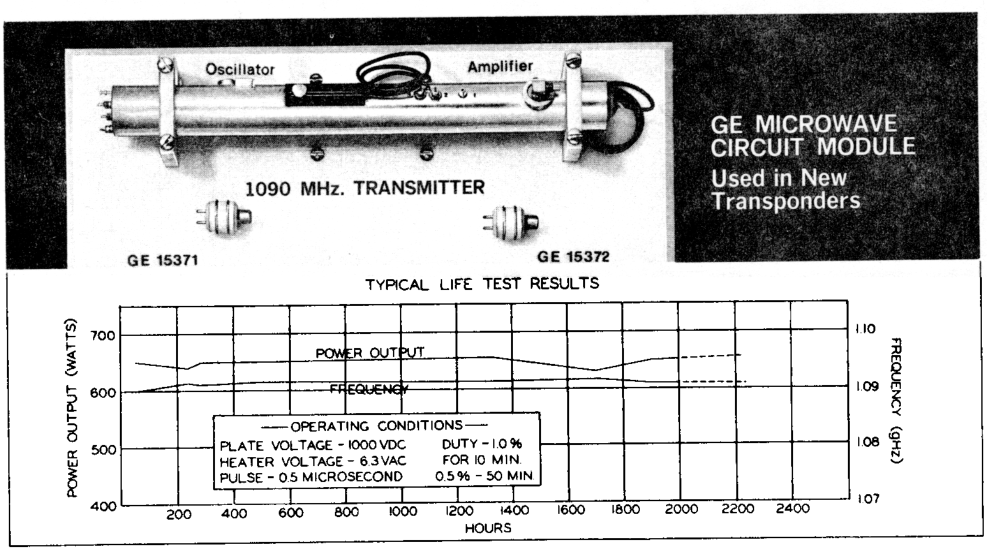


FIG. 15

The 215-260 megahertz telemetry band now being used must be vacated before 1970 to release these frequencies for other services. The new bands are 1435-1535 and 2200-2300 megahertz. Planar ceramic triodes are being used as power drivers and output stages in the new equipments de-

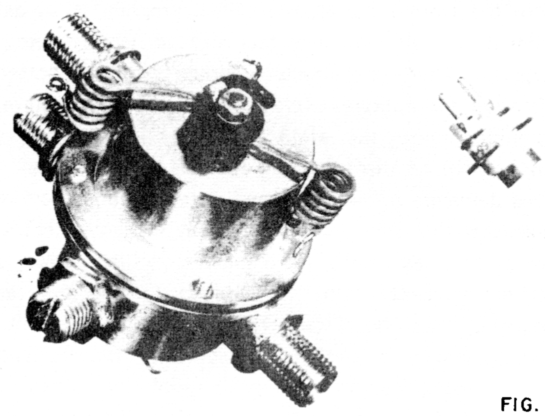


FIG. 16

signed for these higher frequency bands. **Figure 16** is a photo of a small 2200-2300 megahertz transmitter using a Y-1266 triode. This unit delivers 2 to 3 watts of CW output with a large signal gain of over 10 db and an overall efficiency of over 25%, including heater power. The Y-1266 is shown beside the grounded grid coaxial amplifier. Other systems near these frequencies use similar types and circuitry. One of these is a recent collision warning system. This equipment requires narrowband amplifiers with about 35 db of gain and approximately 1 kilowatt of pulsed power output. Only three stages are required if tubes from this new family are used.

The last, but not least, application for the new planar triodes mentioned here is in high frequency signal generators. The small Y-1266 is being designed into two new oscillators by one manufacturer. The almost equal grid to cathode and grid to plate capacitance makes the Y-1266

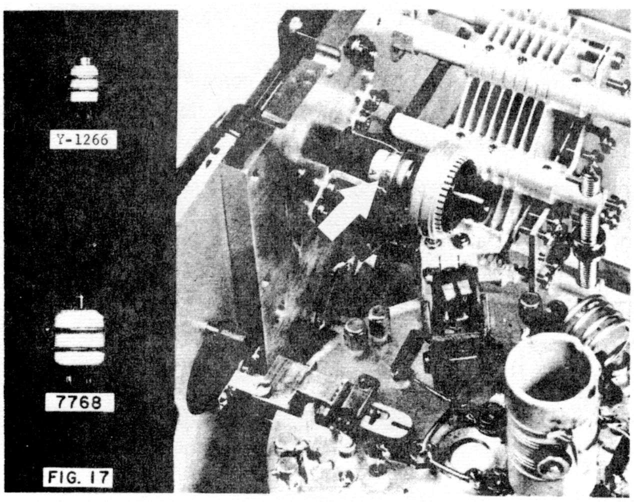


FIG. 17

ideal for butterfly type circuits. The photo shown in **Figure 17** shows the Y-1266 and 7768 being used in a new design signal generator recently released by General Radio. In addition to the wide tuning range available from the Y-1266, the tube was demonstrated to be superior to other competitive triodes in terms of short term and long term frequency stability. The 7768 is used as a broadbanded power amplifier to provide additional signal generator power output and to improve frequency stability under wide variations of load impedance. The Y-1641 bonded heater version of the **7486** is being used in a very stable, local oscillator for a new spectrum analyzer being manufactured by the Tektronic Corporation. Most of the significant new uses for the new ceramic family have been mentioned. There are numerous other uses which cannot be described here. These uses were described in terms of the functions required and the equipments in which they are used. More detailed application information will now be discussed.

APPLICATION NOTES ON PLANAR TRIODES

TUBE CONNECTIONS AND CONTACTS

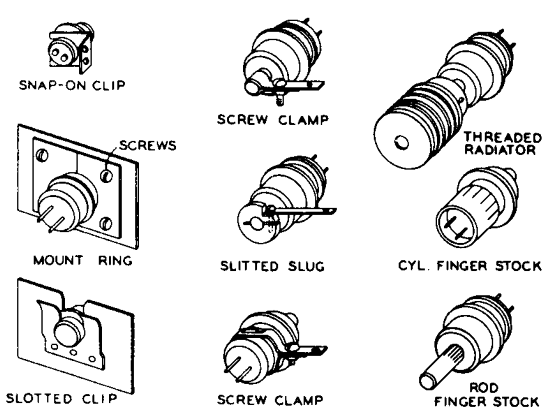


FIG. 18

At high frequencies, good RF connections to the active elements of the tube to be used are fundamental. **Figure 18** shows several sketches of the various methods of connecting to the desired tube element. One additional significant feature of the new tube family is the ability to solder directly to the tube elements. It has been found almost essential to use soldered connections on circuits that must take very high levels of shock and vibration. This method of connection is recommended wherever practical.

Two basic cavity designs have been used most often for the higher microwave frequencies. Most oscillators use a re-entrant type circuit which is basically a grounded cathode amplifier with built-in feedback. The amplifier stage is almost always a grounded grid circuit.

TUBE - CAVITY COMBINATION

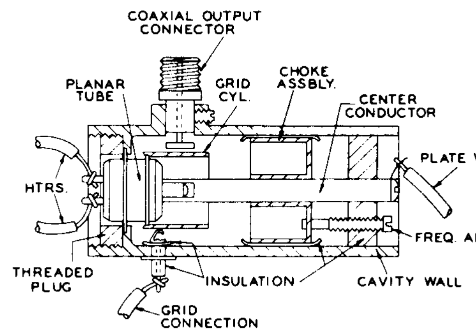


FIG. 19

Figure 19 is an artist-sketch of a common configuration used for oscillator tube-cavity combinations. The most useful tube geometry is the outline which has the cathode as the largest external diameter element. The cathode can then be clamped or soldered to the cavity body with a diameter large enough to accommodate the other cavity elements. The heater voltage can be applied with ease and without consideration of RF bypass or decoupling. A grid cylinder is attached to the grid flange. The length of this cylinder is chosen to resonate as a half-wave resonant circuit. One portion of the half-wave circuit is foreshortened by the grid to plate capacitance of the tube and the other end of the half-wave circuit is open-ended and untuned. This places a voltage maximum at the open end of the grid cylinder. At this point, the voltage is further resonated by tuning the remainder of the plate coaxial cavity to the same desired frequency. This usually is a quarterwave circuit tuned by the placement of the anode choke. This choke can take many forms but the basic purpose is to provide a short-circuit for the tuned RF voltage present on the anode center conductor while providing an open-circuit for the DC applied to the anode. The choke shown in **Figure 19** is a single-tuned, quarter-wave choke. The open circuit seen by the choke looking out of the cavity towards the DC connection is transformed into a short circuit at the inside end of the anode choke. The short circuit at this point is required to prevent RF leakage. Chokes using two or more quarter-wave sections can also be used where extra choking action is required and space is no problem. The oscillator frequency can be changed up to about 10% in frequency by moving the position of the anode choke inside the cavity body. Further frequency range can be obtained if the grid cylinder length can be varied at the same time. The design of the cavity circuit from the end of the grid cylinder looking back toward the cathode end of the cavity is important. This length most often must look like a three quarter wavelength circuit to provide proper phasing at the end of the grid cylinder. Feedback is provided, since the basic circuit resembles a Colpitts oscillator circuit. Resonance is established between the grid and anode and feedback is provided by the voltage developed across the grid to cathode capacitance. Power output can be extracted by inserting a capacitive probe near a high RF voltage point inside the tube-cavity combination or an inductive loop near a high RF current point. This is usually done along the grid cylinder for mechanical reasons. In some cases, a combination loop-probe is used when, for mechanical reasons, a current or voltage maximum point is not easily located.

COAXIAL CAVITY AMPLIFIER

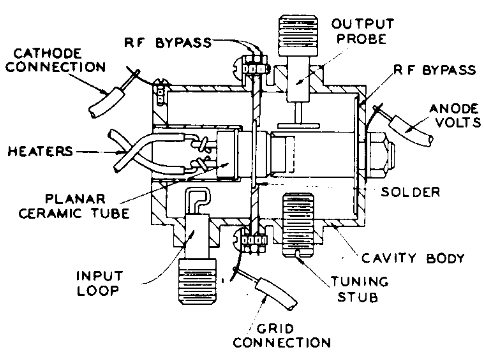


FIG. 20

Figure 20 is a cutaway sketch of a typical amplifier circuit. This is the basic circuit used for the Y-1266 tube-cavity shown in **Figure 16.** At 2300 megahertz, the capacitances of the Y-1266 are sufficiently low to permit use of quarter-wave resonators in the cathode and anode circuits of this grounded grid configuration. Quarterwave circuits produce, among other things, smaller size and weight devices but limit the upper useful amplifier frequency. In the arrangement shown in **Figure 20,** the grid is DC isolated using mica by-passes. Bias can be fixed using a DC value of grid voltage, or a grid leak or cathode resistor can be used for variable bias. The input signal is applied using an inductive loop. The input capacitance is usually larger than the output capacitance, and it is more difficult to obtain a high RF voltage point inside the cathode cavity. The output is taken from a voltage probe in the anode cavity. The cathode cavity is loaded heavily with the low impedance of the grounded grid input and is usually tuned near the desired frequency. Further tuning is not necessary over a relatively wide frequency range. The anode circuit must resonate the input frequency, and in this amplifier the anode cavity is tuned by susceptance loading of the output cavity. Brass slugs are inserted which in effect raise the resonant frequency. Two slugs were necessary to tune the desired range of 2200 to 2300 megahertz. In some cases, the plate circuit can be tuned to a frequency much higher than the cathode circuit. In this case, where higher frequencies are desired, a half or three-quarter wavelength cathode circuit is used. This lengthens the cavity length and increases the size and weight.

In most amplifier applications, bandwidths as well as other RF performances are important. For maximum bandwidth, only quarterwave circuits should be used as suggested by

BANDWIDTH SHRINKAGE EFFECTS OF A CAVITY VS. THEORETICAL AVAILABLE BANDWIDTH

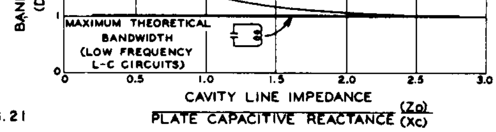
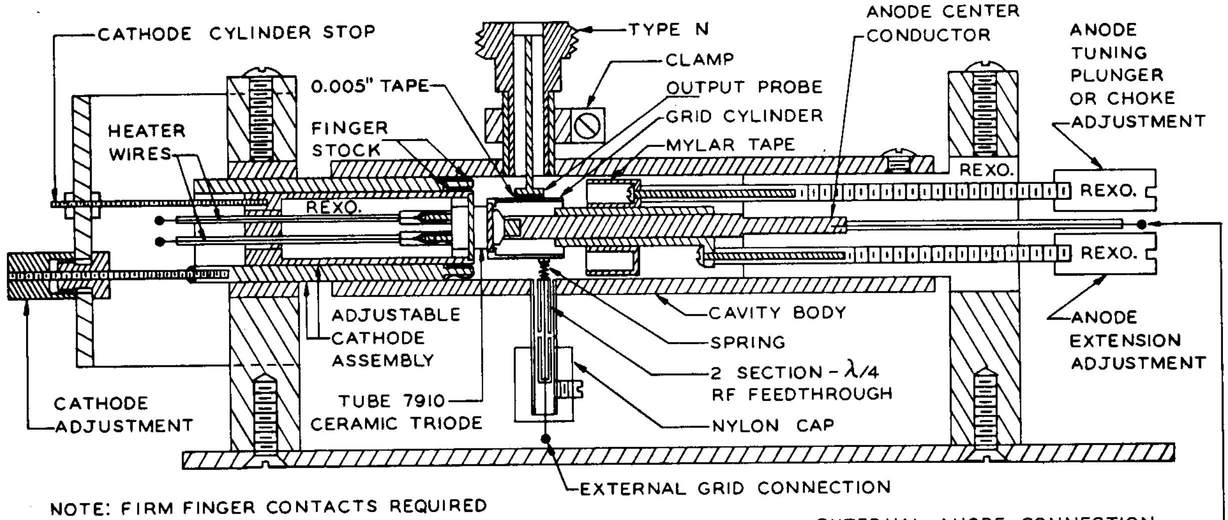


FIG. 21

Figure 21. This is particularly true for the anode cavity or circuit. However, for oscillators, multi-wave length circuits are actually recommended for maximum stability and extending usefulness to higher frequencies. The narrowbandedness of the multi-tuned circuit improves stability by providing higher effective Q's, and half-wave circuits provide resonance at higher frequencies. Half and three-quarter wavelength amplifiers are used to extend the upper frequency of some of the large power triodes and tetrodes. In multi-tuned circuits used for broadbanded circuits, it is sometimes impractical to use quarter-wave circuits throughout.

There are many insidious design features in most success-

PLANAR CERAMIC TUBE TEST CAVITY



NOTE: FIRM FINGER CONTACTS REQUIRED ON TUBE CONNECTIONS, ANODE CHOKE, AND EXTENSION.

FIG. 22

ful tube-cavity designs which the designer is usually hesitant to describe, and there are also no sure-fire design equations. For these reasons, original designs require a large amount of trial and error. To provide a maximum number of variables, the cavity shown in **Figure 22** was built. The feedback can be adjusted by adding lengths of coaxial line at the cathode end. Various lengths of grid cylinder can be inserted. The anode choke assembly can be moved to change frequency. Various kinds of bias can be applied, and the output coupling can be adjusted as desired. Using this cavity, the type 7910 was evaluated over the frequency range and cathode current

MICROWAVE PERFORMANCE AT HIGH CURRENT DENSITIES

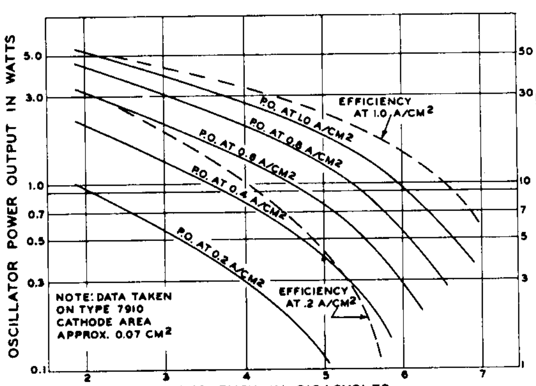


FIG. 23

densities shown in **Figure 23.** These results also show the significance of high current density operation already discussed.

PRESENT STATUS OF NEW TUBE FAMILY

A new family of lug terminal planar tubes has been developed for lower frequency use. The high temperature tolerance, extreme mechanical ruggedness and high electrical performance available from the internal dimensions of the new tube fabrication techniques result in their usefulness

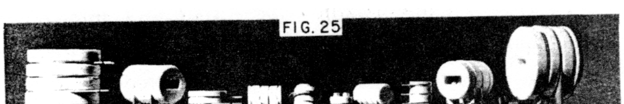


FIG. 25

at low frequencies. **Figure 25** is a photo of most of the available external outlines. These tubes are well suited for socketed circuits, printed circuits and socketless circuits. Some of the outlines use a "T" bolt which is attached to the heater-end ceramic. The tube can be mounted to any supporting surface, and the tube serves as its own terminal strip. This method of tube mounting is particularly useful

LOW FREQUENCY APPLICATIONS

- POWER SUPPLIES - HIGH-VOLTAGE AND IN ADVERSE ENVIRONMENTS
- VIDEO AMPLIFIERS - HIGH-TEMPERATURE, SHOCK AND VIBRATION USE
- I.F. PRE-AMPS - LOW NOISE AND S.T.C. CIRCUITS
- ION AND STRAIN GAGE PRE-AMPS - HIGH-Z AND VIBRATION USE
- PULSE MODULATORS AND AMPS - HIGH-VOLTAGE AND FAST RISE TIME NEEDS
- DETECTOR PROBES - HIGH-Z AND BROADBAND INSTRUMENTS
- DC, AUDIO AND SERVO AMPS - IN ADVERSE ENVIRONMENTS
- DIFFERENTIAL AMPS - TEMPERATURE AND TIME STABLE CIRCUITS
- MOBILE TRANSMITTERS - REDUCES SIZE AND WEIGHT
- RECONAISSANCE RECEIVER PRE-AMPS - ELINT

FIG. 26

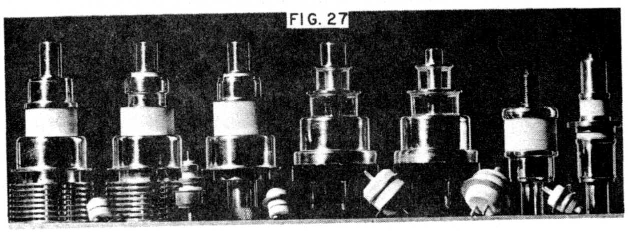


FIG. 27

where wire-wrap joints are used. **Figure 26** is a brief list of successful low frequency applications. **Figure 27** is a photo showing the available high frequency or microwave outlines. These are the types most discussed in this paper but some of the types shown are older designs using conventional sealing techniques. **Figure 28** is a photo of the latest developmental types. Only a portion of these tubes is available. The most significant of these are the two larger tubes shown at the center of the photo. Up to one kilowatt of CW power output at 1.3 gigahertz has been obtained at about 65% efficiency. Transconductance over 500 ma per volt has been obtained. The smaller tubes have been operated as oscillators to frequencies up to 16 gigahertz.

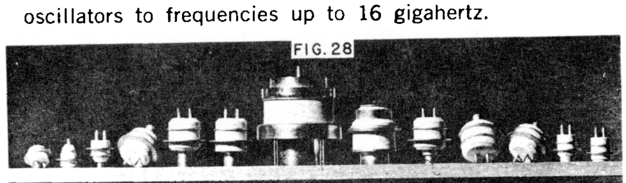


FIG. 28

Notes on the performance and application of the Microwave Circuit Module, MCM

THE MICROWAVE COMPONENT AND ITS CIRCUIT

It should be obvious to the reader that a planar ceramic tube, transistor, tunnel diode, avalanche diode and other active components must be applied to some circuit arrangement. General Electric Company has and is active in the manufacture of most of these devices and has developed a lot of "know-how" in the normal routine of evaluating, testing and specifying gridded tubes, back-diodes, tunnel diodes, etc. In many cases, the completed circuit is demanded by the customer who knows the importance of careful "mating" of the active component to its circuit. Continuing efforts are being directed towards circuits using the new generation of small planar ceramic tubes. These tubes have been discussed earlier in this brochure. This line of packaged components is being expanded to include solid-state active components, isolators and circulators.

The term Microwave Circuit Modules, MCM, is used to describe various circuit arrangements. The well-known tube-cavity combination using coaxial resonators are used along with strip-line configurations. Lumped-constant coils and capacitors are used at lower frequencies. The choice of circuit configuration and active component depends upon the application requirement. A large percent of these applications are discussed earlier in this brochure with exception of the applications served only by solid-state components. A brief suggestion of the more significant solid-state applications are:

- Lower frequency low voltage local oscillators using transistors
- More reliability in less demanding applications
- Small size and weight where only few active components are satisfactory
- Higher CW power outputs at higher frequencies as available from varactor multipliers
- Lower noise figures
- Maximum overall efficiency where tube heater powers are much larger than the signal powers

Future plans for the MCM will be directed towards these applications.

MCM PERFORMANCE CAPABILITY

The overall tubed MCM capability as a function of frequency is shown in **Figures 10** and **11.** In almost all cases the power outputs are above the figures available from single-component solid-state devices. The exception to this would be varactor multiplier and certain active diodes which produce more CW power at frequencies of X-band and over.

There are, however, experimental MCM results that yield ¼ to ½ watt CW outputs at 9.6 gHz, using a single tube-cavity combination.

When an active component is added to a circuit, certain additional performance criteria are necessary. The MCM can be designed to meet these requirements with typical performance as follows:

- ± 1½ mHz. pulling at C-band for VSWR's of up to 1.5 to 1 depending upon the circuit
- 100 KHz. per volt plate voltage pushing about a normal operating condition at C-band
- ±5 KHz. per degree centigrade frequency drift over a temperature range of −55° C to +125° C at C-band
- Down to 1 oz. weight and 0.5 cu. in. volume at higher frequencies
- Survival at shock levels of over 15,000 g's
- Down to 3 secs. or less warm-up— 90% of steady state currents

APPLICATION OF DC AND MODULATING VOLTAGES TO THE MCM

There are several methods of applying the necessary voltages to a pulsed amplifier or oscillator. Pulsed voltages must be applied with caution, because the tube cannot tolerate the usual pulsed levels on a continuous basis. The

VARIOUS TUBE-CAVITY PULSING CIRCUITS

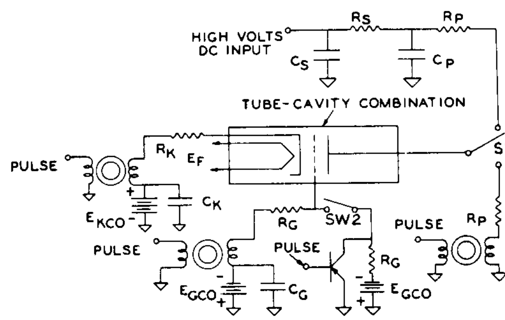


FIG. 24

circuits shown in **Figure 24** show the various methods of applying the pulsed voltages. All of these are not used at one time and are presented here to show all possible combinations individually.

The most common pulsing circuit uses a pulse transformer to supply a large positive voltage on the tube anode only during the on-time. In this case, SW1 is switched from the position shown. The cathode is grounded and a grid leak resistor, Rg, is used. The current levels can be adjusted by changing Rg. This is the circuit for maximum reliability and performance. Since the tube has high voltages applied only during the plate pulsed periods, no serious arcing can normally occur, since in most cases the voltage is removed after a microsecond or so. Higher voltages and currents can be used, and more pulsed performance results. This type of pulsing requires the largest amount of pulser power.

Where lower pulser powers must be used, grid or cathode pulsing is used. The plate is connected to a high voltage DC source through a filter network. For grid pulsing, the cathode is grounded and a positive pulse is applied through some value of grid resistor, Rg. A bias voltage, Egco, sufficient to cut the tube off during the pulse-off period is used.

The pulse level and Rg are adjusted to give the desired current levels. Cathode pulsing can be applied in a similar fashion by grounding the grid and applying a negative pulse with the tube cut off with a positive cathode voltage, Ekco. SW1 would be in the position shown. In grid pulsing, the pulser must supply only the pulsed grid current, and the pulser can be a relatively high impedance device. In cathode pulsing, the pulser must supply the pulsed tube cathode currents. For this reason, the pulser must be a high current source of relatively low impedance. A value of Rk can be chosen to limit the peak currents drawn from the pulser and provide the DC degenerative effects of a cathode resistor.

Cathode and/or grid pulsing usually results in less reliable operation. Unless care is taken, the slightest tube arc can destroy the tube. All the energy available from the DC source can be "dumped" into the tube, and severe damage can result. Several things can be done to minimize this effect. A suitable value of Rp can be added to limit the plate current. At the high currents associated with arcing, a large drop will appear across Rp and more reliable operation results. Rp is usually about 100 ohms depending upon the allowable plate voltage drop for normal operation. Another method often used is to add a relatively large value of series plate resistance, Rs, and use a relatively small value of filter output capacitance, Cp. Cs should be a much larger value to provide a low impedance source looking towards the DC supply. The RC constant of Rs and Cp is chosen to prevent serious pulse-droop. The value of Rs should be large enough and Cp small enough to essentially discharge Cp under arcing conditions. This method of reducing serious arcing effects can also be used to reduce the voltage and current levels when higher duty factors are used. The duty factor is the ratio of the on-time to the time between pulses.

The two methods of applying cathode and grid pulsing using pulse transformers permit operation with pulse modulation into the positive grid region. The pulse levels need only to surpass the levels of cut-off bias used. If this condition is not required, a simpler pulsing circuit can be used. If SW2 is switched closed and the pulse transformer removed, a cut-off bias can be applied through a series resistor, Rg. If a PNP transistor of suitable collector voltage rating is placed across the grid terminal and no pulse is applied to the base, as shown, the tube will draw no current. If a negative going pulse of sufficient level is applied to the transistor base, the transistor will short out the bias voltage and the cut-off bias is lost. In this case, the tube can be pulsed-on to a level that corresponds to zero bias. In this case, some series value of Rk or Rg, not shown, could be used to limit tube currents during the pulsed-on period.

Plate pulsing is usually used if the pulse durations are long enough and the voltages and powers small enough to permit use of SCR's. Very short pulses are not possible because of the storage times associated with SCR's. Very high voltages and average power levels are usually limited by the SCR's voltage and power handling capabilities. Transistor pulsers have not been able to supply useful levels of voltages and currents reliably. Grid or cathode pulsing is used when the power output levels are available within the tube's maximum ratings for this service. The grid-cathode pulsed ratings are often only half the plate pulsed ratings. Very short pulses can be obtained, because transistor storage times are much shorter than SCR storage times. Transistors of sufficient capacity are available for grid-cathode pulsing. Very short pulses can be generated using avalanche diodes as a switching element. RF pulses from a diode moduated tube-cavity combination of less than 50 nanoseconds are easily obtained with rise and fall times of less than 10 nanoseconds.

For maximum performance, reliability and life the following check list is recommended:

Plate pulsing
- Most reliable method yielding maximum performance
- Requires maximum modulation power
- Low level of catastrophic failures
- No pulse stretching or CW moding
- Very short pulsing and rise-times difficult

Cathode pulsing
- Fastest rise-time capability
- Lower modulating powers
- Less tendency to arc than grid pulsing
- Less tendency to pulse stretch and CW mode
- Catastrophic arcing can exist
- Long pulsing more practical
- Low impedance modulators required
- Sensitive to load mismatch

Grid pulsing
- Most subject to arcing and failure
- Lowest modulating powers required
- Fast rise-time capability
- Most sensitive to load mismatch
- Less tolerant to tube changes
- Higher impedance modulators usable
- Most subject to pulse stretching and CW moding
- Requires most careful servicing

Summary

Modern electronics has placed new requirements upon the active devices and circuitry used at microwave frequencies. The gridded vacuum tube and other active devices have undergone significant redesign to provide the new levels of performance required. A new planar ceramic tube family has been designed using higher temperature seals, closer mechanical spacings, improved RF constructions and high performance grids. Proven long life at high current densities has provided new levels of power output and efficiency at frequencies never before reached with tubes. Extreme mechanical ruggedness of the active component and its circuit permits the application of the extra microwave performance to all known weapon systems and other adverse environment applications.

Several new equipments using the new family of planar tubes and microwave circuit modules show significant improvements over their earlier prototypes. The most significant of these are the new broadbanded TACAN-DME transponders, high performance radar altimeters and new lightweight aircraft identification transponders for both military and commercial use.

Planar ceramic tubes and their circuitry can be used in a variety of concepts at frequencies up to 10 gigahertz and are well suited for broadbanded amplifier applications. Several internal and external geometries are available and also a variety of microwave functions normally relegated to more complex and expensive microwave devices.

Figure 29 is an attempt in a very general way to compare the performance and features of the various microwave devices. The reader must realize that it is very difficult, for example, to compare a traveling wave tube with a gridded tube, since these devices are used in radically different applications. This chart provides a first-look, best choice selection of the active microwave device. The type of circuitry must then be used to best fit that choice.

MICROWAVE DEVICES

	T.W.T.	T.D.A.	C.F.A.	KLYS.	V.T.M.	S.S.	TUBE
POWER	KW'S	MW'S	KW'S	MW'S	WATTS	MW'S	WATTS
FREQ.	MM	X	X	KU	C-X	KU	X
EFF.	35%	-	60%	55%	70%	-	70%
SIZE	MED.	SMALL	MED.	LARGE	MED.	MED.	SMALL
WT.	LBS.	OZS.	LBS.	LBS.+	LBS.-	LBS.	OZS.
B.W.	WIDE+	WIDE	WIDE+	WIDE-	-	WIDE	WIDE-
G-B.W.	HI++	MED.	HI+	HI	-	LOW	MED.
DYN. RNG.	MED.	LOW	-	HI	-	MED.	HI
N.F.	LOW	LOW+	-	HI	-	MED.	MED.
RELIB.	GOOD	GOOD+	POOR	GOOD	AVE.	GOOD	AVE.
W/$	LOW	LOW--	HI+	HI	AVE.	LOW-	HI+

FIG. 29

NOTES

PLANAR TRIODE LIFE AND RELIABILITY SUMMARY

Results from Adverse Environment Tests

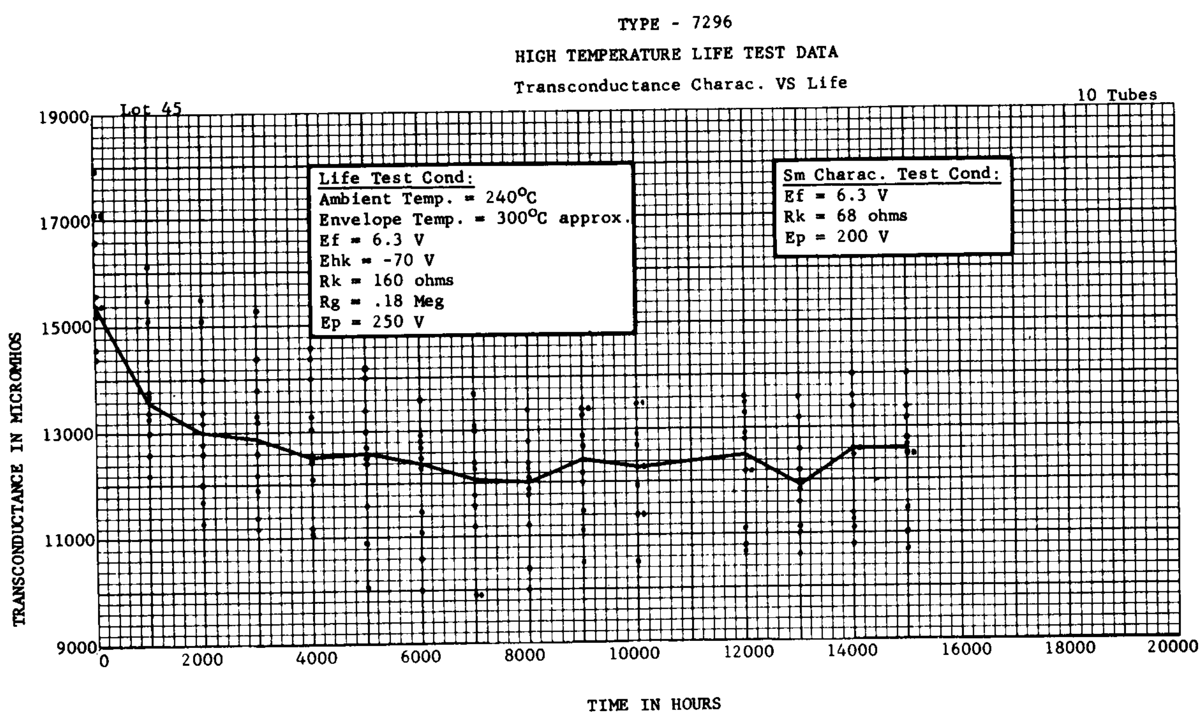
Temperature and Humidity

While it is generally recommended that the published temperature ratings not be exceeded where emphasis is on long and reliable life, some interesting long-life evaluations at higher-than-rated temperatures have been made as a matter of design capability study. A summary of these tests is as follows.

Type	Lot	Amb. Temp.	Env. Temp.	Ef*	LT.Duration	n
7296	472	400°C	450°C	5.4V	2000 Hr.	10
7296	305	500°C	550°C	4.3V	4000 Hr.	10
7296	45	240°C	300°C	6.3V	15000 Hr.	10
7296	46	240°C	300°C	6.3V	15000 Hr.	10
Z-2354	253	400°C	450°C	5.0V	17000 Hr.	10

* Note that lots 472 and 305 of the 7296 and lot 253 of the Z-2354 were life tested at reduced heater voltage. This was done to obtain longer tube life by keeping the cathode temperature within bounds.

These data demonstrate the capability of reliable operation at higher than rated temperatures provided that due considerations are given to proper heat sinking and commensurate derating of the heater voltage. As an example of these tests, a Transconductance vs Time graph of 7296 (Lot #45) is in the following graph.



In addition to the high temperature evaluations, the effects of high humidity environment have been investigated with regard to absorption of moisture into the ceramic and seal areas. The test consisted of a sample of type 7768 tubes subjected to steam vapor of approximately 100°C and 95-100 percent relative humidity. These conditions were in accordance with MIL-STD 1311A, Method 1011 with the exception that the duration was extended to 1000 hours. At the completion of this test, the tubes were checked for electrical characteristics and found to have withstood the steam bath with no deleterious effects.

Mechanical

Planar tubes ability to withstand severe mechanical stresses, such as might be encountered in missile applications, is included in the regular acceptance criteria of the test specifications. Vibration fatigue testing is performed through the range of 30-2000 Hz at acceleration levels up to 30 g for a duration of 6 hours to assure that the tubes are free from mechanical resonances. In addition, tubes are subjected to mechanical shock at a typical level of 450 g for 1 millisecond duration. Test experience has shown the design capability of these tubes to be generally well in excess of the actual test requirements. For higher levels of shock and vibration, the bonded heater versions of the planar triode family is recommended.

II. Results of Production and Engineering Quality Control Tests

(a) Shelf Life or Storage

It may be appropriate here to make an observation about shelf life. Although normally taken for granted, this can be especially important in certain applications where the tubes are held non-operating for long periods of time but expected to function properly when the equipment is finally turned on. One such evaluation was made on a group of 65 type 7077 tubes which were held in storage for nearly 8 years from 1/25/60 to 12/22/67. Test data of the electrical characteristics were recorded before and after this holding period and the tubes were found to have remained essentially unchanged. Similar investigations on planar tubes have likewise shown that degradation during extended storage periods is not a significant problem.

(b) Operation Life

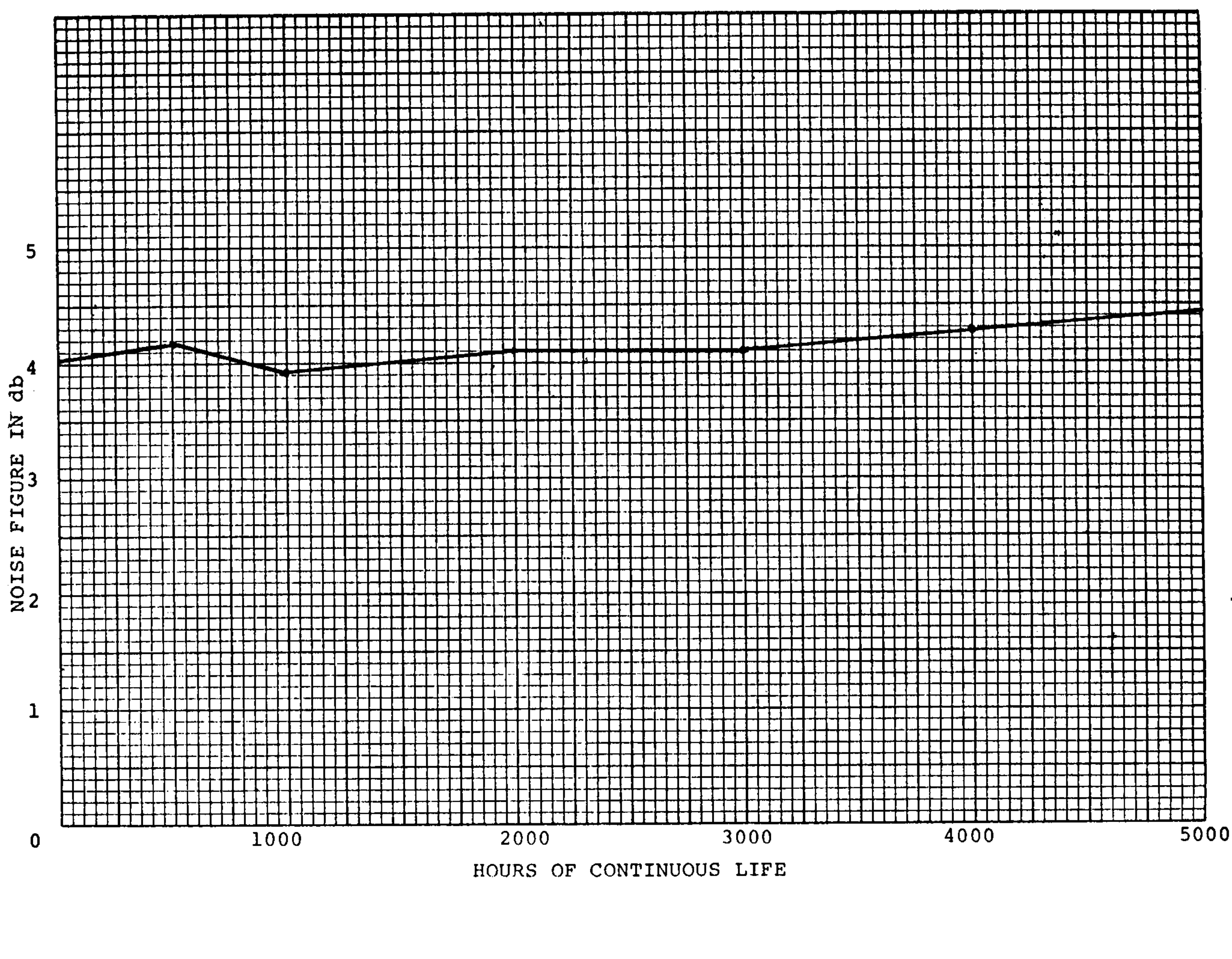
As a part of the regular lot acceptance testing, each lot is sample tested under operating conditions which are typically set at the maximum rated values for plate dissipation, cathode current, and plate voltage for 1000 hours duration. Exemplary failure rates determined from accumulations of these data are as follows.

Type 7077 (Small Signal RF Amplifier)

Results of 7077 life tests have consistently indicated a very good reliability. Cumulative 1000 hour test data during a recent production period show a failure rate of 0.4%/1000 hrs. (4 defectives out of 960,000 tube hours) giving an MTBF of 250,000 hours. This low failure rate is typical of that experienced over several years production.

The above data was taken under DC conditions with various performance criteria determined at down period intervals. One of the most important criteria of the 7077 is noise figure. The following graph is a plot of this recorded rf performance on 50 tubes run to 1000 hours, 25 of which were extended out to 5000 hours life test. Noise Figure was measured at a frequency of 450 MHz.

TYPE 7077 NOISE FIGURE LIFE



TYPES 7911, GE13971, GE18651 (PULSED AMPLIFIER OR OSCILLATOR TYPES)

Cumulative results of life tests under plate pulsed oscillator operating conditions are as follows:

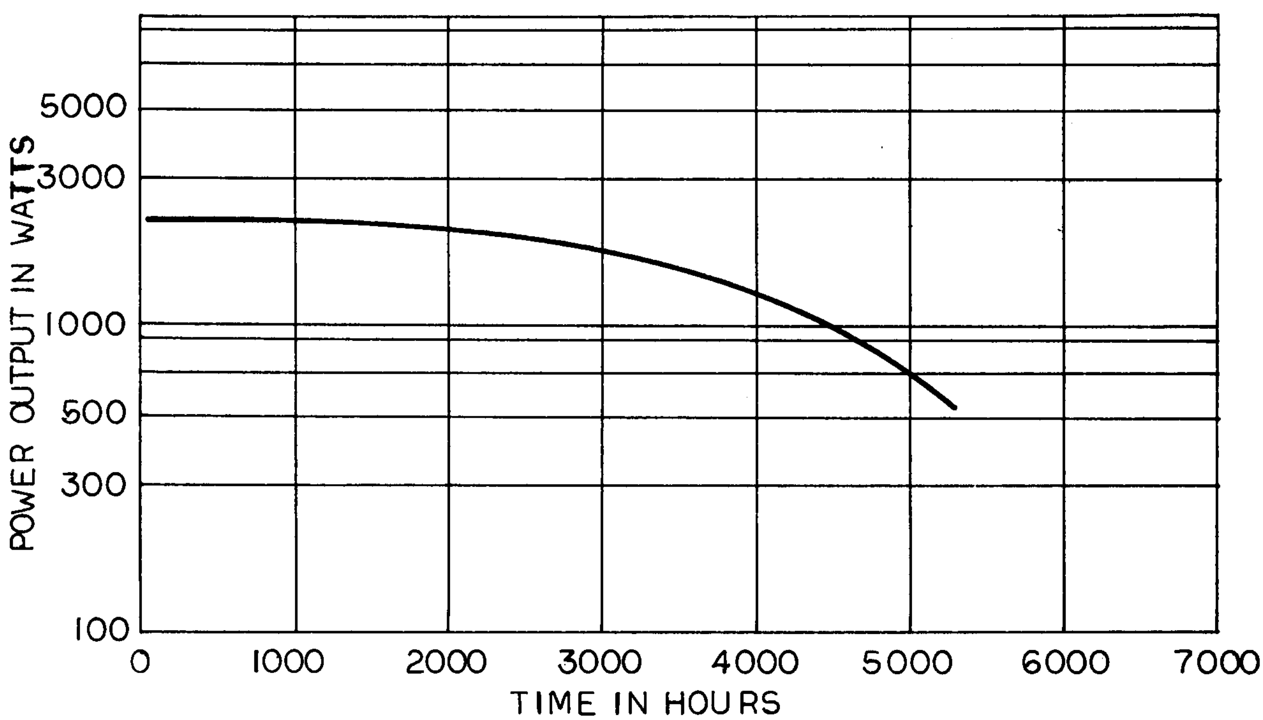
TYPE	LOT	SAMPLE	HOURS	FAIL
7911	68-09	5	5000	-
	68-06	5	5000	-
	68-01	5	5000	-
	67-50	5	5000	-
	67-46	5	5000	-
GE13971	X6	4	4000	-
	X7	3	3000	-
	X8	2	2000	-
	X9	3	3000	1
	X10	3	3000	-
	X11	3	3000	-
	X12	4	4000	-
	X13	4	4000	-
	X14	3	3000	-
	X15	3	1000	-
	X16	4	4000	-
GE18651	A	4	4000	-
	A6-7	4	4000	-
	A8	4	4000	-
	A11	4	4000	-
	B	4	4000	-
	C2	4	4000	-
	C3	4	4000	1
	D	4	4000	-
	9E	3	3000	-
	69-49	4	4000	-
	TOTAL:	100 tubes	100,000 tube hrs.	2 defectives

Failure Rate = 2%/1000 Hrs. (MTBF = 50,000 Hrs.)

IN–SERVICE LIFE RESULTS

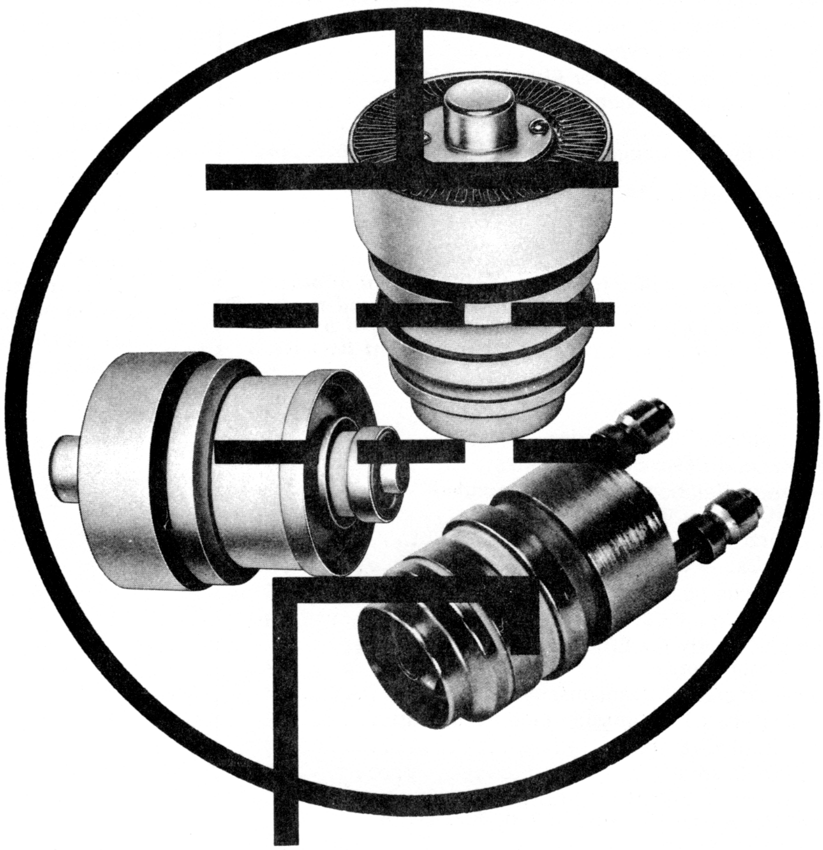
Recent life tests were conducted in two transmitter-amplifier chains for a new DM⌣ design. This amplifier was part of a Distance Measuring Equipment life tested under simulated field conditions. This amplifier chain had a bandwidth of 13% centered around 1100 MHz. In this equipment, acceptable performance is defined as a minimum power output of 500 W.

TRANSMITTER TEST



medium power vhf-uhf coaxial tetrodes

**BASIC CONSIDERATIONS
PERTINENT TO BROADBAND
RF POWER AMPLIFIER CIRCUITS
FOR TRANSMITTER APPLICATIONS**



Tube Department

GENERAL ELECTRIC

Table of Contents

ILLUSTRATIONS:

Introduction

The evolution of the modern vacuum tube covers a long and fascinating period of technological achievement. In fact, it has often been considered that the first real vacuum tube was made and studied by Thomas Edison as long ago as the year 1883. At that time he was doing intensive work on the development of the incandescent lamp, and he noticed that if an additional terminal were placed within the vacuum, a current would flow to the filament when this terminal was made positive. Subsequently, other renowned scientists performed experiments with these Edison bulbs, and they formulated the fundamental laws of emission from incandescent materials in a vacuum. Dr. Lee DeForest is usually credited with use of a third electrode or grid, thereby creating a new tool which he called the "Audion." This marked the beginning of the "negative-grid tube" and the start of a long and enduring era in electronics and the field of electrical engineering. The years after saw great strides in the evolution of the vacuum tube as new concepts for envelopes, electrodes, and emitters were introduced and applied. Today the "negative-grid tube" continues to find wide acceptance and use in a variety of modern equipments, some of which are extremely vital to our nation's defense and livelihood. One of the more noteworthy areas of comparative recent achievement is the application of these tubes in extremely broadband RF circuits for radar transmitters. Electronic bandwidths on the order of 20 percent or more are being achieved now in the VHF-UHF frequency range. The objectives of this brochure are to acquaint the equipment designer and user with the General Electric Company's line of "negative-grid" tubes for pulsed transmitter service and to present some of the basic considerations pertinent to their application in broadband, RF power amplifier circuits.

Coaxial Tetrodes: what are they?

High performance VHF-UHF transmitting tetrodes by General Electric satisfy peak power output requirements in the approximate range of 1 to 50 kilowatts at frequencies up to the 1500 MHz region.

These tubes (shown typically in Figure 1) are of coaxial design with concentrically aligned screen-grid and control-grid structures surrounding a cylindrical, unipotential, oxide-coated cathode.

To illustrate the general configuration and design features common to the line, Figure 2 depicts General Electric's ZP-1065 tetrode, and its basic subassemblies. Because of its typical nature, the ZP-1065 and its associated characteristics will serve as a basis for much of the discussion which follows.

Figure 1 — GE High Performance Pulse Tetrodes

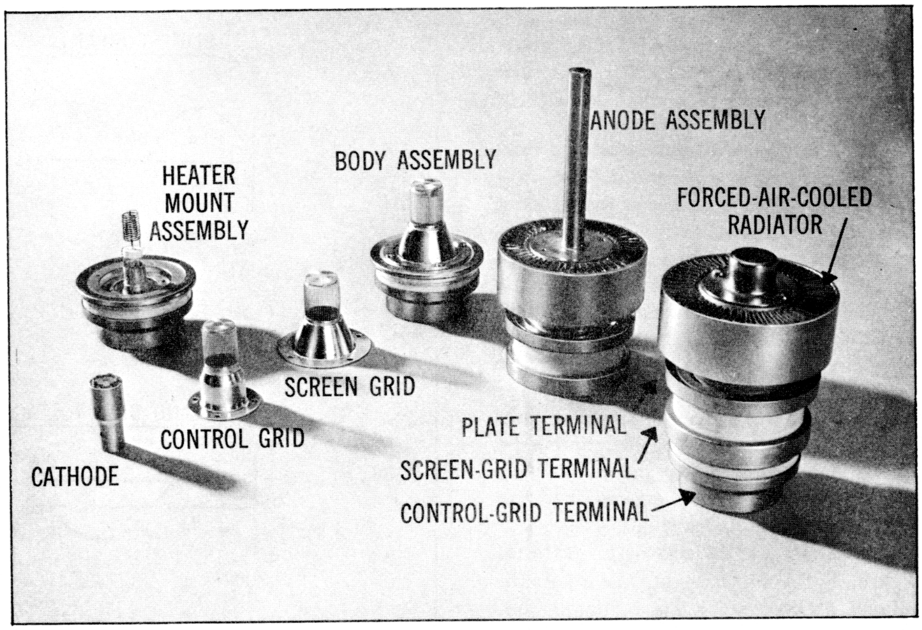


Figure 2 — ZP-1065 Tetrode and Basic Subassemblies

Component Parts of the Coaxial Tetrode

The high performance VHF-UHF transmitting tetrode consists of these elements:

Cathode: Just one of the proven features of this tube, the cathode system used in the ZP-1065 is typical of the oxide-coated cathodes used in GE's pulse tetrodes which have demonstrated excellent service life in the field. The nickel-base emitter is coated with a mixture of barium, strontium and calcium carbonates that change into oxides during the manufacturing process. Operation of the oxide emission system depends on the controlled production of barium through the reduction of its oxide by an active ingredient in the nickel-base metal. Operating temperature and concentration of the reducing element determine the rate of reduction. Oxide-coated emitters are characterized by their ability to supply extremely high emission levels . . . while operating at relatively low cathode temperatures. This design feature provides high peak current for RF pulsed service, while the low thermal requirement is conducive to long life expectancy.

Control-grid and Screen-grid: Wires in the grid structures are made of molybdenum to provide mechanical strength . . . an important consideration in avoiding small deformations that would otherwise tend to alter tube characteristics. Moreover, molybdenum displays good heat conductivity as well. The grid wires are gold-plated to minimize grid emission and are welded into copper supporting cones, which furnish good thermal conductivity to the exterior surfaces of the tube.

Careful attention is given during manufacture to precise alignment of the control-grid and screen-grid wires to realize uniform characteristics, minimize electron interception, and provide high performance from tube to tube.

Insulators: Low loss, high-purity alumina ceramics are used for the electrical insulators throughout the tube structure. Metalized by a special processing technique which GE developed and patented, these ceramics are brazed to concentric ring seals. Special care given to the design of the screen grid-to-anode ceramic results in a low value of associated capacitance. The straight-sided dimension is sufficiently long to allow reliable operation with the high levels of anode voltage required under pulsed RF service.

Anode: Oxygen free, high conductivity copper is the standard material used in anode fabrication for all tubes in this product line. Of the common metals, copper has the highest heat and elctrical conductivity and is readily formed by spinning, drawing and machining. The anode cooling configuration can take a variety of forms, depending on the requirements of the equipment designer or user. For example, forced-air cooling can be provided by either a transverse or axial flow radiator, and liquid or heat-sink-cooled anode designs exist for special applications where these cooling techniques may be preferred.

Tube Terminals: The use of ring-type tube terminals facilitates adaptation to cavity circuit configurations. Ample surface for spring-finger contacts is provided to assure positive electrical connection. The concentric ring-seal construction has successively larger diameters for the control-grid, screen-grid, and anode terminals to facilitate their insertion into circuits comprised of coaxial-line or waveguide cavities.

RF Circuitry: How applied!

Grounded-Grid Service: Each tube has been designed to operate as a grounded-grid amplifier with the screen-grid tied to the control-grid as far as RF potentials are concerned.

In this form of operation, the input circuit is connected between the control-grid and the cathode, and the output circuit between the screen-grid and anode. The control grid is held constant at zero a-c potential, while the cathode voltage varies about the zero potential line. During the amplification process, a negative-going cathode causes the plate voltage to drop while the plate current increases. RF voltage and current waveforms are shown in Figure 3 to aid in visualizing their phase relationships in a grounded-grid circuit.

RF Cavities: While any conventional RF circuitry may be employed with these tetrodes, their construction is ideally adapted to either coaxial or waveguide lines which are appropriate for the range of frequencies within their general application area. This covers frequencies from approximately 200 MHz to the upper frequency limit of the tubes, as high as 1500 MHz in most cases.

In a line-type cavity, the tube and stray circuit capacitances generally form most of the capacitive reactance of the resonant circuit. The "cavity" is configured to provide an equal inductive reactance when viewed from the tube terminals. Coupling arrangements are provided for the input and output resonators to match the impedances of the tube as determined by specific operating conditions of voltage and current for a given application. Suitable bypass capacitors for interrupting d-c continuity are typically made by using thin sheets of dielectric material between two plates or cylindrical surfaces.

Very Broad Bandwidth Applications: These applications require use of multi-tuned resonators for the output circuit. Design of the input circuit usually is less critical and more straightforward.

A broadband, waveguide "cavity" is shown in Figure 4 to illustrate typical design and configuration features. Developed by the Bendix Communications Division, this circuit is deployed in the transmitter of an advanced phased array radar system. The main body of the cavity is made of flat plates and formed sheet metal, joined by self-tapping screws. Approximate outer dimensions are 9 x 7 x 1¾ inches for a quarter-wave (λ/4), double-tuned output circuit and 1 x 1 x 9 inches for the single-tuned input configuration.

The output section with its cover or anode connecting plate removed is shown in Figure 5. A sliding-short fitted with contact fingers provides for an output tuning range through the adjustment of line length.

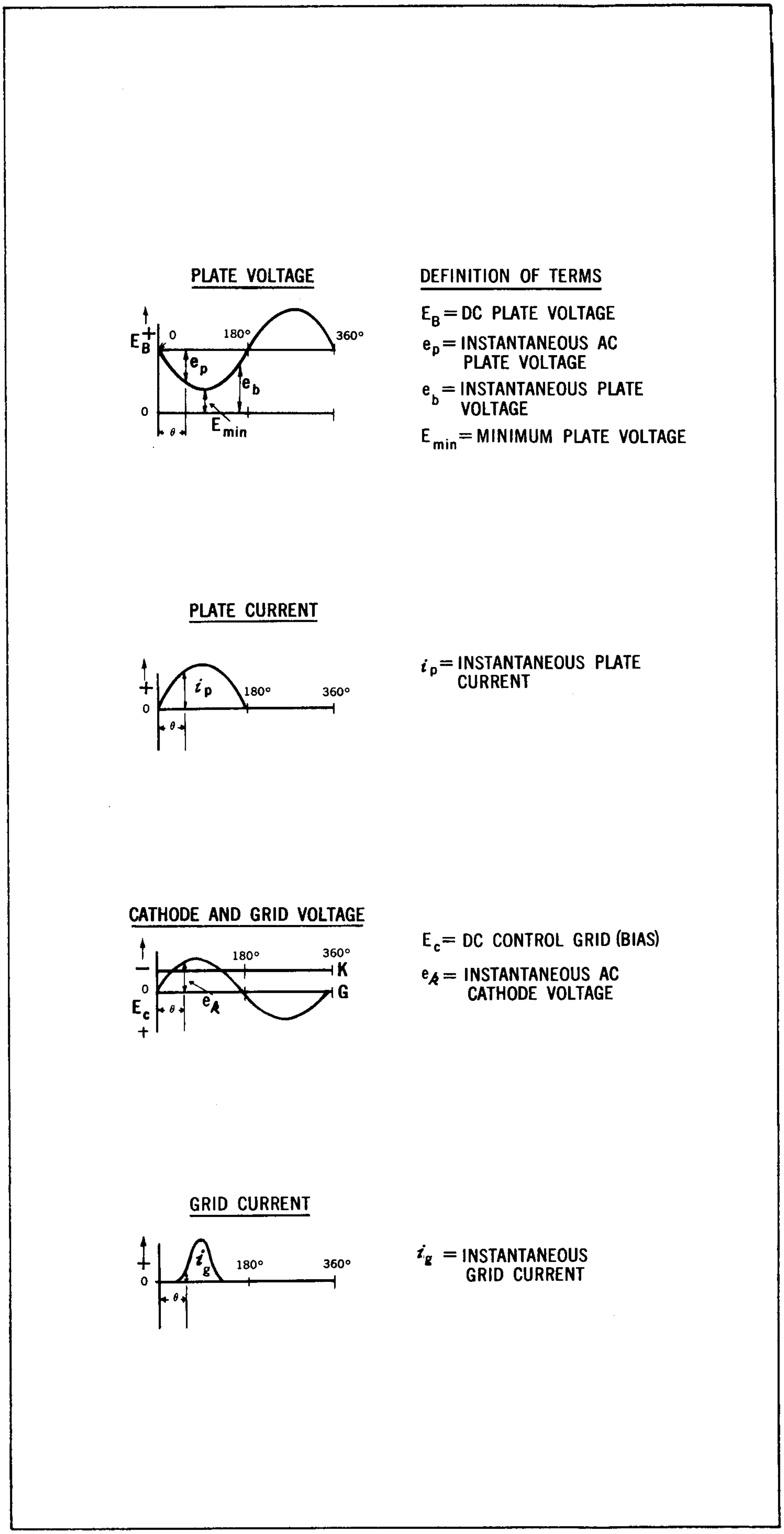


Figure 3 — RF Voltage-Current Relationships in a Grounded-Grid Amplifier (Class B Operation Shown)

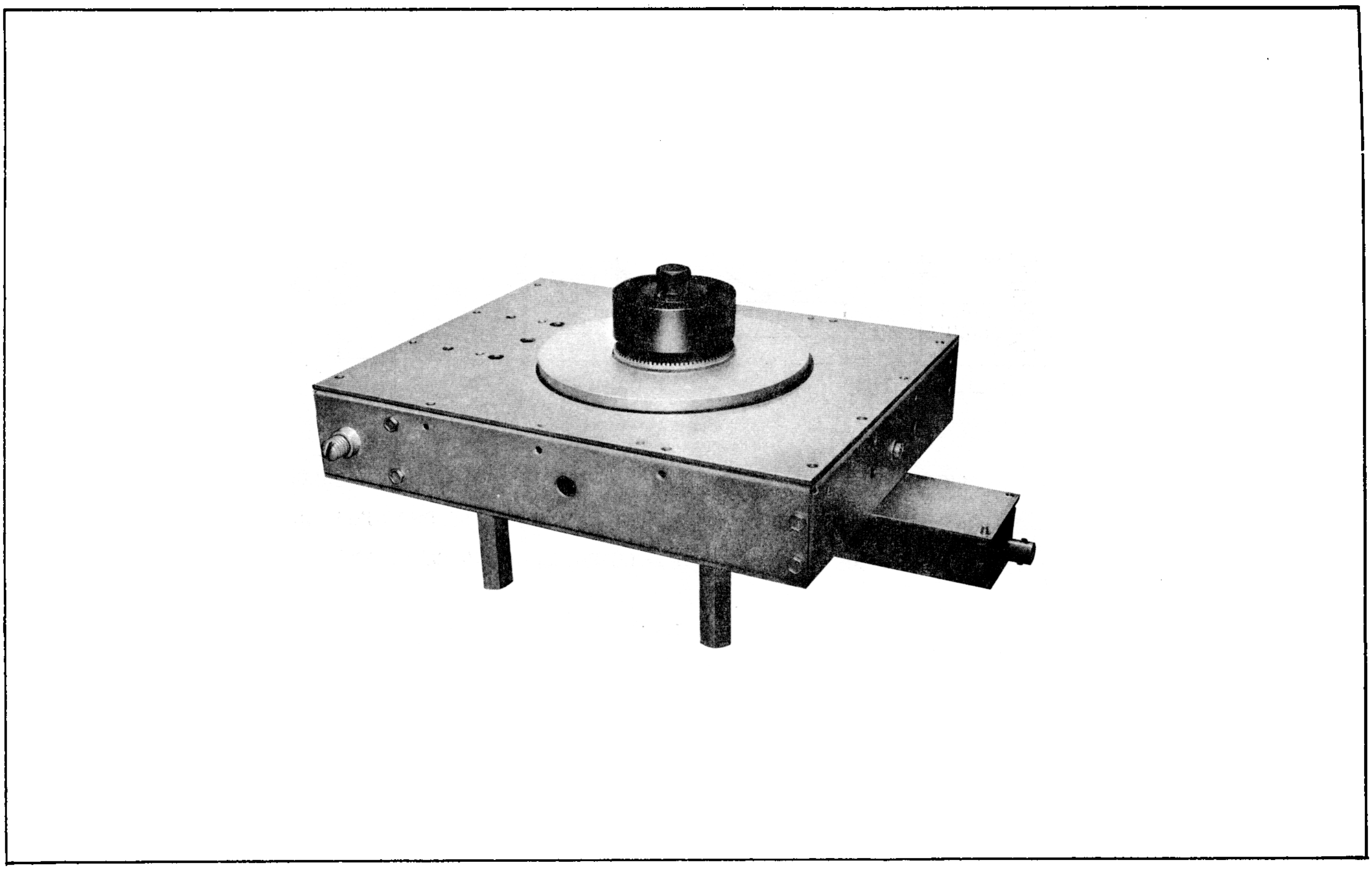


Figure 4 — Typical Broadband RF Cavity

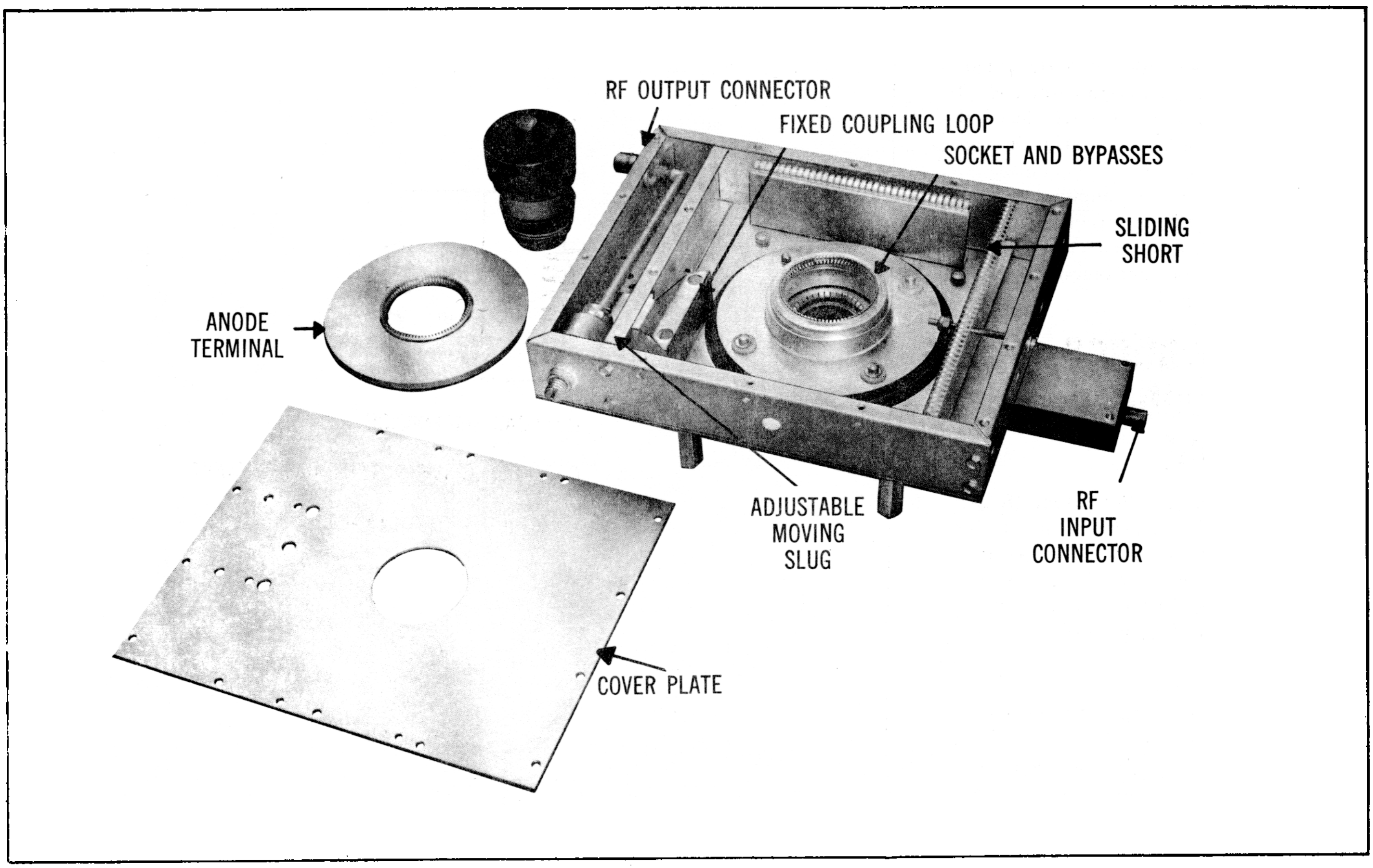


Figure 5 — Internal View of Cavity Showing Output Section

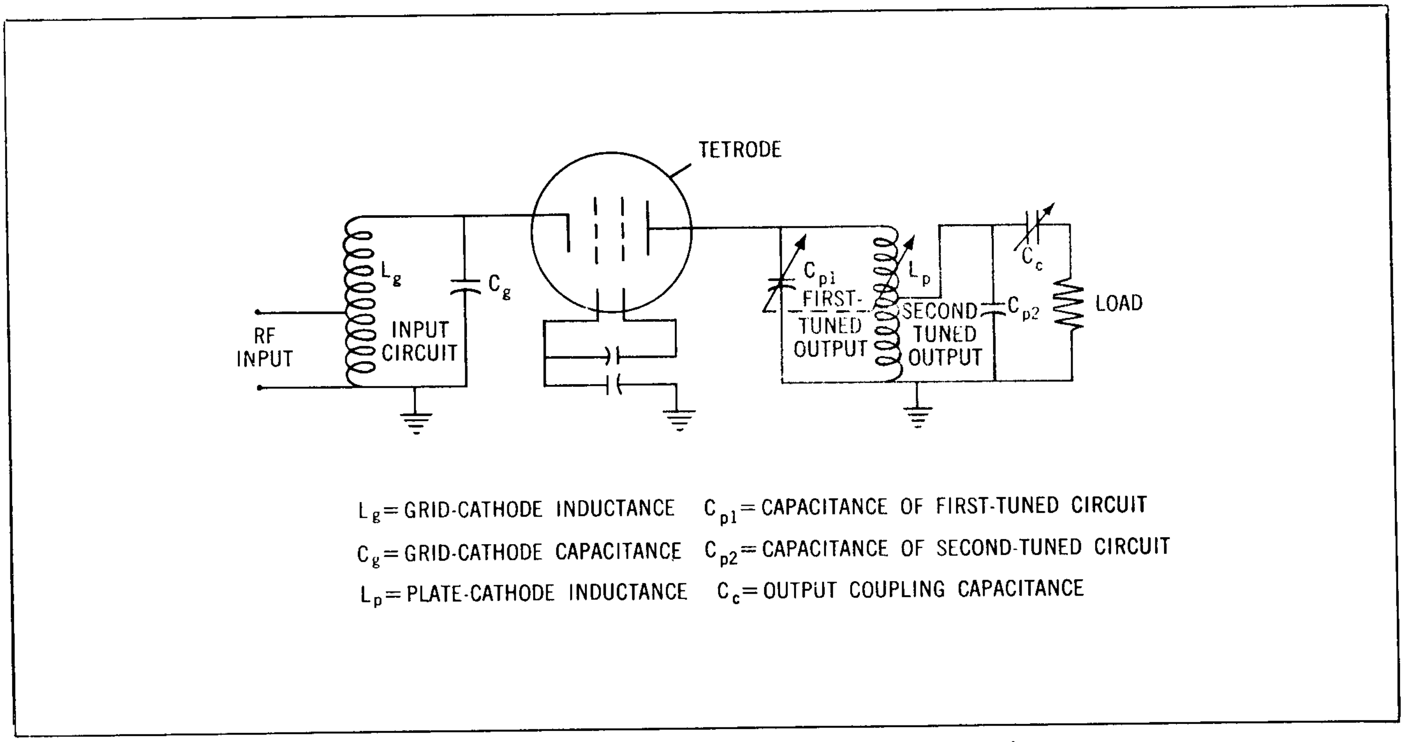


Figure 6 — Simplified RF Schematic for Broadband Cavity

A fixed loop arrangement couples the first tuned circuit (which incorporates the tube) to the second tuned output circuit. Coupling from the output to the load is accomplished by a shunt capacitance formed by an adjustable, moving slug.

The cavity is designed to operate at a center frequency of 542 MHz with a nominal tuning range of 500-560 MHz and an electronic bandwidth of approximately 50 MHz at the 3-dB point. The RF electrical equivalent for the cavity is given in Figure 6 for purposes of clarification.

Nominal RF power amplifier performance includes 30 kilowatts of useful peak power output, 0.005 duty factor, 250-μsec pulse length, 50-percent output efficiency and 10 dB of gain.

Considerations for broadband operation

Impedance Bandwidth Product ($R_L \Delta f$): To satisfy broad bandwidth requirements in power amplifiers, the RF circuitry must be designed to effect low values for both the effective output capacitance and the operating load impedance presented to the tube.

The relationship of these parameters is demonstrated in Figure 7:

$Q = 2\pi f_o R_L C_{eff}$ . . . for a parallel resonant circuit

and . . . $Q = \dfrac{f_o}{\Delta f}$

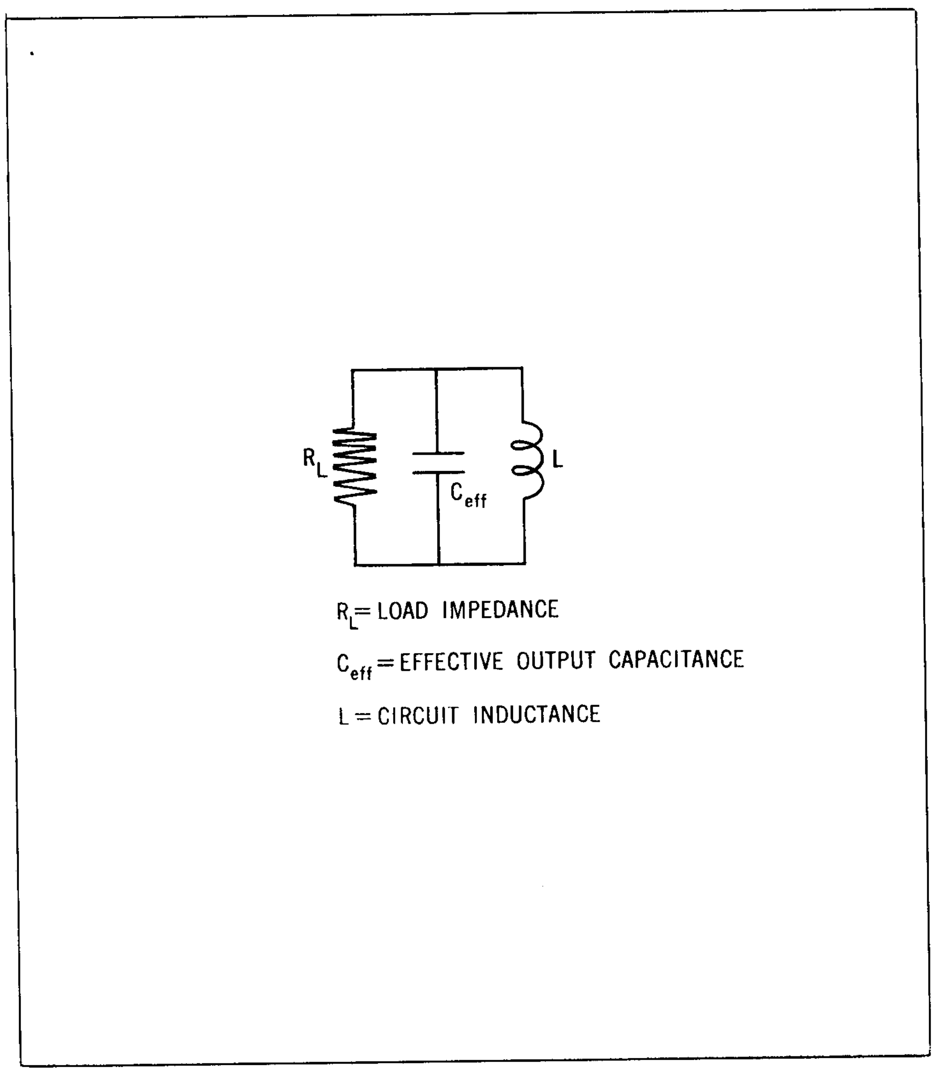


Figure 7 — Parallel Resonant Circuit

A parameter $R_L\Delta f$ can be derived from the above equations according to . . .

$$R_L\Delta f = \frac{1}{2\pi C_{eff}}$$

. . . where R_L is the load impedance in ohms, Δf is the required singly-tuned 3-dB bandwidth in hertz, and C_{eff} is the effective output capacitance in picofarads. Additional consideration is given to the $R_L\Delta f$ or "impedance-bandwidth product" in following paragraphs.

Measurement of Effective Output Capacitance (C_{eff}): The value of C_{eff} is frequency dependent and is related to the electrical length of the internal metal parts of the tube, the size and shape of the ceramic seal, and the electrical length of the circuit external to the tube envelope. The effective capacitance decreases as frequency increases and an increasing portion of the tube circuit lies within the tube envelope. The converse is true when frequency is lowered . . . so the value of C_{eff} will then approach the d-c capacitance of the tube. This relationship may be explained further as follows:

Referring to Figure 7, the expression for the operating center frequency at resonance is presented as . . .

$$\omega_0^2 = \frac{1}{LC} = \frac{1}{L} \times C^{-1}$$

Differentiating the above with respect to ω_0, we have:

$$2\omega_0 = -\frac{1}{L} \times C^{-2} \times \frac{dC}{d\omega_0}$$

Upon re-arranging terms . . .

$$C^2 = \frac{1}{2\omega_0 L} \times \frac{dC}{d\omega_0}$$

and, since $\omega_0 L = \dfrac{1}{\omega_0 C}$, and $\omega_0 = 2\pi f_0$

then, $C = C_{eff} = -\dfrac{f_0}{2} \times \dfrac{dC}{df_0}$

This derivation provides a practical method of determining effective circuit capacitance for a given frequency and tube output configuration in a given resonating circuit. A perturbation technique is used to obtain measured test data for the value of dC/df_0. This information is obtained through the use of a specially constructed dummy tube . . . such as the one illustrated in Figure 8. The structure shown incorporates a micrometer centrally located at the tube's active anode-grid region where the effective and d-c capacitances are equivalent.

This mock-up is placed in an appropriate external circuit or cavity, as shown in Figure 8, to resonate at f_0, the desired center frequency. By changing the amount of projection of the micrometer probe, the output capacitance is perturbed to provide the term dC or ΔC.

A low-frequency bridge is used to measure the value of capacitance for each position of the probe. The resonant frequency of the tube and cavity combination is also measured for each setting of the micrometer to yield data for df_0 or Δf_0.

Readings of capacitance and frequency for common positions of the micrometer probe are then plotted. The negative slope of the resulting line provides the value of $\dfrac{dC}{df_0}$ or $\dfrac{\Delta C}{\Delta f_0}$. The effective capacitance is then calculated by substituting these experimentally determined values in the original expression of:

$$C_{eff} = \frac{f_0}{2} \times \frac{dC}{df_0}$$

Effects of Foreshortened Cavity Resonators on $R_L\Delta f$: Consider the "impediance-bandwidth" product ($R_L\Delta f$) and its relationship to transmission line or cavity resonator circuits of the type used at the higher frequencies. The basic circuit under consideration has a characteristic impedance Z_0, a length l, and a foreshortening capacitance C_{eff}, as shown in Figure 9.

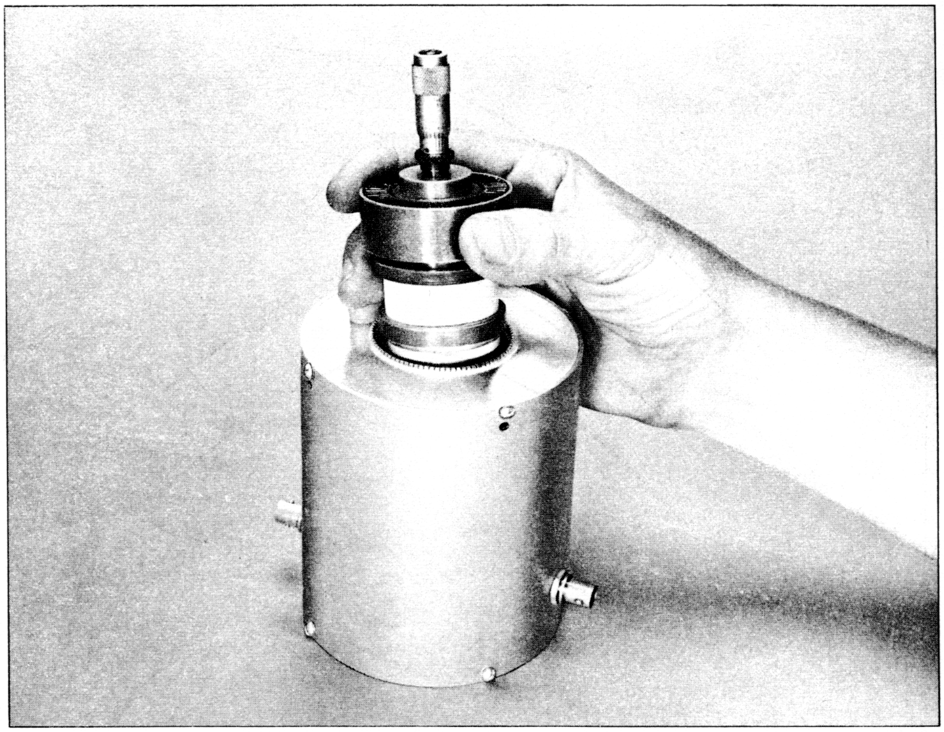


Figure 8 — Dummy Tube and Cavity for Effective Capacitance Measurement

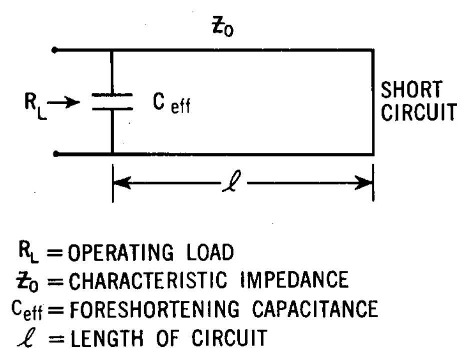


Figure 9 — Capacitively Foreshortened Transmission Line

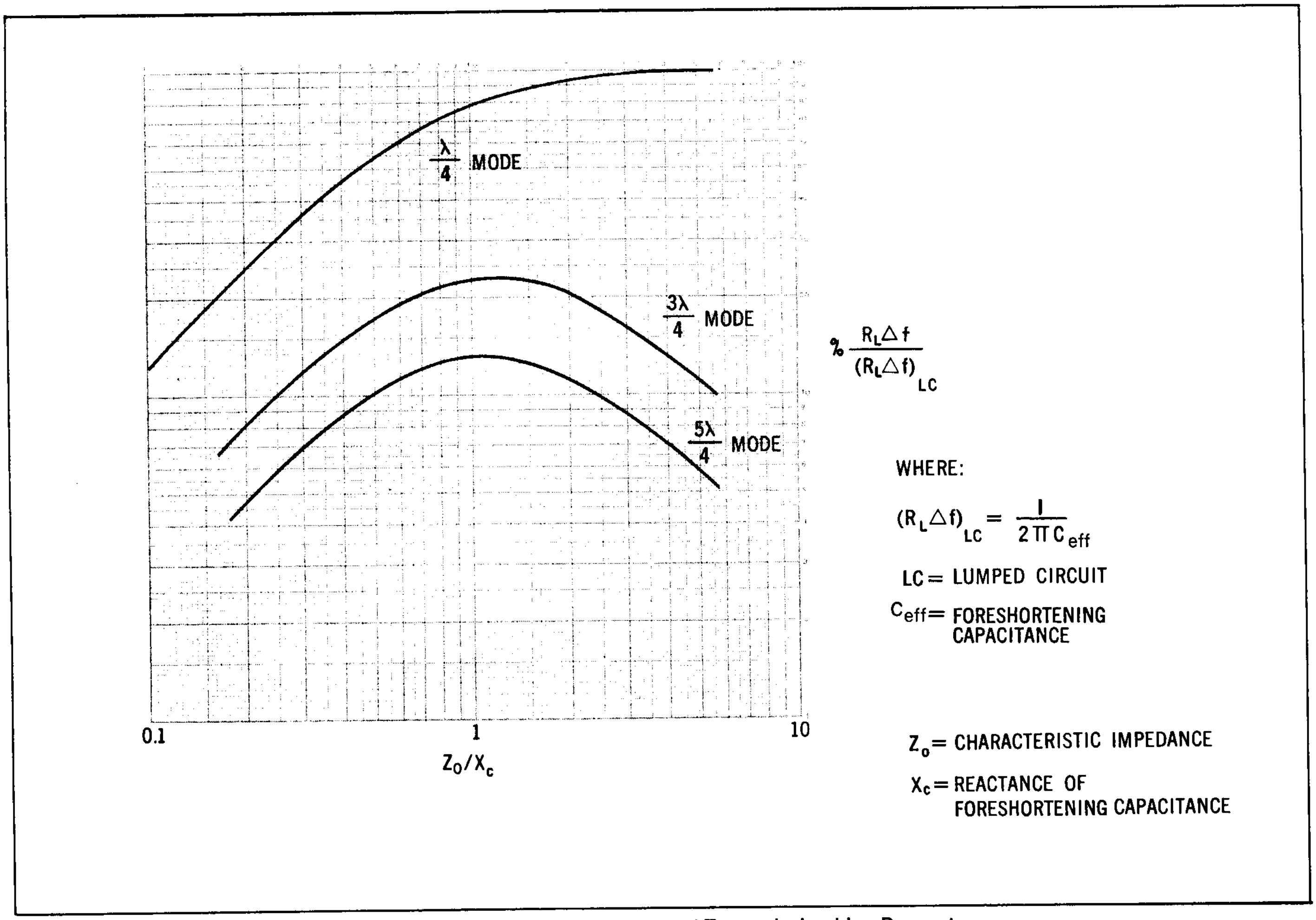


Figure 10 — $R_L\Delta f$ of Foreshortened Transmission Line Resonators

By computing the Q of this transmission line on an energy basis, the $R_L\Delta f$ can be determined and results summarized, according to Figure 10. Since the lumped circuit case has maximum $R_L\Delta f$ for a given C_{eff}, the curves are shown as a percentage of the $R_L\Delta f$ obtainable with $(R_L\Delta f)_{LC}$ taken as 100 percent. The "impedance-bandwidth" product is a function of the degree of foreshortening. Therefore, the abscissa scale is given as Z_0/X_c, where X_c is the reactance of the foreshortening capacitance at the resonant frequency.

For the quarter-wave ($\lambda/4$) resonator, $R_L\Delta f$ asymptotically approaches the lumped circuit value as Z_0 increases. Generally, this dictates that high characteristic impedance transmission lines be used in broadband cavities.

The Figure 10 curves also illustrate that $R_L\Delta f$ decreases very considerably with the higher order modes of resonance. For the same operating load impedance, a $\frac{3}{4}\lambda$ resonator yields less than one-third the bandwidth, while the $\frac{5}{4}\lambda$ resonator can supply less than one-fifth of the bandwidth of the $\lambda/4$ circuit. To achieve maximum broadband operation only the $\lambda/4$ mode should be used for the output cavity circuit which includes the tube.

Bandwidth (Δf) Improvement with Multi-Tuned Circuits: The continuing trend toward the use of broad electronic bandwith in advanced VHF-UHF radar transmitters is the result of special system requirements such as frequency agility, pulse compression, and high data rates. It is becoming common practice to use multi-tuned output circuits in these applications to achieve maximum utilization of the bandwidth and performance ability of a given negative-grid tube type.

Several technical papers have dealt with rigorous design of filter and impedance matching networks consisting of chains of coupled resonant circuits capable of producing the exact amplitude characteristics* desired. Results of some of these studies and mathematical analyses will be used to indicate the relative bandwidth improvement factors of multi-tuned . . . as compared to single-tuned circuits. This same information will also be used in an example to demonstrate the analysis of the RF power amplifier performance of a GE tetrode in a typical broadband radar transmitter.

Consider the multi-tuned circuits in Figure 11 with "n" branches . . . where n=1, 2, 3 etc. for single-, double-, and triple-tuning, respectively. The exact response shape considered here is the "Butterworth" type, also known as "maximally flat," "critically coupled," or "transitionally coupled." For the type of circuits under consideration, a series of power-bandwidth curves has been developed in Figure 12 to show the relative frequency response obtainable for a given power output. Figure 12 also includes plots of "Amplitude" of response in both "Volts" and "dB" as the ordinate, versus "Bandwidth" as the abscissa. The unit or reference point on this graph is the 3-dB bandwidth for the example of the single-tuned circuit (where n=1). The relative bandwidths of the different circuits for any level at the edge of the band can be read directly from these curves.

*See for example:

(1) "Amplitude-Frequency Characteristics of Ladder Networks" by E. Green— published by Marconi's Wireless Telegraph Co. (1954)

(2) "The Design of Dissipative Band Pass Filters Producing Desired Exact Amplitude Frequency Characteristics"— Proc. IRE, Vol. 37 (September 1949)

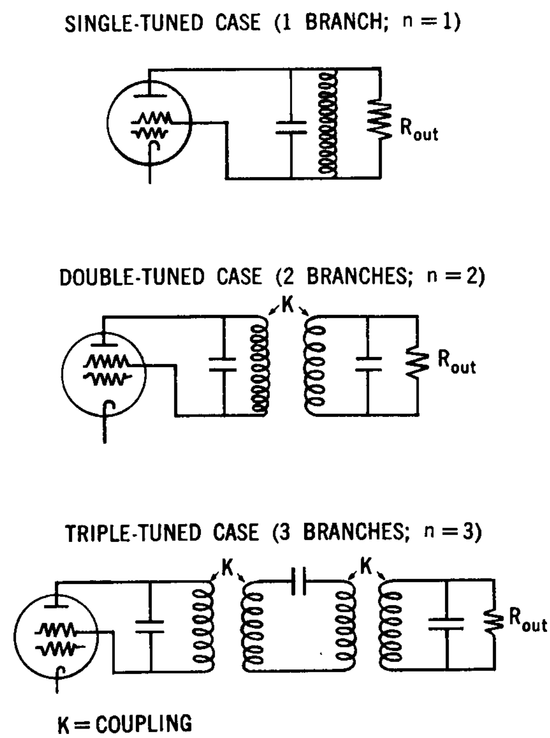
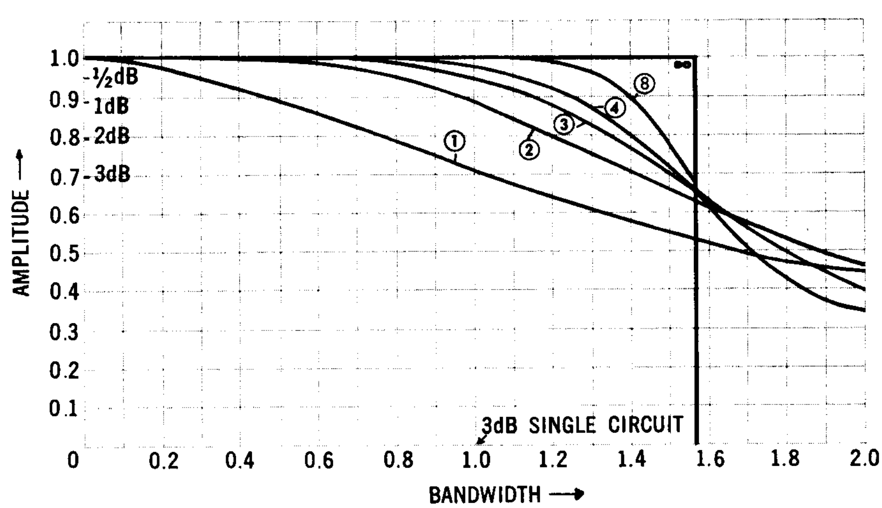


Figure 11 — Multi-Tuned Output Circuits



Extracted from E. Green "Amplitude Characteristics of Ladder Networks" published by Marconi's Wireless Telegraph Co. Ltd. (1954).

Figure 12 — Power-Bandwidth Curves

LEVEL AT BAND EDGE	S-T RELATIVE TO △f	D-T RELATIVE TO △f	T-T RELATIVE TO △f	Q-T RELATIVE TO △f
⅙ dB	0.2X	0.62X	0.86X	1.0X
½ dB	0.34X	0.82X	1.0X	1.15X
1 dB	0.5X	1.0X	1.2X	1.28X
3 dB	Δf	1.41X	1.5X	1.53X

LEGEND

S-T = SINGLE-TUNED OUTPUT CIRCUIT
D-T = DOUBLE-TUNED OUTPUT CIRCUIT
T-T = TRIPLE-TUNED OUTPUT CIRCUIT
Q-T = QUADRUPLE-TUNED OUTPUT CIRCUIT

Figure 13 — Relative Bandwidth of Multi-Tuned Output Circuits ("Butterworth" Response), Referred to 3-dB Bandwidth of Single-Tuned Circuit (Δf)

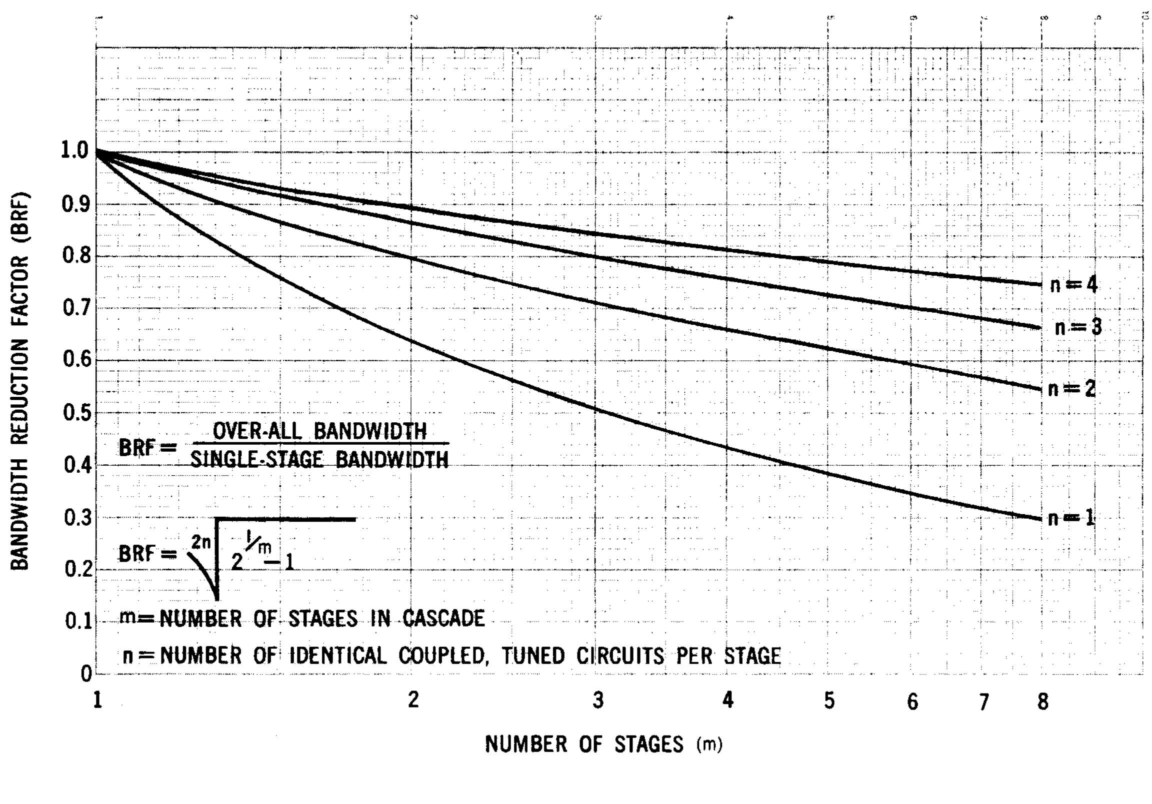


Figure 14 — Bandwidth Reduction Factor Versus Number of Stages

A representative set of the more commonly used values is given in Figure 13 for easier reference. Significant advantages are readily apparent from Figures 12 and 13. For example:

By designing a triple-tuned output circuit, the 3-dB bandwidth of a single-tuned RF amplifier may be improved by as much as 50 per cent. This improvement is obtained: 1) while delivering the same power output at band center, and 2) while providing a more constant impedance over the band.

Where some in-band ripple may be allowed, even greater bandwidth improvements can be obtained. Further, the phase-frequency function may be tailored to a given requirement. These cases are not treated here.

Bandwidth Reduction Factor (BRF): In a typical radar transmitter comprising several cascaded tuned amplifiers, the overall or system bandwidth is less than that of the individual stages in the line-up. An analytical expression for the effect of cascading multi-tuned, transitionally coupled amplifiers is . . .

$$\frac{\text{Overall Bandwidth}}{\text{Single Stage Bandwidth}} = 2n\sqrt{2^{1/m}-1}$$

where m = the number of cascaded stages of identical bandwidth

and n = number of coupled, tuned circuits per stage.

To maintain a given system bandwidth, it is therefore necessary to increase each stage's bandwidth . . . as the number of stages in the amplifier chain is increased. A graph of this function . . . to be called the "bandwidth reduction factor" . . . is presented in Figure 14.

Consider three stages of triple-tuned, transitionally coupled circuits. From Figure 14, the overall bandwidth is 0.8 times that of the single-stage case. This reduction is **considerably less** than the corresponding reduction of three stages of single-tuned circuits. Generally, the greater the number of tuned circuits per stage, the lower the amount of reduction of system bandwidth. The primary objective of reviewing "bandwidth reduction factor" here is to emphasize that attention must be given to defining the bandwidth requirements of each stage to provide for the proper overall system response.

illustrative example for calculating operating conditions

Introduction

The trend toward the use of broadband radar transmitters makes it desirable to have a method for calculating the RF power amplifier performance of tubes applied in multi-tuned circuit arrangements. The accurate prediction of operating conditions is of particular interest to the equipment designer who must assess a tube's suitability for a given application, and who must also obtain information relative to the design characteristics of associated circuitry. Operating conditions can be calculated according to a semi-graphical analysis that utilizes the static characteristics of the tube. The method used may be briefly described with reference to Figure 15 in which the plate, control grid, and screen grid currents are presented as functions of the plate and drive (signal) voltages on a typical constant-current curve. During operation of the tube, the RF signal and plate voltages vary sinusoidally about the d-c grid bias and d-c plate voltage, respectively. Instantaneous values of these waveforms determine simultaneous currents or "operating points" such as is given at point "P" for a 45-degree angle. As the signal and plate sine waves vary through their successive phase angles over a full RF cycle, the resulting operating points describe a straight line called the "operating load line". Class B operation is given here, which means that no current is drawn during the third and fourth quarters of the voltage cycles (this would also be true for a Class C RF amplifier).

In practice the "operating load line" is drawn on the constant-current characteristic by connecting the points A and A′, and the voltage waveforms need not be drawn. The quiescent operating point, A, is determined by the d-c plate voltage and the d-c grid bias that are applied to the tube. Point A′ is obtained from the intersection of the minimum plate voltage and the maximum instantaneous plate current that occur during the RF cycle. The distance of each operating point from point A is determined by multiplying the sine of its associated phase angle by the overall length of A to A′. This has been done in Figure 15 and the various instantaneous operating points have been located. The use of 10-degree intervals provides satisfactory accuracy. Current values of each of these points are read directly from the curves, interpolating where required. This information may be used to plot the current waveforms, although it isn't necessary to do so. The objective is to determine the average and fundamental components of current, and this is accomplished through the use of formulae that have been developed through Fourier analysis (illustrated in the example which follows). These current components are then used to give the power output, power input, output efficiency, and other parameters of interest in describing amplifier performance. To illustrate these techniques for calculating tube operating conditions, the following example will be used, making reference to various considerations presented previously whenever appropriate.

Example

Consider first that there is a requirement for a broadband, RF power amplifier stage having the basic application objectives outlined below:

Frequency	:	425 MHz
Electronic Bandwidth	:	50 MHz at 1 dB
Peak Power Output	:	10 KW nominal
Pulse Width	:	10 μsec
Duty Factor	:	0.01

Knowledge of available tube types suggests using GE's ZP-1065 metal-ceramic tetrode. This dependable GE tube type offers features and characteristics that lend themselves well to the needs of the service indicated. Tube performance is to be calculated according to the semigraphical analysis of current waveforms, as discussed in the Introduction to this section. Basically, the approach involves these factors:

1) estimating the required resistive load (R_L) to be presented to the tube
2) defining the boundary conditions for the load line which is to be drawn on the constant-current characteristics of the tube
3) performing calculations by a series of successive approximations to yield 10KW of useful peak power output with the required load (R_L).

Complete tube amplifier performance under matched load conditions will be determined and operating conditions tabulated in the following paragraphs. Definitions for the various parameters and related identifying symbols are:

E_B	:	DC plate voltage
E_{c2}	:	DC screen-grid voltage
E_{c1}	:	DC control-grid voltage
E_{min}	:	Minimum plate voltage
E_{p1}	:	Peak value of the fundamental component of plate voltage
E_{g1}	:	Peak value of the fundamental component of control-grid voltage
I_B	:	DC plate current
i_p	:	Instantaneous plate current
I_p	:	Maximum instantaneous plate current (or peak RF plate current)
I_{p1}	:	Peak value of the fundamental component of plate current

I_{c2} : DC screen-grid current
i_{g2} : Instantaneous screen-grid current
I_{g2} : Peak value of the fundamental component of screen-grid current
I_{c1} : DC control-grid current
i_{g1} : Instantaneous control-grid current
I_{g1} : Peak value of the fundamental components of control-grid current
R_L : Resistive tube load
P_o : Plate power output
P_{in} : DC plate input
P_p : Plate dissipation
P_{g2} : Screen-grid dissipation
P_{g1} : Control-grid dissipation
η : Output efficiency
η_{ckt} : Circuit efficiency

1) Determination of Resistive Load (R_L): The first step is to determine the value of the resistive load, recalling that...

$$R_L \Delta f = \frac{1}{2\pi C_{eff}}$$

where Δf = Single-tuned 3 dB bandwidth and C_{eff} = Effective output capacitance

From the application objectives, the bandwidth for the stage under analysis is given as 50 MHz at 1 dB. This assumes that the bandwidth reduction factor has already been taken into account. Since this is an extremely broadband condition, use of a multi-tuned output circuit is indicated.

Triple-tuning is the approach chosen to maximize the load resistance that is presented to the tube. Undue complication of the circuit design will be avoided at the same time. For a Butterworth or transitionally coupled type response, it becomes apparent from Figure 13 that...

$$\text{Single-Tuned 3 dB Bandwidth} = \frac{\text{Triple-Tuned 1 dB Bandwidth}}{1.2}$$

$$\text{or } \Delta f = \frac{50 \text{ MHz}}{1.2}$$

$$\Delta f = 41.6 \text{ MHz}$$

Using the perturbation technique described earlier, the effective capacitance experimentally is found to be...

$$C_{eff} = 9.9 \mu\mu f$$

Substituting the values of Δf and C_{eff} in the formula for impedance bandwidth...

$$R_L = \frac{1}{2\pi \Delta f C_{eff}}$$

$$R_L = \frac{1}{6.28 \times 41.6 \times 10^6 \times 9.9 \times 10^{-12}}$$

$R_L = 390\Omega$, approximately

It shoud be noted that at 425 MHz, the GE ZP-1065 operates in the quarter-

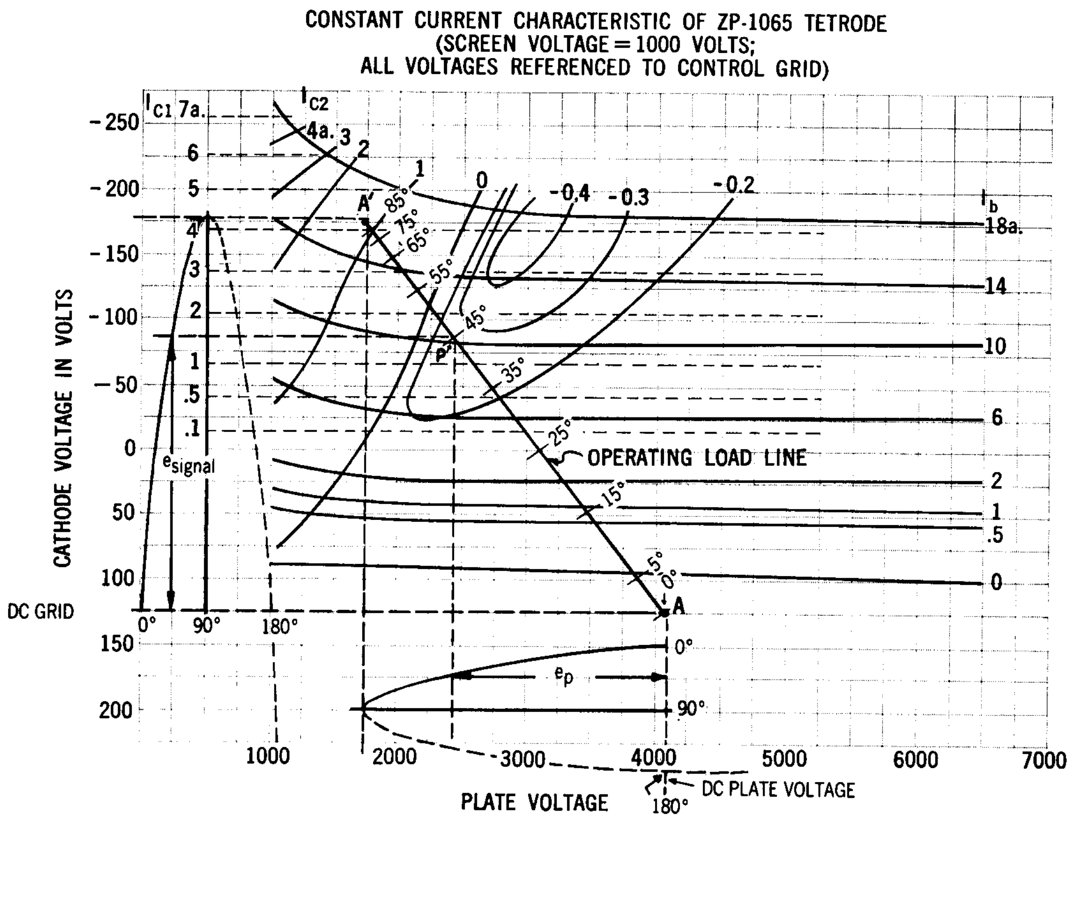


Figure 15 — ZP-1065 Tetrode Constant-Current Characteristic

wave ($\lambda/4$) output mode to effect maximum broadband operation. (Eariler, information was presented concerning the affects that foreshortened cavity resonators and higher order modes have on $R_L\Delta f$.)

Calculating operating conditions is now possible, having determined the value of resistive load (R_L) indicated for the application.

2) Defining Boundary Conditions for First Load Line

(a) **Approximation of Current Levels**

Taking into consideration that the peak power output objective is 10 KW and the indicated resistive load is 390Ω, the peak value of the fundamental component of plate current (I_{p1}) is obtained from . . .

$$P_o = \frac{I_{p1}^2 R_L}{2} \times \eta_{ckt}$$

Estimating η_{ckt} = 90 percent and substituting and solving for I_{p1}, we obtain . . .

$$I_{p1}^2 = \frac{2 \times 10 \times 10^3}{390 \times .9} = 57 \text{ a}^2$$

$$I_{p1} = 7.5 \text{ a}$$

Class B operation is used, since it generally leads to the best overall balance of tube performance relative to bandwidth, output efficiency and power gain. Approximate current relationships for Class B service are:

$I_{p1} = 1.7 \times I_B$, and

$I_p = 3.5 \times I_B$

Therefore,

$$I_B = \frac{I_{p1}}{1.7}$$

$$I_B = \frac{7.5}{1.7}$$

$$I_B = 4.4\text{A}$$

Also:

$I_p = 3.5 \times I_B$

$I_p = 3.5 \times 4.4$

$I_p = 15.4$ a

(b) **Approximation of Voltage Levels**

From: $E_{p1} = I_{p1} \times R_L$,

$E_{p1} = 7.5 \times 390$

$E_{p1} = 2920$ v

Also:

$E_{min} = E_B - E_{p1}$

Setting:

$E_B = 4100$ V

Then $E_{min} = 4100 - 2920 = 1180$ V

Also, set

$E_{c1} = -125$ V (approximate value for cut-off)

and $E_{c2} = 1000$ V (for high transconductance)

(c) **Summary of Conditions Chosen for First Load Line**

$E_B = 4100$ V

$E_{c2} = 1000$ V

$E_{min} = 1180$ V

$E_{c1} = -125$ V

$I_p = 16$ a (increased from 15.4 a)

The load line is now drawn on the constant-current characteristic and the graphical method of Fourier analysis is applied. Referencing Figure 15: point A is determined by the d-c plate voltage (4100 V) and the grid bias (−125 V). Point A′ is obtained by the intersection of the minimum plate voltage (1180 V) and the maximum instantaneous plate current (16 a).

3) Step by Step Calculation of Operating Conditions for Final Load Line

Instantaneous values of the tube currents are determined at intervals of 10 degrees along the operating line A-A′. These values are recorded in tabular form below and are combined in arithmetic operations to yield the desired parameters that define amplifier performance. The preceding calculations were for the first or approximately correct load line. The calculations following are based on the final load line. In this instance, the approximate and final load lines are the same.

θ	$\sin\theta$	i_p	$i_p \sin\theta$	i_{g2}	$i_{g2}\sin\theta$	i_{g1}	$i_{g1}\sin\theta$
5	0.087	—	—	—	—	—	—
15	0.258	.8	.2	—	—	—	—
25	0.423	4.0	1.7	—	—	—	—
35	0.574	7.8	4.5	−.2	−.2	.5	.3
45	0.707	10.3	7.3	−.3	−.2	1.4	1.0
55	0.819	13.0	10.6	.2	.2	2.4	2.0
65	0.906	14.5	13.1	.6	.5	3.2	2.9
75	0.966	15.5	15.0	.9	.9	3.8	3.7
85	0.996	16.0	15.9	1.2	1.2	4.2	4.2
Σ		81.9	68.3	2.4	2.4	15.5	14.1

The average or d-c value of plate current becomes:

$$I_B = \frac{\Sigma i_p}{18}$$

$$I_B = \frac{81.9}{18}$$

$I_B = 4.5$ A, DC during pulse

The peak value of the fundamental component of plate current is obtained from:

$$I_{p1} = \frac{\Sigma i_p \sin\theta}{9}$$

$$I_{p1} = \frac{68.3}{9}$$

$I_{p1} = 7.6$ a, during pulse

The output load impedance is determined by:

$$R_L = \frac{E_{p1}}{I_{p1}}$$

where $E_{p1} = 2920$ v

or

$$R_L = \frac{2920}{7.6}$$

$$R_L = 384\ \Omega$$

This value compares favorably with 390 Ω, or the amount of load impedance indicated to achieve the desired bandwidth. The peak power output may now be calculated from the expression:

$$P_o = \frac{E_{p1} \times I_{p1}}{2}$$

$$P_o = \frac{2920 \times 7.6}{2}$$

$P_o = 11.0$ kilowatts, peak

The values of screen-grid and control-grid currents are calculated according to the following expressions:

$$I_{c2} = \frac{\Sigma i_{g2}}{18}$$

$$I_{c2} = \frac{2.4}{18}$$

$I_{c2} = 0.1$ A, DC during pulse

and

$$I_{g2} = \frac{\Sigma i_{g2} \sin\theta}{9}$$

$$I_{g2} = \frac{2.4}{9}$$

$I_{g2} = 0.3$ a, during pulse

also

$$I_{c1} = \frac{\Sigma i_{g1}}{18}$$

$$I_{c1} = \frac{15.5}{18}$$

$I_{c1} = 0.9$ A, DC during pulse

and

$$I_{g1} = \frac{\Sigma i_{g1} \sin\theta}{9}$$

$$I_{g1} = \frac{14.1}{9}$$

$I_{g1} = 1.6$ a, during pulse

With the use of the various currents and voltages found above, further calculations may be made to determine a complete set of operating conditions and parameters as follows:

$$P_{drive}(\text{gnd-cathode}) = \frac{E_{g1} \times I_{g1}}{2}$$

$$P_{drive}(\text{gnd-cathode}) = \frac{305 \times 1.6}{2}$$

P_{drive}(gnd-cathode) = 244 watts, peak

Although the foregoing calculation for grounded-cathode driving power does not represent actual required grounded-grid driving power, it is useful in determining grid dissipation, bias loss, and feedthrough power. Continuing, therefore:

Bias Loss = $E_{c1} \times I_{c1}$

Bias Loss = $125 \times .9$

Bias Loss = 113 watts, peak

Since

$P_{g1} = P_{drive}$(gnd-cathode) − Bias Loss

$P_{g1} = 244\text{-}113$

$P_{g1} = 131$ watts, peak

and

$P_{g1}(\text{Avg}) = P_{g1} \times \text{Duty}$

$P_{g1}(\text{Avg}) = 131 \times 0.01$

$P_{g1}(\text{Avg}) = 1.3$ watts

also

$$P_{g2} = (E_{c2} \times I_{c2}) + \frac{E_{g1} \times I_{g2}}{2}$$

$$P_{g2} = (1000 \times .1) + \frac{(305 \times .3)}{9}$$

$P_{g2} = 202$ watts, peak

$P_{g2}(Avg) = P_{g2} \times Duty$

$P_{g2}(Avg) = 202 \times 0.01$

$P_{g2}(Avg) = 2.0$ watts

Additionally,

$P_{in} = E_B \times I_B$

$P_{in} = 4100 \times 4.5$

$P_{in} = 18.4$ kilowatts, peak

and from

$P_p = P_{in} - P_o$

$P_p = 18.4 - 11.0$

$P_p = 7.4$ kilowatts, peak

$P_p(Avg) = P_p \times Duty$

$P_p(Avg) = 7.4 \text{ kw} \times 0.01$

$P_p(Avg) = 74$ watts

For anticipated operation under grounded-grid conditions:

$$P_{drive}(\text{gnd-grid}) = \frac{I_{p1} + I_{g2} + I_{g1}}{2} \times E_{g1}$$

$$P_{drive}(\text{gnd-grid}) = \frac{9.5}{2} \times 305$$

$$P_{drive}(\text{gnd-grid}) = 1450 \text{ watts, peak}$$

Also, in grounded-grid service, a substantial portion of the driving power appears as useful feedthrough power in the output circuit. Thus:

$P(\text{feedthrough}) =$
$P_{drive}(\text{gnd-grid}) - P_{drive}(\text{gnd-cathode})$

$P(\text{feedthrough}) =$
$1450 - 244 = 1206$ watts, peak

In order to indicate useful power output as it would be measured in a resistive load, the circuit efficiency must be applied to the "electronic" output of the tube. At the operating frequency of interest, practical efficiencies of approximately 85-90 percent are typical.

Therefore:

$P_o(\text{useful}) = [P_o + P(\text{feedthrough})] \times \eta_{ckt}$

$P_o(\text{useful}) = [11.0 + 1.2] \times .85$

$P_o(\text{useful}) = 10.4$ kilowatts, peak

$$\text{Since } \eta = \frac{P_o(\text{useful})}{P_{in}} \times 100$$

$$\eta = \frac{10.4}{18.4} \times 100$$

$$\eta = 56.5\%$$

If $\eta_{input} = 85\%$

$$\text{then Power Gain} = \frac{P_o(\text{useful})}{P_{drive}} \times \eta_{input}$$

$$\text{Power Gain} = \frac{10.4}{1.45} \times .85$$

Power Gain $= \times\ 6.1$, or 7.9 dB

The calculated RF power amplifier and tube operating parameters are summarized in the following tabulation:

Summary of Calculations for Broadband RF Power Amplifier

Pulsed Operation; Class B RF Power Amplifier

RF Grid-Pulsed Service; Grounded-Grid Operation

Quarter-Wave ($\lambda/4$) Triple-Tuned Output Circuit

Matched Load Conditions; 425 MHz Center Frequency

DC Plate Voltage	4100 Volts
DC Screen-Grid Voltage	1000 Volts
DC Control-Grid Voltage	−125 Volts
DC Plate Current, during pulse	4.5 Amperes
DC Screen-Grid Current, during pulse	0.1 Ampere
DC Control-Grid Current, during pulse	0.9 Ampere
Drive Power, during pulse	1.7 Kilowatts
Power Output (useful), during pulse	10.4 Kilowatts
Power Gain	7.9 dB
Output Circuit Efficiency	85%
Input Circuit Efficiency	85%
Output Efficiency	56.5%
Output Impedance (R_L)	384 Ohms
Duty Factor	0.01
Pulse Width	10 μsec
Electronic Bandwidth at 1 dB	~50 MHz

a new, high performance broadband cavity for the ZP-1065

An RF cavity (Figure 16) has been developed for the General Electric ZP-1065 tetrode with the application objectives and operating conditions presented and analyzed in the preceding section. The performance achieved represents a highly significant contribution to the technological progress in extremely broadband, power amplifiers utilizing negative-grid tubes.

The RF cavity, developed by Microwave Cavity Laboratories, Inc., of LaGrange, Illinois, incorporates specialized multi-tuned circuit design techniques. The electronic bandwidth achieved is appreciably greater than that usually obtained with broadbanded high power tetrode cavities. A demonstrated 1 dB bandwidth of 50 MHz has been attained at a center frequency of 425 MHz . . . under the following operating conditions for 10 KW of peak power output:

Pulsed Operation; Class B RF Power Amplifier

RF Grid-Pulsed Service; Grounded-Grid Operation

Quarter-Wave (λ/4) Triple-Tuned Output Circuit

Matched Load Conditions; 425 MHz Center Frequency

DC Plate Voltage	4100 Volts
DC Screen-Grid Voltage	1000 Volts
DC Control-Grid Voltage	−150 Volts
DC Plate Current, during pulse	4.5 Amperes
Peak Power Output (useful), matched load	10 Kilowatts
Power Gain	>7 dB
Output Efficiency	54%
Duty Factor	0.01
Pulse Width	10 μsec
Electronic Bandwidth at 1 dB	50 MHz

The foregoing operation illustrates some of the capabilities and features of the ZP-1065 RF cavity package which are of particular interest and benefits to the transmitter equipment designer and user. For example, in addition to the extremely broad electronic bandwidth and high peak power output, a stage gain of greater than 7 dB, and a high output efficiency of approximately 55% are provided. This performance is achieved with the ZP-1065 tetrode operating under RF grid-pulsed amplifier conditions.

The tube receives special high voltage seasoning and testing in order to provide assurance of reliable operation and consistent performance from tube to tube under the indicated high levels of d-c steady-state and screen-grid voltages. This form of operation eliminates the need for a high-level modulator (for plate and screen pulsing) thereby conserving space and weight.

Microwave Cavity Laboratories' latest advances in coaxial cavity design and fabrication techniques have also resulted in a significantly small and lightweight package. With reference to the cavity shown in Figure 16, the overall body dimensions are approximately 6 inches (O.D.) x 7 inches long. The model is constructed of silver-plated aluminum and it weighs approximately 5 pounds, which is of interest to applications where light weight is important. Modifications, including different operating conditions, may be effected in order to custom design an RF power amplifier package to a particular application's needs.

Literature, sales information, or technical assistance on the application of General Electric medium power VHF-UHF Coaxial Tetrodes can be obtained from any of the following Electronic Component Sales Operation regional offices or by contacting the Microwave Tube Business Section, General Electric Company, Building 269, Schenectady, New York 12305. Telephone: (518) 374-2211, Extension 5-4273.

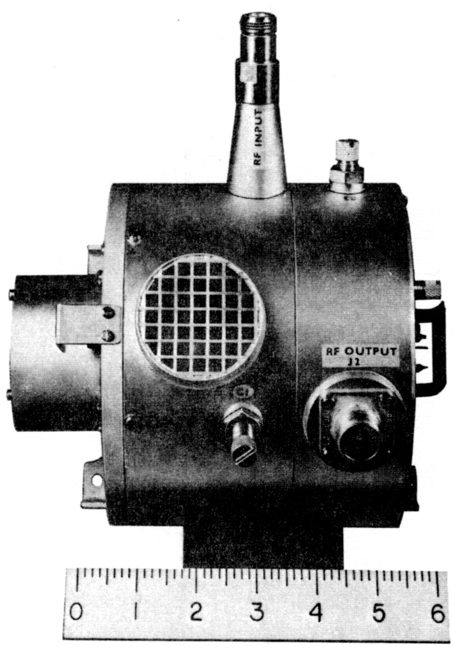
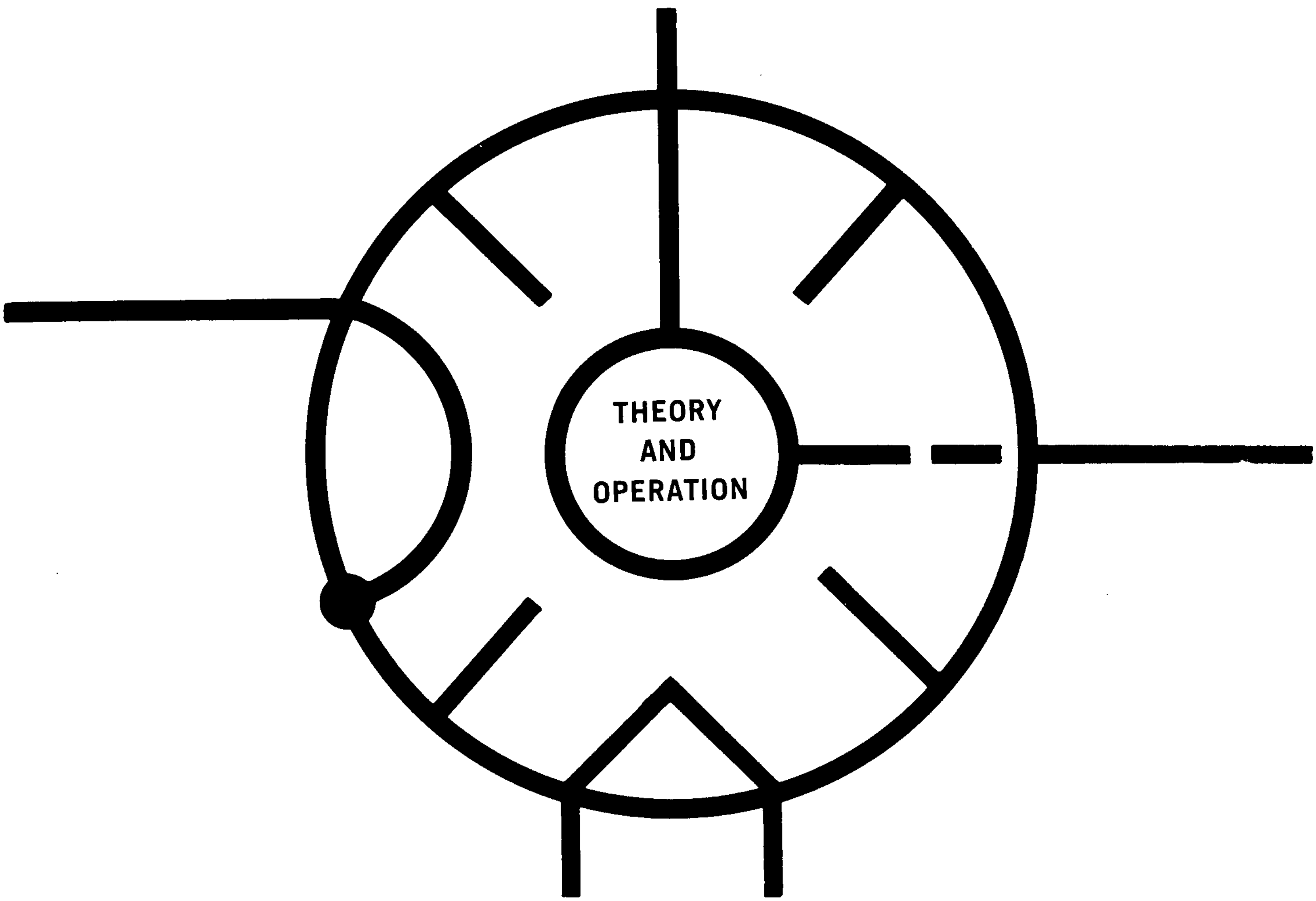


Figure 16 — High-Performance, Broadband Cavity for ZP-1065 Tetrode

NOTES

ELECTRONIC
INNOVATIONS
IN ACTION

voltage tunable magnetrons



TUBE DEPARTMENT

GENERAL ELECTRIC

Table of Contents

ILLUSTRATIONS:

VTM's: what are they?

Voltage Tunable Magnetrons (VTM's) are high frequency, continuous wave oscillators operating in the microwave region. General Electric VTM's cover a wide range of frequencies—from a few hundred to over 5,600 megacycles, and their capability has been demonstrated up into the X-band region.

Power output of voltage tunable magnetrons begins at tens of milliwatts and can be extended through hundreds of watts. General Electric has, in fact, attained 500 watts of power in the laboratory, and even higher levels are feasible depending on the center frequency and bandwidth being used.

A packaged voltage tunable magnetron consists of three elements:

(1) a basic vacuum tube wherein a conversion of d-c power into radio-frequency power occurs in the interaction region.

(2) an r-f circuit or cavity which presents the required impedance to the tube over the desired bandwidth.

(3) a permanent magnet which provides the required magnetic field.

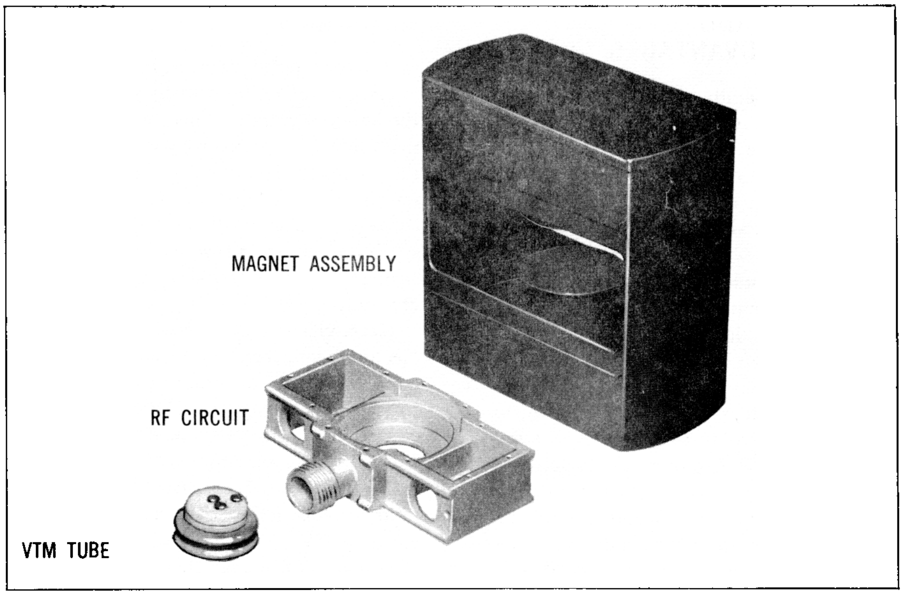


Figure 1—Elements of a Typical VTM

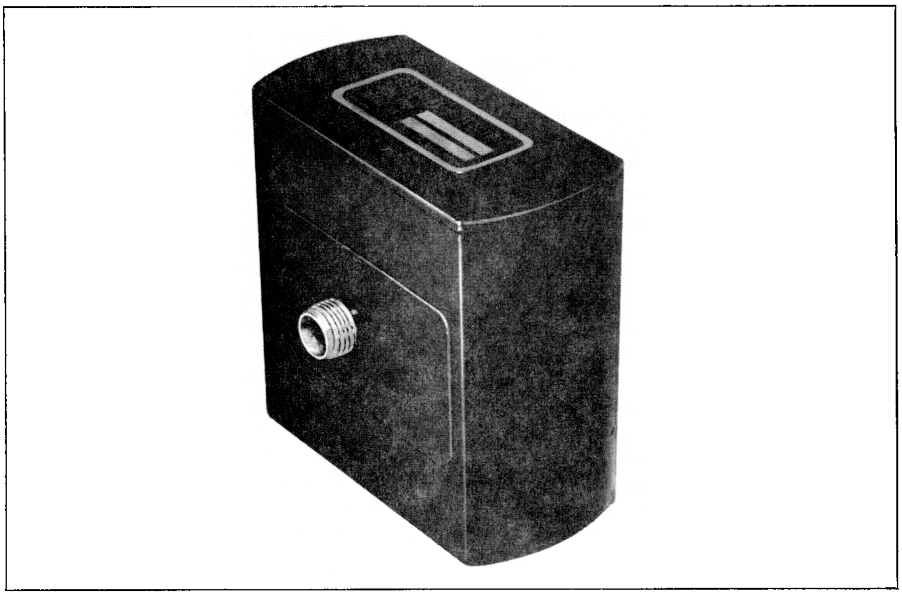


Figure 2—Typical VTM Assembly

KEY FEATURES AND ADVANTAGES

Essential features which distinguish a VTM from a conventional magnetron are:

(1) an r-f circuit loaded down to a very low Q.

(2) an electron current limited to a value less than the normal space-charge-limited (BRILLOUIN) current.

When these conditions are met, the oscillation frequency becomes a function of the anode voltage, rather than of the circuit resonant frequency.

Specific advantages of VTM's, in addition to their electronic tuning features, are:

(1) Rapid Modulation—VTM's are capable of being frequency-modulated at high rates. Sweeping rates up to 20,000 mc per microsecond have been attained.

(2) Linear Tuning—the VTM has a tuning characteristic (or frequency vs. anode voltage) which is not only linear, but also proportional. This means the tuning line passes through, or close to, the origin; hence, a good approximation to the tuning sensitivity (mc per volt) can be achieved simply by dividing the center voltage into the center frequency. Since this proportionality is an intrinsic characteristic of a VTM, one cannot specify center frequency, center voltage, and tuning sensitivity independently. For octave band tubes, the actual tuning curve is normally within $\pm 1\%$ (in units of center voltage) from the best straight line, and this line will pass within two or three per cent of the origin.

(3) Low Noise—the VTM can be constructed for low noise operation. IF noise, 30 mc from the carrier, may be approximately 95 db/mc below the carrier signal level.

(4) High Efficiency—high powered VTM's (75 watts and up) attain conversion efficiencies of 65 to 70 per cent.

(5) Size and Weight—VTM's operating at 10 watts over 35% bandwidths are available in one-pound packages. Size is 1⅝ inches in diameter and 1⅝ inches high in a cylindrical shape. At other levels, weight varies roughly as the square root of the power.

(6) Power Variation Across the Band—VTM power variations across an octave band can be restricted to four decibels with the use of a matched load or adequate isolation.

In addition to these features, the VTM can be made adaptable to airborne and space application environments where extremes in shock, vibration and temperature all may be encountered.

VTM TYPES AND GENERAL ADVANTAGES

The VTM family is divided into three major groups: the low power group of tubes up to one watt in power output, the intermediate group with a power output of one to ten watts and the high power VTMs ranging from tens to hundreds of watts.

The low power group is most often used in low noise applications for local oscillators, electronically tunable signal sources, test equipment such as signal generators and on wide band receivers requiring frequency agility.

The intermediate power VTM is an excellent device for fusing, altimetry, telemetry and parametric pump applications.

High power VTMs are used in ECM barrage jamming, broadband transmitters and missile and aircraft applications where their high efficiencies can be exploited.

VTMs are usually custom developed to perform one particular function in one specific application. Experience on past programs has shown that when pertinent system knowledge is obtained prior to VTM construction the result is an economic, well integrated system device.

Part of the construction procedure used to obtain optimum VTM performance for a given application lies in correctly orienting the vacuum tube-cavity combination with the magnetic field generated by the magnet. Emphasis is placed on gaining the best performance for those parameters most important to the application. This is how a VTM is customized for maximum performance at the factory. The importance of tailoring the VTM to its specific application cannot be overemphasized. Careful discussion and compiling of specifications for VTM operation is the only logical first step toward obtaining a satisfactory device; hence the potential user of a VTM is urged to follow this procedure.

general theory

CROSSED FIELD ACTION

The conventional high Q magnetron is a cylindrical diode wherein the electron current from the cathode is influenced by a magnetic field parallel to and coaxial with the cathode, and acting at right angles to the applied radial, electric field. When electrons travel in a direction perpendicular to the magnetic field, the field imposes a force at right angles to the direction in which the electrons are moving. This causes the electrons to spiral into orbit at a velocity directly proportional to the electric field applied between cathode and anode, and inversely proportional to the magnetic field. An illustration of this effect appears in Figure 3.

BUNCHING

Random noise present in the tube induces some radio-frequency voltage on the anode segments. This, in turn, tends to modulate the electron beam and build up the intensity of the radio-frequency fields on the anode structure. Figure 4 depicts three electrons rotating in the interaction space at an instant when adjacent anode segments are negatively and positively charged.

Here, electron A is moving in a reduced electrical field region caused by the radial component of the radio-frequency electric field acting in opposition to the d-c electric field. Thus, its velocity is decreased. Electron B is passing through an area of unmodified radial electric field; therefore, it maintains its initial velocity. Finally, electron C's velocity is advanced since it is moving in a higher electrical-field region where the radial radio-frequency field augments the applied d-c field. For this reason, electrons A and C tend to close in on B; furthermore, the same effect occurs at every position around the anode where the field orientations are the same as at B. Thus, the electrons form into a number of bunches equal to one-half the number of anode segments in the interaction space.

At each of these bunch positions (such as at B) there is a tangential r-f field, tending to retard the bunch. However, just as the radial field causes electrons to move tangentially in a magnetron (as in Figure 3), so a tangential field causes them to move radially. Thus the effect of the retarding force on the bunch is not to slow it down; rather it is to make the electrons move out towards the anode, lose potential energy, and contribute energy to the r-f field as they do so.
(Note that the average angular velocity of an electron around the cathode is proportionate to the d-c radial electric field on that electron.)

R-F POWER GENERATION

The continuous passage of these electron bunches past the anode induces radio-frequency fields on the anode structure. For voltage tunable magnetrons, the frequency of these fields is controlled by the average electron velocity; hence, by the anode voltage.

Interactions in a voltage tunable magnetron are similar to those in a high-Q magnetron in that an emitter produces electrons which enter into an interaction space, become bunched, and induce radio-frequency fields in an anode structure. In wide band voltage-tuned operation however, the r-f beam reactive current component is used to tune the circuit; therefore, the magnitude of the circulating beam reactive current component must be of the same order of magnitude as the circuit reactive current. This condition may be satisfied, and reasonable output powers produced, by using a heavily loaded (or low Q) radio-frequency anode circuit. Also the number of electrons injected into the interaction space is limited in order to facilitate the bunching process.

LOW Q CIRCUIT

Figure 5 shows an equivalent circuit of a magnetron. X_e is the reactive effect of the electron beam in the interaction space. C and L represent all capacitances and inductances in the magnetron and external circuit. R represents purely resistive loading and tube losses (although in many cases a reactance would be included in series with the R).

With fixed frequency magnetrons, the stored energy in the resonant circuit is very large; consequently, the effect of I_c and I_L is very large with respect to I_b, and the beam reactance, X_e, will have little effect on the change of frequency. For this reason, in a high Q magnetron, an increase in anode voltage causes considerable in-

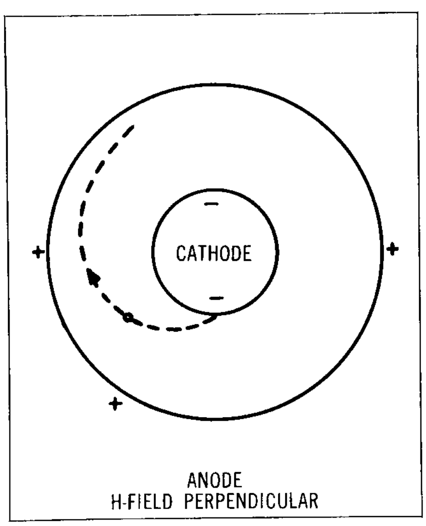


Figure 3—Electron Orbit

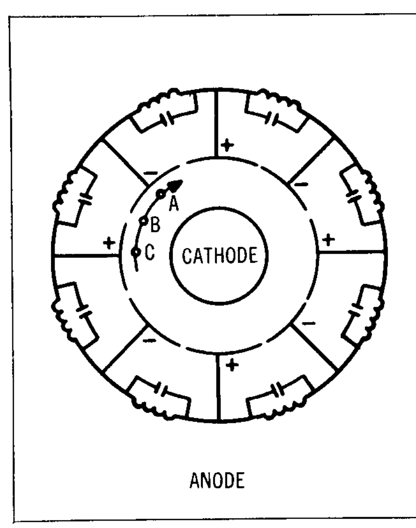


Figure 4—Electron Bunching

crease in anode current and power output, but only a slight change in frequency. As the Q of the external circuit is lowered, a corresponding decrease occurs in I_c and I_L with respect to I_b. Space-charge reactance, X_e, then has a continually greater effect on frequency determination.

One condition for oscillation is that all reactive current components must add up to zero. The reactive components in the interdigital and external portions of the circuit can accomplish this only at discrete frequencies; hence, the reactive portion of the beam current must be sufficiently large to satisfy this condition over the entire tuning range. One practical way to express the result of this phenomenon would be to say that the frequency is determined by the rate of rotation of the electron bunches around the cathode post. Their average angular velocity is controlled by the ratio of the d-c anode voltage to the magnetic field strength, V/B. By maintaining magnetic field strength at a constant value, the average angular velocity of the electron bunches past the interdigital fingers of the anode structure may be varied by changing the anode voltage. A linear relationship is then established between frequency and applied d-c voltage.

LIMITING ELECTRON INJECTION

To obtain wide-band voltage tuning, the circuit reactive current is reduced to the same order of magnitude as the circulating beam reactive current by operating at a low r-f voltage. A low Q anode circuit is then used to obtain power output over wide bandwidths. Since low r-f operating voltage increases the difficulty in bunching the electrons, the number of electrons injected into the interaction space is limited. Without this limitation, the excess space charge would saturate and prevent the low r-f electric fields in the anode interaction space from properly bunching the electrons.

INJECTION SYSTEM

The injection system for a voltage tunable magnetron is represented in cross section in Figure 6. The filamentary cathode is the original source of electrons. The injection electrode acts to accelerate and control the number of electrons entering the interaction space. The cold cathode in conjunction with the anode forms an interaction region where the d-c electron energy is converted into r-f power. In addition, the cold cathode plays an important part in the electron injection system, as illustrated in Fig. 7.

Electrons injected into the tube enter the interaction space. Those entering in the incorrect phase absorb a small amount of energy and are immediately collected on the cold-cathode to produce the relatively high current from the hot to the cold cathode. This current is collected at such low voltage, however, that it represents a negligible power loss (typically about 1% of the anode power in the 75-watt S-Band tubes). Those electrons which do enter in correct phase then constitute bunches which remain focused as they give up a large portion of their energy to the circuit; and they are then collected on the anode.

CATHODE BACK BOMBARDMENT

Not all the electrons in the interaction space contribute to the radio-frequency power output. Depending on their position and on the phase of the radio-frequency voltage on the anode segments, some electrons absorb radio-frequency energy which increases their velocity and causes them to bombard the cold-cathode post.

These electrons dissipate energy producing backheating at the cathode post. They also contribute to the cold-cathode current (referred to in the discussion of the injection system). This energy is relatively small, however, compared with the total generated radio-frequency power.

Since the emissive cathode area is removed from the interaction area, this surface is not exposed to the full back-bombardment current as in a conventional magnetron. Some electrons are directed, however, so that they do collide with the emissive cathode area. In specific instances, it may be necessary to reduce cathode power in order to compensate for added back-bombardment heating. A more detailed discussion will be found in the section on filament supply on page 6.

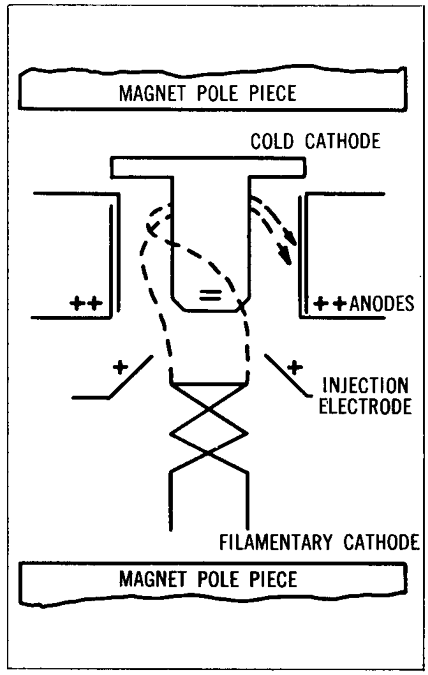


Figure 6—VTM Injection System

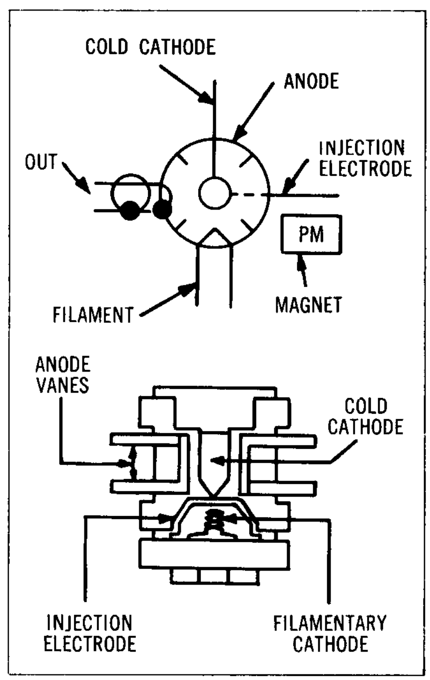


Figure 7—Tube Cutaway View

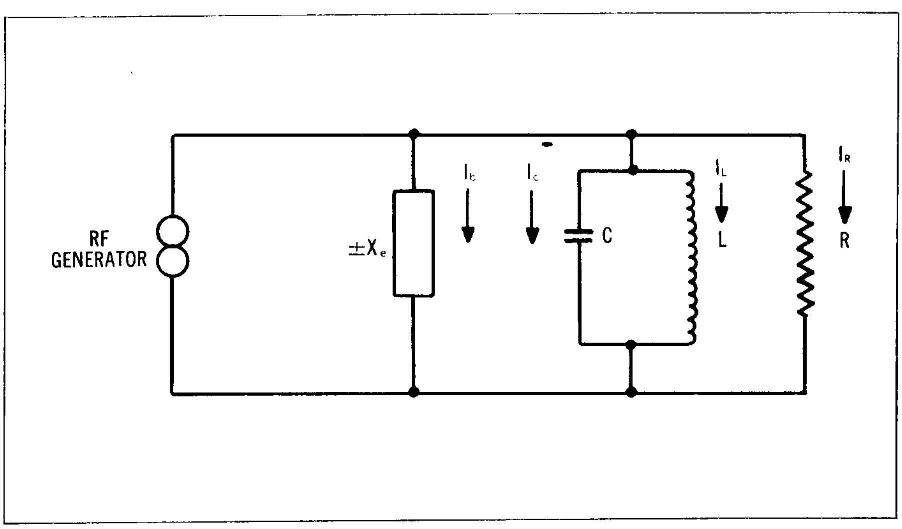


Figure 5—Equivalent Magnetron Circuit

operation and power supply

For satisfactory VTM operation, specific attention to ripple and regulation in the design of power supplies is most important. Permissible ripple can be determined, when the VTM tuning sensitivity and the amount of incidental f.m. allowed by the application both are known, by the following equation:

permissive ripple (volts)=incidental f.m. (mc)/tuning sensitivity (mc/volt)

The tuning sensitivity can be found with adequate accuracy for this purpose by dividing the center voltage into the center frequency, as explained in page 9.

The power supply requirements for the VTM include a filament (emitter) supply (low voltage a-c or d-c), an anode voltage supply (high voltage d-c with adequate current output and good regulation), an injection electrode supply (high voltage d-c very low current drain) and a modulation voltage supply (a-c) to swing the anode voltage about the d-c value and thereby modulate the output frequency.

For test purposes, the circuit of Figure 8 is normally used; the separate supplies afford good flexibility in testing, and the low modulation frequency (usually 60 c/s) allows one to disregard the capacitance of the anode supply unit.

For operation, one can economically derive the Injection Electrode supply from a bleeder across the Anode supply. When the operational modulation frequencies are high, as is usually the case, the modulation supply must follow the anode supply to avoid swinging the capacity of the latter. A coupling capacitor is then required to apply the modulation signal to the Injection Electrode also. This circuit is shown in Figure 9.

FILAMENT SUPPLY

The voltage tunable magnetron (VTM) is capable of long life when operated under proper electrical and mechanical conditions. In addition to the obvious cooling requirements and power limitations, the regulation of the VTM filament-cathode power is extremely important.

Figure 10 shows that the back heating ratio increases very rapidly with frequency so that a low power VTM operating at 4000 mc will have a d-c input to the heater approximately 10% higher than that at 2000 mc, and 6% higher than that at 3000 mc. The leveling off of the solid line is due to a decrease in power level at the higher frequencies. The dashed line indicates the theoretical back-heating ratio at power levels essentially the same as those at the bunch frequency.

Figure 11 shows that a reduction of filament current below 2.0 amperes for the tube operating at 2160 mc brings a rapid fall off in power output due to temperature limited emission from the filament-cathode. At 3160 mc this fall off in power does not occur until 1.9 amperes due to the higher back heating of the filament-cathode. This condition is even more pronounced at 4160 mc where the heating of the filament-cathode due to back heating is more severe and the fall off in power does not occur until 1.7 amperes.

If the VTM is to be operated at spot frequencies or with a very slow sweep (less than 60 cps), then a constant d-c voltage filament supply regulated to ±5% is advised for all VTM's with bandwidths of 50% or greater. This will provide temperature compensation for the filament-cathode by decreasing the d-c input power when the back heating ratio increases.

When using a constant voltage supply, the filament current should be adjusted to the specified value (usually 2.0 amperes) while the tube is operating continuous-wave at the lowest specified frequency for that particular tube. This will provide adequate cathode emission at the lowest back heating ratio. Adjustment of the filament current while the VTM is operating at other than the lowest operating frequency will cause the filament-cathode to operate at higher temperatures than are necessary for adequate emission, and thereby shorten tube life.

If the VTM is to be operated under swept conditions only, and the sweep speed is 60 cps or higher and covers the full band, then the variation of back heating is averaged so that either constant d-c voltage or constant d-c current may be used. Constant current (regulated to ±3%) is advisable in this case as it will tend to decrease the rate of emissive material depletion with tube operation, and thereby help to extend VTM life.

The filamentary cathode is the anode current source. Since the VTM is susceptible to pushing (see separate section on Pushing, Page 13) a ±3% current regulation of a constant current filament supply, or ±5% regulation of a constant voltage filament supply will control this effect. As shown in Figure 12, a change of ±3% in filament current will cause a frequency change of approximately 0.018%; however, it must be realized that the rate of change will vary from tube type to tube type and will depend on what filamentary cathode is used for the particular type.

Use of an a-c filament supply, or of a d-c supply with appreciable ripple, will cause some degree of incidental F.M. Figure 13 shows some typical curves. Information should first be sought from the manufacturer, however, for specific cases with regard to incidental F.M. as well as with susceptibility to pushing.

When specifying the filament power supply, refer to Figure 14 for the volt-ampere characteristics of several G-E VTM filaments.

INJECTION SUPPLY

VTM injection electrode voltage controls the number of electrons injected into the r-f interaction region and thereby

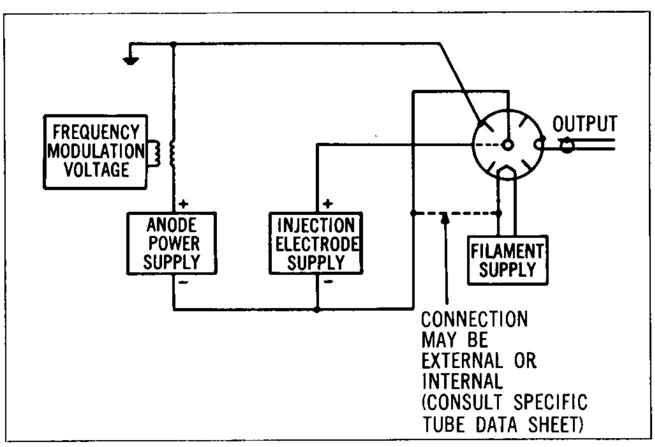


Figure 8—Power Supply Connections for Testing with Low Frequency Modulation and Independent Injection Electrode Supply

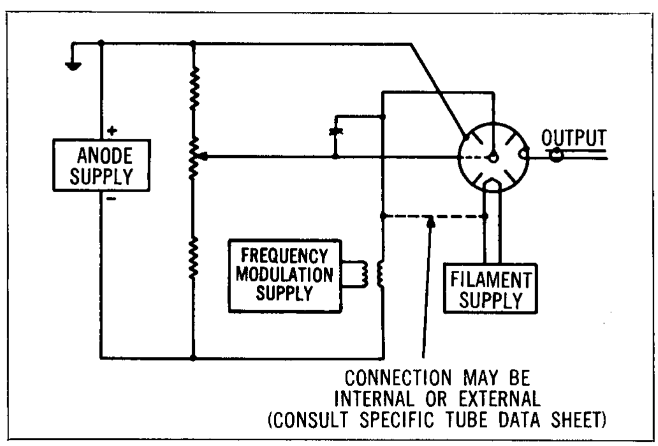


Figure 9—Power Supply Connections for Operation with High Frequency Modulation and Tapped Bleeder Supply for Injection Electrode

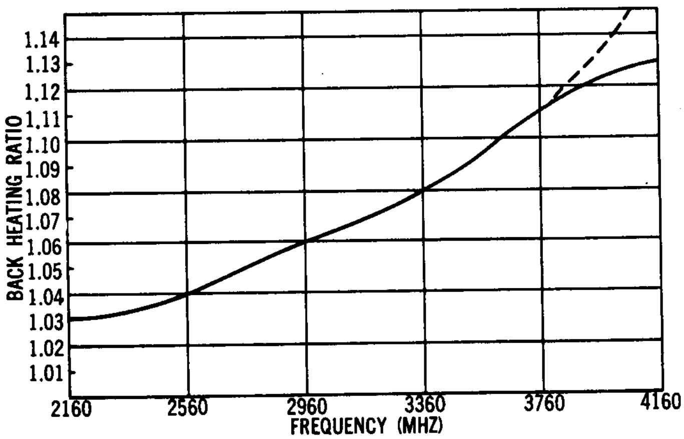


Figure 10—Back Heating Ratio vs. Frequency

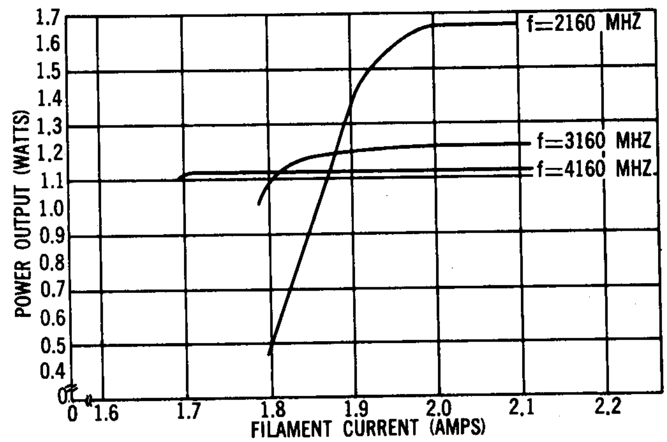
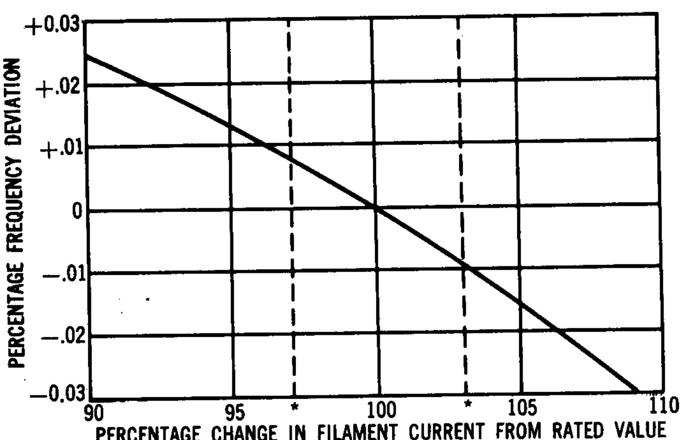


Figure 11—Power Output (emission) vs. Filament Current



* Recommended current regulation ±3 percent. This corresponds to a voltage regulation of ±5 percent.

Figure 12—Pushing Effect of Filament Current

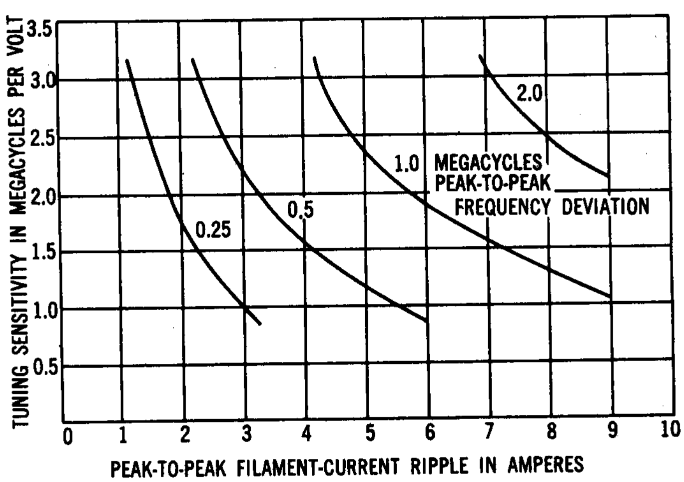


Figure 13—Peak-to-Peak Frequency Deviation Due to Filament

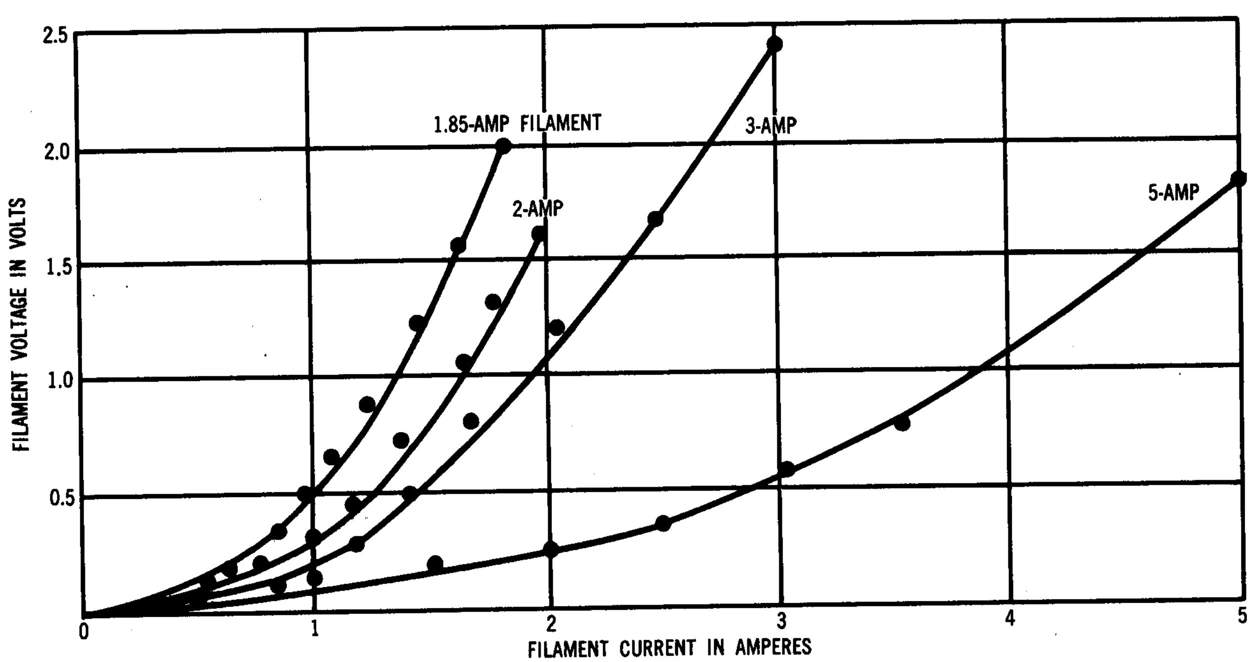


Figure 14—Filament Voltage-current Characteristics

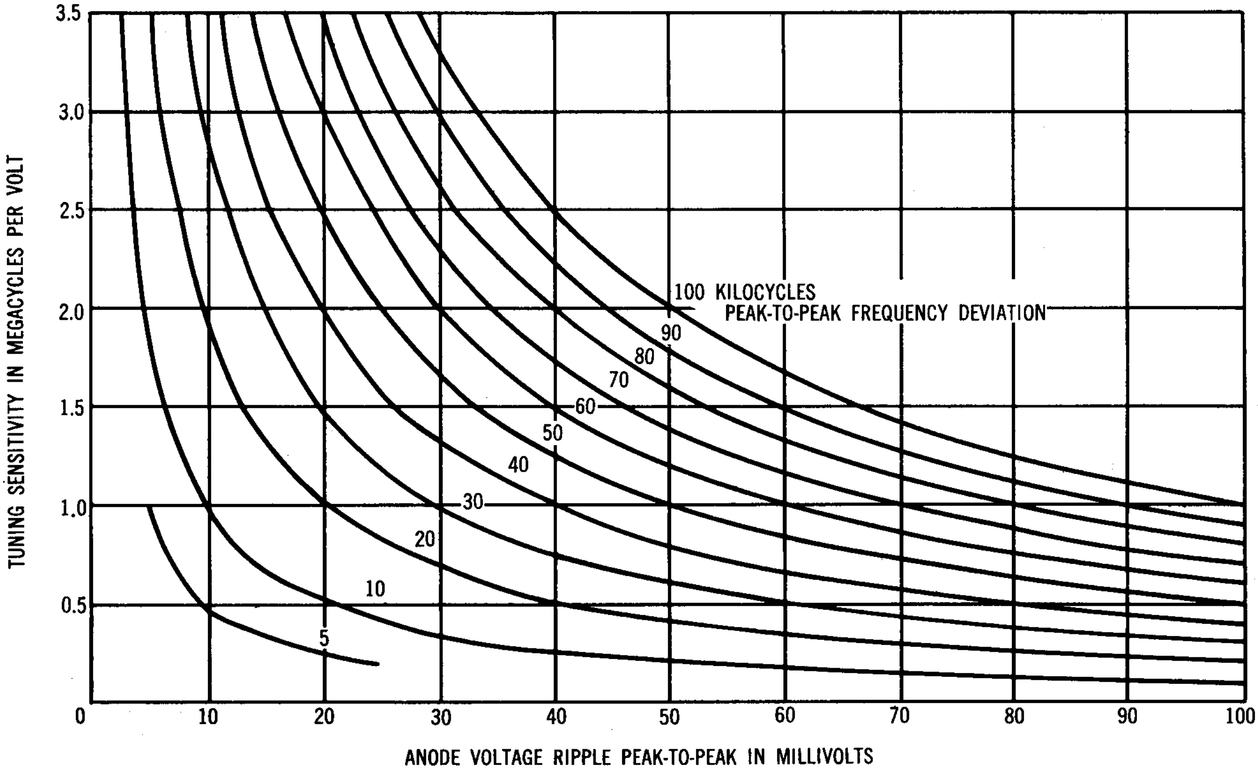


Figure 15—Peak-to-peak Frequency Deviation due to Anode Voltage Ripple

determines the anode current and power level at which the VTM will operate. This change of power with change in injection voltage is essentially a linear function, but its rate will vary from one VTM type to another depending primarily on the normal power output of the VTM at a particular frequency. As discussed in the section on Amplitude Modulation (See Page 16), the VTM is limited in both pulsed and amplitude modulated operation by the small power variation available. This is due to the requirements of electron current for coherent oscillation. When the injection voltage is set too low, too few electrons are injected into the interaction region to permit VTM oscillation. When the injection voltage is set too high, too many electrons are injected into the interaction region to permit the required bunching action to take place. The VTM spectrum breaks up and the tube then becomes noisy or unstable, or drops out of oscillation entirely. In high power tubes the power output may reach a saturation level without break up of the spectrum.

A 3- to 6-db power variation capability appears to be a practical limit for broadband tubes. Such a variation will generally result in a less than one per cent frequency shift due to the pushing effect.

ANODE SUPPLY

The anode-to-cathode voltage (often referred to as the anode voltage) controls the frequency of oscillation of the VTM. One of the VTM's advantages is that its change in frequency with the change in anode-to-cathode voltage is a linear function. Anode-to-cathode voltage actually controls the angular velocity of the electron beam in the interaction area and thereby controls the frequency. In most applications the anode is operated at ground (as shown in Figure 8) with the cathode at a B-minus setting. Modulation is applied between ground and the cathode to vary the velocity of the electron beam, and thus sweeps the tube between prescribed band limits. Further discussion on this operation may be found in the section on Modulation (See Page 15).

The electronic tuning feature places a firm regulation requirement on the anode to cathode power source in order to keep the incidental frequency modulation to a minimum. The peak-to-peak voltage ripple will cause a peak-to-peak frequency change which depends on the tuning sensitivity of the tube as well as on the magnitude of the ripple. Figure 15 indicates what deviations may be expected. Select the tuning sensitivity for which the VTM has been designed and, by intersecting this value with the power supply ripple value, one can determine the peak-to-peak deviation. In addition to frequency and power level control, the VTM is also sensitive to power supply characteristics for starting conditions. Starting can be defined as the ability of the VTM to assume immediate coherent oscillation upon application of all required voltages. The voltage sequence and rise characteristics play an important part in starting the VTM. Should any starting problems arise, experience has shown that the best solution is to operate the VTM with the pertinent power supply while the VTM is being aligned at the factory. A further discussion on starting is presented in the section on Starting (See Page 16).

VTM WITH B+ SUPPLY

Many equipment manufacturers have designed power supplies which operate tubes from a B-plus rather than a B-minus source. In such cases the VTM can be adapted to operate with a B-plus supply as shown in Figure 16. Use of a d-c block will allow the VTM to be operated with a B-plus supply while the r-f hardware is at ground potential. Further modification of the VTM will allow the VTM case (dashed lines) to be operated at ground potential. (If the equipment is such that the VTM case can be "floated" then this latter modification is unnecessary.)

After these modifications are made, the VTM will operate in the conventional manner as it would with the B-minus supply. As shown in Figure 9, the injection electrode supply may be replaced by a tapped bleeder across the anode supply. In this case, a coupling capacitor is not strictly necessary although a bypass to ground may be helpful.

tuning characteristic

The tuning characteristic of the VTM is the curve relating frequency-to-anode voltage. A major advantage of the VTM lies in the fact that this characteristic is very nearly a straight line passing through the origin (i.e., frequency is proportional to voltage).

However, the first essential condition for voltage tunability (See Page 4)—namely that the anode circuit be loaded down to a low Q—implies that the performance of the tube is load-sensitive. It is therefore convenient to discuss departures from the ideal tuning characteristic and load mismatch effects at the same time.

Because of this inherent load sensitivity, the majority of VTMs are built with integral load isolation either in the form of an attenuator (in low power tubes) or of a ferrite isolator (in high power tubes). When a ferrite device is used it is physically a circulator, but with its third port matched so that it is functionally an isolator.

The following paragraphs discuss the effects of load variations applied directly to the VTM. To understand the nature of the problem, one should read them bearing in mind that for VTM packages with the integral isolation the effects will be similar in nature but numerically one or two orders of magnitude smaller. Under these conditions the load tolerances of the VTM is as good as that of other voltage-tunable devices.

The tuning characteristic is described by the following terms:

Tuning Sensitivity: defined as the slope (df/dv) of the best straight line through the observed frequency vs. voltage measurements.

Linearity: defined as the deviation in frequency of the actual tuning characteristic from the best straight line, expressed as a percentage of the center frequency.

Slope Deviation: defined as the deviation of slope of the actual tuning characteristic from the average tuning sensitivity expressed as a percentage of the tuning sensitivity.

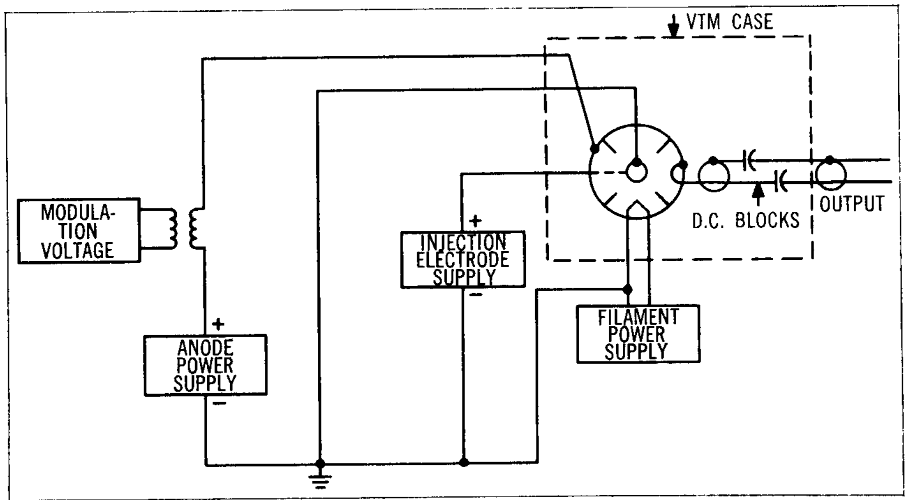


Figure 16—VTM with D-c Block Operating with a B+ Anode Power Supply

The last two terms are not independent: the Linearity is the integral of Slope Deviation normalized to center frequency. Both terms are in use, however—Linearity being a more convenient concept for some applications and Slope Deviation for others. Slope Deviation is sometimes referred to as fine grain Linearity.

TUNING SENSITIVITY

Since the tuning characteristic extended downwards passes close to the origin, the Tuning Sensitivity is closely equal to f/V. Differences from this value result mainly from the resonant properties of the anode circuit.

For octave band tubes this effect is negligible, but when high power tubes have Q values of 10 or more, this causes their Spot Tuning Sensitivity (measured over a small portion of the tuning range) to vary from about 10% below the average value at the low frequency to about 10% above at the high frequency end. Average Tuning Sensitivity (over the whole band) for these tubes is still close to f/V measured at band center. Normally the tuning sensitivity cannot be specified by the user. The basic requirements of frequency and power determine f and V; therefore, Tuning Sensitivity is fixed also. However, this value is considerably higher than the Tuning Sensitivity of a Backward Wave Oscillator (whether O or M type) operating at a comparable voltage. As a result, the modulation power at high modulation frequencies is much less for the VTM than for the other voltage tunable sources.

Figure 17 shows the tuning characteristics of two typical tubes: a wide band low power tube, the ZM-6223 with 2.65 mc/volt average tuning sensitivity; and the 75 watt ZM-6047 with 1.09 mc/volt across a 13% band.

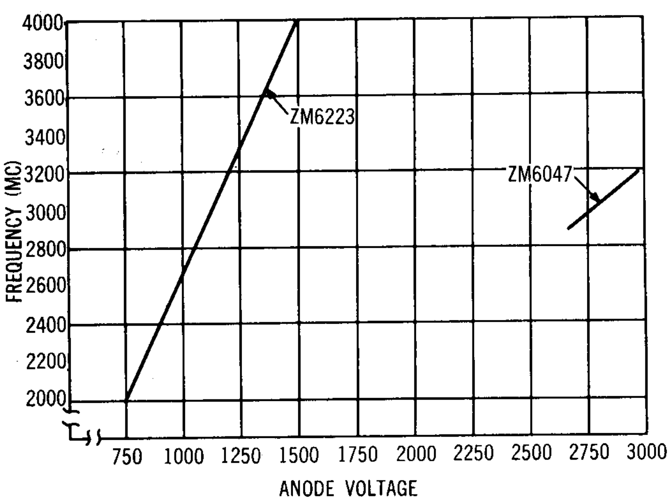


Figure 17—Tuning Characteristics of Broadband VTM and High Power VTM

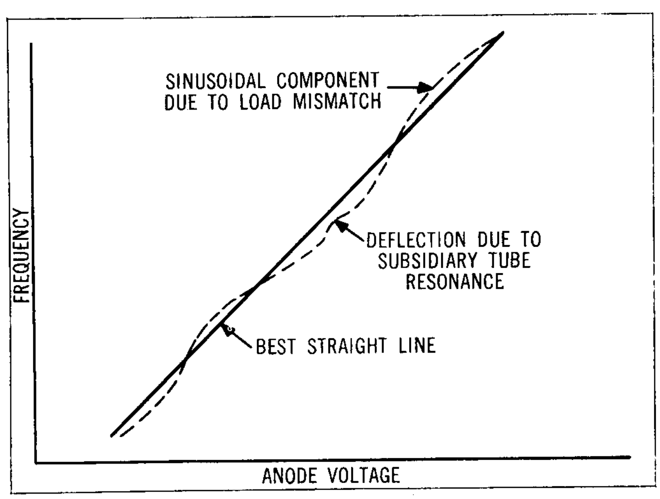


Figure 18—Deviation (exaggerated) of Actual Tuning Curve from Best Straight Line

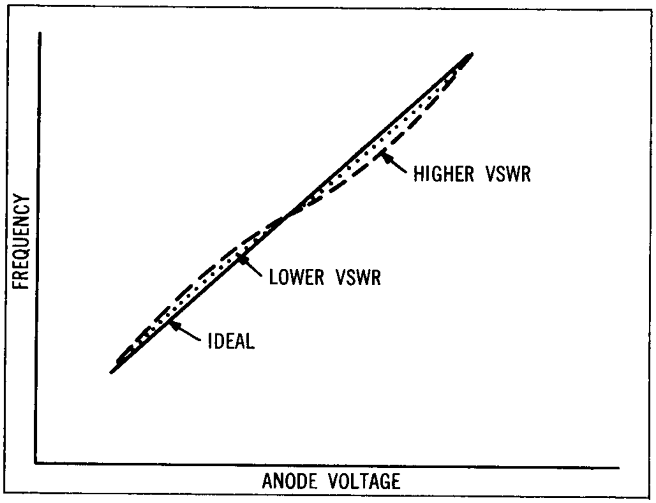


Figure 19—Effect of VSWR on Tuning Characteristic

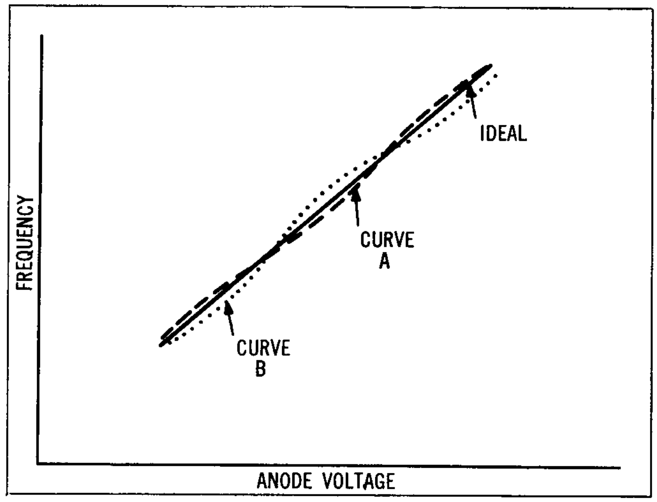


Figure 20—Effect of Phase Change on Tuning Characteristic

LINEARITY

The actual tuning curve departs from the best straight line as illustrated in Figure 18. Sinusoidal variation is associated with reflections from a mismatch in the output line, while isolated deviations may occur due to subsidiary resonances within the VTM. In a well-designed tube, the load reflection effects are the dominant ones; the amplitude of the deviations from the straight line is determined by the load VSWR (See Figure 19) and the direction of deviation (i.e., to higher or to lower frequency) by the phase of the reflection at each frequency.

Figure 20 shows a tuning characteristic (Curve A) measured when operating into a small mismatch whose phase varies slowly with frequency: If the load is then shifted so that the VSWR remains constant while the phases are changed through 180° the tuning characteristic will shift to curve B. Intermediate phase shifts will introduce corresponding small undulations in the tuning characteristic.

The periodicity of the sinusoidal variations is determined by the distance to the reflecting element. A smooth curve with low periodicity is obtained (See Curve A, Figure 21) if the line length is kept as short as possible. As the reflecting element moves further away, the "waves" will slide down the tuning characteristic and become shorter in length and, therefore, steeper. (See Curve B, Figure 21.) When the tangents at the steepest points become vertical, the curve breaks up into discontinuous segments with missing frequency bands (or "holes") between them. This is, of course, an unacceptable situation and the load VSWR must be kept low enough to prevent it. For a VTM without isolation this means a load VSMR must typically be held to 1.2:1 or less across the band—a very tight requirement. Thus the tube should either have the integral isolation or should look into a well-matched pad or load.

Linearity as defined here refers only to the absolute deviations of frequency produced by these effects from the best straight line. Figure 22 shows typical linearity limits of ±1% of center frequency imposed on the tuning characteristic. For narrow band, low power tubes, linearity limits of ±0.5% can be obtained.

SLOPE DEVIATION

The curve of Figure 18 can also be described by the variations in slope relative to the best straight line. This aspect is of greater significance when a problem of following a swept signal with an AFC loop exists; too great a slope may exceed the loop's gain limits. It becomes apparent that slope deviation is affected by load VSWR, and is affected much more than is Linearity by a distant load mismatch with its attendant rapid phase variations (see in Curve B, Figure 21).

For an octave band tube working directly into a 1.2:1 VSWR within a few wavelengths, slope deviation may be typically ±15%. For high power tubes with integral isolators, slope deviation due to load effects is very small but the consistent variation across the band due to the circuit resonance is about ±10% as mentioned under Tuning Sensitivity. (See Page 9.)

The low values of linearity and slope deviation mean that the problem of linearizing the tube by controlling the voltage sweep is much simpler than it is for tubes with inherently non-linear characteristics whose correction voltages are correspondingly large. This is most important to the design of equipments where precise calibration of the voltage with the actual operating frequency of the VTM is of considerable significance. This precision demands that a change in voltage produce the same change in frequency along the entire tuning characteristic and suggests that the slope deviation must be reduced to a minimum.

INITIAL ACCURACY

The VTM tuning characteristic can be held to a high degree of uniformity from tube to tube. In typical production types, the VTM frequency versus anode voltage characteristic does not deviate from the design value by more than ±2.5%. This is of particular importance to manufacturers involved in production quantities of equipment. It facilitates the calibration of the equipment, helps to standardize on manufacturing procedures, and, in general, reduces manufacturing time [illegible] costs.

REPEATABILITY

The tuning characteristic of the VTM is repeatable to within ±0.1% of its initial value when the entire prescribed frequency range is swept. Such repeatability insures high accuracy on successive sweeps and better precision on resetting equipment for operation over portions of the band or for fixed frequency operation.

power variation

With proper loading (1.2 to 1 VSWR or better), an octave band VTM can limit its power variation to ±2 db over the band without any additional leveling equipment. A poor mismatch will cause considerable variations in the power-versus-frequency spectrum and, in extreme cases, may cause a break up of the spectrum. (See Figure 23.) The mismatch actually produces variations in the impedance presented to the VTM's r-f current. This, in turn, causes a sinusoidal variation in the normal power

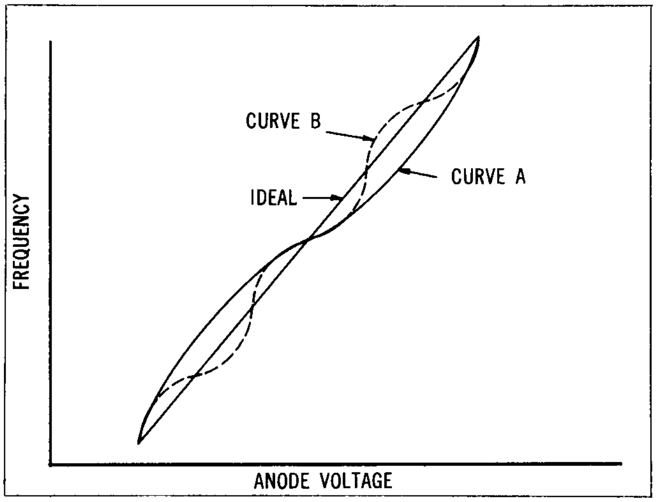


Figure 21—Effect of Load Position on Tuning Characteristic

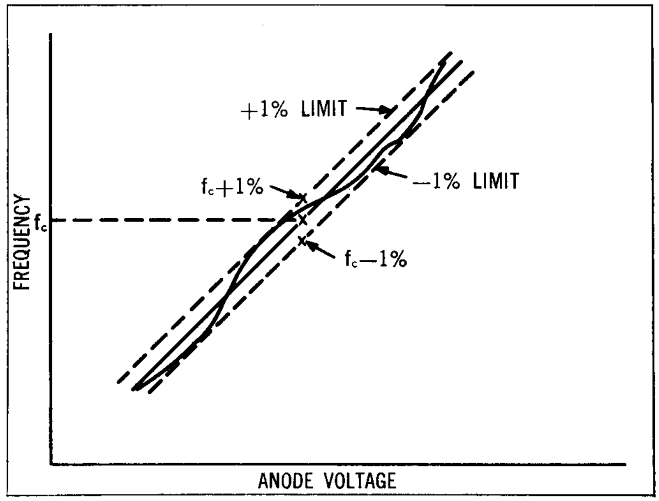


Figure 22—Tuning Linearity Limits

output spectrum. The worse the match, the more severe the variation becomes until a spectrum break occurs. Changing of the load phase will also affect this parameter, and this becomes especially important when the change in phase is coupled with a high VSWR (over 1.2 to 1). Should the loading to the VTM be poor, then isolation in some form is required.

The VTM is capable of being leveled by a feedback loop which controls the voltage on the injection electrode and, in turn, controls the power level. Average power amplitude variation of 3 to 6 db via the injection electrode is available with broadband VTM's, and narrow band VTM's will have a greater variation capability depending on the percent bandwidth. During VTM alignment at the factory, the power spectrum is monitored so that no abrupt changes in power level are present. This characteristic, coupled with the lower over-all power variation, places fewer demands on the leveling system and particularly on the amplifier. Since the injection electrode impedance is in the order of several megohms, a high impedance feedback system can be used.

power output and efficiency

General Electric is producing VTM's with power levels ranging from tens of milliwatts to hundreds of watts.

High powered (75 to hundreds of watts) VTM's with 15% bandwidth have practical conversion efficiencies of 65%, and developmental VTM's (500 watts) have operated at conversion efficiencies of over 70%. (Conversion efficiency is defined as the power output divided by the product of the anode voltage and anode current). These high power, high efficiency devices have found successful application in active electronic countermeasures and can also be used as high level, injection locked oscillators for telemetry and communications.

Intermediate power VTM's (approximately 1 to 10 watts) have efficiencies (which are a function of both power and bandwidth) ranging from 15% to 40% when operated over 30% to 50% bandwidths in L or S band.

Low power, 100 mw VTM's, operating over octave bandwidths, will have efficiencies ranging from 5 to 15%.

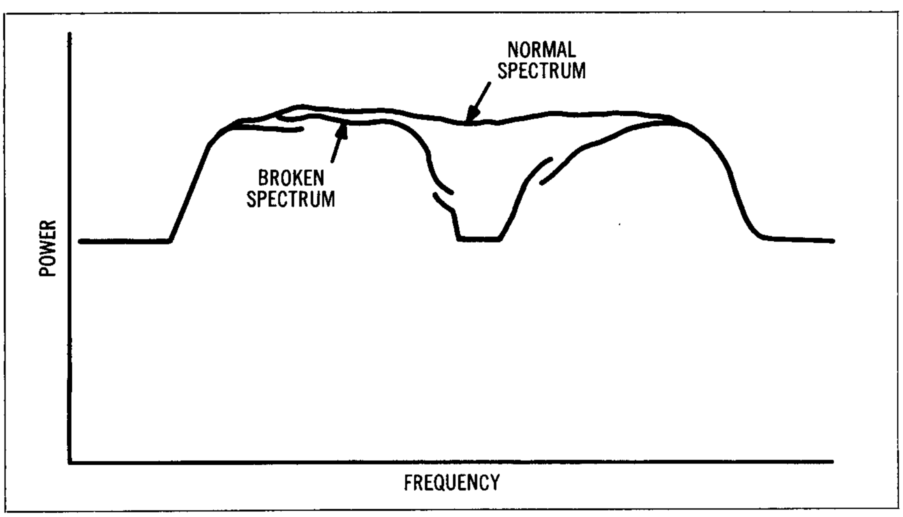


Figure 23—Breakup of Power Spectrum Due to Mismatch

effects on operating frequency

PULLING

A change in the VTM operating frequency caused by external effects such as load variations is often referred to as pulling.

Load variations in the form of changes in VSWR, as well as changes in phase, will cause deviations in the VTM frequency. An S band VTM, operating into a 1.2-to-1 VSWR which is varied through all phases, can change frequency by ±0.5%. A VSWR of 1.05-to-1, will decrease this change to 0.09% when operated through all phases. Thus the preference for a low VSWR becomes evident; furthermore, a load with a fixed phase will also decrease the pulling of the VTM. In addition to the change in frequency caused by reactive variations in the impedance, the pulling phenomena will produce variations in the power output because of changes in the load resistance. (See the Power Variation Section, Page 10 and the load sensitivity section, Page 13.)

f₀=1700 MC

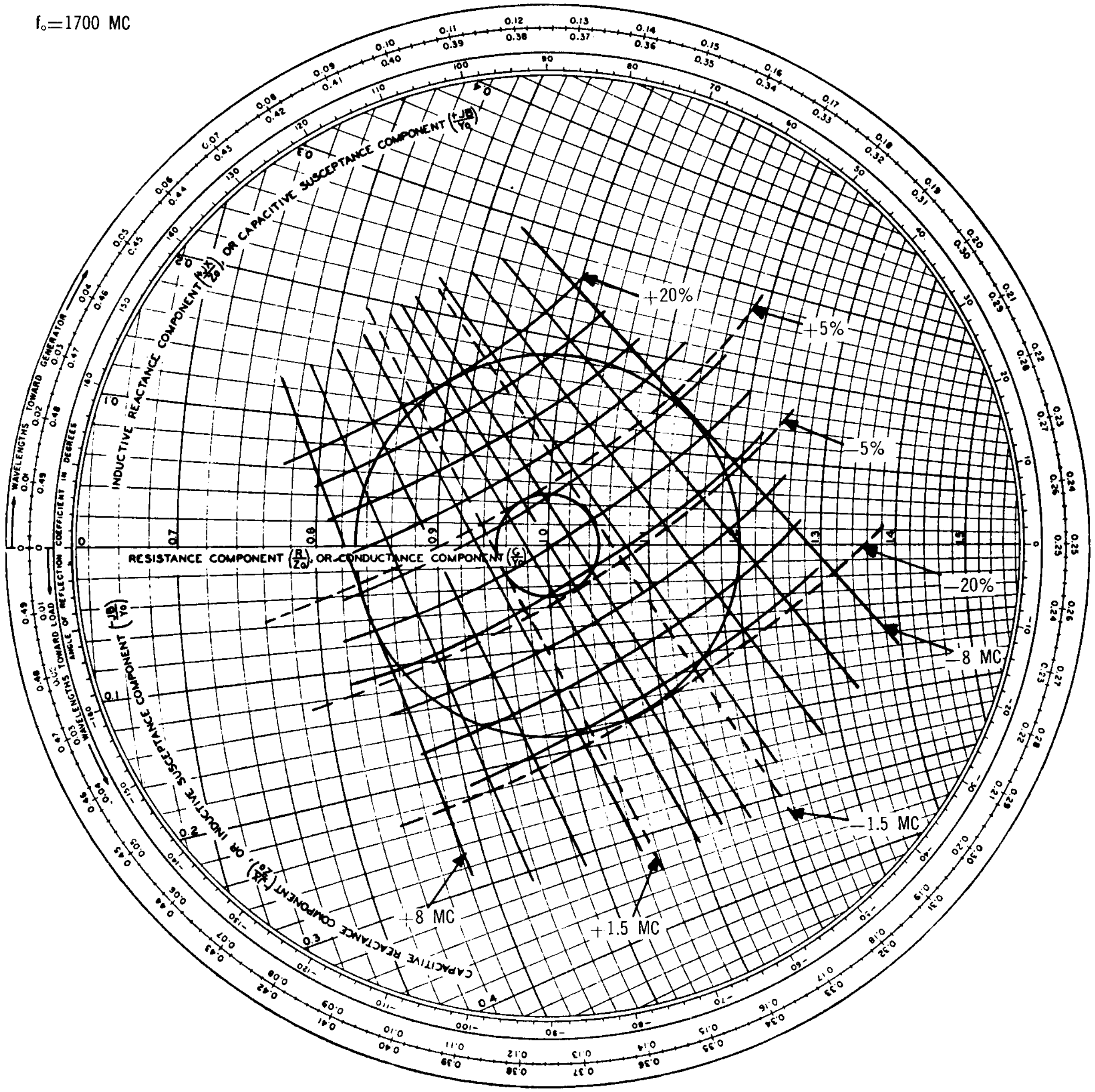


Figure 24—Effect of VSWR and Phase on VTM Power and Frequency

Another effect must also be considered in frequency pulling—the long lines effect present when the load is many electrical wavelengths from the VTM. This causes variations in the tuning characteristic (as discussed in the sections on Linearity and Slope Deviation), and consequently in the VTM operating frequency. Ideally the load should be as close to the VTM as possible.

In many cases where mismatches are part of the system, VTM packages built by General Electric contain an integral attenuator, isolator or circulator to reduce loading sensitivity.

PUSHING

This magnetron characteristic can be defined as a change in operating frequency due to internal effects on the VTM. Two main internal sources of pushing are changes in filament temperature and changes in injection voltage. Both cause variations in anode current and operating frequency.

The rate of frequency change with heater current depends on the filament being used in that particular type of VTM.

In low power VTM's the injection electrode can cause pushing at a 0.2 mc/v rate; hence, changes of 50 volts which may change the power output by 3 db will shift the frequency by 10 mc. Thus in S-band, with a nominal injection voltage of 200 volts a 25% change in injection voltage will produce a frequency shift of approximately 0.3%.

load sensitivity

The voltage tunable magnetron is a load sensitive device. Its parameters—such as the tuning characteristics, power output and operating frequency—depend on both phase and VSWR of the load. Some of these effects have been presented previously.

An indication of the effect of mismatch and change of phase can be seen by consulting the Reike Diagram in Figure 24. Assume you are operating a 3 watt VTM at one frequency (in this case $f_0=1700$ mc). A mismatch of 1.2-to-1 VSWR will produce a set change in frequency and power depending on the phase being reflected back to the VTM. If the load undergoes a 360° change in phase (represented by traveling completely around the 1.2-to-1 VSWR circle) then the VTM frequency and power will be pulled continuously by the amount shown on the orthogonal lines representing percentage changes in power and absolute changes in frequency. Orientation of the orthogonal frequency and power lines depends on a combination of the load and operating frequency. A change in the operating frequency, with the load remaining fixed, will rotate the entire set of the orthogonal lines to a different position. Furthermore, the entire representation does not necessarily have to be centered on the Reike Diagram as has been done here for simplicity purposes. This example assumes that these conditions exist on the anode vanes of the VTM and that the resistance is equal to the characteristic impedance (Z_0). The entire presentation will move away from the center for a normalized resistance other than one.

Reduction of the VSWR will decrease changes in frequency and power considerably (see VSWR of 1.05-to-1). This indicates the importance of using a well matched load or isolating the VTM with a properly matched attenuator, isolator or circulator.

VTM noise

Noise is generally put into two categories with respect to VTM's: IF noise and spurious output. IF noise is that which is integrated over a prescribed bandwidth at a specific center frequency above and below the carrier. The noise level is referenced to the carrier power and is expressed as a signal-to-noise ratio in db/mc.

Noise in narrow band VTM's has been measured at 100 db/mc below carrier at 30 mc from carrier. Wide band VTM's are capable of noise 90 db/mc below carrier at 30 mc or 60 mc from carrier. Broadband IF noise integrated from 100 KC to 100 mc from carrier has been measured 65 decibels below the carrier level. A typical noise measurement system is shown in Figure 25. Here the output of the VTM is fed into a mixer where the noise of the VTM beats against the carrier. The IF amplifiers will pass the noise components whose frequencies are within the particular IF bandwidth. These are displayed on the oscilloscope. A calibrated signal, generally from a signal generator with a calibrated attenuator, will also be presented on the scope. The level of the calibrated signal is adjusted until it is equal to the noise, whereupon a reading of the attenuator dial will indicate the noise level with respect to the signal generator's unattenuated output. This is then referenced to the carrier level of the tube fed into the mixer.

Spurious output results from an interaction of the electron beam with narrow band impedance of adequate magnitude to produce appreciable signal levels. These signals may be of sufficient strength to produce false indications in a sensitive, low noise receiver such as those used in radar surveillance. Spurious VTM output, which includes harmonics as well as other extraneous noise, is measured at about 60 decibels below carrier. On octave band tubes, the second harmonic is approximately —45 db. The measurement of spurious output is accomplished by noting the spurious signal level across the entire bandwidth and in specific cases, the harmonics, when the VTM is operated at a number of equally spaced, fixed frequencies within the specified bandwidth. A substitution method employing a calibrated attenuator

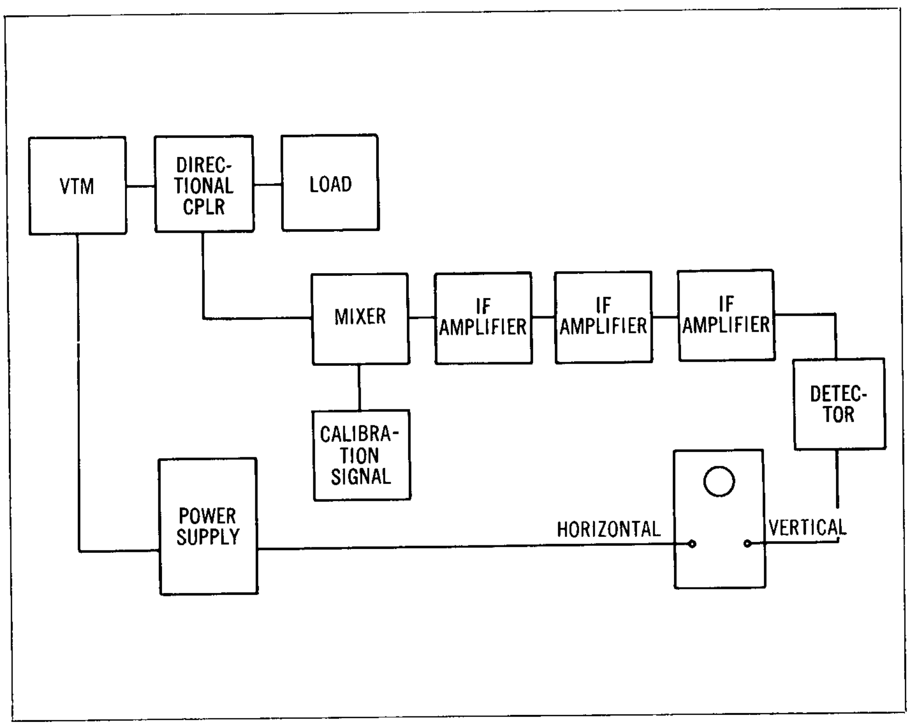


Figure 25—IF Noise Measurement System

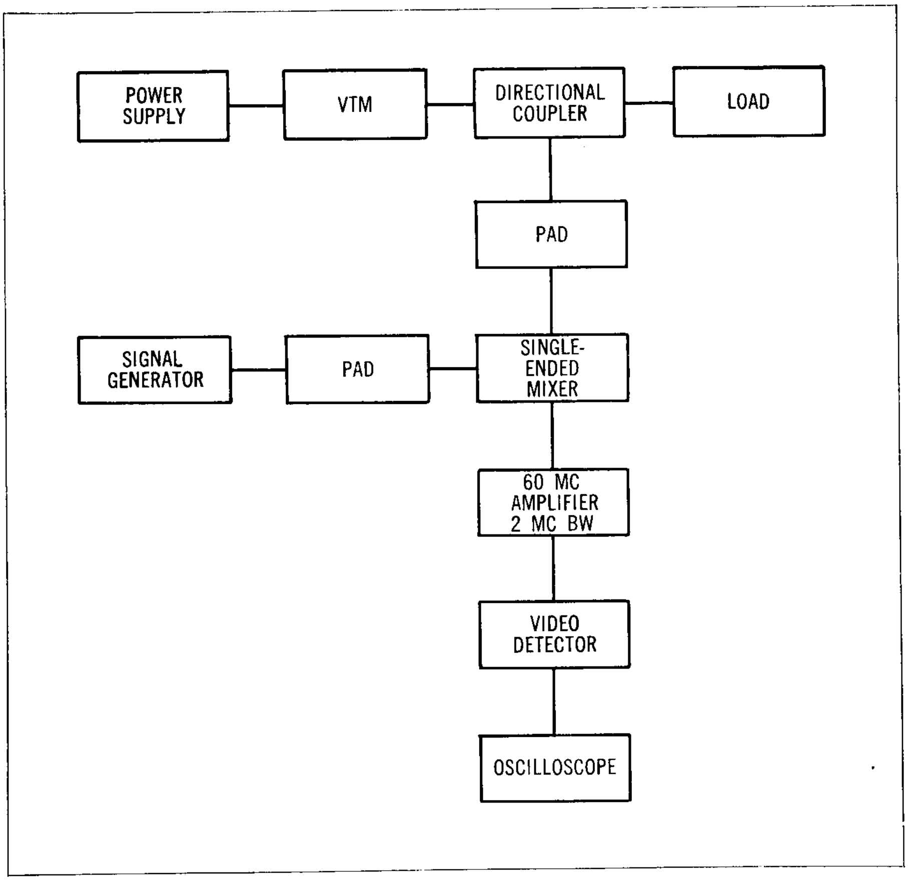


Figure 26—Block Diagram of Spurious Signal Measurement

on a signal generator and a suitable superheterodyne detector are used. (See Figure 26.)

VTM noise only tens of kilocycles from the carrier is important in many test equipments such as sweepers and spectrum analyzers. One basic problem in aligning the VTM for low noise close to carrier has been the lack of a dynamic method for measuring the noise as the VTM is being swept over the prescribed band. The previously used IF method of measurement breaks down. (See Figure 25.) Response of an IF amplifier operating at these low frequencies is so slow that the sweep modulation rate on the VTM must drastically be reduced. Unfortunately this slow sweep rate does not provide an adequate scope presentation of the swept noise characteristics which must be monitored while the VTM is being aligned. Thus the ultimate capability in this area is relatively unknown when compared with the noise levels measured further from carrier. All present evidence points to a higher noise content close to carrier—approximately 50 db/mc below carrier at 10 KC from carrier. Reduction and flattening of the noise level takes place further from carrier. From 100 KCS out to 100 mc and beyond, IF noise levels of 90 db/mc on wide band VTM's are practical; and apparently, at frequencies greater than 100 mc from carrier, there is very little improvement in noise performance.

For optimum, low noise performance a VTM should be factory aligned with the actual loading into which the tube will be operating.

environment

Temperature-compensated tubes will limit their frequency change to 0.2% over the range from $-20°C$ to $+80°C$. Thus, a VTM operating in S-band will not shift frequency by more than 6 mc during a 100°C change in temperature.

RADIATION RESISTANCE

On-site testing at a pulsed reactor facility proved that VTM's are capable of withstanding high levels of gamma and neutron radiation. Repeated exposures to gamma rates up to 1.68×10^7 rads per second and neutron intensities up to 2.55×10^8 rads per second did not affect VTM operation. The threshold of radiation levels that might affect the General Electric VTM's, in fact, could not be determined at this pulsed reactor facility. VTM magnets containing cobalt exhibited no induced radiation activity after the repeated exposures and were not considered a personal hazard.

VIBRATION

The hard mounted VTM will operate at 10g vibration levels from 5 to 2000 cps. When isolation-mounted, the maximum FM from a VTM can be held to 0.1% at levels of 7g from 200 to 2000 cps.

SHOCK

VTM's shocked at 1600g levels have continued to operate normally. One test type had been shocked 45 times—with 30 of these shocks above the 1000g level—and its operation after these tests remained normal.

ALTITUDE

General Electric VTM's have been designed and produced to operate in missile as well as airborne environments.

TEMPERATURE

Depending on its power requirements, the VTM may operate at $-55°C$ to $+125°C$ with only conduction cooling required.

shielded VTM's

VTM operation depends on maintaining the same magnetic field used when aligning the VTM at the factory. During this alignment, the tube's parameters must be carefully monitored on oscilloscopes and meters, and recordings must be made of the operating voltages and currents required to produce a package which meets specifications. Any subsequent change in, or distortion of, the magnetic field destroys the careful factory alignment and degrades VTM performance; also failure to keep ferro-magnetic materials at suggested distances from the VTM packages or use of ferro-magnetic tools and dynamic fields (such as those generated by transformers) can adversely affect the magnetic field of the VTM.

MAGNETICALLY SHIELDED VTM'S

Magnetically shielded VTM's will provide a solution to the above problems. General Electric has developed a VTM package with improved magnetic circuitry and new (but inexpensive) magnetic materials which lend themselves to shielding techniques never before possible. Shielded VTM's can be stacked one on top of the other with no degradation in performance, and this package will be unaffected by transformer fields normally found in many electronic systems. Such a decrease in degaussing susceptibility allows the shielded

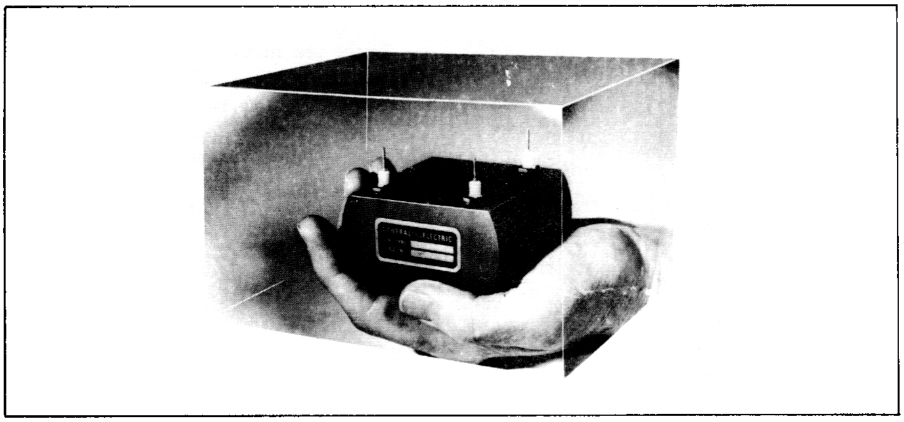


Figure 27—Shielded VTM Compared with Space Requirements for Conventional VTM

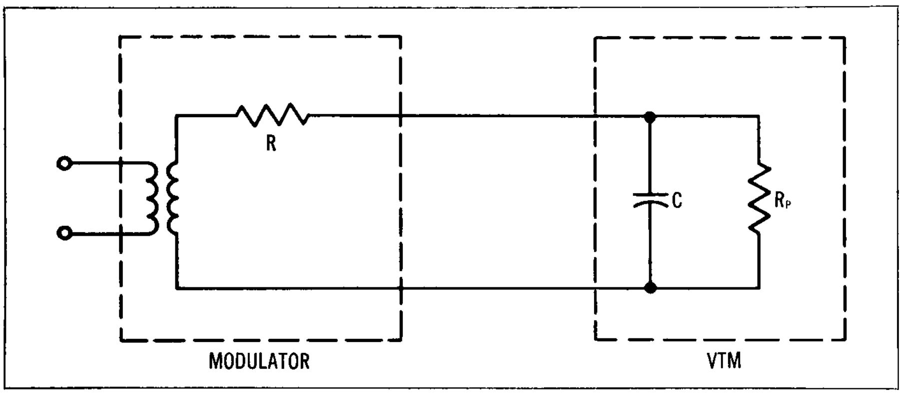


Figure 28—Equivalent Circuit for Frequency Modulating the VTM

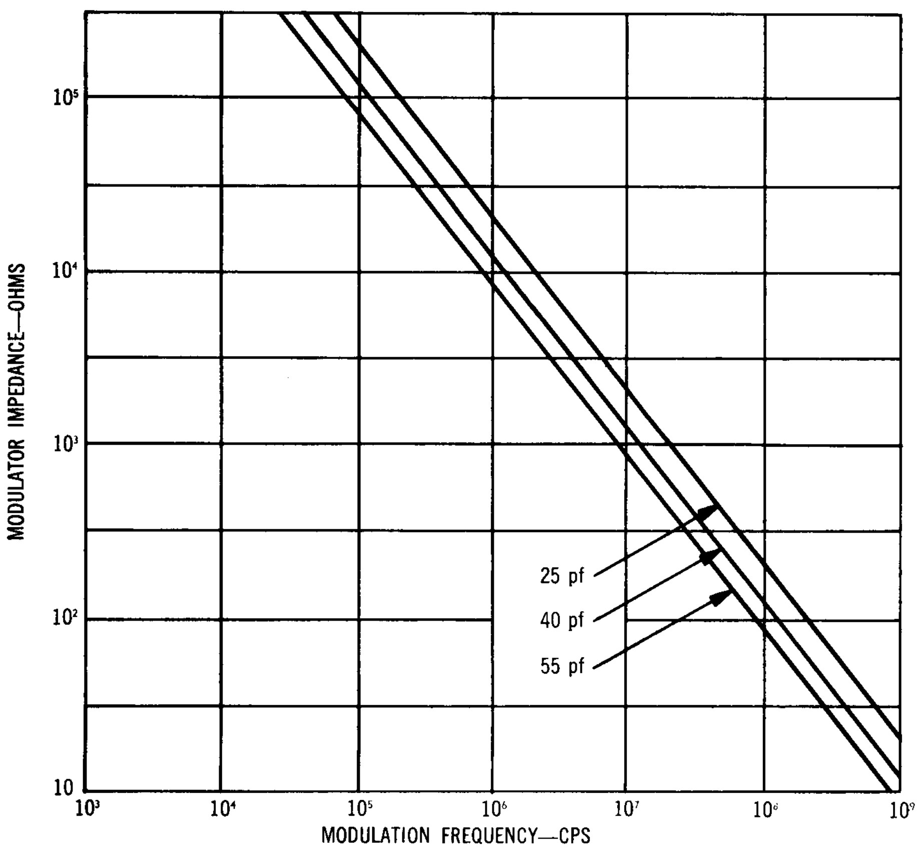


Figure 29—Maximum Modulator Impedance vs. Modulation Frequency

VTM to be used in compact, high density equipments where passive magnetic devices must come in direct contact with the tube. Previous requirements for minimum spacing or protective boxes are eliminated. Figure 27 indicates the reduction in space requirements now possible through integral magnetic and RFI shielding.

RFI SHIELDING

The magnetically shielded VTM also incorporates RFI shielding to attenuate stray RF on the d-c leads. This extraneous radiation is annoying as it can produce unwanted modulation, degrade receiver sensitivity and decrease accuracy of the system in which the VTM is operating. RFI shielding reduces stray radiation on the d-c leads to below minus 30 dbm. This attenuation—provided as an integral part of the VTM shielded package—will eliminate the radiation screens, shields and cages normally required with conventional, electronically tuned oscillators employing magnetic fields.

VTM modulation

General Electric VTM's have been modulated at 20,000 mc per microsecond rates thereby, making the VTM a candidate for frequency agility equipments such as broad band, surveillance receivers and electronic countermeasures systems. VTM's are frequency-modulated by changing the anode to cathode voltage. The voltage-frequency relationship is linear (as discussed in the section on Tuning Characteristic). In regard to modulation, the VTM can be presented as a capacitance and resistance in parallel. (See Figure 28.) At high modulation rates, the internal impedance of the modulator and lead impedances assume greater importance while VTM plate resistance (Rp) can be ignored. To increase the frequency, an increase in anode voltage is required and is obtained by charging the tube capacitance C. The time t for charging would be equal to RC where R is the modulator impedance. Thus "t=RC" is an approximation for increasing the VTM frequency at a constant rate. Since the time constant of the RC circuit represents approximately one-half a sine wave, the time for one period of oscillation would be 2t. Values calculated for several VTM types are shown in Table 1 (page 16). Figure 29 is a presentation of modulation impedance as a function of modulation frequency for three different values of VTM capacitance. Approximations were used to arrive at the maximum modulator resistance and a factor of 0.5 or 0.3 should be used to avoid modulation distortion.

MODULATION DATA FOR TYPICAL VTM'S

Tube Type	C Pf	R_P Kilohms	Maximum Modulator Impedance for Various Modulation Rates: 1 mc Kilohms	10 mc Kilohms	100 mc Kilohms
ZM-6046	35	50	30	3.0	0.30
ZM-6047	35	50	30	3.0	0.30
ZM-6085	40	300	25	2.5	0.25
ZM-6205	110	100	9	0.9	0.09
ZM-6211	40	72	25	2.5	0.25
ZM-6222	140	90	7	0.7	0.07
ZM-6223	40	130	25	2.5	0.25
ZM-6222	140	90	7	0.7	0.07

Table 1

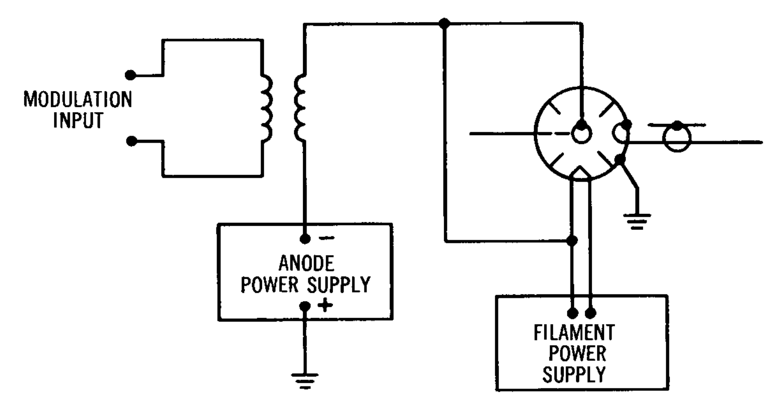


Figure 30—Series Transformer Modulation

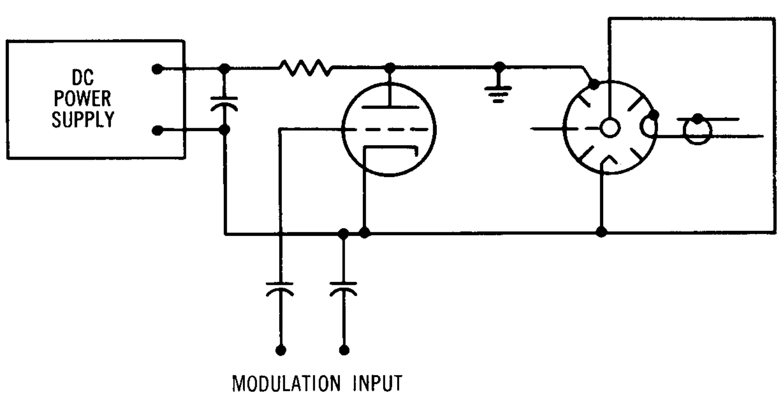


Figure 31—Series Resistor Modulator

FREQUENCY MODULATION

There are many methods for frequency modulating the VTM, but the simplest involves the use of a series modulated transformer where the transformer is connected in series with the power supply. (See Figure 30.) Another method utilizes a series resistor. (See Figure 31.)

AMPLITUDE MODULATION

Amplitude Modulation of the VTM must be limited to changes in power levels of from 3 to 6 decibels depending upon the power output and the bandwidth of the VTM being used. Thus pulsing or square-wave-modulating in the VTM is limited due to both small amplitude modulation capability and frequency pushing considerations.

STARTING

Another factor involved in pulsing and square wave modulating is the starting characteristic of the VTM; that is, the ability of the VTM to assume immediate coherent oscillation as soon as all required voltages have been applied. Broadband low power VTM's are most suspectible to starting problems. At the low end of the frequency range—normally the "hard starting" portion of the band—the space charge is close to the cold cathode and the circulating current and the r-f fields are small. All are poor conditions for starting. To improve them, it is necessary to fill up the interaction space between the anode and the cathode by dispersing the space charge away from the cold cathode. This increases both the r-f fields and circulating current. One way to accomplish this is to use the following voltage sequence for turning on the tube:

(1) apply the heater and injection voltage.

(2) turn on the anode-to-cathode voltage. For best starting results, one should first perform evaluation tests on the VTM with the power supply the VTM will be using in the equipment. Another approach which has produced excellent results is to perform the starting tests with the pertinent power supply while the VTM is being aligned at the factory.

Coupled with the r-f voltage and circulating current considerations for starting is the impedance presented to the current. If the impedance is low, even a moderate amount of current will not provide an adequate condition for starting. Furthermore, the impedance over the prescribed bandwidth has two restrictions in that (1) the power variations across the band must generally be kept to a minimum and (2) the tuning characteristic must be as linear as possible. Thus the impedance must satisfy power output, power variation, linearity and starting requirements. The cavity and circuitry must essentially shape the impedance characteristic across the band to meet all these requirements.

Once again, factory alignment of the VTM using the specific power supply involved will produce a VTM with excellent starting characteristics.

fixed frequency operation of VTM's

While VTM's are used predominantly in swept, broadband applications, they have also found use in fixed frequency operation. VTM's are also capable of being electronically switched from one method of operation to the other.

With a well regulated power supply, frequency variation can be held to ±0.03 percent and, if tighter limits are required, a feedback approach can be used to provide more precise control.

The following discussion of feedback circuits includes frequency comparison, phase comparison and injection locking. The general characteristics of each circuit are summarized in Table 2.

FREQUENCY COMPARISON

In a frequency-modulated telemetry system, frequency tolerances are small. Response time of the feedback loop must be slow enough to retain the lowest frequency components of the modulating signal. Within these requirements, the frequency comparison circuit in Figure 32 will provide satisfactory control.

In this circuit, alternate samples of the tube frequency and a frequency standard (such as a crystal oscillator) are compared by means of a trigger circuit at a rate determined by a square wave generator. Switching rates must be well below the lowest frequency-modulation rate of the system. The sampled signals are amplified and converted into voltage by a discriminator. This voltage is then amplified and oriented by a synchronous detector which transmits a correction signal to the modulator or power supply.

VTM's with this type of feedback circuit have been used successfully in a transmitter for communication in space. The critical center frequency is held to within 0.002 percent. To retain the lowest frequency modulation of 700 cycles per second, a 200 cps switching rate was used. In a frequency comparison circuit, the output voltage of the power supply must be relatively stable over one complete switching cycle since the circuit cannot sense rapid frequency changes. If the ripple frequency approaches or exceeds the frequency of the square wave generator, suitable power-supply filtering will be necessary.

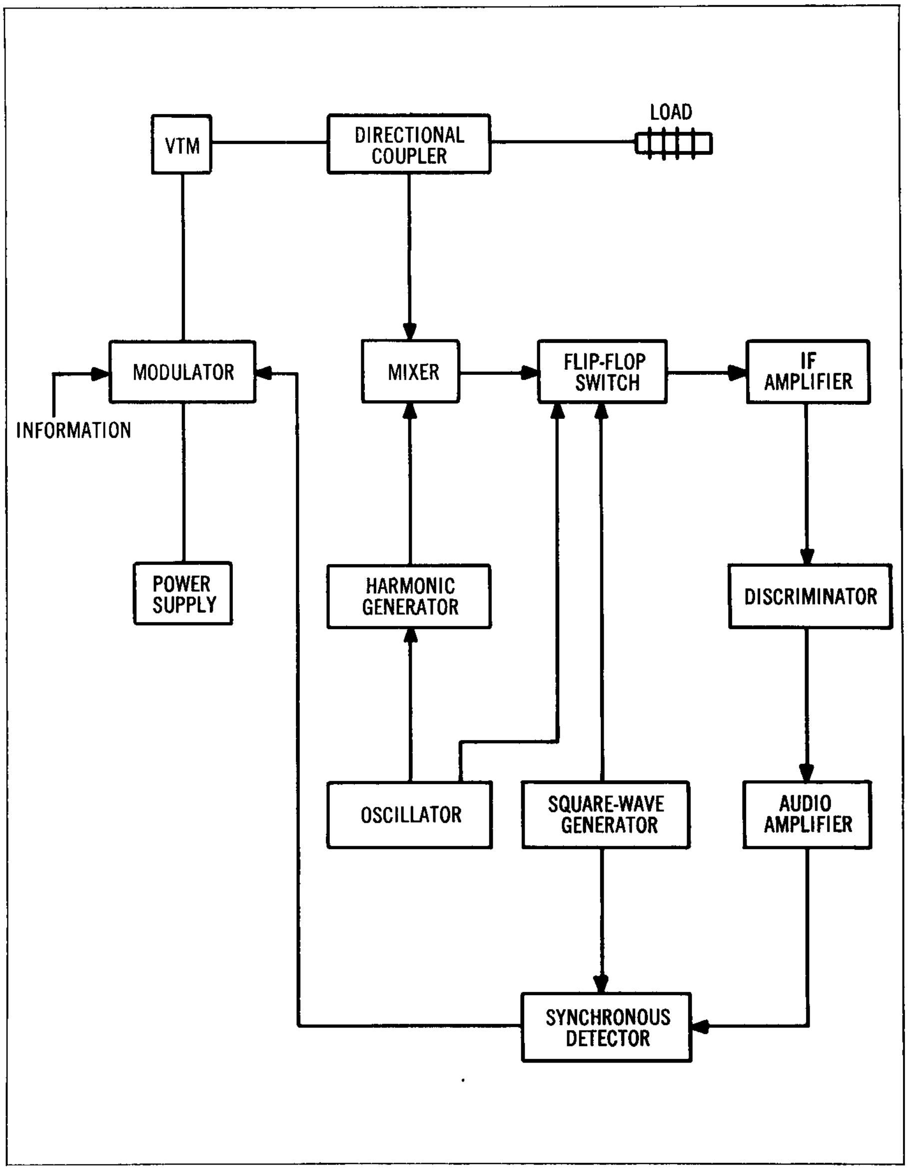


Figure 32—Frequency Comparison Chart

CHARACTERISTICS OF VTM FEEDBACK CIRCUITS

Circuit	Frequency Error	Ease of Modulation
Frequency Comparison	0.002%	Good at high frequencies. Limited at low frequencies by comparison rate
Phase Comparison	Crystal accuracy	Good at high frequencies. Limited at low frequencies by response speed of network
Injection Locking	Same as injection frequency within locking range	Good, by modulating injection frequency at any rate. The frequency deviation must be within the lock-in frequency range
No feedback	1.0%	Good

Table 2

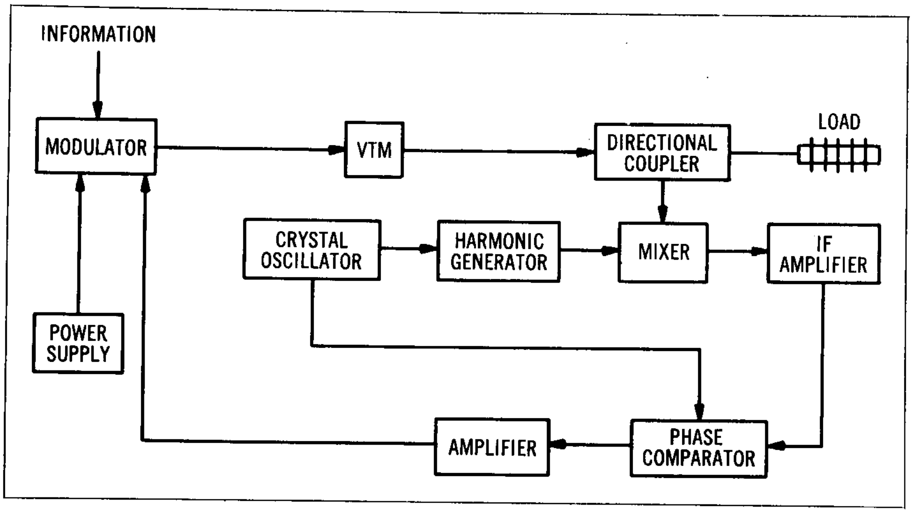


Figure 33—Phase Comparison Chart

An unbalanced condition may result when the frequency modulation rate approaches an odd harmonic of the switching frequency. This condition can be eliminated by a filter at the output of the synchronous detector.

PHASE COMPARISON

The phase comparison circuit in Figure 33 mixes a portion of the magnetron output with a harmonic of a crystal oscillator. The resulting signal is fed into an intermediate-frequency amplifier of the same frequency as the oscillator. Next, the amplified signal is phase compared directly with the crystal-output frequency, and the error signal is then amplified and fed back to the tube for frequency correction.

This circuit maintains the magnetron frequency at crystal accuracy. This accuracy can be maintained at regular intervals in the tuning range determined by the harmonics of the crystal; thus it is possible to phase-lock onto one frequency or step-tune the tube across its entire frequency range. In this service, the response time of the feedback circuit determines the lowest frequency modulation rate.

The allowable power-supply variations are determined by the crystal frequency and by the tuning sensitivity of the VTM. For example, a 60 megacycle crystal with a harmonic generator produces a signal every 60 megacycles in the desired frequency range. Here, a tube with a tuning sensitivity of 3 megacycles per volt will limit the power supply voltage variation to ±10 volts, and a greater voltage variation will cause the system to lock onto an adjacent harmonic.

INJECTION LOCKING

The VTM can be "slaved" to the frequency of a low level signal by injection locking. The effect of this method of operation on the normal tuning curve is shown in Figure 34. Figure 35, meanwhile, shows the trade-offs between lock-in range and gain. The locked frequency range depends on the injected power level, the power output of the VTM and its tuning sensitivity. Increased power output or tuning sensitivity will decrease the lock-in range of the VTM while an increased level of injection signal will increase the lock-in range, but at the expense of gain.

Without modification, a voltage tunable magnetron can be frequency-locked by simply feeding an injection signal through the output connector of the tube. This can be done with a circulator, a directional coupler or a tee. Advantages and limitations of each injection method are shown in Table 3 on page 19.

The preferred and most efficient method is with the circulator. The insertion loss is less than 1 decibel and the loss of in-

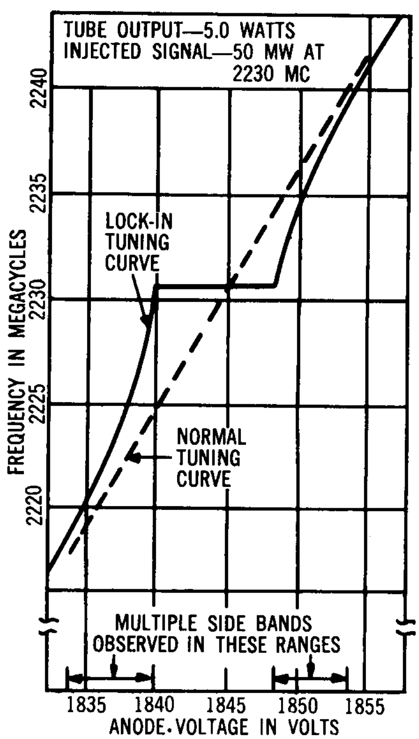


Figure 34—Effect of Injection-locking on Tuning Characteristic

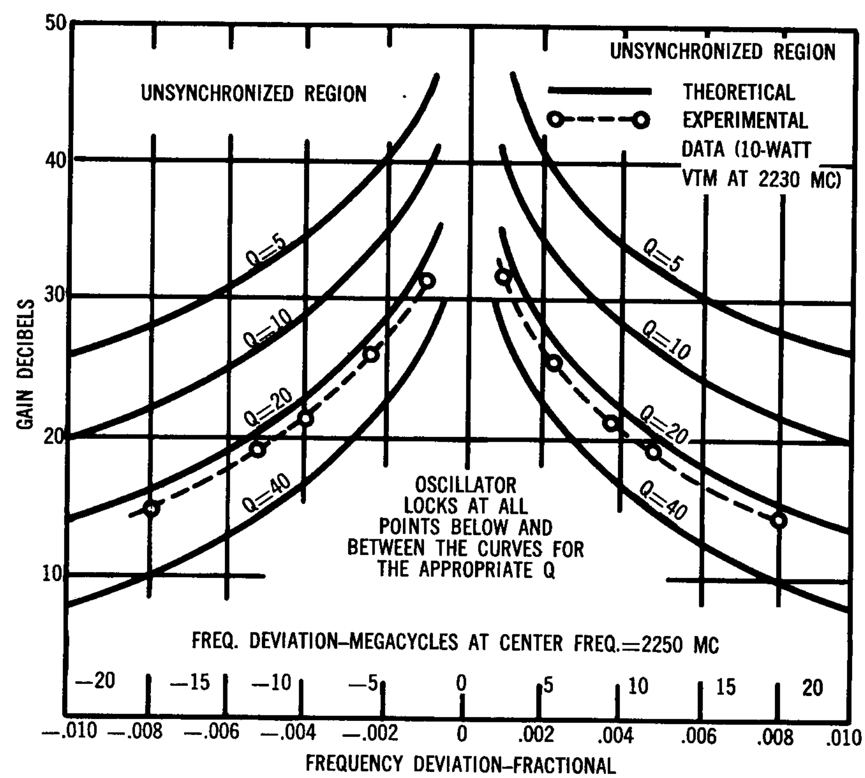


Figure 35—Injection Locking Capabilities

jection power only 2 decibels. Bandwidth is limited by the circulator—a particularly significant problem where temperature extremes are involved.

The directional coupler circuit offers octave bandwidth and low insertion loss, but the injection power loss is high. If a 6-db coupler is used, injection loss is 7 db; however, the insertion loss is only 1.6 db in a given octave of bandwidth.

The tee circuit has the widest frequency range of the three circuits although the insertion loss is high. In this circuit, an insertion loss of 3 db and an injection loss of 4 db can be expected.

typical telemetry performance

When injection-locking the VTM with the circulator (as shown in Figure 35), a typical telemetry VTM may have a center frequency (f_o) of 2250 megacycles, a Q of 10, a power output of 100 watts, a gain of approximately 25 db, and a lock-in range of 25 megacycles so that Δf_o equals 12.5 megacycles. Thus, a signal of one watt would be entirely adequate, and this performance would be at a conversion efficiency of approximately 65 per cent.

FREQUENCY RESPONSE

Another point of interest is the modulation capability of an injection-locked VTM. For a VTM with 20-db gain, center frequency of 2250 megacycles and Q of 10, the "pull-in" time is about 0.05 microseconds. This is the time required to sweep across the entire lock-in range and corresponds to one-half cycle of modulation. Thus, a maximum modulation frequency of approximately 10 megacycles is possible.

MULTIPLEX OPERATION

In addition to the VTM's low Q and high efficiency, another outstanding characteristic is its linear voltage tuning. The tunability feature, which will serve for drift correction, can also be used when a number of information channels are to be transmitted on a time-multiplex basis. Instead of sequentially modulating them on one carrier, they may be given separate carriers within the telemetry band being used. The VTM voltage can be stepped so that it locks to each carrier in turn for an appropriate time. (See Figure 36.) The time taken to re-lock to a new channel depends on the input capacity of the VTM and the current capability of the power supply. If the capacity is 35 pico-farads and the supply is capable of 100 milliamperes momentarily, a 300 volt-step can be completed in 0.1 microseconds. Thus, a one-megacycle switching rate is possible.

COMPARISON OF INJECTION CIRCUITS

Circuit	Insertion Loss	Injection Power Loss	Octave
CIRCULATOR	Less than 1 db	2 db	Octave
DIRECTIONAL COUPLER	20 db Coupler, 0.3 db 10 db Coupler, 0.9 db 6 db Coupler, 1.6 db	21 db 11 db 7 db	Octave Octave Octave
TEE	3 db	4 db	Greater than Octave

Table 3

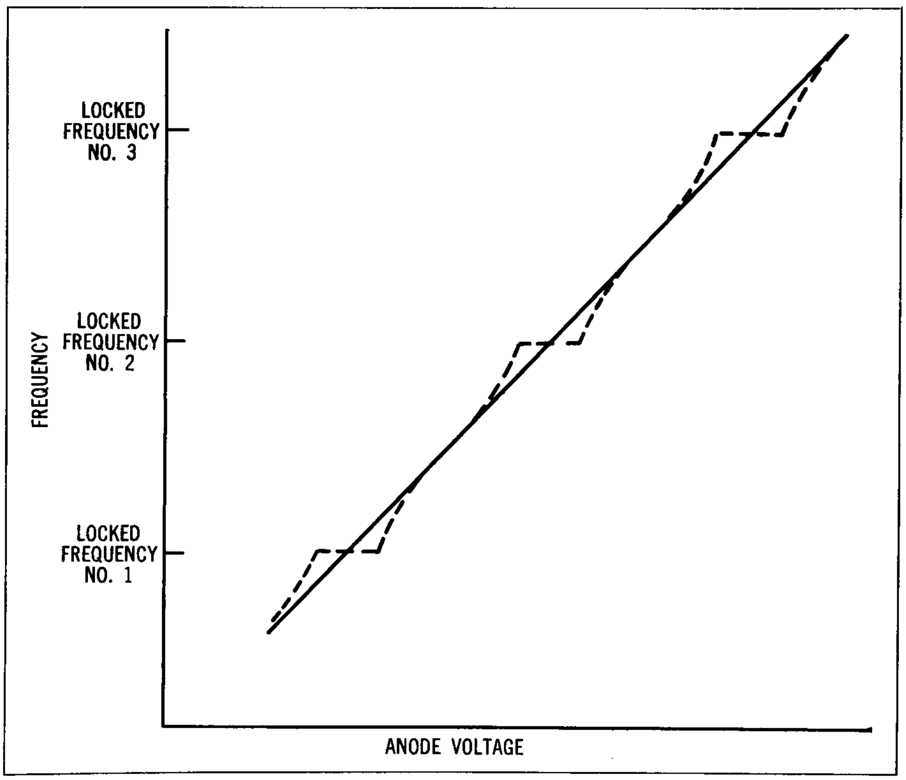


Figure 36—Step Injection Locking the VTM

specific applications

LOCAL OSCILLATORS

The VTM is used in low noise, broad band receivers as a local oscillator. The linear, electronic tuning simplifies calibration and equipment requirements. In addition the minimum power variation over the prescribed frequency reduces the demands on leveling circuits. Its broad band and rapid modulation make the VTM an ideal component for surveillance radar.

TEST EQUIPMENT

Power output in the watt region and octave tuning make the VTM attractive for signal generators and swept signal sources, or as a swept signal oscillator for test equipment.

ELECTRONIC COUNTERMEASURES (ECMs)

G-E VTMs with power levels approaching 500 watts and conversion efficiencies of 65% possess all the specifications for active ECM equipment requiring high efficiency, high power density, rapid tuning and low power variation.

The VTM's low noise, wide bandwidth, flat power spectrum and frequency agility meet the requirements of sophisticated ECCM equipment.

RADAR ALTIMETER & PROXIMITY FUSES

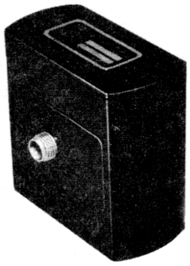
Accuracy in measurement results from the linear tuning characteristic and flat power spectrum of the VTM. Radar altimeters will also find that the VTM's electronic tuning overcomes the limitations of mechanically tuned components. High power and high efficiency VTMs further serve to reduce equipment size and weight without sacrificing long range capability.

TELEMETRY AND COMMUNICATIONS

Injection locking the VTM suggests its use as a frequency modulated amplifier. The high efficiency and broad controllable frequency characteristics make the VTM suitable for communications.

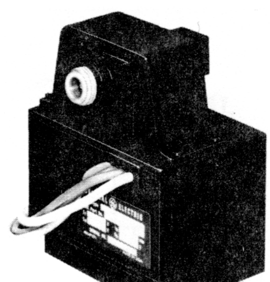
Its tuning linearity, flat power spectrum and electronic tuning make the VTM a precise and flexible telemetry component. (A more detailed discussion of telemetry applications appears on page 19.)

typical package designs



MAGNETIC AND RFI SHIELDING

Becoming the standard for low and intermediate power VTM's. Rapidly replacing the conventional and unshielded E magnet VTM package. Package weights range from 1.5 to 3.0 pounds and nominal package dimensions are 3"x3"x2". Both weight and size depend on power, bandwidth and center frequency.



MAGNETIC SHIELDING AND INTEGRAL ISOLATOR

This design is being used primarily on high power VTM's although it is adaptable to low and intermediate power packages as well. The integral isolator allows the systems designer wider latitude in regard to VTM loading and eliminates an extremely important tube-systems interface. Typically a 100-watt, S-band VTM with 20% bandwidth will weigh 3.0 lbs. and measure 3" x 3" x 4" excluding isolator.

MINIATURIZED, SHIELDED VTM

Many applications place a premium on package size. Use of special magnetic materials enables a 10 watt, S-band VTM with 30% bandwidth to be packaged in a 1" x 1¾" x 1¾" size. As with the other shielded VTM's, this package lends itself to high density, compact equipments, where passive magnetic materials may be in contact with the VTM. The weight of this package is less than 1 lb.

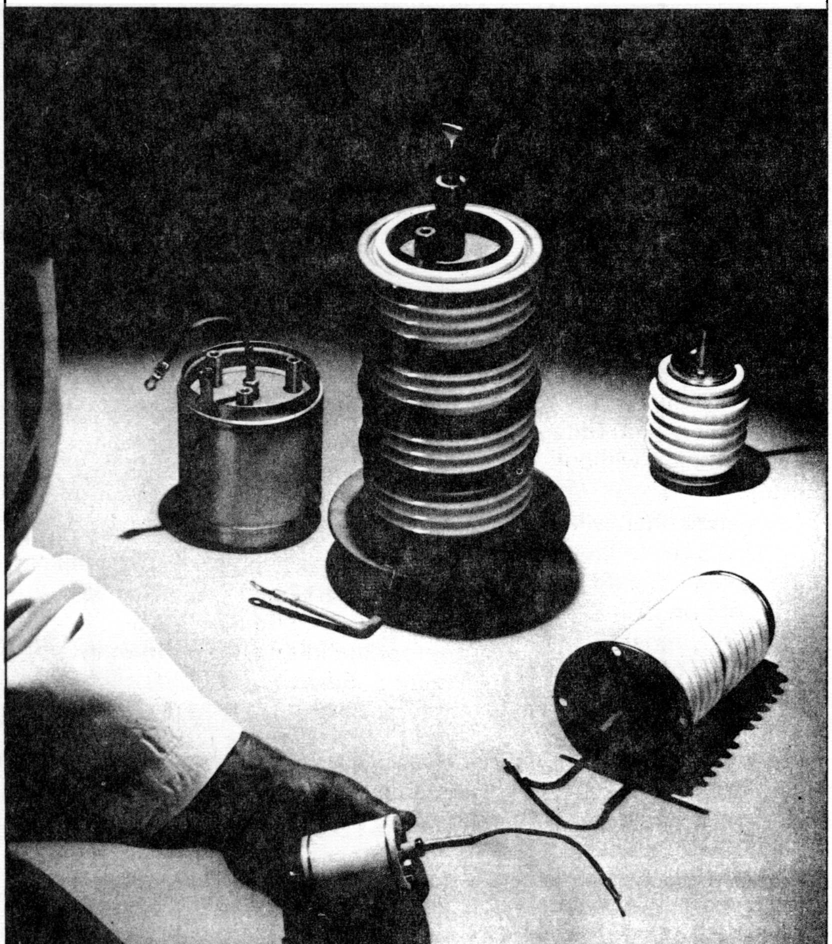
For more information on VTM's, consult your nearest
General Electric Electronic Components Sales Office, or write to:

Microwave Tube Operation
General Electric Company
Building 269
Schenectady, New York 12305
or telephone: (518) 374-2211 Extension 5-3433 or 5-4273

GENERAL ELECTRIC

Triggered Vacuum Gaps

ADVANCED TECHNOLOGY AT WORK TO SOLVE YOUR SWITCHING AND COMPONENT PROTECTION PROBLEMS.



General Electric Triggered Vacuum Gaps feature plasmoid triggers and advanced electrode materials to provide the most effective switching and protective devices you can buy. These construction innovations give you more long-lasting and responsive gaps than ever before.

GENERAL ELECTRIC

Versatility in high power electronic switching with new General Electric TVG's

RELIABLE HOLDOFF, RAPID SWITCHING

A Triggered Vacuum Gap, or TVG, is an electronic switch that is closed by applying a pulse to its novel trigger electrode (Figure 1). In the non-conducting state, it is a high vacuum tube, exhibiting the high electrical holdoff capability of a vacuum device. Yet, when the gap is triggered, it instantly becomes a vapor tube, conducting current through a metal-vapor plasma.

General Electric employs hydrogen plasmoid injection in its TVG's to produce firing times of a fraction of a microsecond. And modern refining processes now make it possible to provide electrode materials of copper and copper alloys which are not only free of absorbed or chemically trapped gases, but also free of chemically combined impurities which could evolve gas when decomposed during the operation of the tube. These characteristics make TVG's ideal for high-voltage service as crowbars in protective circuits and as switches for capacitor discharges, for metal processing, exploding bridgewire applications, and many others where consistent holdoff reliability and long life are required.

UNIQUE PLASMOID TRIGGERING

The new triggering device contains hydrogen impregnated metal (titanium hydride). When a voltage surge is applied to the trigger, it creates an electric field which results in a spark being formed across a small groove, as shown in Figure 1. A plasmoid of hydrogen ions, electrons, and hydrogen gas is thus created. The plasmoid is accelerated into the main gap space, finally causing a cathode spot to be formed. This sequence occurs with much less electrode erosion than with any other vacuum gap triggering technique.

WIDE-RANGE PERFORMANCE, FEATURES

Easily triggered GE TVG's fire rapidly, in fractions of a microsecond, over a wide voltage range and recover their vacuum state quickly after each triggering to give your circuits an extra margin of protection. The recovery capability of these tubes is characteristically in kilovolts per microsecond.

Hydrogen-plasmoid injection and gas-free copper alloy electrodes permit the tube to fire easily without degenerating its voltage holdoff capability. While GE TVG's can withstand rated voltage with a considerable safety margin, they also fire reliably at levels as low as 300 volts. Full fault protection is offered at any power supply voltage setting. This includes crowbar-initiation at very low voltage, and high voltage operation without danger of prefiring.

Laboratory tests show that reliable General Electric TVG's retain their rapid-firing, high-voltage capabilities through thousands of firings. And these units function efficiently regardless of operating position to offer you more design flexibility than many other switching devices. Small in size, GE TVG's are of rugged metal-ceramic construction designed to withstand high levels of shock and vibration, wide swings in ambient temperature, and exposure to nuclear radiation.

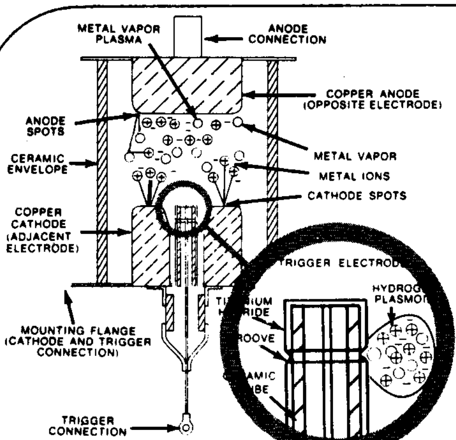
VERSATILE IN APPLICATION

As an energy diverter, the TVG acts as protector for vulnerable high-voltage equipment by short-circuiting the direct-current supply within a microsecond after initiation of the trigger pulse. Referring to the basic circuitry of Figure 2, when a load fault occurs, a sensing circuit delivers a small signal to the firing circuit which triggers the TVG into conduction. If the fault is not self-clearing, the gap can be fired repetitively until the problem is corrected or power into the circuit is interrupted in the conventional manner by electrical contactors.

In cases where energy diversion is needed only momentarily to permit a fault to clear, or to perform a capacitor-discharge function, a single firing of the TVG is usually sufficient to initiate the sequence characterized in Figure 3. A typical firing circuit for this case, Figure 4, functions as follows: A signal from the fault sensor first actuates a small switch such as an SCR, allowing a capacitor to discharge. This discharge, in turn, delivers a pulse to the trigger of a small triggered gas gap (ZR-7514). When the gas gap fires, energy from its firing capacitor (or pulse-forming network) delivers the proper pulse to the trigger of the crowbar TVG, thus switching the TVG to its conducting state. The entire process can be completed in less than one microsecond.

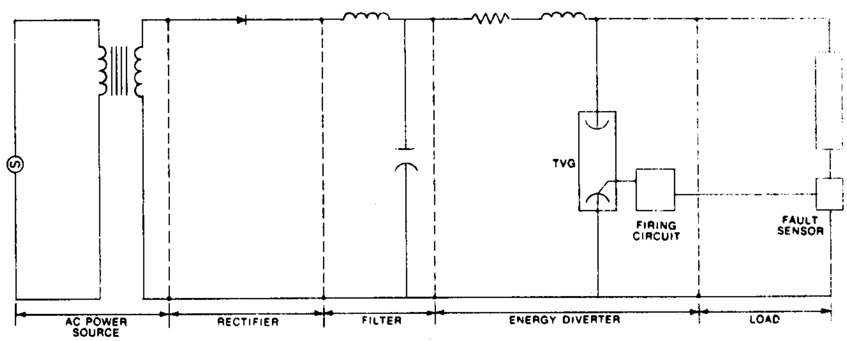
In cases where fault protection must be of the sustained type — e.g. until a-c contactors open — it becomes necessary to refire the TVG at appropriate intervals, as shown in Figure 5. Since a small TVG such as General Electric's ZR-7513 can be repetitively fired, it is utilized in a multiple-firing role for triggering the crowbar TVG. A typical circuit for the sustained crowbarring case is presented in Figure 6. The initial pulse from the fault sensor fires a small switch, or SCR, directly as was done in the momentary diversion case. It simultaneously initiates a "burst" of signals which are delivered by the SCR so that the small TVG (ZR-7513) is caused to trigger the crowbar TVG at the proper intervals for the required protection period.

When the trigger of the crowbar TVG is at a high negative potential, it is often necessary to use a transformer to couple the triggering impulse into the TVG. This is readily accomplished without sacrificing pulse rise time or amplitude characteristics through recent innovations in pulse transformer design.



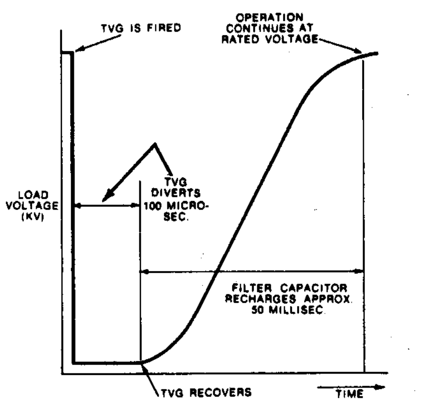
Fundamental Triggered Vacuum Gap Cross Section

FIGURE 1



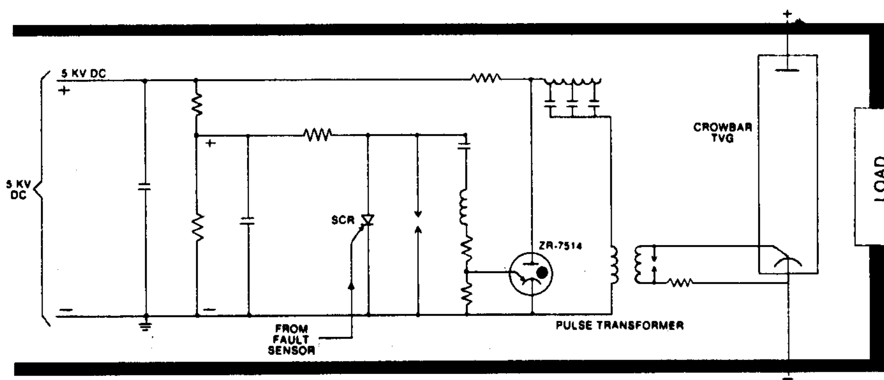
Equivalent Circuit of Power Supply, Energy Diverter, and Load

FIGURE 2



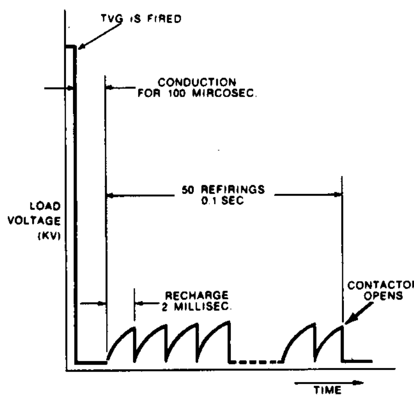
Typical Momentary Crowbarring Characteristic

FIGURE 3



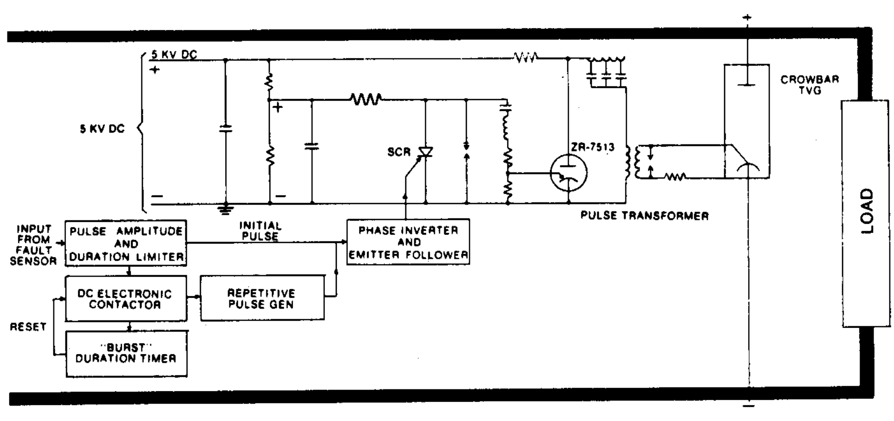
Typical "Single-shot" Firing Circuit for Momentary Crowbarring

FIGURE 4



Typical Sustained Crowbarring Characteristic

FIGURE 5



Typical "Multiple-Shot" Firing Circuit for Sustained Crowbarring

FIGURE 6

Triggered Vacuum Gap Characteristics

TYPE (SEE NOTES)	ZR-7512	ZR-7516	ZR-7517	ZR-7513
MAX. RATINGS, MAIN GAP				
DC Voltage, Max. (KV)	45	25	15	6
DC Voltage, Min. (V)	300	300	300	150
Peak Current (KA)	50	40	20	4
Total Conducted Charge (Coulombs)	0.7	0.6	0.4	0.05
Delay Time† (μsec)	0.1	0.1	0.1	0.3
TRIGGER DRIVE REQUIREMENTS				
Applied Voltage§ (KV)	5	5	5	1
Short-Circuit Current, Typical (A)	40	40	40	12
Pulse Width, 50% level, Typical (μsec)	1	1	1	0.5
PHYSICAL CHARACTERISTICS				
Envelope Dia. (inch)	3½	3½	2	¾
Envelope Height (inch)	8	5	3	1
Net Weight (lbs)	4	2	1	.03

NOTES: (1) Information on other types for higher voltages available on request.

(2) General Electric's line of high-voltage pulse ignitrons also fulfill many crowbar and capacitor-discharge switching needs. Information available on request, or write for GE publication No. PT-57A, "Ignitrons-Capacitor Discharge and Crowbar Service."

† Measured at rated voltage, time from trigger-gap breakdown to beginning of main-gap breakdown, typical.

§ Magnitude of open circuit voltage of trigger drive circuit. Trigger will fire typically at 500 to 1500 volts on the leading edge of the pulse. Rise time should be as fast as is consistent with the firing speed required. All voltage must be removed from the trigger in the intervals between firings.

For Further Information Contact Your Local GE Electronic Components Sales Office, Or:
Microwave Tube Operation
Building 269
General Electric Company
Schenectady, New York 12305
or Phone: (518) 374-2211, ext. 5-4421

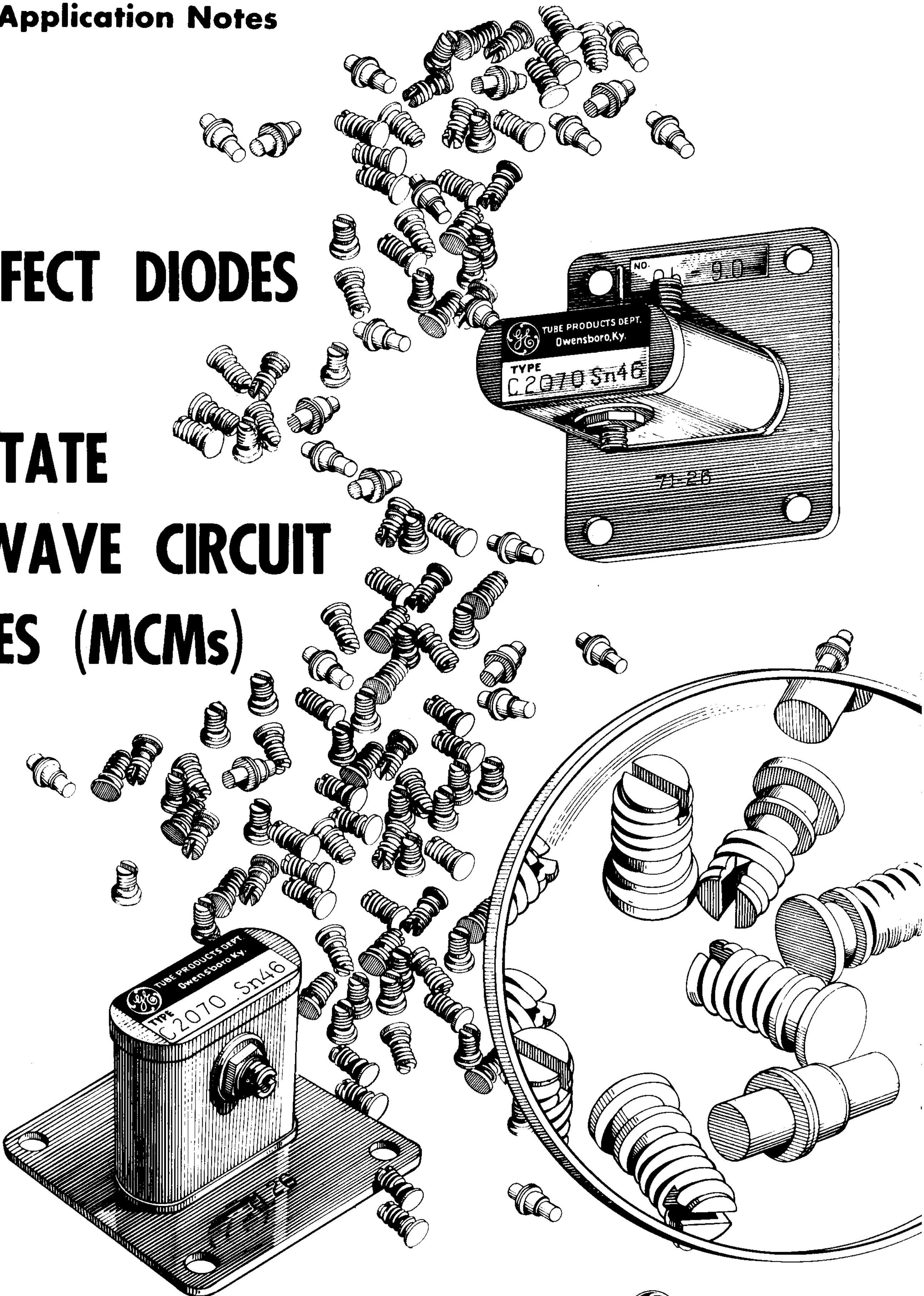
GENERAL ELECTRIC

ETD-4397B

TPD–6104

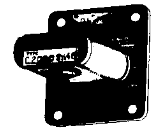
Application Notes

BULK-EFFECT DIODES AND SOLID STATE MICROWAVE CIRCUIT MODULES (MCMs)



GENERAL ELECTRIC

INDEX



I. WHAT IS A BULK-EFFECT DIODE?

The bulk-effect diode is a microwave negative resistance solid-state diode constructed of gallium arsenide (GaAs) semiconducting material. The active region is lightly doped, about 10 to the fifteenth power carriers per cubic centimeter, with heavily-doped ohmic contacts.

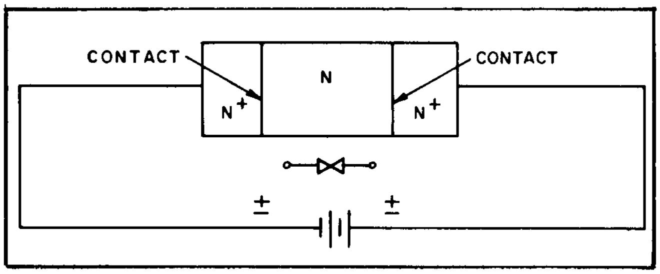


FIGURE 1

1. MECHANICAL & DC CHARACTERISTICS

Negative resistance at microwave frequencies results from the phase shift of the carriers in the active region moving at the saturated drift velocity of about 10 to the seventh power centimeters per second. The carriers consist of low mobility electrons transferred from a high mobility state in the presence of voltage gradients of about 3KV per centimeter and higher. Maximum negative resistance is obtained when the carrier currents are shifted 180 degrees in phase with the applied voltage. This simplified theory of operation leads to the following basic equation;

$$\text{Length of the active region} = \frac{\text{velocity}}{\text{frequency}} = 10\ \mu m$$

or 10 microns, ten millionth of a meter, at a 10 giga hertz frequency. This is a typical thickness of the active N-layer in Figure 2.

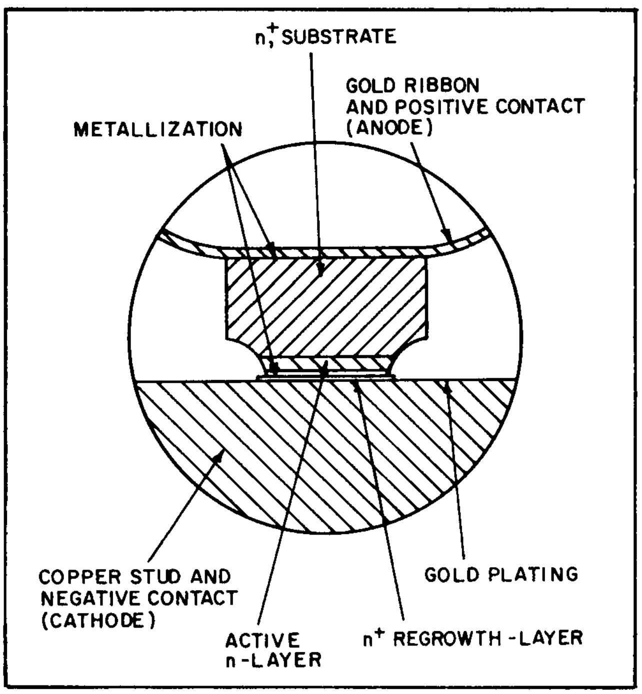


FIGURE 2

To reduce the value of capacitive reactance at microwave frequencies and the absolute value of heat generated, the active cross-sectional area of the diode must be very small. GaAs is a relatively poor conductor of heat and extreme care must be exercised in the construction of the diode. The N+ regrowth must be very thin and well bonded to the copper heat sink.

One of the significant advantages of the bulk-effect or transferred-electron effect diode is its low voltage operation. The diode is typically operated at three times its threshold voltage. The threshold voltage provides sufficient gradient to transfer the electrons to their low mobility state and for this gradient, the threshold voltage is about 3 to 4 volts. The operating voltages are accordingly in the 9 to 10 volt region. At twice the frequency the active layer would be about 5 microns thick and the operating voltages would be about 5 volts. At lower frequencies the reverse is true. In practice an effective N-layer of 10 microns will function over about an 8 to 12 GHz range. Wider ranges are practical if the variation in power output with frequency is not a critical requirement.

Many of the diode characteristics can be measured at DC or low frequency. A typical DC plot of diode voltage and current is shown in Figure 3.

TYPICAL DC CHARACTERISTICS

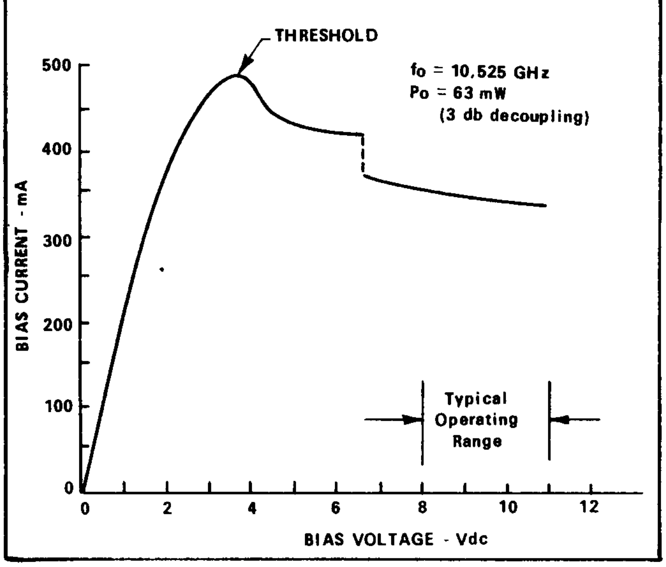


FIGURE 3

In practice it has been found that actual RF testing is required after proper DC or low frequency parameters have been measured.

II. APPLICATION OF THE BULK-EFFECT DIODE

1. HEAT SINKING AND BIAS POLARITY

To reduce the probability of damage due to over voltage, transient bias circuit oscillation and reversed polarity the bulk-effect diode should be biased as shown in Figure 4.

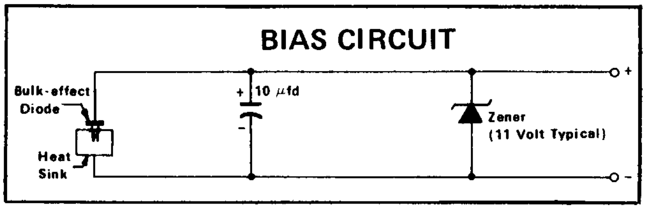


FIGURE 4

The polarity and heat-sink shown applies to the basic construction shown in Figure 2. This construction is a flip-chip mesa configuration.

To reduce the input power or bias current, smaller area mesas can be used but are mechanically difficult to flip and bond. A lower-cost reduced-efficiency low-power diode can be constructed by bonding the substrate directly to the heat sink. In this case the heat must be conducted through the substrate rather than through the thin regrowth layer. This unflipped-chip is less critical with respect to the thermal resistance of the heat sink but the chip will still be damaged if the bias polarity is not retained as before, substrate positive and regrowth layer negative. The effective thermal resistance of the low-power unflipped-chip is lowest when the anode or positive terminal is connected to the heat sink.

2. NOISE CHARACTERISTICS

The AM and FM noise characteristics of the bulk-effect diode at frequencies near the carrier are important in evaluating the capabilities of the diode as a doppler transmitter or as a local oscillator in an FM communications system. A complete FM and AM evaluation of the diode requires extensive testing. For a homodyne or zero IF CW doppler system such as police radars or intrusion alarms, the radar sensitivity is a function of the AM noise near the carrier. This oscillator AM noise is at the same frequency as the returning target echoes and will mask these signals. Diodes intended for doppler applications are evaluated for AM noise in the simplified circuit shown in Figure 5. Typical values of relative noise (for the General Electric diodes) are 115 db below the carrier. The bandpass of the test audio amplifier was chosen to represent doppler speeds of 10 to 100 mph.

The variable waveguide attenuator is set to give 0.5 VDC of detected output, edc, when driven by the diode/circuit under test. The 0.5 VDC bias is developed by the detector diode in a commercial diode mount. The audio output, e_O, of the amplifier represents the noise around the carrier and is measured with a rms reading voltmeter. The double sideband noise to carrier power ratio is then calculated by the formula: AM S/N (db) = 20 log (e_i (rms) /edc).

The value of e_i is determined by the gain of the test amplifier. Care must be taken to maintain at least 15 db in the attenuator to prevent burnout of the detector diode.

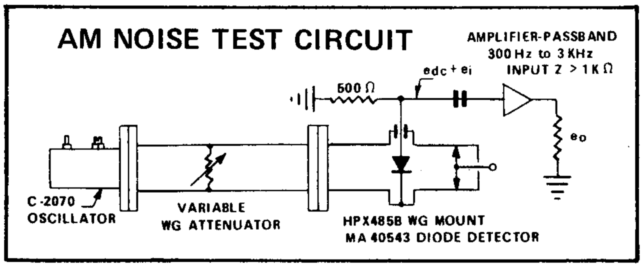


FIGURE 5

III. SELF-DETECTING OPERATION

The bulk-effect diode can also be used in a simple self-detecting doppler radar as suggested by Figure 6.

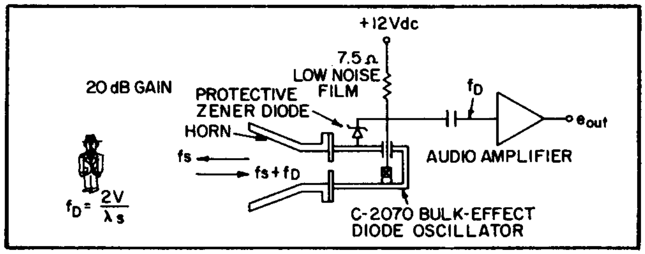


FIGURE 6

The transmitter signal, fs, is shifted in frequency as it strikes a moving target by fd, the doppler frequency. The return echo (fs + fd) mixes with the outgoing fs in the bulk-effect diode. The nonlinearities of the diode generate the sum and difference frequencies of which fd is filtered out and fed to the audio amplifier. For applications using this target detection system, an oscillator figure of merit, the self-detecting sensitivity, SDS, test can be made. Figure 7 shows a sketch of this test.

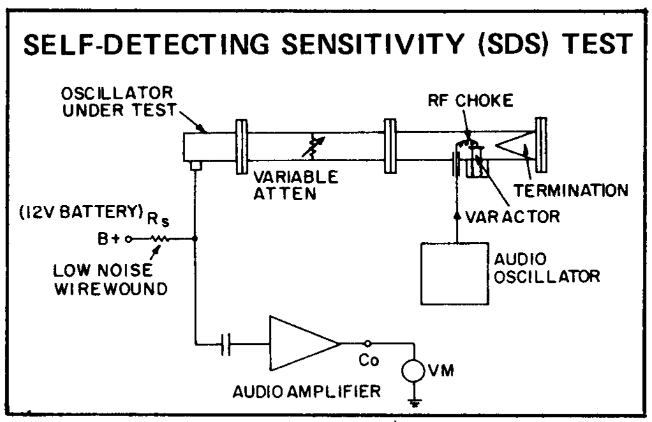
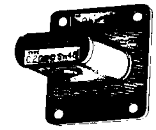


FIGURE 7

The audio oscillator varies the voltage on the varactor and thus changes its capacitance at a constantly varying rate. The varactor coupling into the waveguide is adjusted to present a small constantly changing reactance which simulates a moving target. The variable attenuator is adjusted for a given signal to noise ratio at the output of the amplifier. All General Electric oscillators and diodes designed for this SDS service are given a standardized test which has been correlated to actual radar range data taken on a walking man. Figure 7A shows the expected range on a walking man for oscillators of various SDS sensitivities.



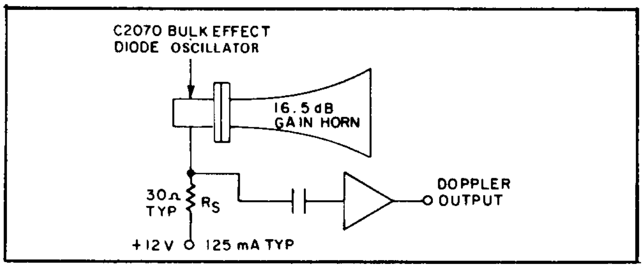


FIGURE 7A

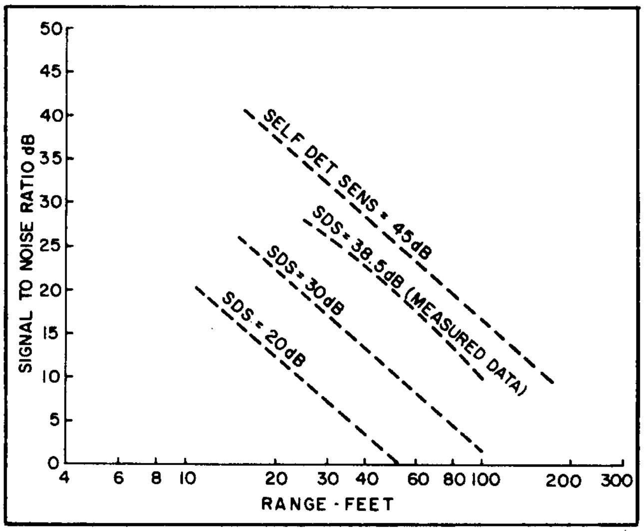


FIGURE 8

The self-detecting sensitivity test is designed not only to establish the noise output of the oscillator but to also establish its conversion efficiency as a doppler detector. These tests can be also used to evaluate the diode oscillator under all conditions of voltage, temperature and the value of the series resistor.

1. PULSE OPERATION

Bulk-effect diodes lend themselves to moderate levels of pulsed operation above CW conditions. They are more useful in this respect than transistors but less useful than the gridded vacuum tube. The primary limitation for the bulk-effect diode is the thermal time-constant of the chip. For this reason pulsed operation is limited to about one or two microseconds. The maximum duty factor is determined by the average heating effect which must be limited to the CW capabilities of the diode. A general guide line suggests maximum voltage of about ten times threshold, maximum pulse widths of one micro-second and maximum duties of about 1%. Specific maximum ratings and typical operating performances are available upon request.

2. POWER SUPPLY AND REGULATOR CONSIDERATIONS

The bulk-effect diode has no unusual requirements. A check list might be:

*Suitable DC regulation to prevent excessive pushing of the oscillation frequency.
*Suitable ripple reduction to prevent undesirable AM or FM modulation.
*Choose the operating voltage well above the threshold value to reduce spurious outputs, sufficiently high to start at cold temperature and low enough to reduce power output drop at higher temperatures.

IV. WHAT IS A MICROWAVE CIRCUIT MODULE (MCM)?

The MCM is a microwave circuit module with its active component constructed to perform a specific function. Most of the examples discussed in this brochure are bulk-effect diodes functioning as free running oscillators in a waveguide circuit. The diode is also being considered in amplifier circuits which require special treatment and operating considerations. In a few words, the amplifier circuit is usually of the reflective ferrite—isolated configuration biased and terminated to prevent any instabilities within the desired bandwidth. Early results show promise as broadbanded medium power amplifiers to compete with TWT and other similar devices. Best performance will also require minimum reactance packaging of the diode itself.

1. CIRCUIT CHOICES

Figure 9 is cutaway sketch of a typical X-band oscillator.

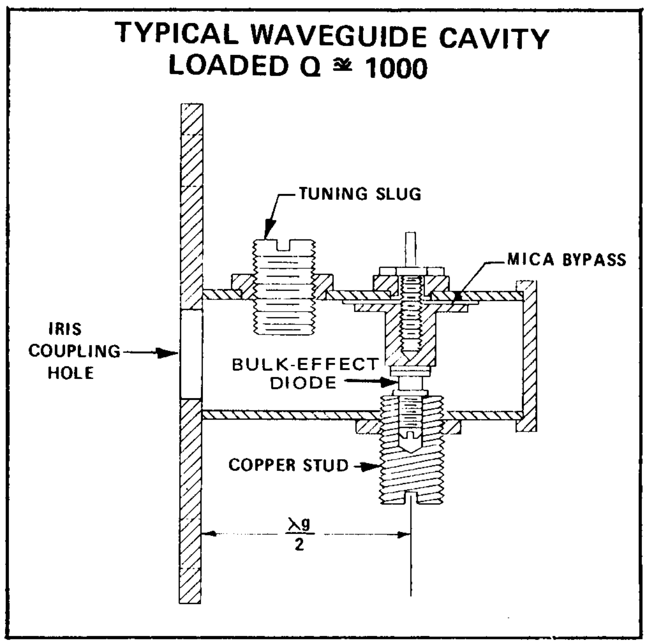


FIGURE 9

Coaxial resonator constructions are also useful particularly at lower frequencies where waveguide structures become quite large. Strip-line resonators are also used and an over-simplified advantage of each might be:

*Coaxial resonators for size reduction at lower frequencies.
*Strip-line for lower cost and size reduction.
*Waveguide for higher frequencies and performance.

Figure 10 shows a typical power output versus bias voltage at room temperature.

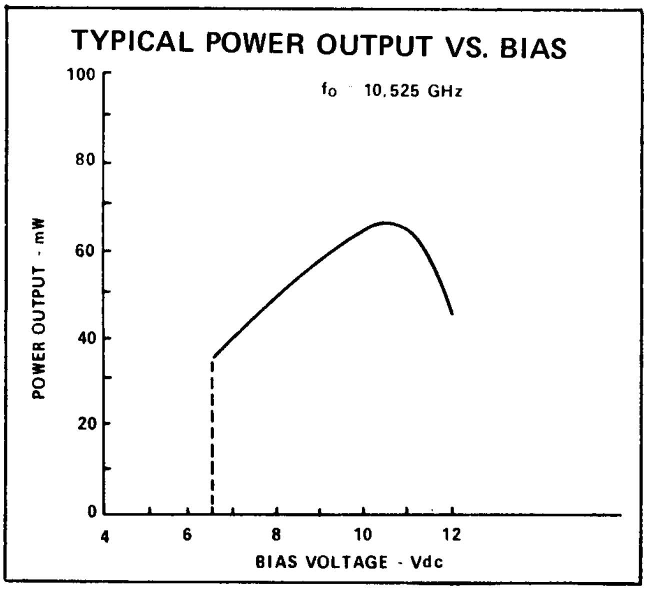


FIGURE 10

V. APPLICATION OF THE MCM

1. WAVEGUIDE VS. COAXIAL OUTPUTS

Figure 11 uses iris coupling to a waveguide system. Coaxial outputs are desirable in some systems but extra care must be taken in the choice of components. The behavior of the load in all cases must be established. Figure 11 shows results on three coupling techniques. The microminiature coaxial system used a right angle coaxial connector connected to the load through a section of a .08" coaxial cable. The miniature coaxial system used a straight-through SMA coaxial connector connected directly to a low VSWR pad before the wattmeter. The SMA 17 connector is designed to work with .141" coaxial cable.

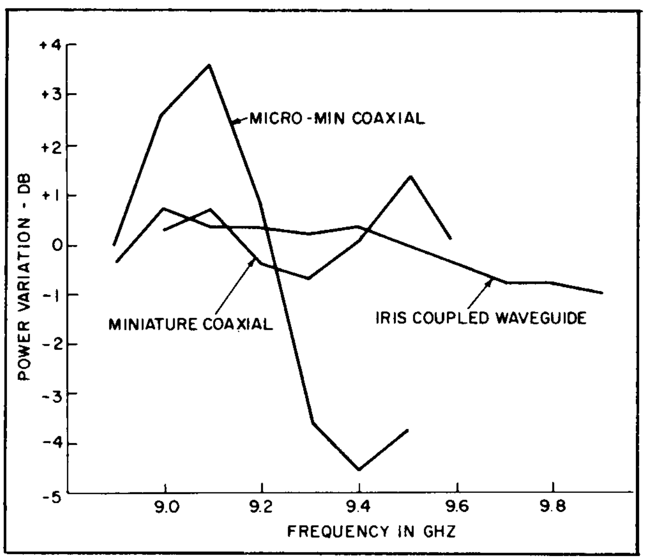


FIGURE 11

2. IMPROVING STABILITY

It has been determined that in a high loaded "Q" circuit the frequency depends primarily upon the resonator and to a lesser degree upon the diode. If the high "Q" circuit is moderately decoupled from its load frequency, stability approaching the stability of the cavity metals/materials can be obtained. This statement assumes voltage regulation to reduce pushing and stable loads to reduce pulling effects. A typical copper/brass construction yields stabilities of about 250 kilohertz per degree of centigrade of temperature change. Figure 12 is a typical stability plot for a circuit similar to Figure 9.

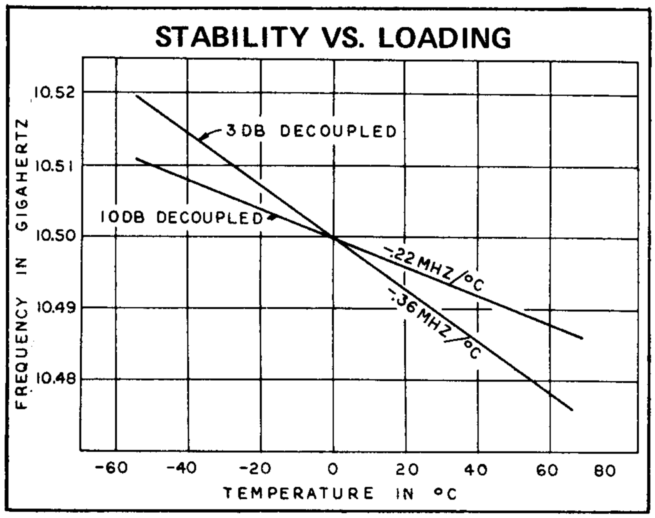
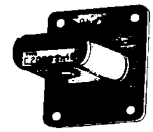


FIGURE 12

The GaAs bulk-effect diode is more efficient at lower temperatures and the circuit design must be optimized for minimum parasitic resonances near the desired frequency. At lower temperatures the oscillator may start at a parasitic frequency which may or may not pre-empt proper operation at the desired frequency. This effect plus the power and frequency changes at high temperatures usually determine the optimum operating voltage and/or the acceptable level of regulation. Figure 13 is a plot of a typical family of curves of power output versus frequency for various bias voltages. A similar change in frequency with temperature can result and frequency stability is also a design criteria when operating voltages are selected.

Temperature stabilities of a few parts per million per degree centigrade can be obtained through the use of invar materials, significant decoupling to obtain high to added circuit "Q's" and the proper choice of the operating voltage over the desired temperature range. Further improvements can be obtained by addition of bi-metal compensation. Figure 14 shows the results of an invar cavity local oscillator for X-band radar and radar beacons tunable over a 500 megahertz range around a 9350 megahertz center-tuning frequency. This data was taken with a coaxial output connector.



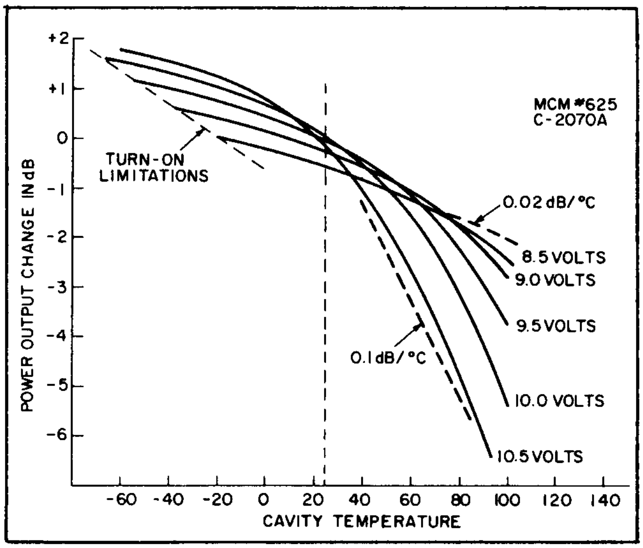


FIGURE 13

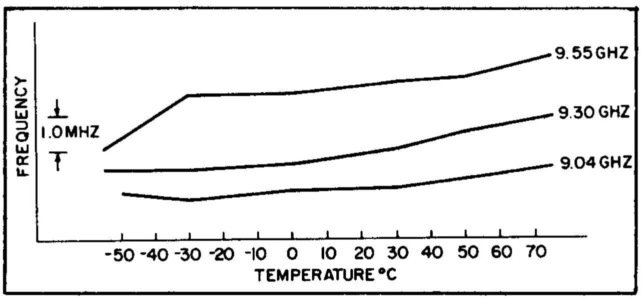


FIGURE 14

3. MECHANICAL TUNING

Figure 15 is a plot of power output versus the frequency change with the insertion of the tuning screw in an oscillator circuit similar to Figure 10 with an iris coupled output and a coaxial connector output. The reduction in power output for the iris coupled case was due to the low frequency cutoff characteristics of the waveguide test circuit.

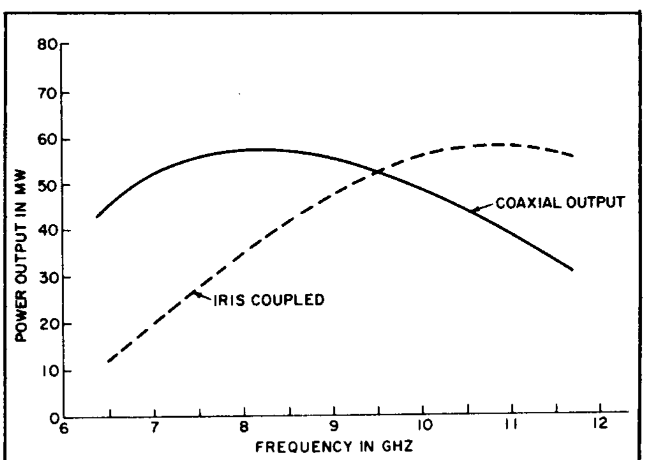


FIGURE 15

4. VOLTAGE TUNING

Varactors and YIG tuning can be used to voltage/current tune the bulk-effect diode oscillator. The range of tuning depends to a varying degree upon the quality of the tuning element, loaded "Q's" and the diode itself. The varactor and YIG element should exhibit highest available "Q's" at the operating frequency commensurate with the required tuning range and the cost of the varactor or YIG element. Further improvements in voltage tuning ranges can be obtained through special diode packaging with reduced series inductance and parallel capacitances. Work continues in this area. Low cost varactors can be used to tune the usual AFC ranges required for single-frequency operation and oscillators of an octave or so have been tuned with well designed circuitry and allowable power variations of several db over the chosen band.

5. PUSHING AND PULLING

Pushing defines the changes in oscillation frequency as the bias voltage varies. Figure 16 shows typical results on an X-band oscillator.

Pulling defines the changes in oscillator frequency with changes in load impedance. Figure 11 shows this to be very dependent on the quality of the connecting circuitry between the resonant element and the load itself. Pushing is more a function of the diode. This pushing effect can be used only for small frequency excursions in AFC applications, bearing in mind that the tuning rate in MHz/volt varies from diode to diode. The rate also varies within one diode as the temperature varies. Figure 16 shows that actual tuning reversals occur at higher voltages. For more linear results in an AFC loop, a varactor or YIG tuning should be used.

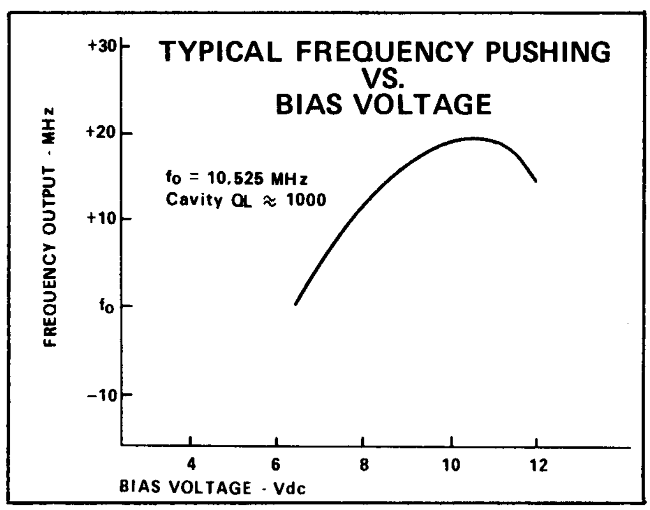
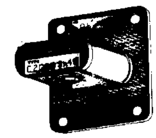


FIGURE 16



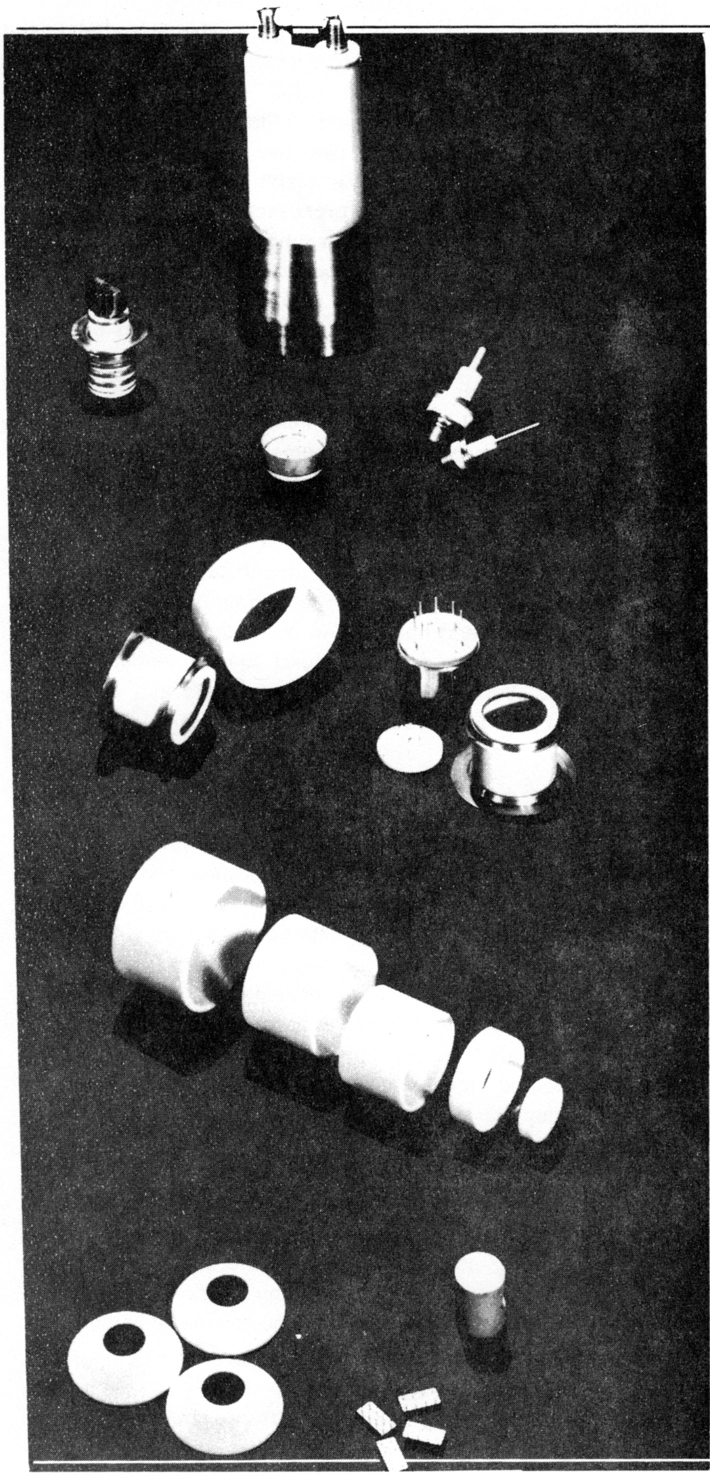
GENERAL ELECTRIC

MICROWAVE DEVICES
PRODUCTS SECTION

Owensboro, Ky. – Schenectady, N.Y.

TPD–6104

Hi-TECH Ceramics



Ceramic and ceramic-metal components custom-engineered and manufactured for the highly technical application

For top-of-the-industry ceramics and ceramic-metal components, specify *Hi-TECH* — a superior-quality line of alumina and forsterites, particularly oriented toward the highly technical application. Manufactured by General Electric's Tube Department, *Hi-TECH* ceramics carry the backing of a highly advanced technology resulting from years of ceramic experience closely allied to the electronics industry. This combination of materials excellence and competence in the design and manufacture of complex technical ceramic components is ready to work for you in solving your most difficult component problems.

General Electric's acknowledged position as a leader in the electron-tube field — one of industry's most demanding — offers you these advantages:

- Special processes, techniques, and treatments
- Unique manufacturing facilities
- Exact product-customization
- "Breakthrough" engineering capability

For example, GE pioneered many of the basic ceramic-metal techniques and processes enjoying industry-wide use today. And ceramic-metal assemblies are currently being engineered to perform with long life and reliability in a variety of severe environments. In fact, our reputation stems largely from the successful application of a multitude of such innovations covering the broad range of General Electric products, those of our customers, and the Government and its agencies.

General Electric specializes in overcoming "the difficult job". End use and operating environment are all we need to know in most cases to come up with the optimum configuration, materials, and processing necessary to meet rigorous life and reliability specifications.

Whether your application involves going to the planets, orbiting the earth, riding a nuclear warhead or simply a small feedthrough, let General Electric *Hi-TECH* ceramics help you.

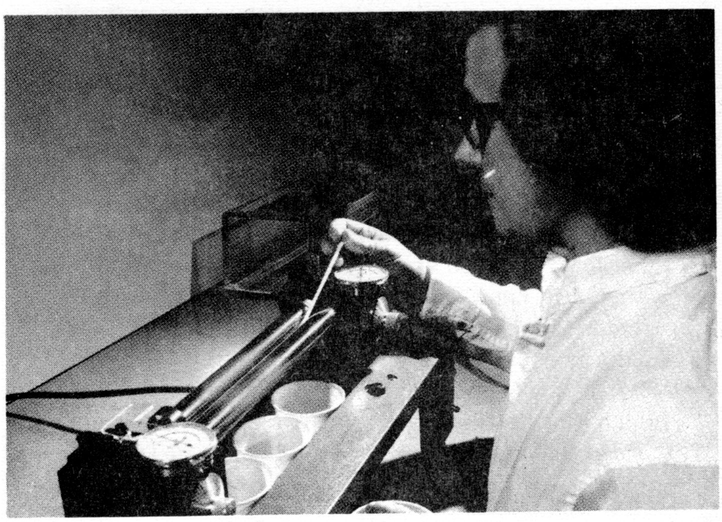
GENERAL ELECTRIC

A Broad Selection of Body Compositions

General Electric's *Hi-TECH* line offers a broad variety of alumina, forsterite, and other special electronic-grade ceramic materials. Each is precisely engineered and manufactured under rigid quality control to meet the demands of your particular application. Experienced consultants — just one facet of our start-to-finish capability — will be happy to assist you in the selection of the optimum composition for your needs, in many cases based merely on projections of where and how the device is to be used. The most widely used compositions, and their respective characteristics, are tabulated in the properties chart in the centerfold.

Hi-TECH forsterite ceramics are dense, fine-grained, vacuum-tight materials originally developed for use in hermetically sealed devices such as electron tubes, rectifiers, and vacuum switches. Their unique combination of properties, however, has led to broadly increased usage. *Hi-TECH* forsterites combine high mechanical strength, low dielectric constant, and resistance to corrosion, with a unique expansion characteristic ideally suited to ceramic-metal sealing techniques. For example, the thermal expansion characteristic of one patented composition matches that of titanium metal over a broad range of temperatures, permitting the manufacture of many devices for critical space-age applications. Forsterite ceramics are readily metallized by "molybdenum-manganese" or "solution metallizing" techniques and are easily bonded to metals by active-metal processes.

Hi-TECH alumina ceramics meet the life, reliability and severe-environment factors imposed by the space age in virtually every respect: exceptional hardness, flexural strength and electrical properties combined with an excellent resistance to highly corrosive agents. Available in a broad range of grades, from "general purpose" to virtually pure sintered forms approaching theoretical density, *Hi-TECH* alumina characteristics are readily matched, in the required combinations, to the needs of each application. The "general purpose" materials range in composition from 94 to 97% alumina, and have a precise balance of additive oxides to control such critical properties as flexural strength, density, dielectric power factor, and metallizeability. Other grades are tailored to have extremely fine grain size and low pore volume where surface finish is important. Still other compositions exceeding 99% alumina are designed to have ultra-low dielectric loss for critical electronic applications, or inertness to the severely corrosive alkali metals used in many new generation equipments. All of these compositions can be joined to metals and non-metals by appropriate sealing techniques. The complete *Hi-TECH* line of aluminas is also available metallized and plated to your specifications—a factor of considerable importance to those applications demanding one of the high-purity aluminas which require special metallizing treatments.



Roller Micrometer capable of measuring and sorting precision-ground ceramic parts to tolerances greater than 0.00001 inch.

Ceramic-metal Seals and Structures

Over the years, General Electric's leadership in the electron-device industry has culminated in a technology conversant with essentially all of the known natural elements and almost every branch of the physical sciences. The two basic techniques for sealing ceramics to metals — "molybdenum-manganese" and "active-metal (including titanium hydride)"—were pioneered by General Electric over two decades ago. Today, this same leadership is making available even newer and more sophisticated techniques and processes to meet the critical demands of aerospace and other advanced systems.

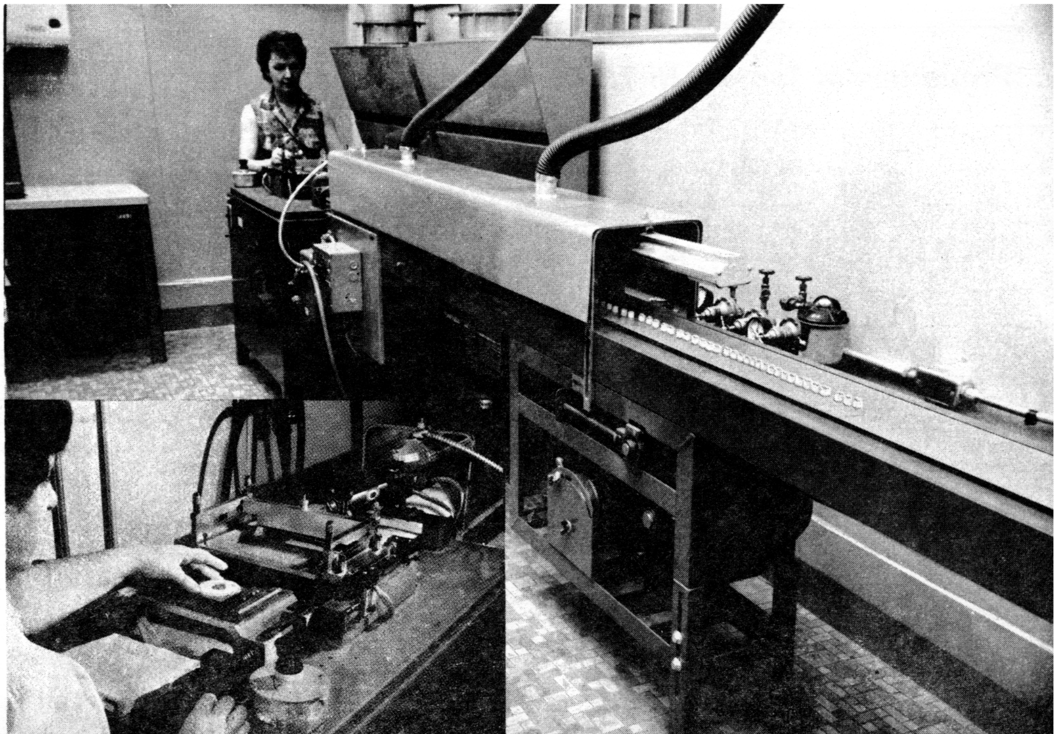
A wide range of ceramic-metal combinations is available, in almost any configuration, for your consideration. *Hi-TECH* alumina and forsterite ceramics are currently being used in assemblies which incorporate seals to copper, monel, molybdenum, tungsten, chromel, alumel, titanium, tantalum, niobium, nickel-iron and nickel-cobalt-iron alloys and certain stainless steels. In fact, depending upon geometry and seal design, virtually all metals can now be satisfactorily sealed to one of our ceramic materials. Through the use of less widely known, as well as several recently developed techniques, ceramic-to-metal seals can also be prepared to withstand duty in severe environments involving ultra-high temperatures, high RF fields, corrosive vapors and liquids, and many others (see "Typical Applications"). Here, again, General Electric engineers are available to assist in selecting the materials and structural design that will serve your application best.

Facilities and Staff

Complete engineering and manufacturing facilities are maintained by General Electric's Tube Department for the design, development and manufacture of *Hi-TECH* ceramics as well as a wide range of components incorporating ceramic-to-metal seals. Our staff includes ceramists, metallurgists, chemists, and physicists, as well as electronic, electrical and mechanical engineers—many with advanced degrees and long years of experience in their respective fields. Over the years, this staff has contributed significantly to the state of the art, as it is known today, in glass, quartz, and ceramic devices, and ceramic-to-metal seals.

Thus, a technologically advanced capability is in place and ready to undertake the manufacture of components and devices of virtually any size and degree of complexity. In our laboratories, experienced technicians and supporting personnel utilize the latest processes and equipments to assure consistent product quality and adherence to your specifications:

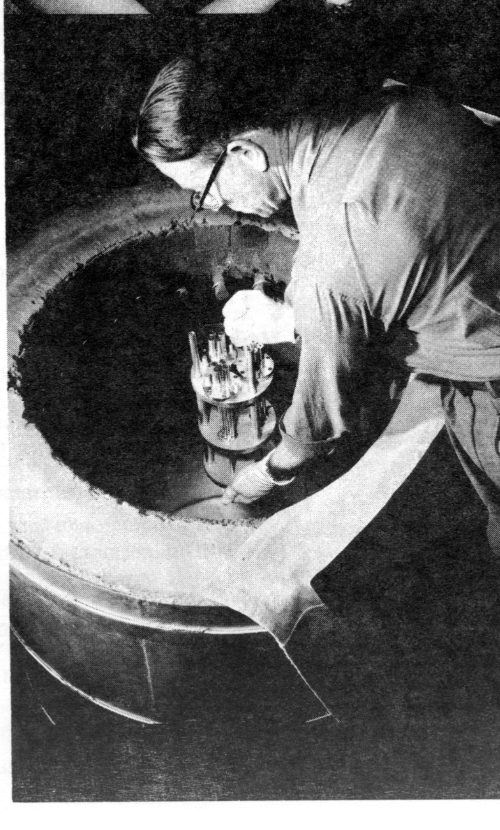
- **Modern Powder Processing Techniques** for batching, milling, drying. Pre-tested synthetic oxides, fluxing additions, grain refining agents and fugitive binders are scientifically batched and vibratory milled to the exact particle size distribution required. Spray drying then ensures free-flowing agglomerates for feeding to the presses.



Applying metallizing by automatic screening machine. (Inset) View of loading end showing operator placing ceramic part in vacuum fixture before being fed into automatic silk-screening machine where it will receive a metallizing coating.

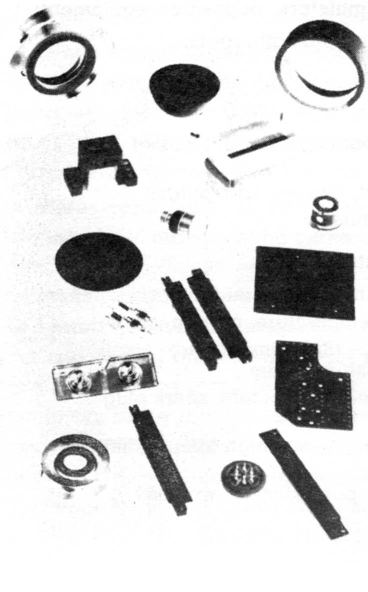
Semi-automatic banding machine permits simultaneous application of several metallizing bands to ceramic cylinders.

Diamond-wheel Centerless Grinding of Alumina Cylinders



Brazing fixture being inserted in large super-dry hydrogen retort furnace. These furnaces will accommodate ceramic-metal assemblies up to 2½ feet in diameter and 3 feet high.

Formed Ceramic Components Prior to Firing

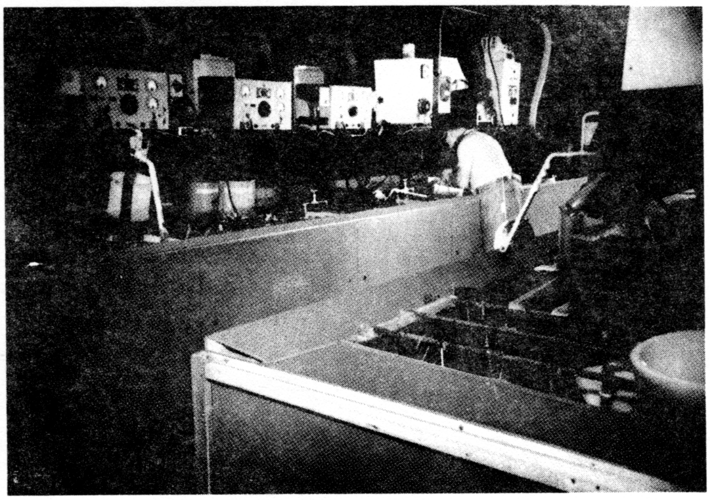


• **Wide Variety of Forming Techniques** including automatic and semi-automatic dry pressing for high-volume production, extrusion, and isostatic pressing for superior uniformity of properties. Intricate prototypes and other complex shapes are precision machined from isostatically-formed blanks.

• **High-Temperature Firing Equipments** ensure that the exact properties are imparted to the formed ceramic article during the critical sintering operation. Electrically heated, gas-fired, or hydrogen atmosphere furnaces are utilized over cycles ranging from a few hours up to several days.

• **Complete Machining and Finishing Facilities** for parts of almost any size and shape. Typical operations include diamond-surfaced finishing equipment for centerless, cylindrical or surface grinding to extremely close tolerances; lapping; and special vibratory techniques for finishing and polishing.

• **Advanced Metallizing Techniques** include a variety of methods such as automatic screening, semi-automatic cylindrical banding, dip coat, brush, and others, depending upon geometry and quantity involved. Unique high-temperature furnaces permit sintering of metallizing on ceramic shapes up to eighteen inches in diameter and nearly two feet high. The metallized areas are subsequently electro-plated to enhance wetting by the molten braze alloy during assembly. Each step of the critically important metallizing process is closely monitored to assure that all coatings are of "electron tube" quality.



Electroplating Ceramic Parts and Assemblies. (This area also includes facilities for electropolishing, wet and dry abrasive blasting, and chemical cleaning.)

• **"White Room" Assembly Conditions** ensure component integrity. Assembly operations are carried out in "white room" areas under close control of temperature, humidity, and airborne dust, using many materials, fixturing techniques and equipments developed for the electron tube industry.

• **Wide Assortment of Furnaces for Sealing and Brazing** of assemblies incorporating ceramic-to-metal seals. Hydrogen-atmosphere furnaces are selected from box, continuous-belt or retort types. And resistance or high-frequency heated bell and retort furnaces are extensively used in vacuum firing, brazing and sealing operations. In our hydrogen furnaces, protective atmospheres range from wet to super-dry, depending upon the alloys being brazed; assemblies up to three feet high and over two feet in diameter can be accommodated in the larger retorts. The resistance and high-frequency-heated furnaces are ideal for use in active-metal sealing, or high-temperature brazing to oxygen-sensitive metals such as titanium, tantalum and niobium. Other metal-joining equipments include a variety of welders—spot, electron-beam, and tungsten-inert-gas types—in various sizes.

• **Thorough Inspection Procedures** ensure product reliability. Closely controlled in-process and final inspection procedures are augmented by a broad selection of modern inspection and testing implements. Typical examples include: automatic gauging equipment for dimensional checks, helium mass spectrometer for leak detection, metallographic examination, x-ray fluoroscopic inspection, chemical analyses, and dye-penetrant testing to aid in the detection of minute flaws. A rigid system of equipment and gage calibration is also in force in our facilities.

	FORSTERITE					
Body Designation	**F-202 Forsterite**			**OW-6 Forsterite**		
Body Type	Low loss, titanium-matching			General Purpose forsterite		
Meets GE Specifications	—			—		
Color	White			White		
Alumina Content, percent	—			—		
Constituent Oxides	MgO, SiO_2 Al_2O_3, BaO			MgO, SiO_2 Al_2O_3		
Porosity[a]	Non-porous			Non-porous		
Gas Permeability[b]	None			None		
Hardness, Mohs' Scale	7.5			7.5		
Density	3.11-3.15			2.8-3.0		
Flexural Strength, K psi	20-25			20-25		
Thermal Expansion Coefficient, cm/cm/°Cx10^{-6}						
25-300°				9.5		
25-600°				10.5		
25-800°	10.8					
25-900°				11.3		
Dielectric Constant	**25°C**	**200°C**	**500°C**	**50°C**	**400°C**	**600°C**
10^2 Hz	6.77	6.99	14.73			
10^4 Hz	6.76	6.96	8.13			
10^6 Hz	6.76	6.94	7.31	5.2	5.6	6.0
10^7 Hz	6.76	6.94	7.28			
8.5×10^9 Hz	6.74	6.92	7.23			
10^{10} Hz						
Loss Tangent (Tan δ)	**25°C**	**200°C**	**500°C**	**50°C**	**400°C**	**600°C**
10^2 Hz	.000515	.00277	4.29			
10^4 Hz	.000240	.00124	.178			
10^6 Hz	.000245	.00067	.00975	.001	.006	.070
10^7 Hz	.00025	.00052	.00394			
8.5×10^9 Hz	.00080	.0015	.0027			
10^{10} Hz						

(a) As determined by water absorption or dye pentration.
(b) Measured using a helium mass spectrometer leak detector and a .010-inch thick specimen.
(c) AT-100 Alumina ceramic is an improved version of A-976.

Properties of Typical *Hi-TECH* Ceramic Bodies

				ALUMINA														
F-118 Forsterite			**A-994 Alumina**	**A-1004 Alumina**			**A-919 Alumina**			**A-923 Alumina**			**A-1000 Alumina**			**AT-100 Alumina**		
Higher expansion than F-202 and OW-6 forsterite			Low loss, calcia-free	General purpose, easily metallized			Low loss, easily metallized			General Purpose, low loss			Fine grained, very small pore size and volume			Ultra-low loss, extremely corrosion resistant, near theoretical density		
A5C2			A5D7A	A5D7A			A5D8			A5D8			--			(c)		
Buff			White	White			White			White			Cream			Translucent, colorless		
---			94	94			97			97			99.8			99.9+		
MgO, SiO_2 Al_2O_3, BaO			Al_2O_3, SiO_2 MgO	Al_2O_3, SiO_2 MgO, CaO			Al_2O_3, SiO_2 CaO			Al_2O_3, SiO_2 MgO, CaO			Al_2O_3			Al_2O_3		
Non-porous			Non-porous	Non-porous			Non-porous			Non-porous			Non-porous			Non-porous		
None			None	None			None			None			None			None		
7.5			9	9			9			9								
3.10-3.14			3.66-3.69	3.63-3.67			3.75-3.79			3.74-3.77			3.91-3.94			3.97-3.98		
20-25			45-50	50-55			45-50			50-55			40-45			35-40		
11.2			7.7	7.7			7.8			7.8			7.8			7.8		
25°C	**200°C**	**500°C**	**25°C**	**25°C**	**200°C**	**500°C**	**25°C**	**200°C**	**500°C**	**25°C**	**200°C**	**500°C**	**25°C**	**200°C**	**500°C**	**25°C**	**200°C**	**500°C**
6.63	6.80	8.78		10.48						10.26			10.08			9.98	10.21	12.60
6.62	6.78	7.20					9.62	9.60					10.07	10.33		9.98	10.21	10.86
6.62	6.77	7.09		9.10	9.21	9.83	9.38	9.59		9.28	9.50	9.95	9.98	10.31	10.80	9.98	10.21	10.63
6.62	6.77	7.08		9.00	9.20	9.74	9.37	9.59	10.03	9.27	9.50	9.91	9.96	10.29	10.76	9.98	10.21	10.62
6.59	6.73	6.98		9.01			9.35			9.24			9.77			9.96		
			8.7															
25°C	**200°C**	**500°C**	**25°C**	**25°C**	**200°C**	**500°C**	**25°C**	**200°C**	**500°C**	**25°C**	**200°C**	**500°C**	**25°C**	**200°C**	**500°C**	**25°C**	**200°C**	**500°C**
.00098	.00134	.421		.00226						.00227			.00048			.000007	.000603	.289
.00013	.00044	.0188					.0206	.00089					.00135	.00101		.000001	.000045	.0237
.000072	.00074	.00329		.0142	.00163	.0133	.00139	.00006		.00952	.00089	.00976	.00664	.00051	.00341	.0000015	.000010	.00082
.00011	.00025	.00149		.00228	.00105	.0071	.00030	.0001	.0035	.00165	.00040	.0037	.00612	.00170	.00135	.000007	.000007	.0002
.00083	.00092	.00119		.00125			.00069			.00067			.00258			.000048		
			.00015															

Typical Areas of Application

Vacuum and gas-filled devices:
- Electron, image, and x-ray tubes
- Switches, spark gaps • Ionization chambers

Hermetically sealed electronic components
- Capacitors • Resistors • Relays
- Crystals • Batteries

Semiconductor devices:
- Transistor headers
- Rectifier and SCR housings
- Flatpaks

Substrates for thin- and thick-film circuitry

Electrical equipment and systems:
- Ordnance devices
- Windows for transmitters, proximity switches, etc.
- Pick-up heads for information storage units

Vacuum equipment, space simulators, deposition equipment:
- Multi-lead and thermocouple feedthroughs
- High voltage bushings

Mechanical devices:
- Dimensionally-stable components for all types of machinery
- Abrasion resistant fixtures, gauges, wire guides
- Brazing and soldering fixtures

Devices in severe environments:
- Thermionic converters — high temperature, cesium vapor
- MHD generators — high temperature, cesium vapor
- Turboelectric generators — high temperature, alkali metal vapor; high temperature steam
- Jet engines — thermocouple connectors, spark plug connectors
- High temperature rectifier tubes — high temperature, alkali metal vapor
- Rechargeable batteries — potassium hydroxide electrolyte
- Fuel cells

NOTES

GENERAL ELECTRIC

TUBE PRODUCTS DEPARTMENT
OWENSBORO, KENTUCKY 42301