TWT's

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ENGINEERING NOTES

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GENERAL INFORMATION ON TRAVELING-WAVE TUBES

R. A. Huggins*

Engineers who design circuits which use conventional low frequency vacuum tubes usually have at least a speaking acquaintance with their theory of operation. In designing circuits, or in deciding which type of tube to use, careful use is made of the various characteristic curves associated with these tubes. The following discussion will similarly acquaint the applications engineer with the simple theory of the operation of the traveling-wave tube and familarize him with the various characteristic curves which describe its operation.

MICROWAVE AMPLIFICATION

As the microwave region of the radio frequency spectrum came into practical engineering use, devices were needed which could function as better amplifiers and oscillators. Above a few hundred megacycles conventional vacuum tubes no longer amplify properly because the period of one cycle of the radio frequency wave approaches the time it takes the electrons to travel from cathode to anode. As a result the amplification factor of the tube decreases with increasing frequency until no useful gain can be obtained. This is known as the transit time effect.

The first device which served as a practical amplifier above a few hundred megacycles actually made use of the transit time effect. This device, known as a klystron, became useful primarily as an oscillator. The klystron, however, is also capable of acting as an amplifier. Klystron amplifiers are capable of giving high values of amplification, but they have two serious limitations for many applications: They are relatively narrow band devices which usually have no more than a few megacycles bandwidth between half-power points, and tuning the amplifier to different frequencies requires a mechanical adjustment. The klystron is, however, the forerunner of a general class of microwave electron devices which utilize transit time effects in electron beams. This general class of devices includes the traveling-wave tube (TWT).

A klystron is a narrow band device because it involves the use of a very sharply resonant, high-impedance tuned circuit. It is necessary for a klystron to use such a circuit because interaction takes place with the electrons over very short physical distances, and it must develop high electric fields to have efficient interaction with the electrons. It was reasoned, however, that one might obtain a very broadband amplifying device by using a low-impedance, broadband circuit which developed relatively small electric fields that interacted with the electrons over a long physical distance. Such is the basis on which the traveling-wave tube works, and this basic concept led to a wide range of devices which are useful as broadband amplifiers and oscillators in the microwave region.

The traveling-wave tube yields high values

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of amplification over wide frequency ranges with no mechanical tuning mechanism or variation of any voltages or currents applied to the tube. Power amplification greater than 10,000 (i.e., 40 db gain) over a 2:1 frequency range has been obtained with this type of tube. The actual number of megacycles covered by such tubes can be a staggering figure. For example, an operating bandwidth of 2000 megacycles has been obtained in one typical tube type; 7000 megacycles has been obtained in another.

Tubes can be designed to emphasize various special characteristics. For instance, tubes are built which have low noise properties, high power output, or controlled variations of gain with frequency. Many special tubes are designed and built for particular applications.

FORWARD-WAVE HELIX TYPE AMPLIFIER

Among the simplest circuit elements which satisfy the broadband low-impedance circuit requirement of the traveling-wave tube is a simple helical transmission line. On this so-called helix, the radio frequency energy essentially travels at the velocity of light along the wire from which the helix is wound. Since the wire and the energy follow this helical path, progress of the energy along the axis of the helix is at some fraction of the velocity of light. This velocity is determined by the geometry of the helix (i.e., circumference and pitch) and by the dielectric loading due to the structure which supports the helix. The fields associated with this "slow-wave" extend outward and inward into the region adjacent to the helix, and it is into this region that electrons are sent to interact with the waves.

If electrons are directed along the axis of the helix at essentially the same velocity as the waves, an interaction between the waves and electrons occurs wherein the electrons become bunched. As the electrons progress along the axis they become packed into tighter and tighter bunches. Simultaneously, the radio frequency fields on the helix increase in magnitude as they progress down its length. The increase in energy of this "growing wave" is just balanced by a decrease in the average kinetic energy of the electrons in the beam. In other words, the electrons, on the average, have been slowed down and have given up just enough energy to account for the increasing energy or power level in the waves on the helix.

In practice, the electron beam is formed by an electron gun and focused down the center of the helix to a collector electrode at the farther end (see Figure 1). The electron beam is maintained in a tight pencil beam throughout the length of the tube by the confining forces of a longitudinal magnetic field which surrounds the tube. Electron velocity is determined by the voltage difference between the cathode in the electron gun and the helix, and this voltage difference is ad-

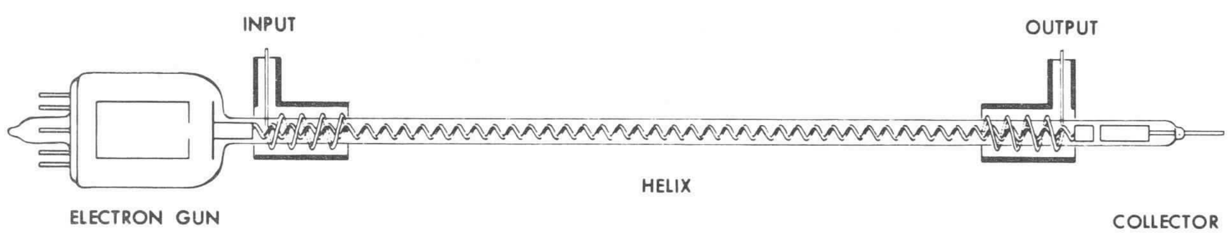

Figure 1. Schematic View of a Traveling-Wave Tube.

justed to give the electrons the right velocity for interaction with the waves. The signal to be amplified is coupled onto the end of the helix nearest the electron gun and propagates along the helix in the same direction as the electron beam. Because of the interaction, the fields on the helix grow exponentially with distance, and these amplified waves are coupled off the helix at the end farthest from the electron gun.

The devices which couple the energy onto and off the helix are special directional couplers which are in themselves helices. These helices are outside the vacuum envelope and are matched to the input and output coaxial transmission lines. This type of coupler is capable of transferring most of the energy from the lines to the main helix, which is inside the vacuum envelope, without making physical contact with the helix. These "coupled helix matches" have the additional advantage of having very wideband transfer characteristic. Further, they are capable of presenting a standing-wave ratio to the coaxial transmission line of 1.5:1 or less.

SPECIFIC CHARACTERISTICS

In the following sections, various characteristic curves that are encountered in traveling-wave tube operation are discussed.

Helix Voltage vs Frequency. A helix is usually designed so that the "slow-waves" travel at a velocity in the range of 1/10 to 1/30 of the velocity of light. If the wave velocity is examined as a function of frequency, it is found that a helix has a general characteristic curve similar to that shown in Figure 2.

In region A, wave velocity varies as a function of frequency; in region B, wave velocity is essentially independent of frequency. A broadband amplifier is normally operated in region B – the "non-dispersive region."

As mentioned before, the electrons must

FIG. 2

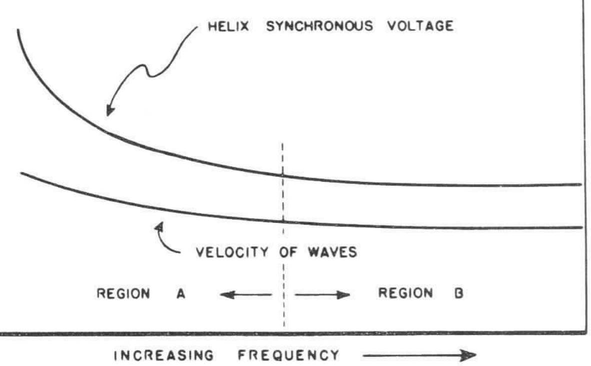

Figure 2. Phase Velocity of the Waves and Synchronous Helix Voltage as a Function of Frequency.

travel at essentially the same velocity as the waves (synchronous velocity) in order to obtain the growing wave interaction. Since the velocity of the wave varies as a function of frequency, the velocity of the electrons must also be varied so that interaction is maintained. Inasmuch as the velocity of the electrons is determined by the voltage difference between the cathode and the helix, there is also a characteristic curve of synchronous helix voltage as a function of frequency (see Figure 2).

The reason a helix type traveling-wave tube can amplify over wide bandwidths without changing helix voltage can now be easily seen. In the non-dispersive region the synchronous voltage is essentially independent of frequency. That is, the helix voltage which gives the desired interaction between the waves and electrons can be fixed at one value, and this will be the correct value to provide interaction and amplification over a wide range of frequencies.

Gain vs Helix Voltage. To obtain the greatest interaction between the waves and the electrons, and thus the greatest amplification, the helix must be operated at synchronous voltage.

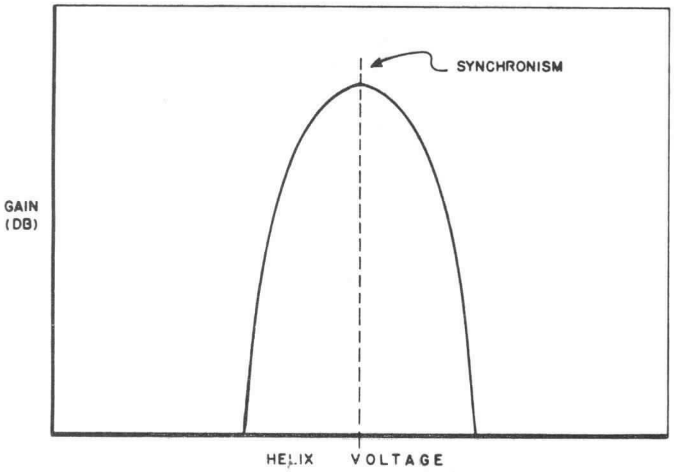

Figure 3. Gain as a Function of Helix Voltage.

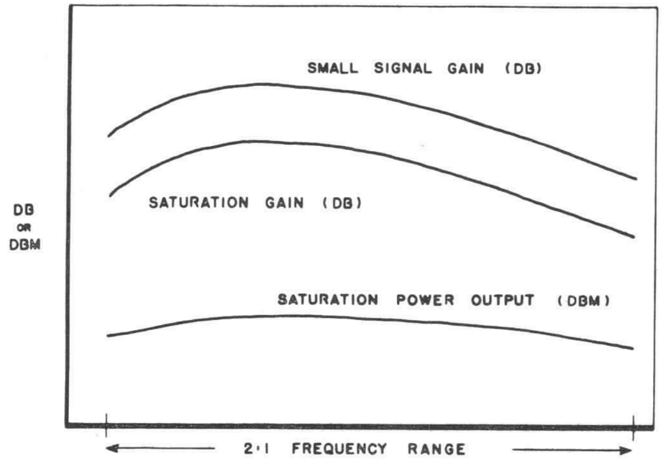

Figure 4. Gain and Power Output as a Function of Frequency.

Should the helix voltage be varied away from synchronism, the gain will decrease as shown by the curve in Figure 3. This characteristic of the TWT is parabolic in shape.

Gain and Power Output vs Frequency. Figure 4 shows a typical plot of gain vs frequency for a fixed value of helix voltage. Small-signal gain is obtained for all values of input signal level except those which drive the tube near its maximum or saturation power output. Saturation gain is the gain obtained when the tube is driven to its saturation power output level. It is seen that the gain and power output curves generally reach a maximum near the center of the band and droop off toward the band edges.

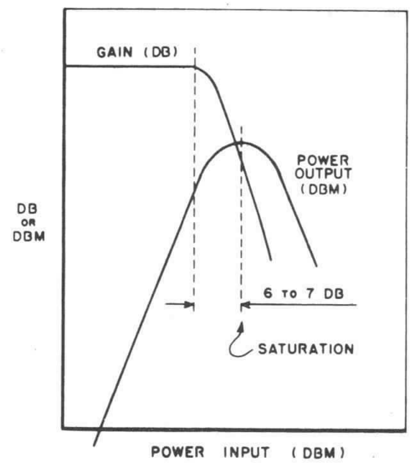

Figure 5. Gain and Power Output vs Power Input at a Fixed Frequency.

Power Output and Gain vs Power Input. Figure 5 shows the variation of power output and gain as a function of power input at a fixed frequency. The amplifier is linear until the power input is within 6 to 7 db of the value which drives the amplifier to saturation. At this point the power output curve begins to droop over, and gain of the tube begins to decrease. At saturation the power output is a maximum, and gain has decreased to the saturation gain level. As the tube is driven beyond saturation the power output drops off, and gain continues to decrease. To obtain maximum power output from an ordinary traveling-wave tube, the right amount of power input must be applied. As will be noted, the saturation region does not exhibit the leveling off or flat saturation characteristic that is found in conventional low frequency amplifier tubes.

Gain vs Beam Current and Grid Voltage. Figure 6 shows that the gain at any fixed frequency is essentially a function of the 1/3 power

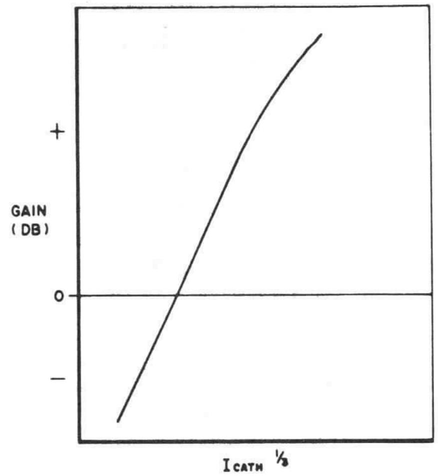

Figure 6. Gain vs Beam Current.

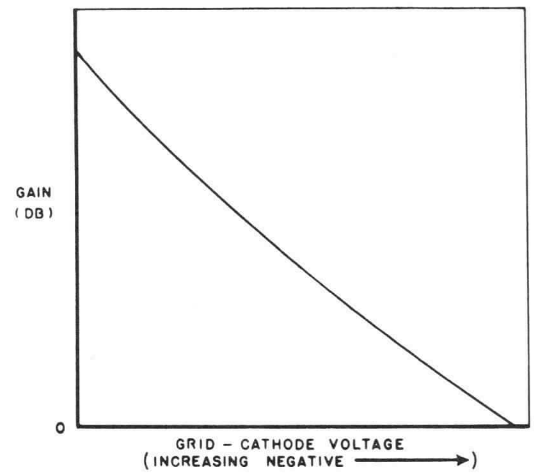

Figure 7. Gain vs Grid Voltage.

of the beam current, and as beam current decreases, gain decreases. Figure 7 shows that the gain of the tube decreases as the grid electrode is made negative with respect to the cathode. This results from the fact that operating the grid negative with respect to the cathode causes the beam current to decrease.

Should the beam current be decreased below a certain value as shown in Figure 6, the gain of the tube measured in db becomes negative, (i.e., the tube is acting as an attenuator). If the beam current is reduced to zero, the net "cold" attenuation of the tube is available. The total change in signal output level as beam current is decreased from its normal value to zero is equal to the sum of the net gain of the tube plus the cold attenuation. This can approach a variation in signal level of about 100 db. Such a change in signal output level is accomplished merely by varying the control grid voltage.

SPECIAL CHARACTERISTICS

Low Noise Amplifiers. A typical low power amplifier may exhibit a noise figure of 20 db. Certain applications require lower noise figures than this. Techniques are available in electron gun design whereby the "shot noise" in the beam can be "de-amplified" before it reaches the beginning of the helix. Special purpose tubes built in this fashion have reached noise figures as low as 6 db.

Narrow Band Amplifiers. Where an application does not require extremely wide bandwidths, tubes can be designed to operate in the dispersive region of the tube characteristic where the velocity of the waves varies as a function of frequency (see Figure 2). In this region, operation at a fixed voltage leads to amplification over only a relatively narrow band of frequencies. Extremely high values of gain (e.g., 60 to 70 db) can be obtained because the problem of controlling spurious oscillations is much simplified since the tube exhibits gain over only a relatively narrow band.

Special Gain Variation. Sometimes special requirements are imposed on the gain variation of the tube as a function of frequency. Tubes can be designed which exhibit an essentially constant value of gain across the band. Others have been built which have linearly increasing gain with frequency. Various other combinations are also possible.

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MICROWAVE MEASUREMENT TECHNIQUES

L. A. Roberts *

As traveling-wave tubes come into more common use, the need becomes apparent for clarification of the various measurement techniques involved in evaluating tube characteristics. In such measurements emphasis must be placed upon broad-band, untuned measuring techniques and test equipment since the traveling-wave tube is a very broadband device. Otherwise, much time would be lost in the repetitive and laborious adjustment and tuning of narrow band devices to make point-by-point measurements across wide frequency ranges.

In this Engineering Note we will concern ourselves with the measurement of small-signal gain, power output, standing-wave ratio, insertion loss, and noise figure. Though automatic and more modern techniques may be available, the methods discussed herein will prove both useful and reliable. Further, these methods require commonly available and relatively inexpensive test equipment.

The types of broadband test equipment commonly used are:

1. Self-tracking tunable signal generator with calibrated power output.
2. Untuned coaxial crystal detector.
3. Untuned coaxial bolometer or thermistor mount with self-balancing power bridge.
4. Broadband coaxial fixed attentuator.
5. Coaxial switch (SPDT).
6. Coaxial slotted section (for VSWR measurements) with untuned detector.
7. Coaxial or waveguide directional coupler.
8. Flourescent noise source.

The emphasis on measuring equipment is placed upon coaxial systems. This is because coaxial test equipment has inherently wideband properties. Waveguide equipment is utilized only when equivalent coaxial equipment is not available or will not do an equivalent job.

SMALL-SIGNAL GAIN MEASUREMENT

A rapid and satisfactory method of measuring gain is to make an insertion gain measurement using the substitution method. This method requires no knowledge of the detector characteristics, nor is absolute power calibration of the signal generator necessary. Accuracy of the measurement depends only on the calibration accuracy of the signal generator output attenuator.

Figure 1 shows a block diagram of the small-signal gain measurement system. With the coaxial switches in position 1, the output of the sig-

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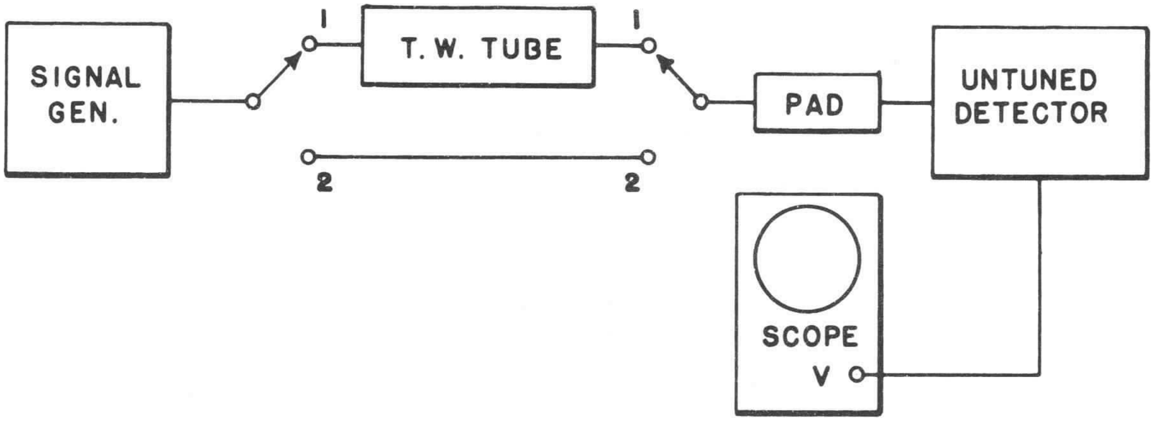

Figure 1. Block Diagram of Gain Measurement System.

nal generator (which has a matched 50 ohm output) is fed to the TWT input through a coaxial single-pole double-throw switch. The output of the TWT is in turn fed through another coaxial switch and a 6 db minimum coaxial fixed attenuator to the untuned detector (If the detector is matched to the coaxial switch, the pad can be eliminated). The detector output is then applied to the vertical deflection channel of an oscilloscope. When the switches are placed in position 2, the tube is by-passed.

Procedure. Apply pulse or square wave modulation to the signal generator. With the tube operating and the switches in position 1, the output attenuator of the signal generator is adjusted to give some convenient deflection on the scope. This is attenuator reading 1. The switches are then placed in position 2 and the attenuator readjusted to give the same deflection on the scope. This is attenuator reading 2. The insertion gain of the tube is then the difference between attenuator readings 1 and 2 measured in db.

Precautions. Both the signal generator and the detector must present good impedance matches as seen by the switch. If they are not matched, reflection losses can give up to several db error in gain. This is the reason for the pad ahead of the unmatched detector, for a 6 db pad guarantees a VSWR of at most 1.5:1.

Also, the switches used must have a low VSWR and very low crosstalk. Crosstalk is stray feedthrough from the input to output terminals of the DPST switch combination through the unused channel. The crosstalk ratio is measured with the TWT removed and is equal to the difference between the signal generator attenuator readings when it is adjusted for equal scope deflection with the switches in position 1 and 2. Regenerative feedback around the switches, which introduces errors in the gain measurement, will be prevented if the crosstalk ratio is at least 20 db greater than the gain of the tube being measured. If the crosstalk ratio exceeds 45 db, it probably cannot be measured with the system in Figure 1. The crosstalk ratio can be measured in such cases with the insertion loss measurement system soon to be described.

Further, care must be taken to insure that the output power level of the tube is at least 10 db below the saturation power level. This eliminates errors due to the decreasing gain characteristic of the traveling-wave tube in the region near saturation.

POWER OUTPUT MEASUREMENT

When making power measurements over wide frequency ranges, it is convenient to use broad-

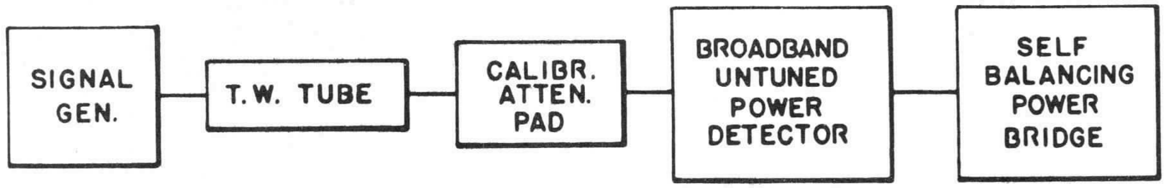

Figure 2. Block Diagram of Power Measuring System.

band (untuned) bolometer or thermistor mounts as power detectors. When these mounts are operated in conjunction with a self-balancing wattmeter bridge, they provide a convenient system for the measurement of absolute power.

If the power level to be measured is greater than the maximum allowable power input to the mount, a calibrated attenuator pad should be inserted ahead of the mount. The most convenient type of pad to use is the resistive film type which has attenuation characteristics that are essentially independent of frequency. Care must be taken not to exceed the maximum input rating of the attenuator.

If power levels to be measured are greater than the rated dissipation of standard attenuator pads, a directional coupler must be placed on the output of the tube. The main arm of the coupler is terminated with a broadband load, and the power detector, with appropriate pads, is placed in the auxiliary arm of the coupler. Broadband coaxial directional couplers with coupling ratios of 10 and 20 db from main arm to auxiliary arm are available.

Should gain measurements be made in conjunction with the power measurements, the power output of the signal generator must be calibrated using its power monitor at each frequency. For accuracy, the power output reading of the signal generator should be initially checked using the bolometer mount and the wattmeter bridge at least over the range of power levels that the bridge can accurately read. Small-signal gain measurements obtained with this equipment should check with the small-signal gain measurements made by the substitution method if the calibration of the attenuator pad and the signal generator is checked.

If all power measurements are recorded in terms of dbm (i.e., db with respect to one milliwatt), the readings of the power bridge, the attenuator (in db) and the signal generator need be merely added and subtracted to obtain power output and gain. Conveniently, most signal generators and power bridges are calibrated in dbm as well as milliwatts.

Figure 2 shows the block diagram of the power measurement system. The power output is given by the equation:

Pout (dbm) = Pad Attenuation (db) + Ratio of Directional Coupler (db) + Power Bridge Reading (dbm)

The gain of the tube is given by the equation:

Gain (db) = Pout (dbm) − Pin from Signal Generator (dbm)

The above gain measurement is valid for both small-signal gain and gain measured in the saturation region and beyond.

STANDING-WAVE RATIO OR REFLECTION COEFFICIENT MEASUREMENT

For direct measurement of voltage standing-wave ratio (VSWR), the coaxial slotted line is the simplest method. It is essential to have an untuned detector on the slotted line, however, in order to speed up broadband measurements.

When it is necessary to examine the VSWR continuously over a wide frequency range, and where fast checks are required such as on production test setups, the slotted line type of measurement becomes prohibitively time consuming and laborious.

For such situations, the measurement of reflection coefficient through the use of directional couplers is recommended. Two directional couplers which are connected so that one gives an output proportional to the incident wave and the other an output proportional to the reflected wave from the load are used. These two outputs can be detected and compared with a ratiometer to give a direct reading of reflection coefficient. The reflection coefficient is then converted to equivalent VSWR if necessary. The greater the accuracy required in the measurement of reflection coefficient, the higher must be the directivity of the directional couplers. In waveguide, directivity greater than 40 db over the waveguide range can be obtained which allows measurement of reflection coefficient of 0.1 with an accuracy of ±.02. This means that a load VSWR of 1.2 will give an indicated reading between 1.15 and 1.25. A waveguide to coaxial transition can be used to adapt the reflectometer to coaxial systems, but residual reflection in the transition limits the accuracy at low VSWRs.

Once the initial calibration is performed in such a reflectometer system, it is only necessary to tune the signal generator across the frequency range and read the reflection coefficient values directly. An excellent detailed discussion of this method, its errors and limitations, is given in the Hewlett-Packard Journal, Volume 6, Number 1-2, Sept. - Oct., 1954.

Even faster measurements can be made when a swept oscillator is employed. Reflection coefficient can be displayed directly on a scope as a function of frequency, and the measurement can be made at a glance.

INSERTION LOSS MEASUREMENT

Measurement of insertion loss in the range of 40 to 100 db (e.g., cold loss measurement in a traveling-wave tube) over a wide range of frequencies is not uncommon. Such a measurement cannot be made with a 1 milliwatt signal generator and a broadband detector, for this type of system does not have the power levels, nor the sensitivity, to measure much more than 45 db insertion loss.

Wideband measurements can be made quickly and easily, however, with the use of one or two traveling-wave tubes as amplifiers. Use of a traveling-wave tube with a noise figure of 20 db ahead of the detector gives a sensitivity into a broadband detector on the order of –65 to –75 dbm. This alone allows insertion loss measurements of 65 to 75 db. Further, if the output of the signal generator is amplified to the power level of one watt (+ 30 dbm) by using a medium power traveling-wave tube, the total insertion loss then measurable becomes 95 to 105 db.

Since traveling-wave tubes are broadband, untuned devices, the only adjustment necessary in this system as frequency is changed across the band is the tuning of the signal generator. The insertion loss measurement is made in the same manner as the insertion gain measurement shown in Figure 1 except for the addition of the amplifiers before and after the switches. Commercially available switches, when connected as shown in Figure 1, have been measured which have a combined crosstalk ratio in excess of 120 db This, of course, eliminates any crosstalk error in insertion loss measurements in the range of 95 to 105 db.

NOISE FIGURE MEASUREMENT

When one is interested in measuring noise

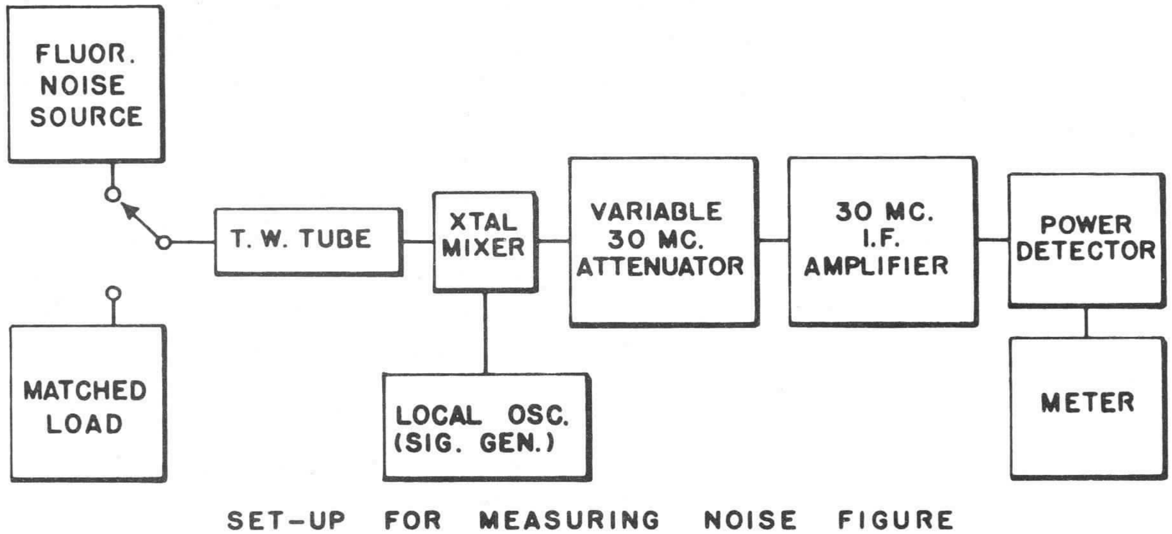

Figure 3. System for Measuring Noise Figure.

figure at a number of frequencies over a wide frequency band, a system which employs a fluorescent tube noise source and a superheterodyne receiver with a square-law second detector (i.e., a power detector) can be used. If noise figure does not vary too rapidly with frequency, a simple receiver can be used which employs an untuned crystal mixer, a local oscillator, a 30 mc IF strip, and a power detector. Note that no preselection or mixer tuning is used.

Without a preselector, the receiver will respond equally well to both the signal and image frequencies. With a 30 mc IF amplifier, these two frequencies will be only 60 mc apart. The measured noise figure will be an average of the true values at two frequencies 60 mc apart providing noise figure is not a rapid function of frequency. This is accurate enough for most purposes. If noise figure varies so rapidly that it will be appreciably different at two frequencies spaced 60 mc apart, then preselection is necessary.

The noise figure of an amplifier is defined as the ratio of the signal-to-noise ratio at the input of the amplifier to the signal-to-noise ratio at the output of the amplifier, That is,

$$F = \frac{(S/N)in}{(S/N)out} \qquad \text{(power ratio)}$$

When the measurement is made according to the method to be outlined, and a noise source whose level is 15.8 db above thermal noise is used, the above equation becomes:

$$F = \frac{37k}{\frac{P_1}{P_2} - 1} \qquad \text{(power ratio)} \qquad (1)$$

where, P_1 and P_2 are as defined in the measurement procedure below, and k, a number smaller than 1.0, is the cable loss in db between the noise source and the tube converted to a power ratio. (e.g., for cable loss = 1db, k = .794)

When a fluorescent noise source is used, it is not necessary to know the effective RF noise bandwidth of the receiver.

Procedure: The tube to be measured is connected into the system shown in Figure 3. The power detector in the receiver may take the form of a bolometer or a thermistor operating into a self-balancing wattmeter bridge, or it may be a thermocouple whose output is connected to a microammeter.

With the TWT operating and the local oscillator tuned to the desired frequency, switch the tube input to the noise source. Adjust the variable 30 mc attenuator until the output meter indicates approximately full scale. This is reading P_1.

Then switch the tube input to the matched load. The output meter now indicates P_2. Insert the values of P_1 and P_2 in equation 1 and compute the noise figure. To determine the noise figure in db, merely convert this number to db.

Precautions: The accuracy of this measurement is based on the assumption that the receiver is linear. This will be true if the local oscillator signal is large enough to produce linear mixing and if the IF amplifier is operated in the linear portion of its characteristics.

The measurements should be taken quickly enough so that power meter drift will not effect readings.

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DOMESTICATING THE TRAVELING-WAVE TUBE*

Peter D. Lacy **

INTRODUCTION

In a report[1] on the status of traveling wave tubes at the Symposium on Modern Advances in Microwave Techniques last November, Dr. Watkins of Stanford characterized traveling wave tubes as "the most advertised, least delivered tubes in electronics history". A major reason for this state of affairs has been their failure to find wide application in the electronics industry. The unique characteristic of TWT's, wide bandwidth, poses a difficult question: How can it be used? This requires careful study of the types of systems in which it will be useful and intensive development of new techniques for handling such bandwidths. Some possible directions that broadband systems may follow will be indicated later.

In this discussion of TWT's, no tubes that are new or that have improved characteristics will be considered since the new advances have been well covered elsewhere[2]. Rather, it is proposed to consider only the most ordinary of TWT's that are advertised and in commercial production now. The gain, relative bandwidth, power output and noise figure of these tubes do not differ substantially from the first tube described by Pierce and Field[3] in 1946. These are tubes with about 30 db of gain, octave frequency coverage through X band, noise figures of 20 to 30 db, and output power of 10 milliwatts to 1 watt.

This power range is well suited to the vital functions of modulation and detection so that microwave circuits may be linked with information handling video circuits. It is suitable for stable oscillator circuits comprised of either microwave oscillators or low frequency oscillators followed by a frequency multiplier chain. The level is quite suitable for most microwave measurements and some new measurement techniques become possible due to the characteristics of the TWT. The upper power limit is adequate for a large portion of the fixed path propagation links in microwave systems. After all, this is about the same power range in which most electronic tubes operate!

As to the broad frequency range of the TWT, one can either take it or leave it. If the application requires a broad bandwidth, then the TWT is without peer. On the other hand, if the application requires only a narrow bandwidth, the TWT may still be used at any point in the wide amplification band provided by the TWT. In such cases, it is often advisable to filter the output of the tube to reduce noise. The narrow band filter is in a passive transmission line for the TWT, as contrasted with the tuned circuits of klystron or space charge control tubes that are in contact with the tube electrodes. In the latter case, the amplification is quite sensitive to the tuned circuit parameters so the design and adjustments become critical.

* Presented at Seventh Region IRE, Technical Conference, Phoenix, Arizona, April 28, 1955

** Hewlett-Packard Company, Palo Alto, California

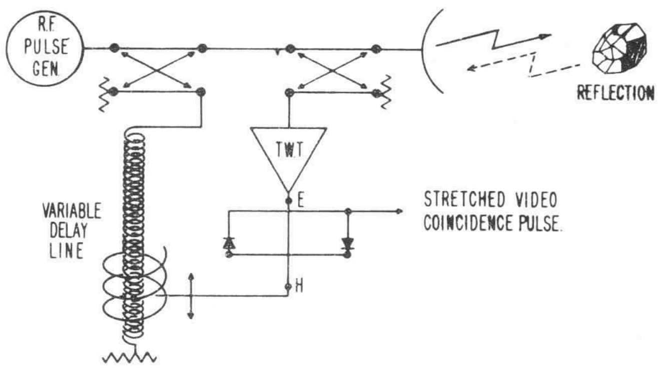


Figure 1. Precision Reflection Location.

NEW FRONTIERS

One of the new frontiers that may be opened up by the bandwidth of the TWT is the transformation of video techniques and functions to the microwave domain. This would mean increasing the effective bandwidth from the tens to hundreds of megacycles now available to thousands of megacycles. This practice would increase by one to two orders of magnitude the resolution or speed of basic physical measurements, electronic measurement, communication, and computation.

The first obstacle preventing such a transfer of video systems to microwaves is the lack of terminal devices, that is, how do we enter such a high speed elec-system and then retrieve the new or processes information? Next, the internal functions of the microwave system, which determine the breadth of application, will demand intensive development. Such problems will require considerable effort before very high speed microwave systems can be realized.

A.C. Beck[4] has recently demonstrated a millimicrosecond pulse system for locating waveguide faults. Another example of a very high resolution pulse echo system that could be achieved with present techniques where the resolution is determined mainly by the center rf frequency will be given. In this system a fast pulse and low phase distortion are also needed. Figure 1 is a block diagram of this type of system. The rf pulse generator, in which the rf carrier is phase locked to the repetition rate, may be of the regenerative type like Cutler's[5] if the phase precession can be eliminated or a beam deflection tube pulser[6] with a frequency multiplier and amplifier driven by the pulse repetition rate frequency and a modualtor. The delayed rf pulse from the generator and the return pulse are fed into a microwave coincidence circuit which consists of a hybrid tee and a pair of crystal detectors with balanced reversed crystals. The microwave coincidence circuit shown in Figure 1 operates as follows: for an input in the E arm, the two crystals conduct during the same half cycle with equal outputs and opposite polarity. For an input in the H arm, one crystal conducts one half cycle and the other crystal in the second half cycle. Again over one complete cycle the average output is zero. For simultaneous inputs to the E and H arms and proper rf phases, the non-linearity of the crystal detectors cause a net output. When the crystals operate square law, the coincidence circuit shown is a cross-correlation detector. Figure 2 shows the resulting waveforms with the rf carrier phase stationary in the pulse. The accuracy of

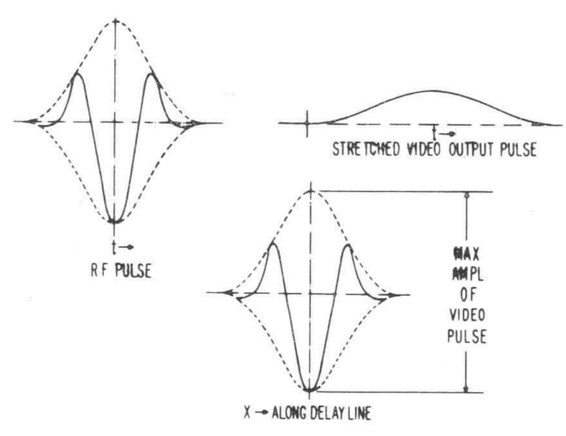


Figure 2. Precision Reflection Location Waveforms.

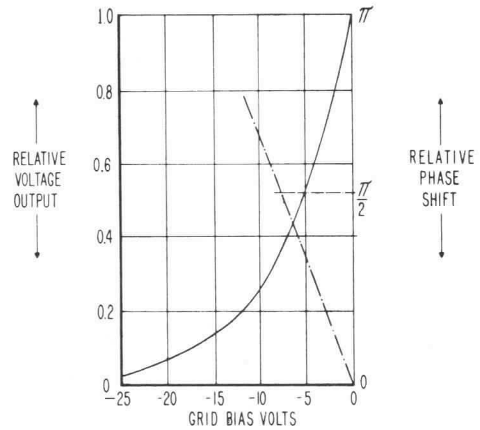


Figure 3. Modulation Characteristics.

locating a minimum near the center of the pulse should be a fraction of an rf cycle, say 10 degrees, so for a carrier of 3,000 mc a spatial resolution of about 1/10 inch could be expected. This is only an elementary example of a high performance system utilizing the bandwidths available from TWT's. This system could also be used in particle emission coincidence studies with an accuracy determined there by the pulse width.

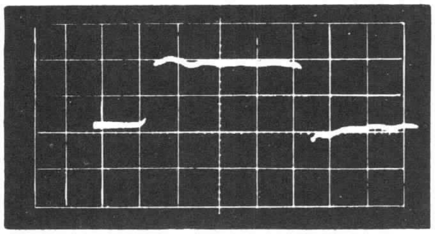
It would be difficult to estimate the ultimate extent of TWT usage in the expanding electronics field. However, it does appear to be a fruitful field for exploration and invention. With the increasing dependence on electronics in industry and science today, certainly there shall develop a demand for greatly increased speed in electronic detection, processing, and control and the TWT is orders of magnitude ahead of any other electronic device.

MODULATION CHARACTERISTICS

Now turning from the broad bandwidth feature, let us consider some more down to earth characteristics. A number of functions may be performed by modulating the TWT electrodes. The two variables are the beam current and beam velocity. The beam current may be varied by means of the potentials applied to one of the electrodes of the electron gun. A typical variation of the rf output voltage and its phase with electrode voltage is shown in Figure 3. The tube output can be amplitude modulated over the linear portion (about 10 db) but the attendant phase modulation is about 90 degrees. The amplitude modulation still can be useful where amplitude detection only is involved in demodulation.

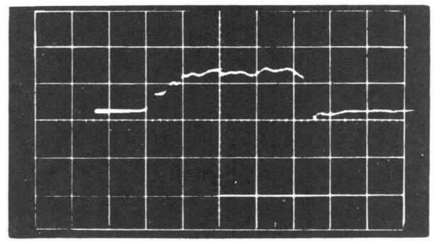
The effect of pulse modulation is shown in Figure 4 for various conditions of tube drive and pulse off-on ratio. In the top oscillogram, the tube is driven to saturation with an off-on ratio of 40 db. The modulating pulse had a rise time of about one millimicrosecond and an amplitude of 40 volts. The resulting rise time is 4 millimicroseconds which was close to the oscilloscope amplifier rise time. In the middle picture, the off-on ratio is still 40 db but the tube is operating well below saturation. It seems that the entrance of the pulsed beam into the helix induces a voltage on the helix that in turn changes the beam velocity. When the electron beam wave velocity is shifted away from the helix wave velocity the TWT amplification decreases. Thus a transient period of reduced amplification occurs at the start of the pulse slowing the rise time to about 20 millimicroseconds. In the lower picture the off-on ratio has been reduced to 18 db by reducing the bias and pulse amplitude to about 15 volts so with this lower current ratio the rise time is good again. It can be seen from these pulse oscillograms that under proper conditions, the TWT may be used for generating low level pulses for testing the receiver portions of systems having fast rise time pulses in the transmitter.

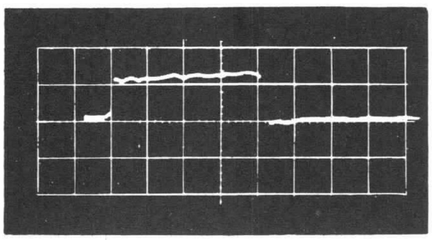
In Figure 5, an amplitude modulated pulse train is shown. This was generated by applying a pulse and sine wave to the TWT grid. By this method, lobing or incidental flutter may be simulated in system testing.



← (a)
PEAK T.W.T. OUTPUT.
40 DB OFF-ON RATIO.

(b) →
LOW T.W.T. OUTPUT.
40 DB OFF-ON RATIO.





← (c)
LOW T.W.T. OUTPUT
18 DB OFF-ON RATIO.

Figure 4. 10 MW, 2-4 KMC TWT Detected Pulse Oscillograms (Sweep 20 Mu Sec Div).

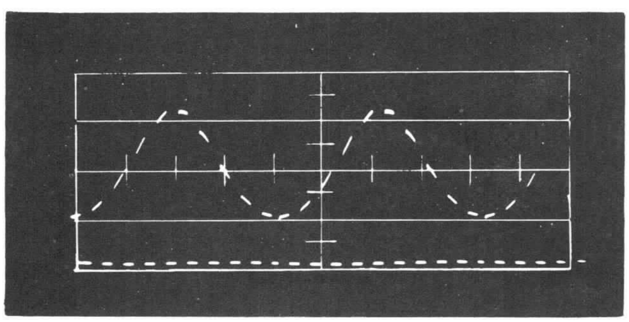


Figure 5. Amplitude Modulated Pulse Train.

Figure 6 shows the effect of varying the helix or beam voltage on the phase of the output of a TWT amplifier. For constant beam current (solid curve) the phase change versus helix voltage is nearly linear. However, the output levels vary, being maximum at some optimum helix voltage and diminishing on either side. The amplitude may be held constant by an amplitude stabilization signal that is fed back to the grid of the electron gun. The resulting phase curve

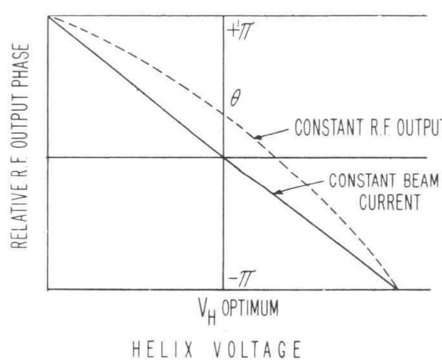


Figure 6. TWT Phase Characteristic.

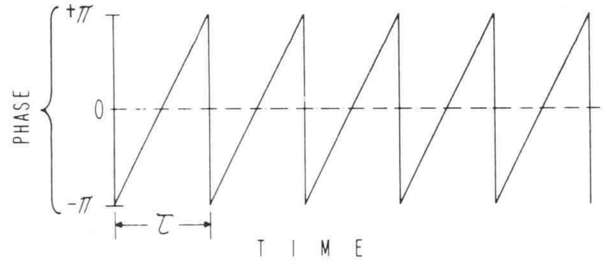


Figure 7. Saw Tooth Phase Modulation.

(dotted) is shown. It is not linear but the advantages of eliminating amplitude variations may outweigh the effect of this phase characteristic distortion.

The amplitude and phase characteristics of TWT's have been presented and the interaction of the two effects due to grid or helix voltage variation. The amplitude modulation characteristics shown are similar to those found in any other amplifier tube, however, the large amount of phase modulation possible (over 360 degrees) can produce some interesting results.

SAWTOOTH PHASE MODULATION

Mr. Ray Cummings[7] of Stanford University has devised a method of unlimited deviation by its stepwise discontinuous equivalent. Consider a uniform phase change, constantly increasing, for instance, it may be approximated as shown in Figure 7 by increasing the phase constantly over a full rf cycle of 2π radians of phase and then quickly jumping back to the starting phase and then commencing to advance again at the previous constant rate. This is a stepwise discontinuous approximation to the continuous phase advance associated with the doppler shift of approaching source. Since one cycle of rf has been added during a period τ, the shift in frequency is just $1/\tau$.

Figure 8 shows oscillograms of the helix

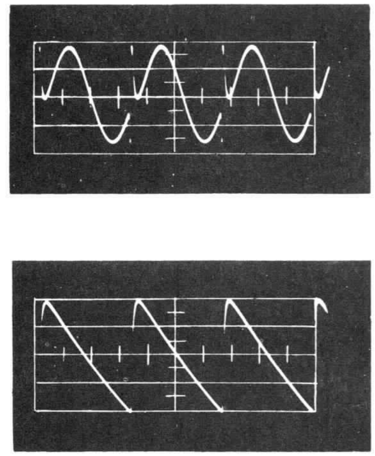


Figure 8. Single Sideband Modulation.

modulation voltage and the mixed product of the original signal and the frequency shifted output signal. Note the switching transient due to the finite flyback time of the sawtooth wave. This flyback time can be reduced considerably and thereby correspondingly reduce the error in doppler simulation. This system becomes a very accurate 2 terminal pair device for simulating doppler shifts from a few cycles per second to about a hundred kilocycles. This should provide a very satisfactory instrument for the design and test of cw doppler and coherent pulse radars.

LINEAR DETECTION IN MICROWAVE MEASUREMENTS

Another powerful application of frequency offset or single side band modulation is the use of linear or homodyne detection which greatly extended the dynamic range of microwave measurements. Figure 9 shows a homodyne measurement system. The TWT provides a frequency offset f_1. This shifted frequency is then applied to the system under test that yields a weak output signal. The weak signal and the strong reference signal or local oscillator are then applied to a crystal mixer. The mixer is operated linearly and the beat frequency f_1, kc possibly,

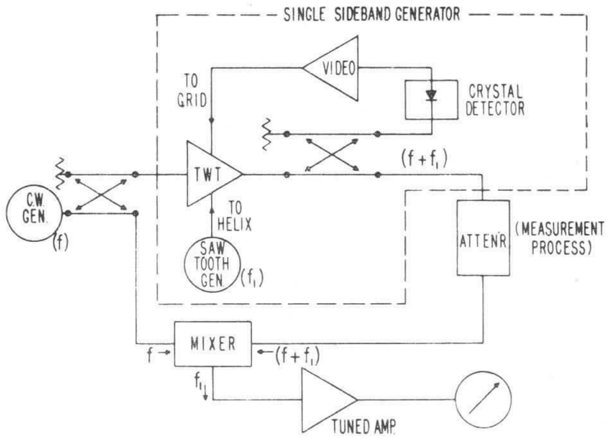


Figure 9. Linear (Homodyne) Detection System For Microwave Measurement.

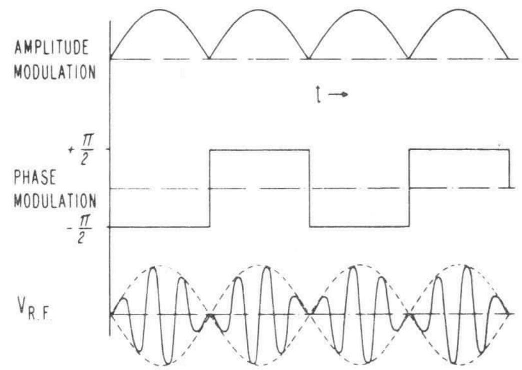


Figure 10. Suppressed Carrier Modulation.

is then applied to a tuned amplifier and meter such as a VSWR detector. Linear mixer output ranges approaching 100 db may be attained compared with an equivalent 50 db range for a square law detector. Thus the sensitivity and dynamic range for measurements is increased by a power ratio of about 10 billion.

The advantages of using the TWT for this frequency shifting function are:

1) low modulated signal out of the input terminals of the TWT. This limits the sensitivity attainable.

2) wide frequency coverage. The helix coupled TWT is useful even beyond the recommended 2:1 bandwidth.

3) gain is provided in the weak signal channel.

Of the possible alternatives, rotating mechanical or ferrite phase shifters and other types of tubes, no one device has all three of the listed advantages.

GENERAL APPLICATIONS

A still further type of modulation available with the TWT is suppressed carrier modulation. Figure 10 shows the required modulations for the control grid and helix and the resulting suppressed carrier rf output. In comparison with a magic tee and crystal balanced modulator, in this one all adjustments are voltages rather than mechanical positions and the TWT modulation characteristics may offer a greater degree of stability than balanced microwave crystals.

For narrowband work, the noise level due to the immense bandwidth of the TWT is often objectionable. The residual noise level and the effective dynamic range (noise to saturation level) may be greatly improved by inserting a band pass filter after the TWT. Under this condition the TWT should then be competitive with either a klystron or triode amplifier as far as dynamic range and gain are concerned, but since a few different TWT types can be adapted to this use anywhere in the microwave range up to 12 KMC, just three or four different tube types could do the job and require only the addition of a single bandpass filter in the output transmission line. The TWT does have greater time delay than its competitors and often the phase distortion has been excessive. The phase distortion is being constantly improved and with the coupled helix circuit on the outside of a tube with excellent internal phase characteristics, the phase distortion can be reduced even further with care in constructing the external microwave circuit.

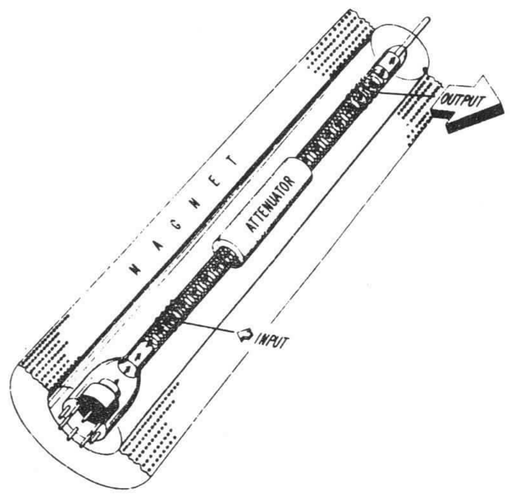


Figure 11. Pictorial View of a Traveling-Wave Tube.

TWT's have been proven excellent as highly stable microwave oscillators. Hetland and Buss[8] of Stanford University have computed noise bandwidths of as low as 10^{-3} cps at 3 KMC. As compared with a stabilized reflex klystron, in which a tube with a relatively low Q oscillator cavity has its mean frequency corrected by a high Q reference cavity, the TWT uses the high Q cavity directly in the feedback circuit of the oscillator to control the instantaneous frequency.

Many other applications of TWT's are possible, since the fundamental component of any electronic system is the signal amplifier. In the role of amplifier, the TWT is rapidly opening up wide regions of the microwave spectrum to a much more flexible approach to system synthesis.

COUPLED HELIX CIRCUITS

Another more speculative role for the TWT lies in the use of helically coupled circuits on the outside of the vacuum envelope. Figure 11 shows a tube with its amplifier circuit in place. The role of the vacuum tube is reduced to an active transmission line in one direction and a passive line in the other direction. Input and output couplers may be arranged along the tube at will, attenuating sections are applied to eliminate any return signal from output to input in the passive backward direction. These are the usual amplifier functions connected with making a stable amplifier tube. Variation of the tube length may be used to adjust the amplifier delay time which might open up the use of the tube for fast switching functions. External reactive and resistive circuit elements can change the frequency response as well as the non-linear characteristics of the active line. Some work has started along these lines; however, there is no estimate yet as to how far or how flexible these external circuitry methods are.

The immediate advantage of external circuitry has been that the internal vacuum tube has been reduced to its simplest form. An electron gun, a uniform helix, and a beam collector. The critical circuit functions of coupling in and out of the tube and stabilizing the amplifier are now made separate and may be adjusted. This can mean a substantial increase in production yield.

The precision machine work required on the capsule has been reduced to a minimum. The coupling helices and the mounting hole require precision while the rest of the manufacturing operations can be simple fabrication methods like sawcuts, stamping, and rolling. Thus with the versatility afforded by the wide bandwidth which also reduces the number of tube types, in conjunction with high yeild manufacturing methods, sufficient demand may be expected to result in low cost-flexible microwave amplification by the TWT.

CONCLUSIONS

The traveling wave tube and other related distributed circuit-electron beam interaction tubes have been the object of intensive research and invention for nearly ten years; however, only in the past year have tubes of a general purpose nature become commercially available. Now it becomes necessary to critically examine the range of usefulness of the TWT. The use of the broadband property will require another research phase before it can be judged and given suitable employment. A broader less critical role can be filled by just a few types of tubes providing narrow band amplification and useful modulation characteristics through most of the microwave range.

It is hoped the modulation characteristics and possible applications presented here may suggest some useful tasks for the TWT.

ACKNOWLEDGEMENTS

The author is indebted to Mr. Ray Cummings of Stanford University for the method of generating a frequency off-set with the TWT and to D. E. Wheeler, G. W. Mathers, and H. C. Poulter of the Hewlett-Packard Company for many contributions and discussions during the preparation of this paper.

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AN X-BAND BACKWARD-WAVE OSCILLATOR*

W. V. Christensen **

Backward wave oscillators can be designed with a number of different rf interaction structures. Some examples include loaded waveguide, interdigital lines, folded lines, helices, and variations of all of these. The one feature that all of these structures have in common is a spatial periodicity along the tube in the direction in which the electron beam flows. The particular choice of circuit may depend on such factors as power output required, ease of construction, shape of tuning curve required, tuning bandwidth, and frequency of operation.

The BWO under present discussion uses a single-filar helix and a glass envelope. This type of construction was selected for several reasons: an X-band helix has reasonable dimensions, techniques for manufacturing all glass helix type traveling-wave tubes were familiar, and a helix type tube leads to a relatively simple tube and package.

PERFORMANCE REQUIREMENTS

The single-filar helix type tube is capable of operating over a 2:1 frequency range. This BWO was designed to operate over the 7.0 to 14.0 kmc frequency range so that the most uniform region of power output would be in X-band (8.2 to 12.4 kmc). The design goal for power output was set nominally at 100 mw across X-band, and it was hoped that the power would not drop too drastically outside of X-band.

PHYSICAL REQUIREMENTS

To cover a 2:1 frequency range, rf power is necessarily coupled out of the tube with a coaxial system. Fortunately, the characteristic impedance of the helix is near 50 ohms, and the coaxial line can be connected directly to the helix with a VSWR of less than 2:1. With a coaxial rf output, the cable can be placed parallel to the axis of the tube and brought out the end of the capsule. This requires that the capsule diameter be only large enough to fit around the gun bulb which is the largest diameter in the tube. In this case the outside capsule diameter is held to 1.0 inch.

The magnetic field required to focus such a tube is approximately 1000 gauss. To supply this value of field it is necessary to build an air-cooled solenoid. The solenoid can be built with open windings in such a way that forced air comes in contact with each layer. This provides efficient solenoid cooling. Air from the same

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** Formerly of Huggins Laboratories, Inc.

blower which cools the solenoid is also used to cool the tube by passing air over the outside of the capsule. This cools the tube sufficiently since the tube elements which need cooling are placed in good thermal contact with the capsule itself.

CONSTRUCTION TECHNIQUES

The fields of the backward wave mode on a helix are associated very closely to the helix wires. Since these fields fall off very rapidly in a radial direction away from the helix, all of the interacting electrons in the beam must pass very close to the helix. This results in the requirement for a hollow beam of very tight dimensional tolerances whose thickness and spacing from the helix is only a few thousandths of an inch. To enhance the backward wave mode, the helix is wound from tape whose width is about four times its thickness. The dielectric loading of the helix by the surrounding glass envelope is kept low by supporting the helix in a fluted glass envelope. Thus, glass touches the helix only along three lines of contact.

The electron beam is formed in a hollow beam gun which has a cathode shaped identically to that of the beam cross section. The gun is a parallel-flow Pierce type, and the cathode is a Philips impregnated cathode which can easily support the required current density of 0.6 ampere/cm^2.

PHYSICAL DESCRIPTION

The tube, which is pictured in Figure 1, is enclosed in a capsule one inch in diameter and eleven inches long. A flexible cable with the dc leads is brought out of the gun end of the capsule. The rf output cable is a flexible coax line with teflon dielectric, and it is terminated either in a special type N coaxial connector which has low residual VSWR up to 14.0 kmc or

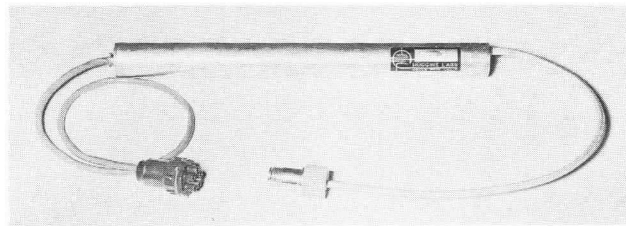

Figure 1. Backward-Wave Oscillator.

in an X-band waveguide adapter if only X-band operation is required.

The BWO fits into a solenoid which is 4¼ x 4¼ inches in cross section. The blower which cools both the tube and solenoid is mounted at one end of the solenoid. The overall length of the unit, including solenoid and blower, is 16½ inches.

TUBE PERFORMANCE

Tuning Curve. Figure 2 shows a curve of frequency vs helix voltage. When plotted on semi-log coordinates, the curve becomes a straight line. This is typical of most backward wave oscillator structures. To cover the full 7.0 to 14.0 kmc frequency range, the helix voltage must vary from 300 to 3300 volts. For X-band operation the maximum helix voltage required is considerably reduced; X-band is cov-

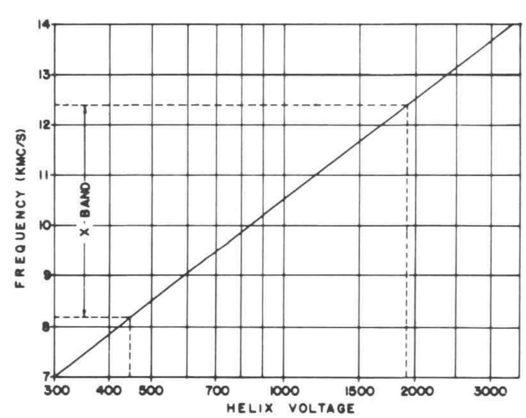

Figure 2. Tuning Curve.

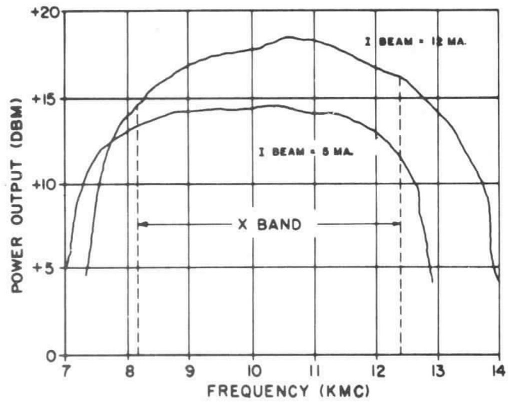

Figure 3. Power Output Curve.

ered by the voltage range of 450 to 1900 volts.

Power Output. The curve of power output vs frequency is shown in Figure 3. Across X-band the power output is greater than 14 dbm, and the minimum power over the 7.0 to 14.0 kmc band is 4.0 dbm.

Spurious Responses. With a load having a VSWR of 3:1 or less, no spurious responses were observed 90 db below signal level. With a load having a VSWR greater than 3:1, frequency discontinuities were noted across the band and spurious frequencies suddenly appeared in the neighborhood of 12.9 kmc. Thus, for extreme VSWR variations of the load, a 3 db pad in series with the output cable will insure satisfactory operation—even for open or short circuit loads.

POWER SUPPLY REQUIREMENTS

Power Requirements for the Tube. Figure 4 shows a schematic diagram of the tube, power supplies, and metering required. The anode supply must be variable and deliver between 300 and 450 volts at 1.0 ma. Both the anode supply and the 7.5 volt ac heater supply must be insulated from ground for the full helix to cathode voltage. The helix and collector electrodes are operated at ground potential, and tuning is accomplished by operating the cathode negative with respect to the helix. The helix and collector supply must be variable from 300 to 3300 volts and capable of delivering 13 ma.

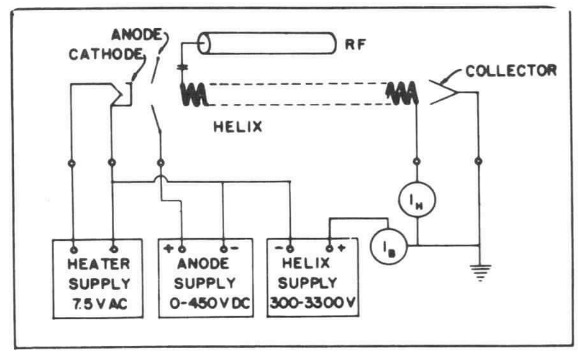

Figure 4. BWO Power Supply.

The tube requires a 1000 gauss magnetic field to properly focus the beam over the 7.0 to 14.0 kmc frequency range. If only X-band operation is required, the magnetic field need be only 750 gauss. For the 1000 gauss field, the solenoid requires 90 volts at 4.1 amps. The blower requires 28 volts at 2.5 amps.

Power Supply Stability and Ripple Requirements. Regulation and filtering of the various power supplies is dependent upon the frequency stability required since all of the voltages affect

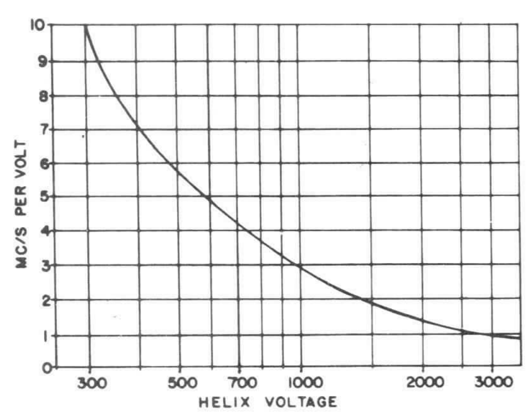

Figure 5. Tuning Curve Slope.

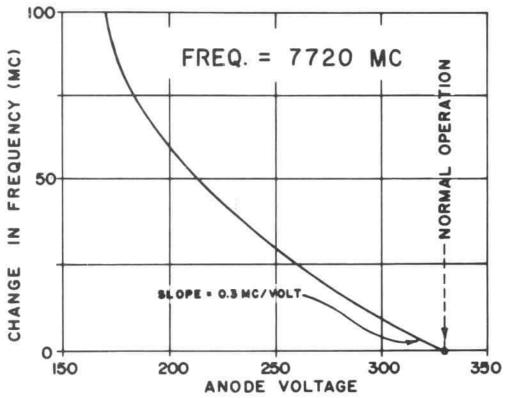

Figure 6. Anode Voltage Frequency Shift.

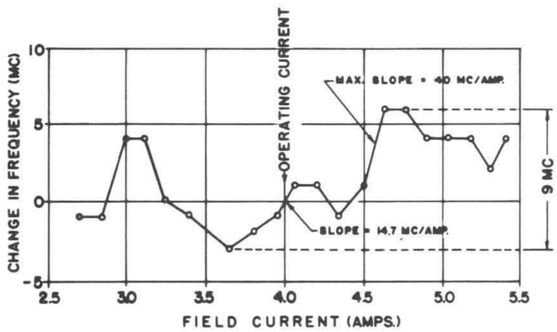

Figure 7. Field Current Frequency Shift.

the frequency in some way. Figure 5 shows the slope of the tuning curve as a function of helix voltage. The large number of mc/volt gives an idea of the stability of the helix voltage required to maintain a fixed frequency.

For example, at 2000 volts the slope is 1.4 mc/volt. A peak-to-peak ripple of 0.1% in helix voltage would result in a variation in frequency of ±1.4 mc. A ripple of 0.1% of any helix voltage in the operating range will result in the same frequency deviation of ±1.4 mc because of the logarithmic variation in the slope.

The anode voltage controls the beam current which has a small effect on frequency. Figure 6 shows the relationship between anode voltage and this frequency shift. As the anode voltage is increased, the frequency decreases. The slope of this increase is 0.3 mc/volt at the normal operating current of 12 ma. A peak-to-peak ripple of 1% on the anode voltage will result in a frequency variation of ±500 kc.

The magnetic flux density of the solenoid affects the shape of the electron beam as well as the amount of intercepted current on the various electrodes. Both of these factors affect frequency. Thus, frequency is somewhat a function of the current through the solenoid.

Figure 7 shows shift in frequency vs solenoid current. The slope of this curve at the operating point is 14.7 mc/amp. A peak-to-peak ripple current in the solenoid of 1% would lead to a frequency variation of ±300 kc. The maximum deviation observed was 9.0 mc, and the maximum slope observed was 40.0 mc/amp. The fine details of this curve will probably vary from tube to tube.

FUTURE TRENDS

The future trend in backward-wave oscillator design will be toward permanent magnet focusing and lower operating voltages. The lower operating voltages will result in a greater slope in the tuning curve and thus more mc/volt. Permanent magnet focusing will eliminate the need for the solenoid, its blower, and the associated power supplies.

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AN X-BAND TRAVELING-WAVE AMPLIFIER*

L. A. Roberts **

As wideband systems and broadband measurement techniques have been developed, the traveling-wave tube has proved to be an extremely useful component. In the case of the forward-wave amplifier, its broadband, untuned amplification characteristics and fast modulation properties make many systems and techniques possible which were previously difficult or even impossible to accomplish.

One of the most widely used microwave frequency ranges is X-band (8.2 to 12.4 kmc). Until the announcement of the tube to be described in this paper, there has been no commercially available general purpose traveling-wave tube amplifier in this frequency range. The tube described herein provides useful operation over an octave frequency range which brackets X-band.

PERFORMANCE REQUIREMENTS

In designing this tube, use of as much of the inherent broadband properties of the traveling-wave tube as possible was desired. Experience in lower frequency tubes has shown that they exhibit their most uniform characteristics over a 2:1 frequency range, even though they can provide useful gain well outside an octave range. The design goal was then set to cover a 2:1 frequency range which centered the region of best operation of the tube in X-band.

The design goals of the tube were:

Frequency Range:	7.0 to 14.0 kmc
Gain:	30 db minimum at a fixed helix voltage
Power Output:	10 milliwatts minimum
Magnetic Field:	300 to 400 gauss
Stability:	No oscillations with these conditions applied simultaneously: 1) total reflections of any phase connected simultaneously to the input and the output, 2) with the helix at any voltage within 200 volts of synchronism, and 3) with the beam current greater than the normal operating value.

Further, the initial adjustments to set the tube into operation had to be simple. There should be no mechanical adjustments to the tube other than a simple alignment in the magnetic field to optimize electron beam transmission through the tube.

PHYSICAL REQUIREMENTS

It was preferable to have a small convection-cooled focusing solenoid which required no external fans or blowers to cool the solenoid or tube. To accomplish this, the inside hole diameter of the solenoid was kept as small as possible to hold solenoid power dissipation and temperature rise to a minimum. The solenoid inside diameter is determined by the diameter of the capsule in which the tube is mounted. The capsule size is in turn determined by the size of the gun bulb. In this case the capsule is one inch in diameter.

With this restriction on capsule diameter, the input and output transmission lines are of necessity coaxial lines. Ease of construction is facilitated by using flexible coaxial cable. The cable connectors required are Type N. When the tube is to be operated directly into X-band waveguide, direct adapters from the cable to the waveguide can be used in place of Type N connectors.

CONSTRUCTION TECHNIQUES

General. The helix is supported in a tight fitting thin-wall glass envelope which also serves as the vacuum envelope of the tube. RF energy is introduced onto and off the helix by means of helical directional couplers which are external to the vacuum envelope.

This type of coupling has become known as the coupled helix match. Such a coupler is a very wide-band device, and almost complete transfer of energy from the coupler to the helix can be accomplished over a 2:1 frequency range. The use of the coupled helix match leads to extremely simple construction of the tube. Within the vacuum envelope there is nothing except the electron gun, a uniform helix, and the collector electrode.

The Coupled Helix Match. The coupled helix match is a co-directional coupler. This means that the power flow in the coupler and the helix is in the same direction. In order to accomplish this, the coupling helix must be wound in the opposite sense to the inner or amplifying helix of the tube. Further, the phase velocity of the inner and outer helix must be matched as closely as possible, and the correct length of the coupling helix must be chosen to give the best power transfer characteristic across the band.

The input and output coupling helices are mounted within the outer conductors of the coaxial cables, and the spacing and dielectric loading between the helix and the outer conductor is arranged so that the helix has a 50 ohm impedance. The couplers are thus connected directly to the 50 ohm coaxial cables. This technique provides low VSWR transition if care is taken to keep the resulting discontinuities small. The VSWR measured in the coxial line due to the transition to the coupled helix and its coupling to the tube is nominally no greater than 1.7:1. With care the VSWR can be made 1.4:1 or less.

Attenuator. The center attenuation on the tube, which keeps it from oscillating, is also introduced by a coupled helix technique. The attenuating helix, which is also external to the vacuum envelope, is actually a lossy directional coupler that dissipates the energy which is coupled into it.

The coupled helix attenuator has several advantages: 1) The acutal application of the attenuation to the tube does not take place until construction and pumping are complete and the tube is ready for test. This means that the attenuator can be checked out on an operating tube. If the attenuator is not exactly correct, the change can be made simply, and the tube

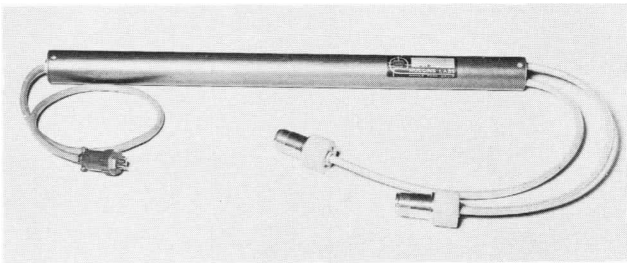

Figure 1. Encapsulated HA-4 Traveling-Wave Tube.

does not have to be torn apart to make the change as is the case where the loss is within the vacuum envelope; 2) coupled helix attenuation is an extremely wideband scheme that can couple in loss which is sufficient to prevent oscillation over the entire range where the tube exhibits any net gain; and 3) the reflection coefficient seen from the helix looking into the attenuator section is extremely low. The latter makes short circuit stability possible and keeps gain fluctuations due to regenerative feedback at a very low level.

The Electron Gun. A parallel-flow Pierce type electron gun is used. The current density required in the beam is 0.5 ampere per square

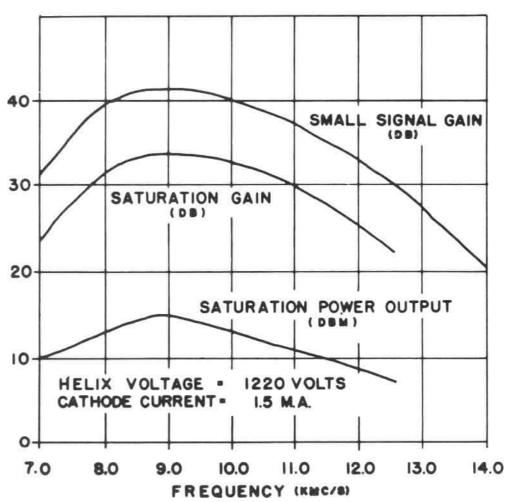

Figure 2. Power Output and Gain vs Frequency.

centimeter. This value of current density is easily obtained with a Philips impregnated cathode, and this type of cathode has the further advantage of being reliable and easy to process.

General Construction Philosophy. In summary, the construction philosophy behind the subject TWT was: 1) A minimum number of parts should be internal to the vacuum envelope in order to simplify tube construction, and 2) all coupling and loss should be accomplished external to the vacuum envelope so that adjustments and changes could be made on an operating tube.

PHYSICAL DESCRIPTION

Figure 1 is a photograph of the encapsulated tube which is one inch in diameter and 13 inches long. The dc leads enter the capsule through the gun end in a flexible cable which is fitted with a high voltage connector. Double braided flexible coax cables with teflon dielectric (RG-142/U) are used for the rf cables, and the rf connectors are special Type N female which introduce only a small reflection up to a frequency of 14.0 kmc. The encapsulated tube fits into a solenoid whose outside dimensions are 3-3/8" x 3-3/8" x 12". This solenoid provides 400 gauss with 60 watts dissipation.

The tube is operated with the collector at ground, and the collector is connected to the capsule. The cathode and anode are operated negative with respect to ground. Thus, there are no exposed electrodes or leads which have a potential with respect to ground. This type of operation prevents any shock hazard to operating personnel.

TUBE PERFORMANCE

Gain and Power Output. Figure 2 shows typical curves of small-signal gain, saturation gain, and saturation power output as a function

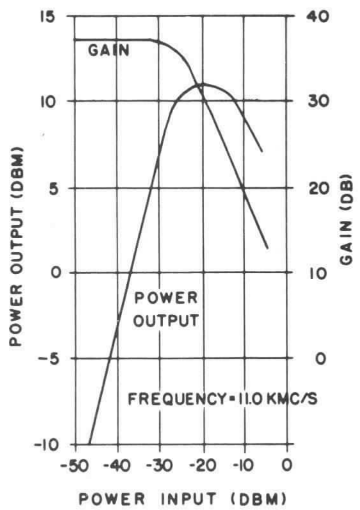

Figure 3. Power Output and Gain vs Power Input.

of frequency. These curves were taken at a fixed helix voltage. The gain meets the design goal of 30 db over the frequency range of 7.0 to 12.5 kmc, and the gain is better than 20 db out to 14.0 kmc. Over the 7.0 to 11.0 kmc range, the power output meets the design goal of 10 dbm, and at 12.5 kmc it drops to 7.5 dbm. Power output beyond 12.5 kmc was not measured due to lack of suitable test equipment.

Power Output vs Power Input. Figure 3 shows curves of power output and gain as a function of power input at a fixed frequency. The output curve exhibits the normal traveling-wave tube characteristic, i.e., the power output is a linear function of power input until the output level is approximately 6 db below the saturation value. Beyond this point the curve levels off and then decreases as the power input increases. The gain curve shows that the small-signal gain remains constant as a function of power input until the power output curve starts to drop. At the power input level corresponding to maximum power output, the gain of the tube has dropped 6 to 7 db from the small-signal value. If the tube is driven beyond the maximum power output point, the gain continues to decrease.

Grid Characteristics. The electron gun contains a non-intercepting grid electrode which has high control action on the gain and power output of the tube. Curves of power output vs control grid voltage are shown in Figure 4. It is seen that the gain and power output of the tube can be varied over an extremely wide range with the use of the grid; thus, the tube can be used as an electronically variable attenuator. The maximum attenuation obtainable amounts to the cold attenuation of the tube with the beam turned completely off. In this tube approximately 80 db net attenuation is achieved. The power output variation from full gain value to maximum attenuation, which is obtained merely by varying the control grid voltage, is something over 100 db.

The control grid is a low capacitance electrode that is capable of turning the beam on and off with millimicrosecond rise time when driven from a low impedance source. Total capacitance between the grid and cathode as measured between their respective pins at the base of the tube is 6.4 mmfd.

Two applications in which the grid control feature of this TWT can be utilized come immediately to mind. With a cw signal applied to the tube input, for example, the TWT can be used as an rf pulse modulator by applying video pulses to the grid. The other application involves using the grid as the control element in the feedback

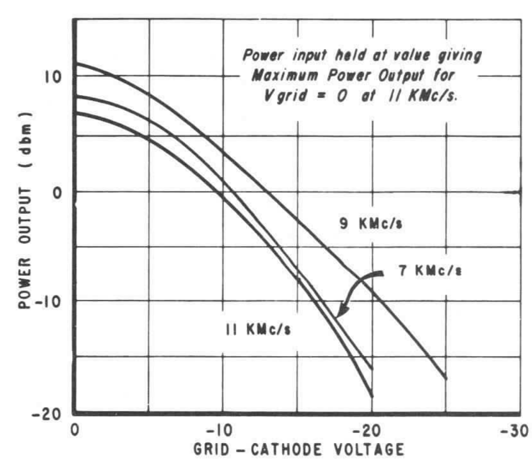

Figure 4. Power Output vs Grid Voltage.

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CASCADING TRAVELING-WAVE TUBES

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In many system applications, over-all traveling-wave tube characteristics are required which are impossible to achieve in one tube. It then becomes necessary to cascade two or more tubes to achieve the desired results. This Engineering Note will point out some of the problems that will be encountered when cascading traveling-wave tubes.

POWER AND GAIN CHARACTERISTICS

Traveling-wave amplifier tubes do not saturate in the same manner as triodes, pentodes, and other conventional tube types. As the power input is increased from a low level in conventional tubes the power output increases linearly and then levels off to a maximum value which is nearly independent of additional power input. This is indicated by curve A in Figure 1.

A traveling-wave tube has a similar output characteristic up to the point of maximum power output, after which further power input will result in a decreased power output. The power output continues to decrease with a further increase of power input. As shown by curve B in Figure 1, excessive power inputs result in further increasing and decreasing variations in the power output which never approach the original maximum power output.

In order to define the significant power and gain parameters of the TWT, let us consider the operating characteristics of a typical traveling-wave tube amplifier. Figure 2 shows the power output and gain plotted as a function of the power input at a fixed frequency.

The gain and power characteristics are divided into the unsaturated or linear region and the saturated or non-linear region. In the linear region the power output is proportional to power input, and the gain of the tube is constant and independent of power input level. The gain in this region is called the <u>small-signal gain.</u>

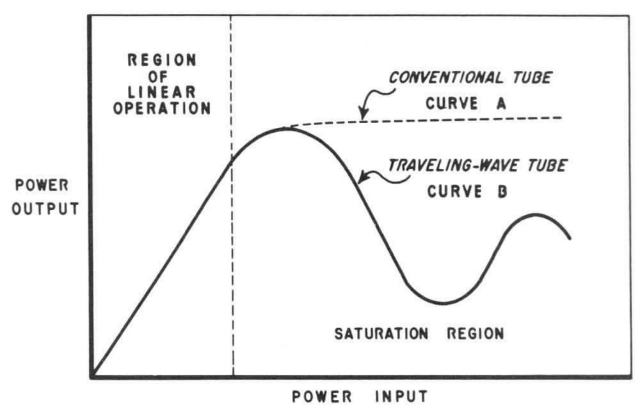

Figure 1. Comparison of Power Output Curves of Traveling-Wave Tubes and Conventional Low Frequency Tubes.

*President, Huggins Laboratories, Inc.

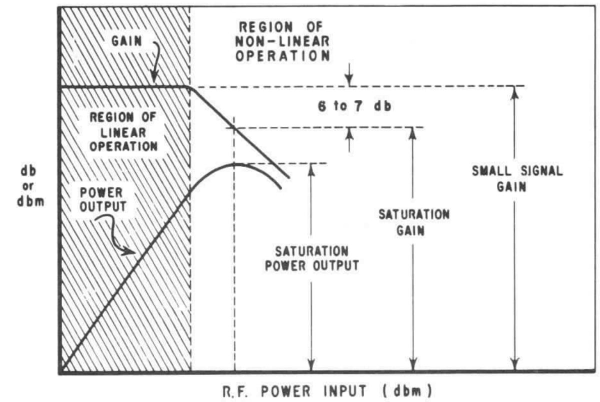

Figure 2. Typical Power and Gain Curves for a Traveling-Wave Tube.

As the input power is increased, the power output is no longer a linear function of power input. That is, the tube has passed into the saturated region. A further increase in power input causes the output power to deviate even more so from a linear curve until it reaches a maximum and begins to decrease. This point of maximum power output is called the saturation power output.

In passing into the saturated region the gain is no longer constant (i.e., the gain becomes a function of input power). As the input level is increased, gain decreases. The gain at the point of saturation is called the saturation gain and is approximately 6 to 7 db less than the small-signal gain.

In Figure 3 we see the power and gain characteristics of the HA-1 (2-4 kmc, low power amplifier) taken at a single frequency. The algebraic difference between the power output and power input measured in dbm is equal to the gain. From these curves the small-signal gain, saturation gain, and saturation power may be ascertained at a frequency of 3.0 kmc.

A continuous plot of these three parameters as a function of frequency is shown in Figure 4 for the same traveling-wave tube. The individual

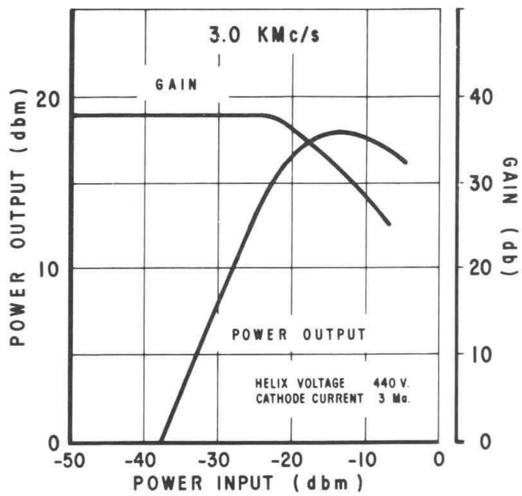

Figure 3. Power and Gain Curves of an HA-1 at 3.0 kmc.

values of these three parameters at the fixed frequency in Figure 3 may be checked at the corresponding frequency in Figure 4.

NOISE CONSIDERATIONS

In addition to the gain and power characteristics of the traveling-wave tube, the role of noise figure in applications where it is neces-

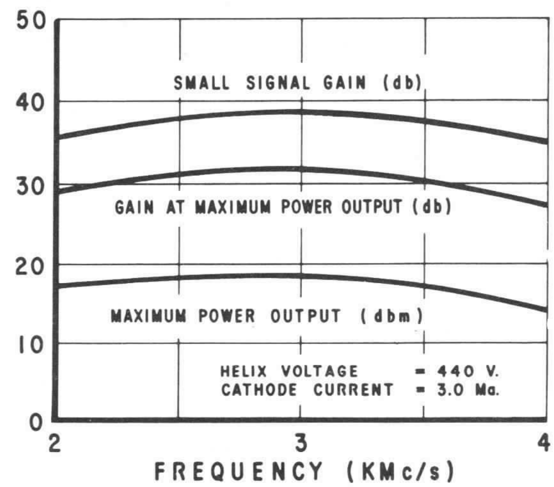

Figure 4. Power and Gain Curves as a Function of Frequency for an HA-1.

sary to cascade one or more traveling-wave tubes must be considered.

Ideally, an amplifier should amplify only the signal which is applied to its input, and it should not produce any signals in the output that do not exist in the input signal. In the practical case, signals are always present at the output which are not applied to the input.

These spurious signals are divided into the categories of noise and hum. Since hum signals are usually a power supply problem which can be eliminated by proper design, noise signals will be taken as the limiting factor in our consideration of traveling-wave tube noise.

The limiting noise levels in a system are those produced by the random fluctuations of the electrons in the conductors. The level of this noise power has a definite relationship to the system's absolute temperature and the bandwidth over which power is accepted. Thus, it is often referred to as thermal noise. In a matched transmission system where maximum power transfer between the source and load occurs, the available noise power at room temperature is 4.0 x 10^{-21} watts per cycle of bandwidth. A more convenient way to express this noise level is -114 dbm per megacycle of bandwidth.

Thermal noise of the system may be considered as a noise generator, the output of which is being amplified by the traveling-wave tube in the same manner and to the same degree that the input signal is being amplified. The output will consist of both the amplified input signal and the amplified system thermal noise.

In addition to the system thermal noise, there is an additional contribution of noise in the output of the tube that is associated with the traveling-wave tube itself. Whereas the thermal noise enters the amplifying portion of the tube by way of the input coupler, the tube noise enters the amplifying portion of the tube by way of the electron beam. An electron beam has noise components that are propagated along it. These noise components are amplified in the same manner that the system thermal noise is amplified.

The measure of the contribution of noise by the tube alone is called the noise figure and represents the power ratio of the signal-to-noise ratio at the input to the signal-to-noise ratio at the output (usually expressed in db). By considering the tube as a perfect amplifier (i.e., introducing no noise), the noise figure can be represented by an equivalent noise generator at the tube input.

Figure 5 is a representation of these signal and noise components as they would exist at the input of a traveling-wave tube amplifier.

The magnitude of the input noise level per

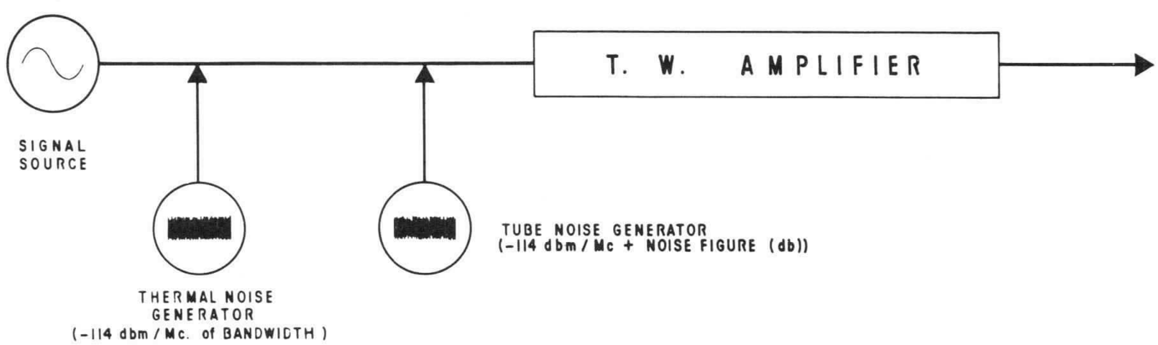

Figure 5. An Equivalent Representation of Noise Sources in a System Containing a Traveling-Wave Tube.

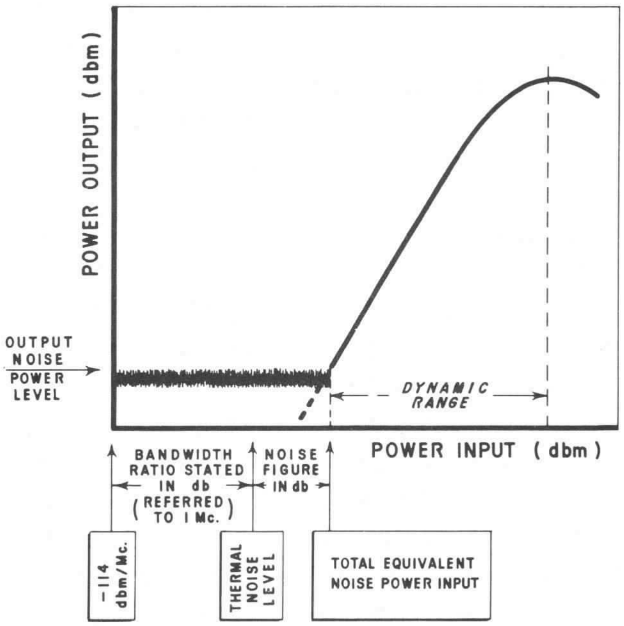

Figure 6. Combined Plot of Noise and Power Characteristics.

megacycle of bandwidth is greater than the thermal noise level by an amount equal to the tube noise figure. As an example, assume that the TWT has a noise figure of 20 db. When this figure is added to the thermal noise level of -114 dbm/mc, we find that the input noise level is -94 dbm per megacycle of bandwidth. As will be noted, the input noise power is predominantly tube noise. This will be true until tube noise figures are reduced to a few db, at which time the thermal noise generator contribution will become appreciable with respect to tube noise.

Further, suppose that a bandpass filter (e.g., an IF amplifier) which is 10 mc wide follows the TWT amplifier. Noise is accepted over a 10 mc bandwidth which is a factor of 10 (i.e., 10 db) greater than 1 mc. The noise level at the TWT's input is then -84 dbm. The preceding, of course, results from a computation using the well-known relationship $N_i = FkTB$. If the tube gain is 30 db, noise power level at the output would be -54 dbm.

Another way of representing these various signal and noise components would be to plot them graphically on the power characteristics of the tube as indicated in Figure 6.

The origin is taken as -114 dbm/mc for convenience. The thermal noise of the system is determined by accounting for the bandwidth of the following system. The equivalent noise power input to the tube is determined by adding the noise figure of the tube (in db) to the thermal noise power in (dbm.) The output noise power level can then be determined from the power characteristic curve.

The dynamic input range of the tube is limited on the low power end to the signal level which is just equal to the equivalent noise input level. It is limited on the high power end by the signal level which just drives the tube to saturation. The dynamic range (in db) is then equal to the difference between these two power levels.

SENSITIVITY

One common measure of sensitivity is known as tangential sensitivity. Tangential sensitivity is the input signal power which is just equal to the total equivalent input noise power of the system. This definition is demonstrated in Figure 7 for an oscilloscope display of the detected output of a pulsed signal. Signal power input is read when the bottom of the noise with the pulse on is tangential to the top of the noise with the pulse off. The noise level which is seen in such a display is a function of the bandwidth of the system as well as its noise figure. Video as

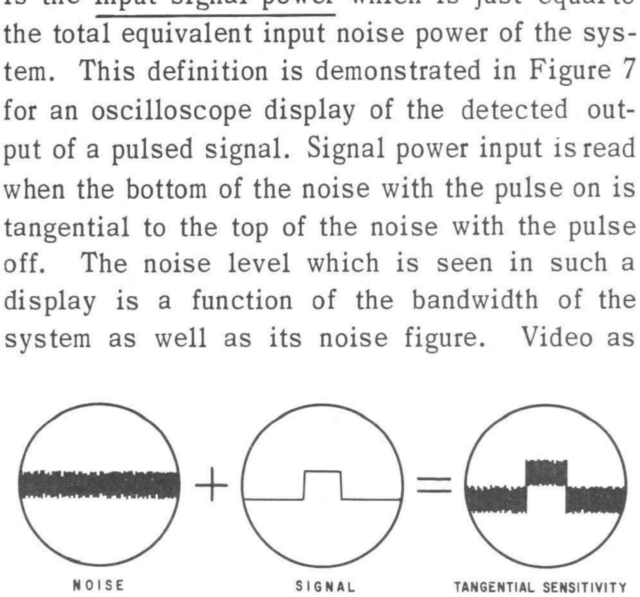

Figure 7. Representation of Tangential Sensitivity.

well as rf bandwidth must be taken into account. If a square-law detector is between the rf and video portions of the system, the calculation of equivalent bandwidth must be modified to account for this.

For example, if the limiting rf bandwidth ahead of the square law detector is 1000 mc wide (30 db with respect to 1 mc) and the video bandwidth following the detector is 10 mc wide (10 db with respect to 1 mc), the sensitivity improvement due to this bandwidth reduction is $\frac{30 - 10}{2} = 10$ db. Dividing by 2 accounts for the square law detector's effect on the noise.

Some systems may be able to handle signals at levels below tangential, and others may require signals much greater than tangential. This criterion, however, does establish a common measure of system sensitivity.

There are two obvious ways to increase sensitivity. One method would be to employ a low-noise traveling-wave tube amplifier. In any chain of amplifiers, it is the input amplifier that determines the noise figure of the complete chain providing the input stage has sufficient gain so that its noise overrides that of subsequent stages. It is for this reason that great care is taken in the circuit adjustment and noise figure of any rf input stage. Therefore, any decrease in noise figure of the individual input tube will directly improve the sensitivity of the complete system. Figure 8 shows the noise figure for the HA-11 (1 mw, 2-4 kmc amplifier) which is essentially an HA-1 with a low-noise gun replacing the standard gun. This low-noise gun design results in an improvement of approximately 10 db in noise figure.

A second way to increase the sensitivity would be to decrease the over-all system bandwidth. Increased sensitivity may be realized with a decreasing bandwidth until the bandwidth becomes so narrow that it will not pass the spectrum that is necessary to accommodate the expected signal.

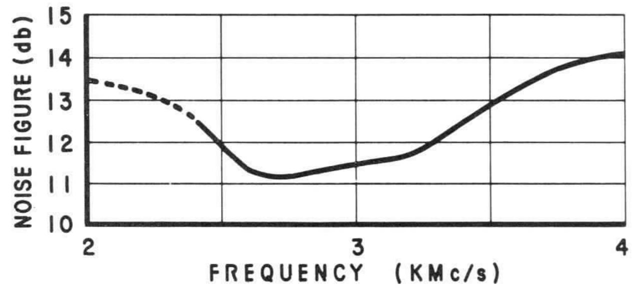

Figure 8. Noise Figure as a Function of Frequency for an HA-11

The noise figure of the tube and bandwidth of the system are two independent variables contributing to the sensitivity of the system. A change in one of these variables does not affect the other. In conclusion, it should be noted that any loss in the rf circuit before the input coupler to the traveling-wave tube represents a decrease in sensitivity. This loss is an additional factor which must be considered with the thermal noise and the noise figure of the tube. The input loss (in db) must be subtracted directly from the sensitivity of the system.

LIMITER CHARACTERISTICS

A device which provides a constant power output over a very large dynamic range of power inputs is desirable in some microwave applications. One method of achieving this involves the utilization of the saturation curves of two dissimilar traveling-wave tube amplifiers. In this application the two saturation curves are combined so as to keep the output tube in a state of saturation over a wide range of input power levels to the first tube. This is best illustrated by considering a numerical example of this principle applied to two of our standard tubes.

Consider the saturation curves of two S-Band production type tubes, the HA-1 and the HA-2.

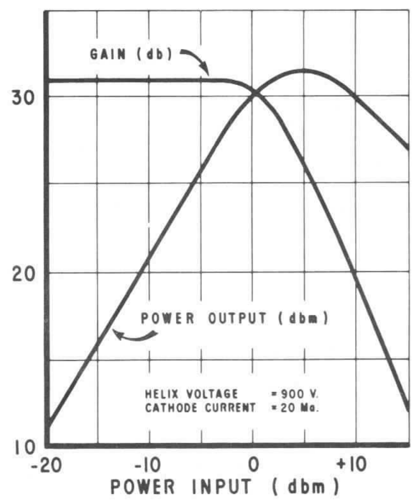

Figure 9. Power and Gain Characteristics of an HA-2 at 3.0 kmc.

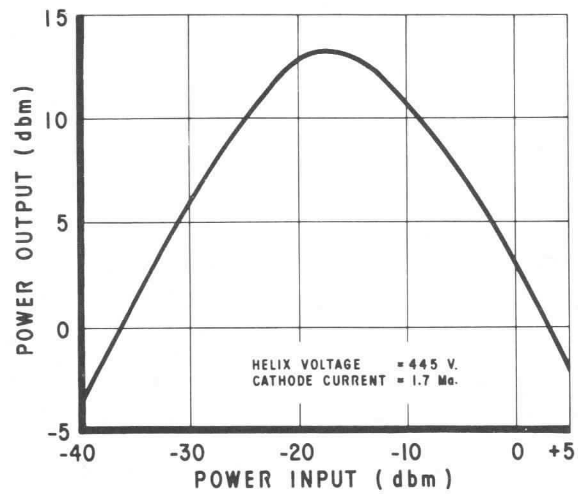

Figure 10. Power Characteristics of an HA-1 at 1/2 Rated Beam Power at 3 kmc.

The HA-1 is a 10 milliwatt amplifier, and the HA-2 is a 1 watt amplifier. Their power and gain curves are illustrated in Figures 3 and 9 respectively. Assume that the specification calls for a constant power output of 1 watt ±1½ db at 3.0 kmc. This output is to be constant over as large a range of inputs as can be easily achieved. It is seen from Figure 9 that if the power inputs were on the order of –2 dbm to +13 dbm, the requirement would be met with the HA-2. Suppose the requirement, however, calls for a wider dynamic range encompassing a lower level of power inputs.

Examing the HA-2 saturation curve further, we note that the input power of +13 dbm results in 28.5 dbm power output. Any further drive would tend to give less power than the specifications call for. It then becomes desirable to have a second traveling-wave tube that will saturate at +13 dbm at 3.0 kmc. Such a condition is met by operating the HA-1 amplifier at approximately one-half its rated beam power. The HA-1 saturation curve under such conditions appears in Figure 10.

The power output scale of the HA-1 is superimposed on the power input scale of the HA-2 in Figure 11. The HA-2 has its beam power adjusted to give a saturated power output of 31.5 dbm at the required frequency of 3.0 kmc. It is seen that when the 28.5 dbm points of the HA-2 are projected on to the saturation curve of the HA-1, the dynamic range over which these power output conditions will be met is –39 to +5 dbm. That is, the input power varies over a range of 25,000 to 1 while the power output varies only over a 2:1 ratio.

The power output for any particular power input is obtained by projecting these individual saturation curves to form a power output curve as a function of power input. As an example, consider any arbitrary input as Point A and project it as shown. Point-by-point projections result in the complete power output curve.

This degree of limiting could be extended by adding a second HA-1 to obtain a power output variation of 30 dbm ±¼ db for 62 db dynamic power input range. Comparable limiting can be achieved with only two tubes when they are specially designed for this type of operation. Depending upon the required power output, such a combination could entail two HA-2's, two HA-

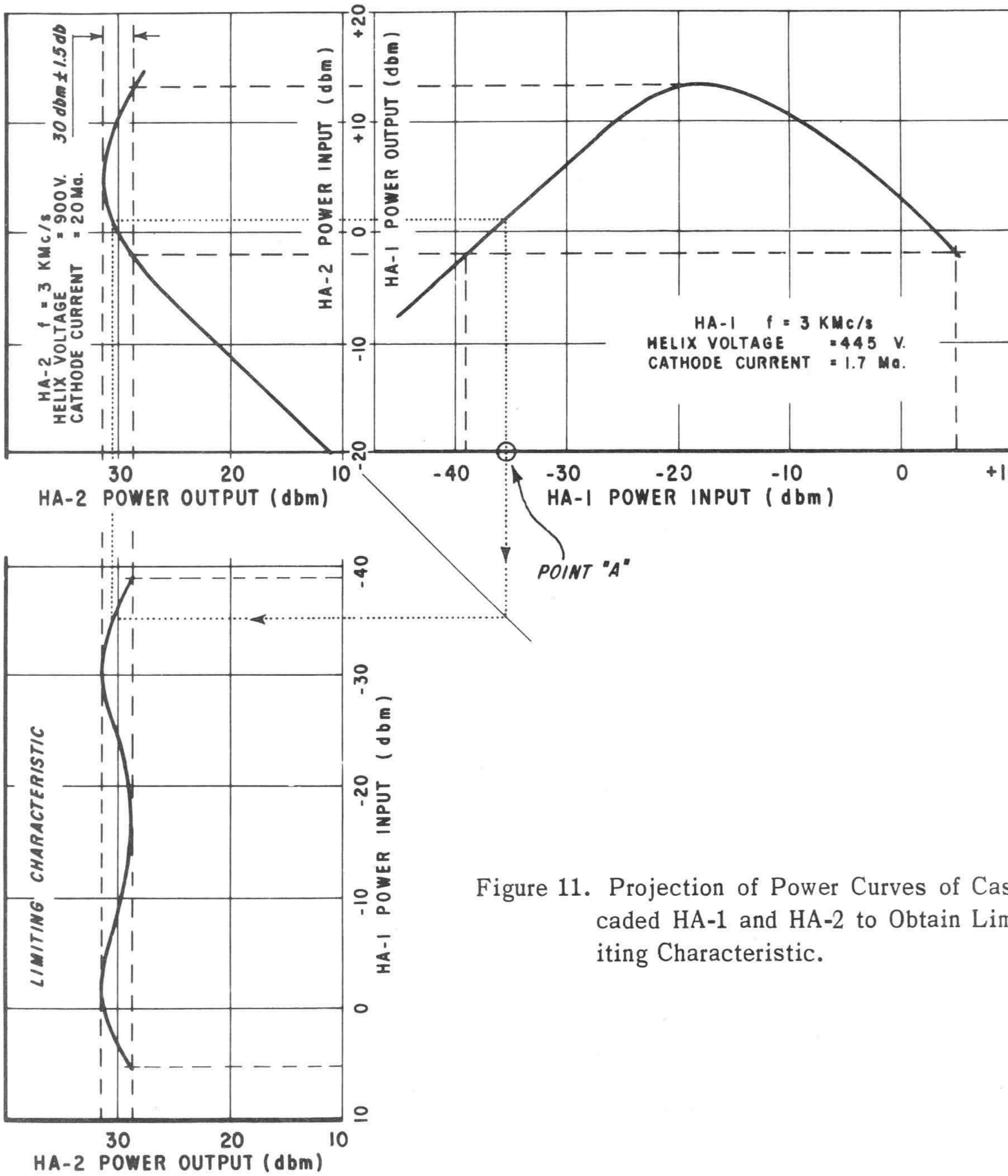

Figure 11. Projection of Power Curves of Cascaded HA-1 and HA-2 to Obtain Limiting Characteristic.

1's, or one of each. Limiting, of course can be performed in other bands with other tube types.

Though this discussion has been concerned with single frequency operation, wide band limiting is possible by using TWTs designed to have flat gain vs frequency characteristics.

CASCADED AMPLIFIER CHARACTERISTICS

Many applications demand more amplification and power output than can be realized with one tube. It is then necessary to cascade two or more traveling-wave tubes to achieve this added performance. The factors to consider are the noise figures of the individual tubes, the saturation characteristics, and how these parameters are interrelated when they are superimposed on each other as they were in Figure 11.

The problem is twofold. The noise of the input tube may saturate the output tube, thus producing an unusable display or output. Also, the signal power output level of the input tube may saturate the output tube which would result in a very small or limited linear dynamic input range. Both of these conditions are dealt with in the following numerical example which illustrates most of the problems that must be solved in this type of operation.

Assume that a 40 db linear dynamic input range with 55 db gain is required at 3.0 kmc. Further assume that the system bandwidth must equal the bandwidth of the tube. Since the required gain is not usually possible with a single tube, it becomes necessary to cascade two tubes. The first apparent solution would be to cascade two HA-1 TWTs, for each would provide in excess of 30 db small-signal gain. The total gain of 60 db minimum should easily meet the assumed requirements.

In Figure 12 the power curves of two HA-1 TWTs are superimposed in the same manner that was followed previously in Figure 11. Noise figure considerations (assuming F = 26 db) and an effective 3 db noise bandwidth of 1.0 kmc are also included.

Examination of Figure 12 reveals that the solution under discussion has failed on two counts: 1) The linear dynamic input range of the first tube is only 34 db (58-24) which falls short of the required 40 db, and 2) the equivalent noise power input of –58 dbm at the input of the first tube, when amplified by 40 db to –18 at the output, will drive the second tube to saturation. The latter produces 16 dbm of noise

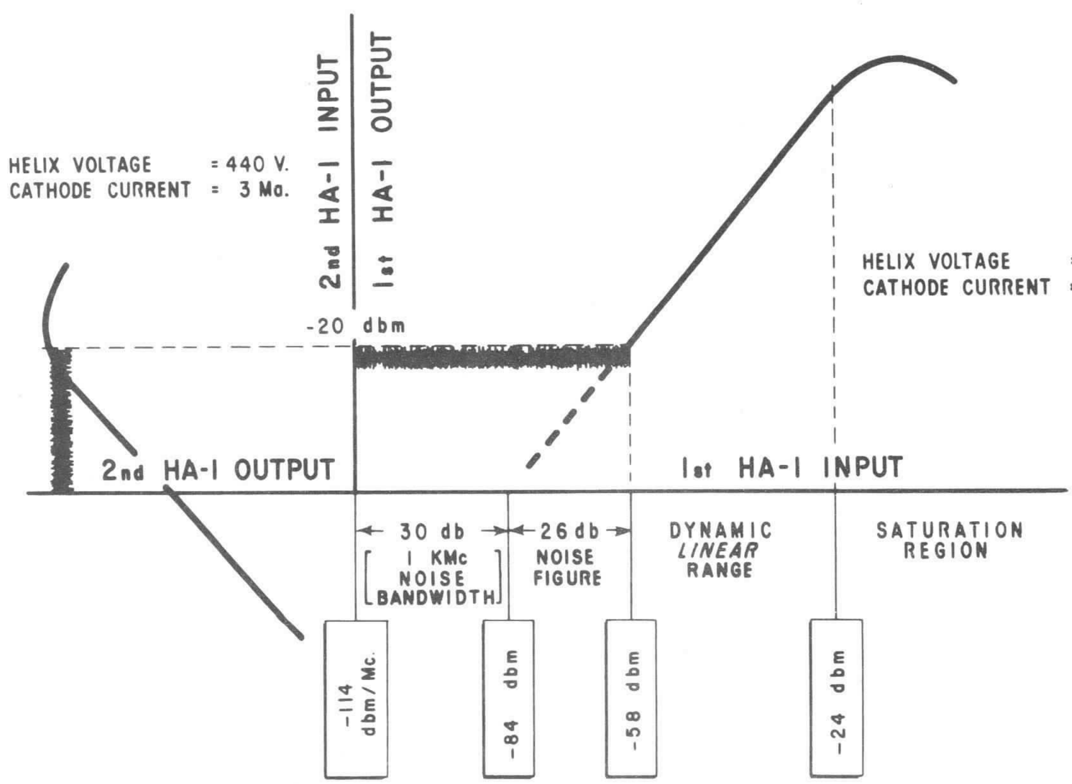

Figure 12. Projection of Power Curves of Two Cascaded HA-1's.

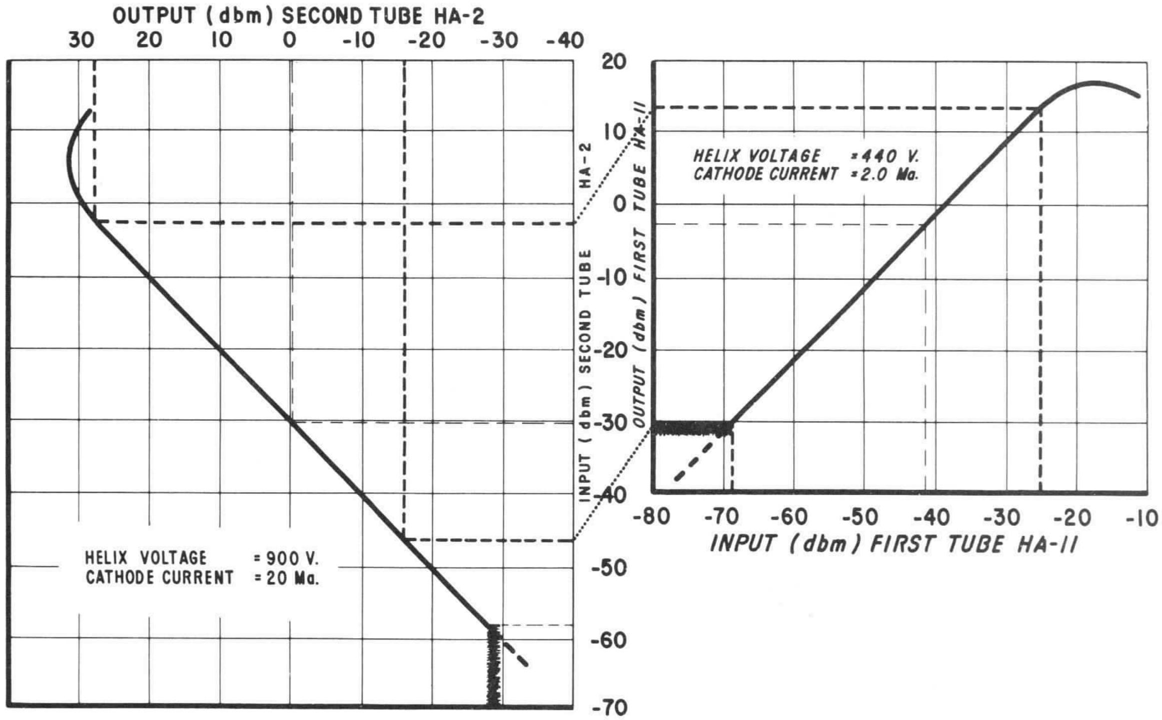

Figure 13. Projection of Power Curves for a Cascaded HA-11 and HA-2.

at the output of the second tube with a corresponding zero linear dynamic range. Any increase in input power drives the second tube further into saturation. Such a cascaded pair may not even work well as a noise source, for operation in the saturated condition produces clipping of the noise peaks.

In summary, the two factors that made this combination fail were the noise figure of the first tube which limited the dynamic input range and the power handling capabilities of the secone tube which allowed it to be immediately saturated by the first tube.

Let us now consider the combination of a low-noise HA-11 followed by a medium power HA-2 as illustrated in Figure 13. The HA-11 has a noise figure of at most 15 db across the 2.0 to 4.0 kmc band. The HA-2 has a maximum noise figure of 25 db over the same frequency band. With a 1 kmc effective noise bandwidth, the equivalent noise power input is –69 dbm for the HA-11 and –59 dbm for the HA-2.

The HA-11 has an inherent linear dynamic range of 44 db (69-25). When operated directly into the HA-2, however, the over-all dynamic range of the cascaded pair is only 27 db (69-42) because the HA-2 is driven into its saturation region at –2 dbm input. The over-all gain for this limited linear range is 69 db (69-0). Both tubes must be made to saturate simultaneously to achieve the greatest dynamic linear input range.

Since 69 db gain is not required, we can put an attenuator between the first and second tube so that the first tube will not saturate the second tube at low input levels as it did in Figure 13. The padded combination is illustrated in Figure 13 where the superimposed first tube output and the second tube input power levels have been moved in respect to each other by the amount of the pad. This shift is indicated by the dotted oblique lines between the first tube's output and the second tube's input. The value of attenuation between the tubes is the difference between the highest linear output of the first tube (+13 dbm) and the highest input resulting in a linear output of the second tube (−2 dbm). In this case, the value for the pad is 15 db.

The final over-all characteristics are a 44 db (69-25) linear dynamic range and a small-signal gain of 53 db (69-16). This closely approximates our original goals.

As these two input-output ordinates were effectively shifted in respect to each other in Figure 13 by the interstage 15 db pad, it should be noted that the equivalent noise power input of the HA-11 came within 11 db (58-47) of falling within the noise region of the second tube. If this had occurred, the limiting sensitivity of the system would have been determined by the noise figure of the second tube and not the noise figure of the first tube.

One other method of operating the tubes which was not examined would be to operate the HA-11 at reduced beam power so that it would saturate simultaneously with the HA-2. Operating in this fashion, the dynamic range of the HA-11 (and thus of the over-all combination) would have been reduced because noise figure does not improve in proportion to a reduction in saturation power output. In fact, noise figure decreases by only 1 or 2 db as beam current (beam power) is decreased.

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PHASE MODULATION OF TRAVELING-WAVE TUBES

R.A. Huggins* and D.R. Bellis**

Although the traveling-wave tube is best known and utilized for its broadband characteristics, it has many narrow band characteristics that are equally important. Frequency shifting by phase modulation (often called transit-time modulation) is one of the more important narrow band applications of the TWT. As a frequency translator the traveling-wave tube finds use in doppler and coherent pulse simulators, homodyne systems, and similar frequency shift or frequency sensistive applications.

Though phase shifts of 2π radians can be obtained by modulating the grid, anode, or helix, the latter proves most satisfactory from the standpoint of amplitude modulation. This discussion, therefore, will be confined to helix modulation.

SERRODYNE OR SINGLE SIDE-BAND MODULATION

The velocity of a signal passing through a traveling-wave tube is determined by the combination of the physical configuration of the helix and the electron beam. Normally, the helix voltage is adjusted so that electron velocity and the wave velocity on the helix are essentially equal. Varying the helix voltage, and thus the corresponding electron velocity, will perturb the velocity of the electromagnetic waves on the helix. Velocity changes tend to advance or retard the phase of the output signal with respect to the phase of the input signal.

In Figure 1, the phase delay or advance is plotted as a function of helix voltage for the HA – 1 at a fixed frequency of 3.0 kmc. An 18 volt modulating voltage, when applied symmetrically about the synchronous value of helix voltage, will shift the phase 360° or 2π radians. This manifests itself as a ±180° (a $\pm\pi$ radian) phase variation from the condition of zero relative phase shift at the synchronous voltage.

Because phase is essentially a linear function of helix voltage, a constant $d\phi/dt$ can be obtained by applying to the helix a voltage that increases or decreases linearly as a function of time. For example, let the voltage begin at a value corresponding to $-\pi$ (433 V); as the voltage decreases to 415 V, the phase will reach the value $+\pi$.

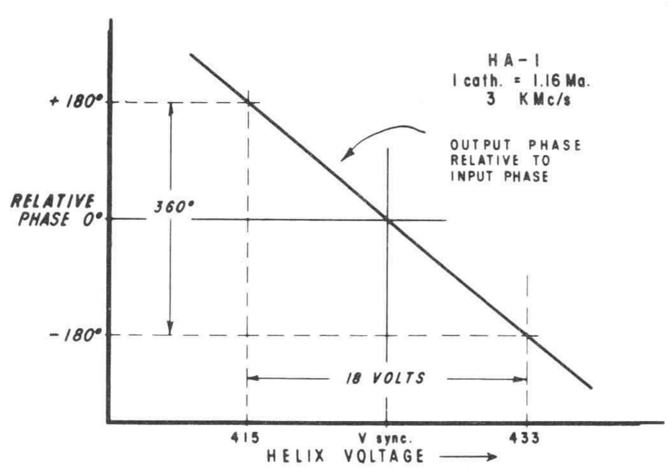

Figure 1. Phase Shift vs Helix Voltage for HA-1 TWT.

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A phase of $+\pi$ or $-\pi$ is functionally the same. Therefore, the helix voltage can be instantaneously returned from 415 V to 433 V with no apparent change in phase. We recognize the composite voltage change as a linear sawtooth with an infinitely rapid flyback time. When this waveform is repeated, the phase at the output of the tube appears to increase indefinitely with time at a constant $d\phi/dt$. This means that the input frequency is shifted negatively by an amount equal to $d\phi/dt$ since the time rate of change in phase equals the frequency shift.

A sawtooth voltage which swings symmetrically about the synchronous helix voltage is usually used in applications where these phase shift characteristics are required. This sawtooth modulating voltage is somewhat analogous to allowing the second hand of a clock to complete 59 seconds of the minute and then quickly turning it backwards in the last second to its initial starting place, after which the process repeats itself. The apparent result is a continuous advancement in phase in which the algebraic summation of the revolutions never exceeds one revolution.

To continue the analogy further, the half-minute point is analogus to the synchronous helix voltage since both variations are a plus and minus change about a center of reference.

The repetitive sawtooth voltage type of modulation just described, in which the phase is changed by 360° per voltage swing, produces single sideband modulation of the original frequency. This modulation is such that the original frequency is shifted by an amount equal to the modulating frequency.

A negative slope on the modulating voltage — as applied to the helix — produces a decrease in frequency output as compared with the original input frequency; a positive slope on the helix modulating voltage results in an increase in frequency output as compared with the original input frequency.

As shown in Figure 2, an 18 volt peak-to-peak, negative-going voltage with a one second period that is superimposed on the dc voltage applied to the helix of an HA-1 TWT will result in an output signal that is shifted in phase with respect to the input signal by 360° per second. In this case the output signal frequency is shifted negatively ($f_{out} < f_{in}$) by one cycle per second, or one cycle per sawtooth period. Should the sawtooth frequency be 1000 cps, the output phase would be shifted 2π radians per 1/1000 of a second. This would result in a negative frequency translation of one cycle every 1/1000 of a second or 1000 cycles per second which again equals the sawtooth frequency.

Figure 2 gives a composite picture of this action wherein a sawtooth voltage of period t with a flyback time of Δt is applied to the curve of Figure 1. In this figure we have taken some liberties with our illustrative projections in that the sine wave shown is an oscilloscope presentation of the difference frequency between the input and output frequencies.

The linear sawtooth voltage input produces a linear increasing or decreasing phase output as shown. Directly below the sawtooth input voltage is plotted the difference frequency signal which would be obtained if the input and output signals were mixed. The difference frequency is made up of two components. The predominate frequency component is equal to 1/t and this side band contains practically all of the available rf power. The other frequency component is equal to $1/\Delta t$ and occurs only during flyback time. Since it is a much higher frequency, little energy is expended during its cycle, and it represents only a small energy component of the total signal spectrum.

In operating one of these phase-modulation frequency-shift systems, an oscilloscope display of the aforementioned difference frequency can be used to monitor over-all system performance.

A simple method of observing the difference frequency is given in Figure 3. This system consists essentially of a signal generator which acts as both an rf source for the TWT and a local oscillator for a crystal mixer. The relative magnitudes of the two signals are proportioned so as to ensure linear mixing. The signal generator presents an LO signal level of 0 dbm to the mixer, and the traveling-wave tube, through a pad, provides an input to the mixer of −30 dbm.

For operating simplicity, the crystal in the slotted line is used as a mixer. Thus, the mixed signal at any arbitrary phase can be examined by adjusting the probe position. The TWT helix is modulated with a sawtooth voltage, and the output is fed through an attenuating pad to the other end of the slotted line. After mixing, the resulting difference frequency is amplified and displayed on an oscilloscope or applied to distortion measuring equipment. Either the oscilloscope display or the distortion analyzer may be used as the criteria for determining the various

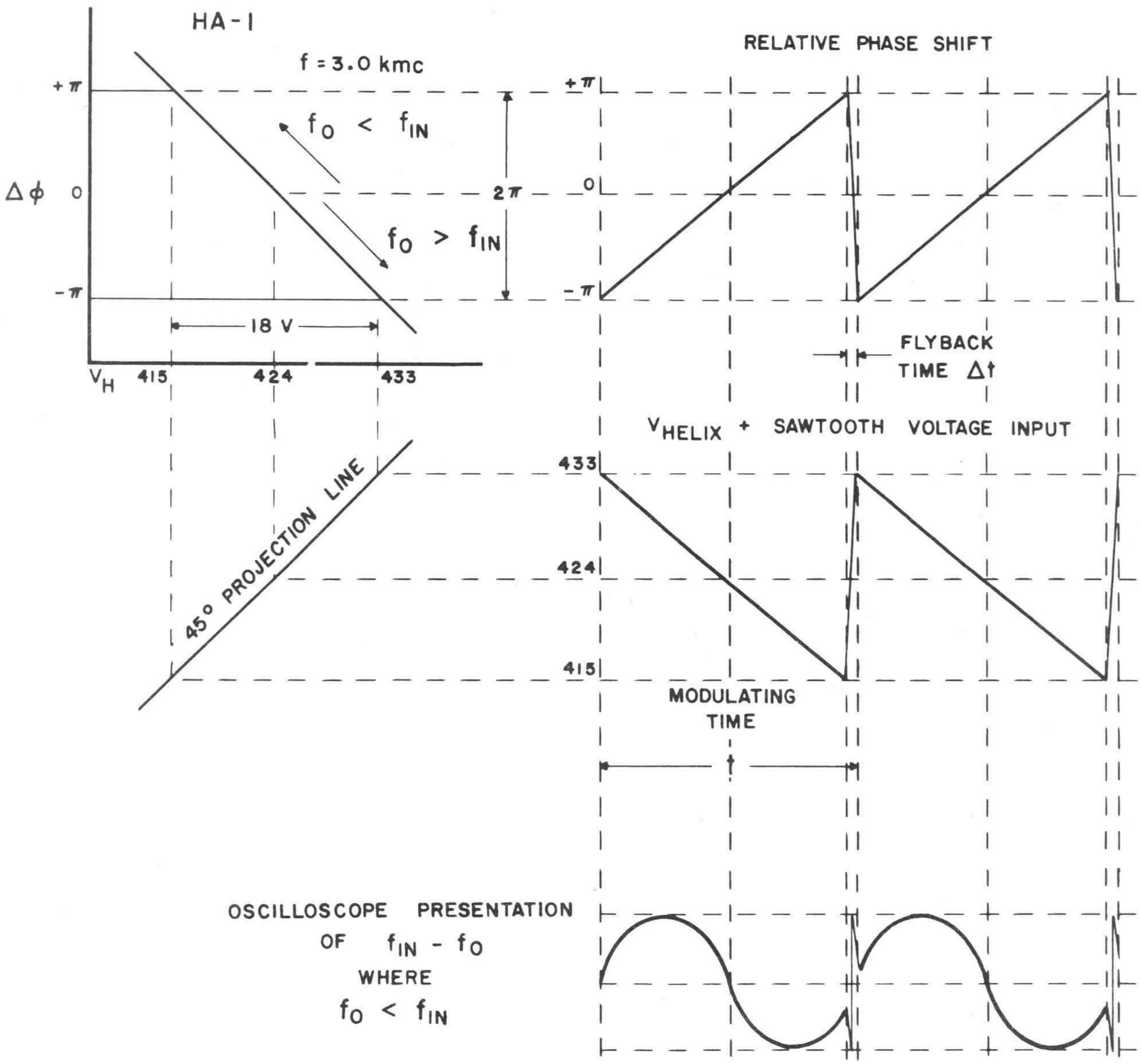

Figure 2. Phase Modulation Time Reference Diagram.

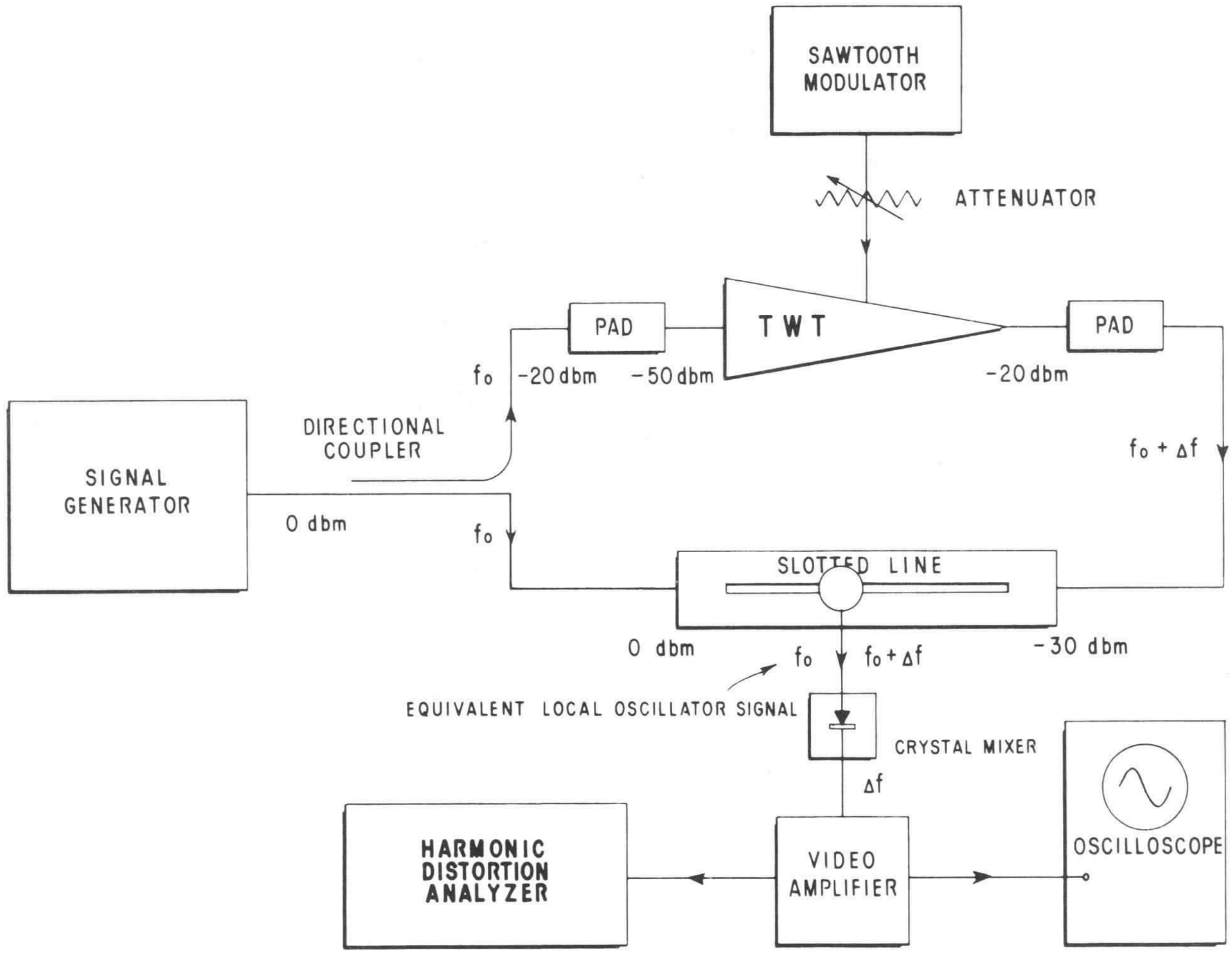

Figure 3. Block Diagram of Measuring Equipment.

TWT voltages required to achieve optimum modulation conditions.

The correct value of voltage swing and its center or average value may be determined by observing the best sine wave on the oscilloscope or by setting the system up for minimum harmonic distortion. The adjustments consist of 1) determining the magnitude of the sawtooth voltage and, 2) adjusting the helix voltage so that the modulating voltage swings about the synchronous value.

A third adjustment, which may be impossible to make due to system considerations, is to adjust the pick-off point on the slotted line for an optimum phase relationship of the flyback time. The difficulty in making this adjustment stems from the fact that the mixer is usually in a fixed position in the rf circuit of a normal system. Even if the mixer is movable for optimum conditions, it would be at the correct position for only selected frequencies across any operating band.

INCIDENTAL AMPLITUDE MODULATION AND PHASE DISTORTION

Since distortion may greatly affect the operation of the system in which these signals are to be used, it is appropriate to consider the types of distortion, their causes, and how they can be remedied.

Spurious signals in the output of the tube can be thought of as distortion. This results from the fact that the unwanted frequencies are harmonically related to the frequency shift, Δf, and can be represented by a spectrum of frequencies centered about the shifted carrier and spaced from one another by Δf. If the output of the tube is linearly mixed with the input signal as shown in Figure 3, and the difference frequency signal is examined, the unwanted frequencies are then merely harmonics of the difference frequency. Their value can then be determined simply by making a harmonic distortion measurement.

Our discussion so far has assumed that phase change in a TWT is a linear function of helix voltage. This is only approximately true since the phase change actually varies as the square root of the helix voltage change. Because the helix voltage change is such a small percentage of the synchronous voltage, however, the approximation of a linear phase characteristic is valid.

For example, a 2π radian shift in phase is obtained in the HA–1 with a helix voltage change equal to roughly 5% of the synchronous value. The slight curvature which exists in the phase characteristic over this voltage range introduces harmonic distortion of 2 to 3% assuming there are no other sources of distortion and the modulating voltage is a perfect sawtooth.

Distortion may also be introduced by not operating V_m (the helix modulating voltage) about the correct helix voltage. Figure 4 depicts three operating conditions – two when V_m is centered on the correct helix voltage and one

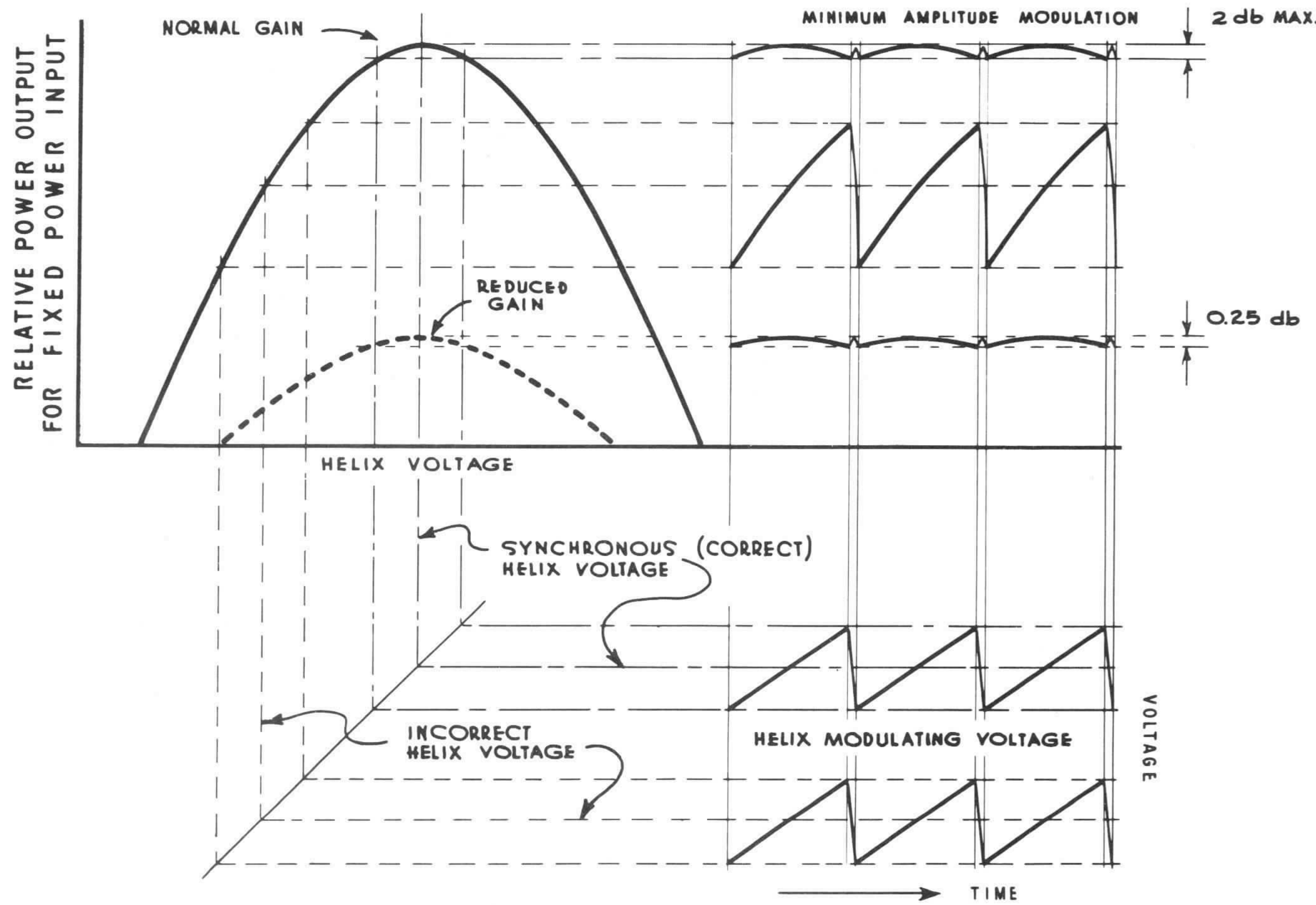

Figure 4. Amplitude Modulation Derived from Voltage Modulating the Helix.

when V_m is operated about an incorrect helix voltage. In this illustration the modulating voltage of the proper amplitude to vary the phase $\pm\pi$ radians is projected on the normal gain vs helix voltage characteristic. As will be noted, the helix voltage which produces the least amplitude modulation for the required helix voltage swing is equal to the synchronous voltage (i.e., the value of helix voltage which gives maximum gain).

The resulting AM is responsible for the harmonic frequency distortion that produces AM sidebands in the spectrum of the shifted frequency. On a standard HA-1 with greater than 30 db gain, the superimposed amplitude modulation is on the order of 2 db. This value of AM is lowered to approximately 0.25 db, as indicated in Figure 4, by reducing tube gain by a factor of 4. Gain reduction can not be effected by decreasing the beam current; it must be lowered by applying the proper cold loss to the TWT during manufacture.

In most traveling-wave tube applications the helix is operated at a fixed value for all frequencies under consideration. This represents truly broadband, untuned operation. Phase shift applications, on the otherhand, require that the traveling-wave tube operate at its point of maximum gain which in turn is determined by the synchronous helix voltage. In other words, the helix voltage must be optimized at each frequency to minimize incidental AM.

Figure 5 illustrates how the optimum helix voltage or synchronous voltage varies as a function of frequency. This curve is known as the dispersive characteristic for the traveling-wave tube.

The curve of phase shift as a function of helix voltage in Figure 1 and the curve of power output as a function of helix voltage in Figure 4 are superimposed in Figure 6 for various fixed frequencies. The required helix voltage for a 2π radian shift in phase is indicated, and the inci-

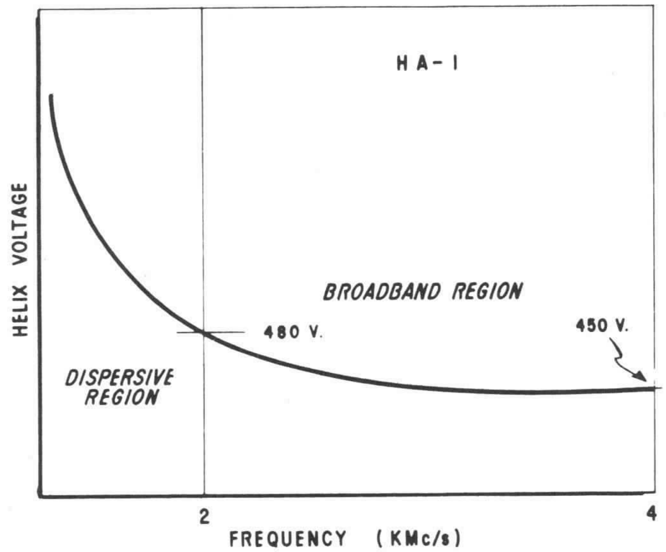

Figure 5. Dispersive Curve for HA-1.

dental amplitude modulation for this phase shift is also noted.

The series of curves in Figure 6 illustrates several important points concerning phase modulation in traveling-wave tubes. Figure 6, for example, shows that the optimum helix voltage increases as frequency decreases. This relationship is also indicated in Figure 5. Too, the magnitude of the modulating voltage required for a 2π phase shift increases with decreasing frequency. As indicated in both Figures 6 and 7, modulating voltage amplitude is approximately proportional to the wavelength.

We also note from Figure 6 that the gain response as a function of helix voltage becomes wider as frequency decreases. Coupled with the inverse relationship between modulating voltage amplitude and frequency, this fact accounts for the observation that the degree of AM, and hence distortion, is nearly independent of frequency.

Distortion that arises from improper adjustment of the amplitude of the sawtooth modulating voltage is caused by the fact that the phase shift through the tube is not advanced or retarded exactly 2π radians. Figure 8 gives various oscil-

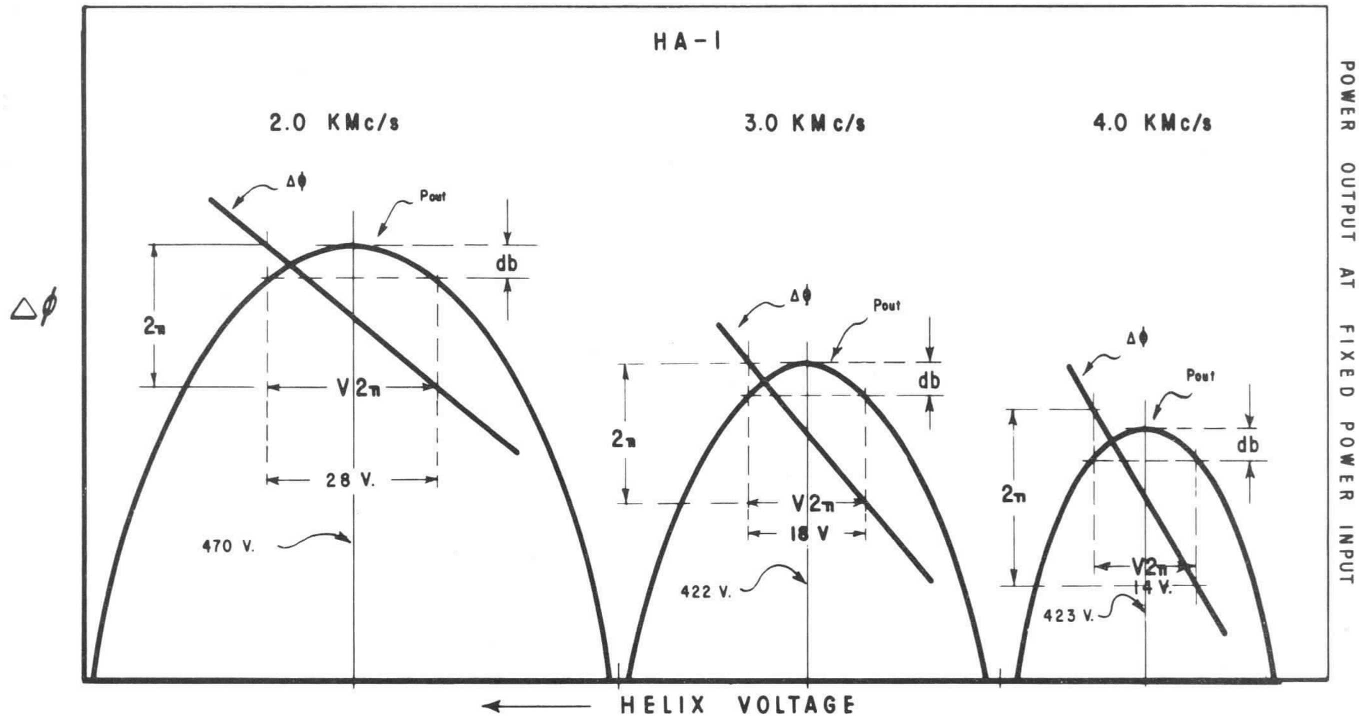

Figure 6. Gain and Δ φ vs Helix Voltage.

loscope presentations of the difference frequency from the mixer in Figure 3 which will be observed as the modulating voltage is varied. As discussed elsewhere in this Engineering Note, adjustment of the modulating voltage can be monitored by visual inspection of the oscilloscope pattern or by use of a harmonic distortion analyzer.

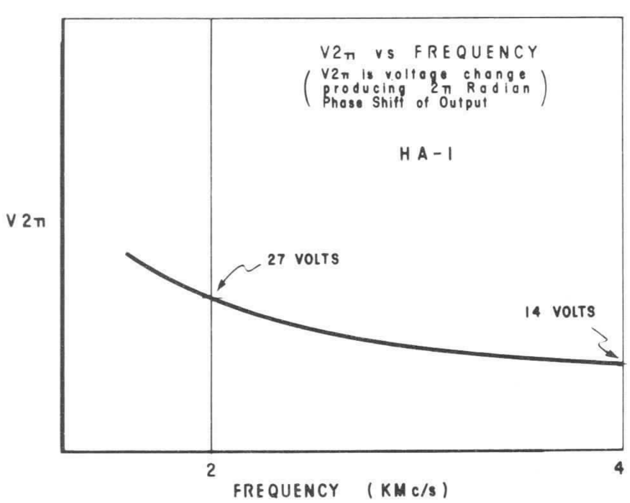

Figure 7. Peak to Peak Voltage Required for a 2π Radian Phase Shift.

Another source of distortion which may be adjusted for a minimum value in the test method illustrated in Figure 3 is the relative phase of the original signal and the altered signal at the mixer. The amount of distortion in the difference frequency output can be varied by moving the probe carriage on the slotted line. In most applications this adjustment would not be available since the usual system would have to perform satisfactorily under conditions of arbitrary phase relationships at the fixed position of the mixer.

Figure 9 shows the position of maximum and minimum distortion. It is readily seen that this distortion is a function of the position of the flyback time retrace on the difference frequency. The difference in percent harmonic distortion between the optimum pick-off condition and the poorest pick-off condition is 3 to 5%.

A final source of distortion could be attributed to any nonlinearity of the sawtooth voltage. Techniques are readily available so that this factor can be practically neglected. The distor-

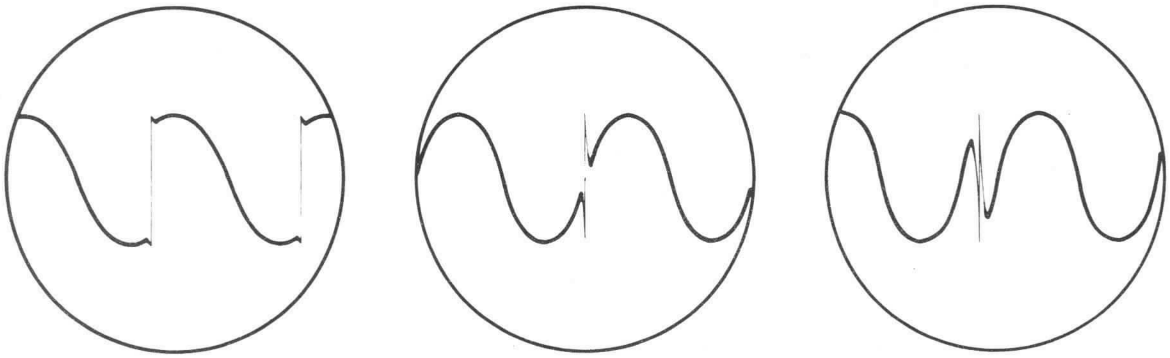

Figure 8. Difference Frequency Presentation for Various Amplitudes of Modulating Voltage.

tion due to the flyback time retrace may be minimized by making this time as short as possible. Such a precaution will result in two beneficial effects: 1) the energy confined to this retrace time will be practically negligible compared with the energy contained in the main trace, and 2) the frequency represented by the retrace time will be so high as to be outside of the frequency range of interest.

Distortions from 5 to 7% were measured in the system of Figure 3. By using TWT's specially made for the low-gain, flat-response characteristic in Figure 4, these distortion figures could be reduced by 2 to 3%. It is thus possible to achieve these single sideband modulations or frequency shifts with a percentage of distortion which corresponds to unwanted sidebands that are 30 to 40 db below the carrier amplitude.

LIMITS OF FREQUENCY SHIFTING

Frequency shifts on the order of tens of megacycles are possible using the phase modulation technique described herein. The only limitation is to design the circuitry that can modulate the helix with a sawtooth voltage at high frequencies.

Distributed amplifiers, as well as some methods which involve charging stray capacitances, are used to modulate the helix in the 10 to 50 mc range. From 50 to several hundred megacycles, frequency shifting with multiple sideband response can be accomplished by sine wave modulation of the helix.

The limiting condition which the tube imposes on such high frequency modulation is encountered when the period of the applied waveform becomes appreciable with respect to the transit time of the electron beam through the tube. Also, the possibility does exist that the modulating frequency would be limited by resonant conditions of the helix as a transmission line.

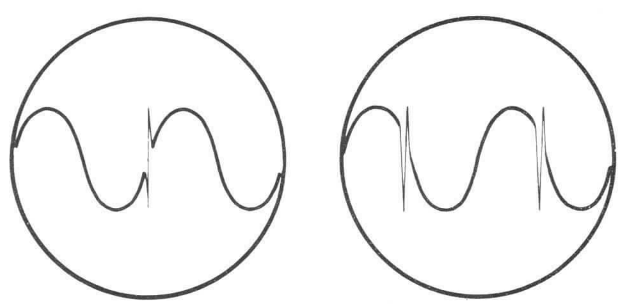

Figure 9. The Phase Relationship of the Flyback Time Retrace for Minimum and Maximum Distortion.

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USE OF THE TWT AMPLIFIER IN CONSTANT POWER SYSTEMS

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Microwave measurement techniques and system circuitry could often be simplified if a signal source with a constant rf power output were available. Simplification would stem from the signal source exhibiting either a constant amplitude as a function of frequency, a precisely fixed and known amplitude at a fixed frequency, or both. Numerous advantages of such a signal source will become immediately apparent to applications and systems engineers.

This Engineering Note briefly discusses those properties of the traveling-wave tube which make it suitable for incorporation in such a constant power system. The external circuitry required for a typical CP system and the results of using the system in several applications are also described.

The method of maintaining a constant power output described herein does not rely on the limiting characteristics of the TWT as discussed in Engineering Note No. 6. In Note No. 6 it was explained how two traveling-wave tubes in cascade could provide a relatively constant (saturated) output power over a large dynamic input range. This Note, however, analyzes a system which operates in the linear (small-signal) region of the TWT's transfer characteristics. Compared to limiter operation, small-signal operation permits power output control over a wider frequency range with a more constant output and less phase distortion.

To achieve small-signal power control it is essential that the traveling-wave tube be capable of being amplitude modulated. That is, the tube must have a beam-control electrode – preferably in the cathode plane. We will call this electrode the grid. (See Engineering Note No. 3).

TWT OPERATING DATA

Consider the transfer characteristics in Figure 1. These curves were obtained from a typical low-level S-band amplifier, the Huggins HA–1, which provides 10 mw minimum saturated power output (Vg=0) with 30 db minimum small-signal gain over the 2.0 to 4.0 kmc band. The data were taken at a frequency of 3.0 kmc, and similar curves are experienced at all other frequencies in the band of operation.

For the purpose of illustration, assume that it is required that the output be unsaturated. This limits operation to the shaded area in Figure 1. The boundries are determined by the curve for zero bias, input noise level, and the point at which the tube ceases to be linear (taken to be 6 db lower than the saturated power output point).

The input noise level is determined by noise figure and noise bandwidth (See Engineering Note No. 6). For the HA–1, a noise figure of 26 db

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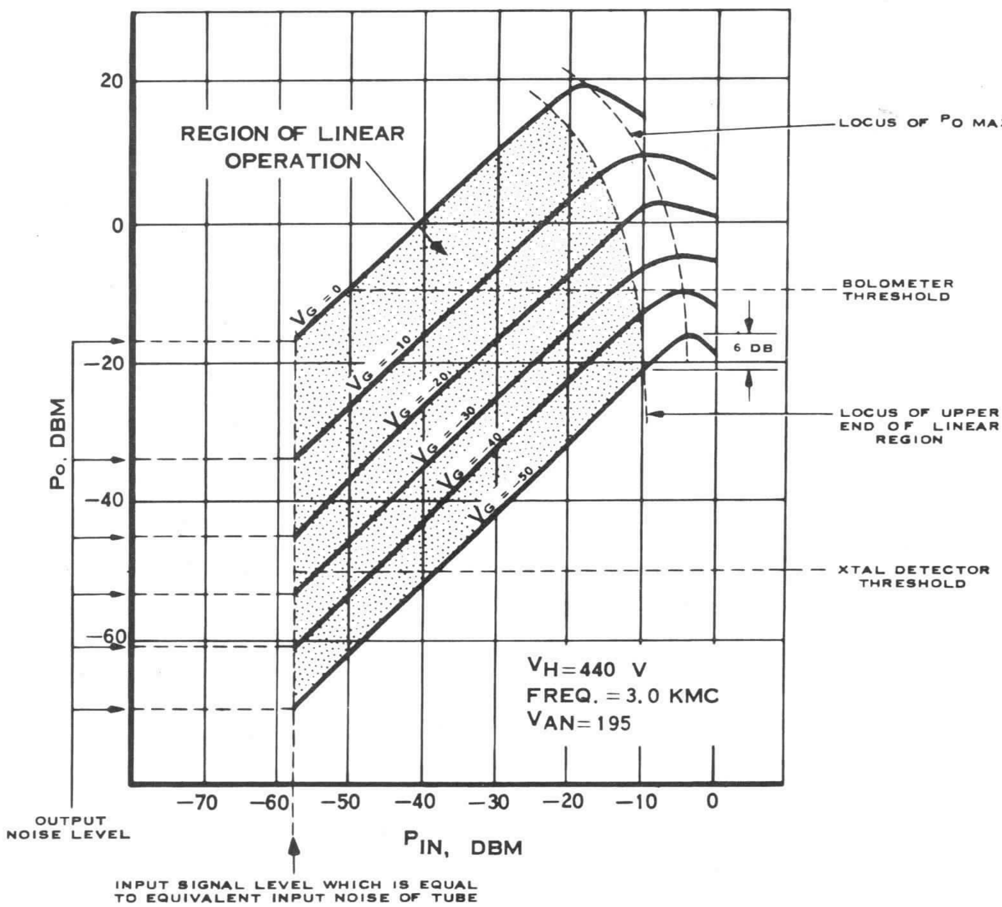

Figure 1. P_O vs P_{IN} for an HA-1 TWT Amplifier With Grid Voltage as a Parameter.

and noise bandwidth of(1 kmc)results in:

$$
\begin{aligned}
KTB &= -\ 114\ \text{dbm/mc} \\
NF &= \underline{\quad 26\ \text{db}\quad} \\
&\quad -\ 88\ \text{dbm/mc} \\
\\
\text{Bandwidth} &= \underline{\quad 30\ \text{db (1kmc)}} \\
\text{Input noise level} &= -\ 58\ \text{dbm}
\end{aligned}
$$

This is the magnitude of the input signal which equals the equivalent input noise, the level of which can be varied only by changing the noise figure or noise bandwidth. In Figure 1 we have assumed that the input noise level is independent of grid voltage. Since noise figure is dependent on beam current, this assumption is not strictly true. The change in noise figure with beam current, however, is not significant enough to be included for the purpose of our discussion.

Close examination of the transfer characteristics reveals that the higher the power output required, the smaller the allowable input power range becomes. For example, the input power can vary from -40 dbm to -14 dbm (a dynamic range of 26 db) for a grid voltage change from 0 to -20 volts if an output power of 0 dbm is required; for a constant power output of -10 dbm, input power can vary from -50 dbm to -12 dbm (a dynamic range of 38 db) for a grid voltage change from -1 to -32 volts.

EXTERNAL CIRCUITRY

From the preceding discussion we see that for all values of power input within the ranges quoted there exists a corresponding grid voltage which gives a constant power output. A system which will provide this operation appears in Figure 2. The system consists of a signal source, varying in both frequency and power output, which drives a traveling-wave tube amplifier through a variable attenuator of fixed value. A portion of the TWT's output is sampled by means of a directional coupler and detected by some convenient means, be it a crystal or bolometer. This detected output is then amplified, compared to a reference, and applied to the traveling-wave tube grid which in turn controls the power output. The choice of the method for sampling, the detecting element, and the feedback unit are dictated by system requirements.

The system in Figure 2 holds the output voltage at Point A constant. The constancy of power output at point B as a function of frequency depends upon the frequence dependence of the directional coupler and the detector characteristics. A recorded plot of the output at point B as a function of frequency is shown in Figure 3. The signal source is a Huggins HO-2 X-band backward wave oscillator, and the TWT, a Huggins HA-4 standard 10 mw X-band amplifier. It is seen that the output is constant within ± 0.7 db from 8.2 to 12.5 kmc. Monitoring is accomplished with a power bridge, thermistor, and recorder.

In making impedance, gain, and other measurements at microwave frequencies, it is often desirable to square wave modulate the radio frequency source. Modulation of the signal source is not always convenient, but a constant power system which uses a traveling-wave tube can be easily square wave modulated.

A directional coupler is the preferred method of sampling. The required microwave equipment would then be a signal source, an attenuator, a gridded TWT amplifier, a directional coupler, and a detector. A standard 1 mw signal generator or a backward wave oscillator which are generally available to cover a 1.5:1 to 2:1 bandwidth can be used as a signal source. Both of these signal sources have a power variation of 7 to 10 db over their operating bands. The 1 mw generators are usually packaged to include a calibrated attenuator.

Whether or not a crystal or boltmeter detector is used is determined by both the operating frequency band and the response of the detecting element. It is possible to have a crystal detector sensitivity of approximately −50 dbm. For low level outputs the secondary arm of the directional coupler may be used for power output, and the primary arm may be used for sampling.

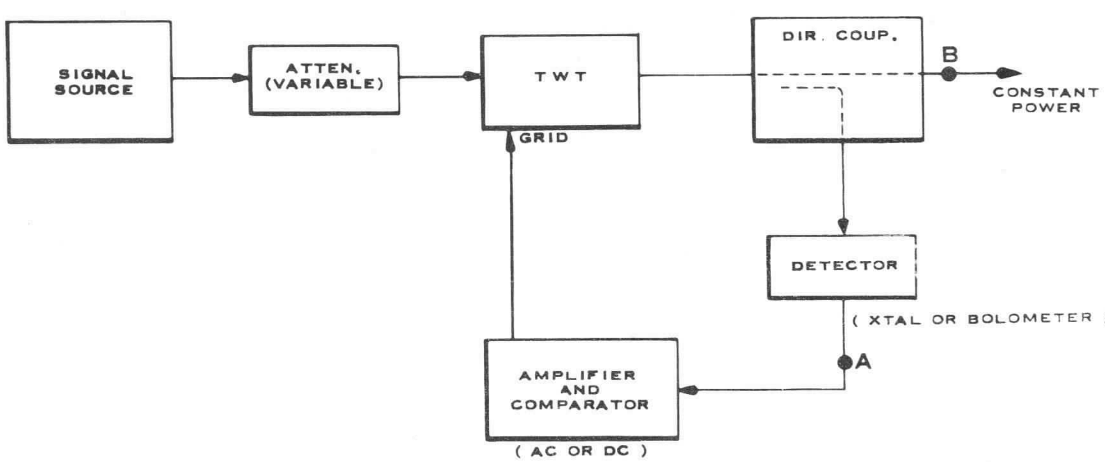

Figure 2. Constant Power System Block Diagram.

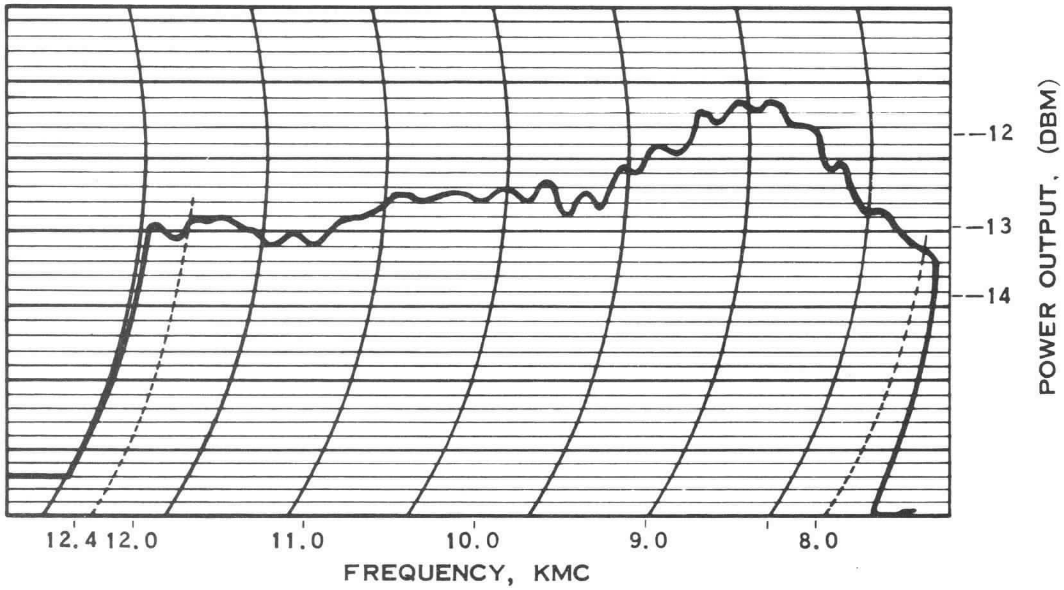

Figure 3. Output of the Constant Power System in Figure 2.

The details of the amplifier and comparator circuitry are not necessary to this discussion. An ac amplifier tuned to 1 kc is used in the feedback loop at this laboratory. This leads to a stable feedback system which can accurately hold rf power levels to very low values. It requires, however, that the rf signal which is fed into the TWT amplifier and feedback system be square wave modulated at a 1 kc rate. The tuned amplifier uses only the fundamental component of the detected square wave signal to derive the dc voltage which is applied to the grid of the traveling-wave tube.

The use of an ac feedback system has the advantage that a bolometer can be used as the detecting element as well as a crystal for signal levels as low as –30 to –40 dbm. It is our experience that certain commercially available coaxial bolometers have extremely flat broadband frequency characteristics as well as an accurate square law response. This leads to accurate control of constant power levels over octave frequency ranges or greater.

If it is desired to apply the feedback system to cw signals, a crystal must be used for the low level detector since a bolometer has serious temperature drift problems at low levels. Also, very stable, low drift dc amplification must be used in the feedback amplifier. The crystal detector does not, in general, have the extremely flat frequency characteristic of the bolometer.

The level of the regulated rf power out of the system is controlled by the amount of gain in the tuned amplifier.

The control voltage applied to the grid of the leveling amplifier is proportional to the logarithm of the TWT's output. This is due to the near-linear relationship between the relative output power of the TWT (in db) and the grid voltage, particularly at lower values of grid voltage.

RECORDING TWT GAIN

Figure 4 shows a constant power system constructed around the Huggins HA–4 X-band TWT amplifier. This amplifier provides a minimum power output of 10 mw and a minimum small-signal gain of 30 db over the 8.2 to 12.4 kmc band. It has an aperture grid in the plane of the cathode which offers the possibility of controlling gain or power output in a manner similar to that of the HA–1 whose characteristics were discussed in conjunction with Figure 1.

The system diagrammed in Figure 4 allows for the recording of small-signal gain as a function of frequency of any amplifier operating over the 8.2 to 12.4 kmc band. A signal source whose frequency can be swept with time is conveniently obtained in an oscillator such as the Huggins HO–2 backward wave oscillator. The BWO is voltage tunable over octave bandwidths, and the Huggins HO–2 oscillator provides a minimum of 10 dbm output over the 7.6 to 13.7 kmc band.

If a backward wave oscillator is used as the signal source, modulation is accomplished by applying a square wave voltage to the anode of the BWO. Power supplies that provide a helix voltage which varies expotentially with time are available for operating the BWO. Due to the logarithmic relationship between frequency and helix voltage, this exponential characteristic of the helix voltage allows for a signal whose frequency varies linearly with time. Such a signal is convenient in interpreting recorded or other data.

A variable attenuator following the BWO is used to regulate the rf drive to the HA–4 leveling amplifier. Since the output from the HO–2 is sufficient to considerably overdrive the HA–4, and therefore increase the complexity of power control, the attenuator reduces the drive to the small-signal input level.

Power out from the HA-4 is sampled by means of a directional coupler and then detected. After amplification and comparison, a dc control voltage is derived and applied to the HA-4's grid to maintain a constant output from the leveling amplifier. The leveled power drives the TWT under test through a precision-calibrated wideband attenuator. This attenuator provides still another means of varying the level of the constant power drive.

The output of the amplifier under test is monitored through another directional coupler, detected, passed through a log amplifier, and recorded by whatever means available. Detecting element B should have as nearly as possible the same response with respect to frequency as detecting element A so that the output-input power-ratio of the tube under test can be measured accurately. These crystal detectors are commer-

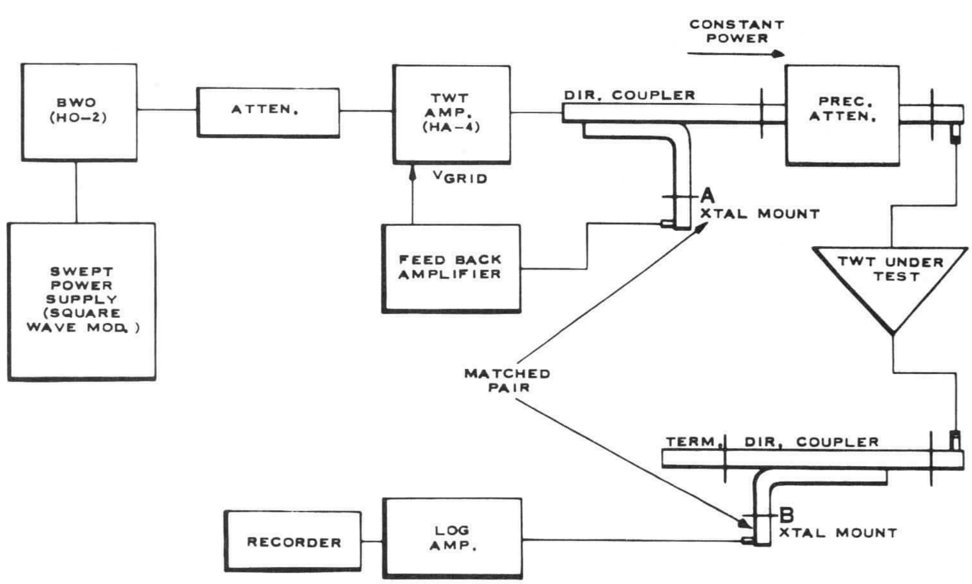

Figure 4. Use of Constant Power System for Recording Gain of TWT Amplifier.

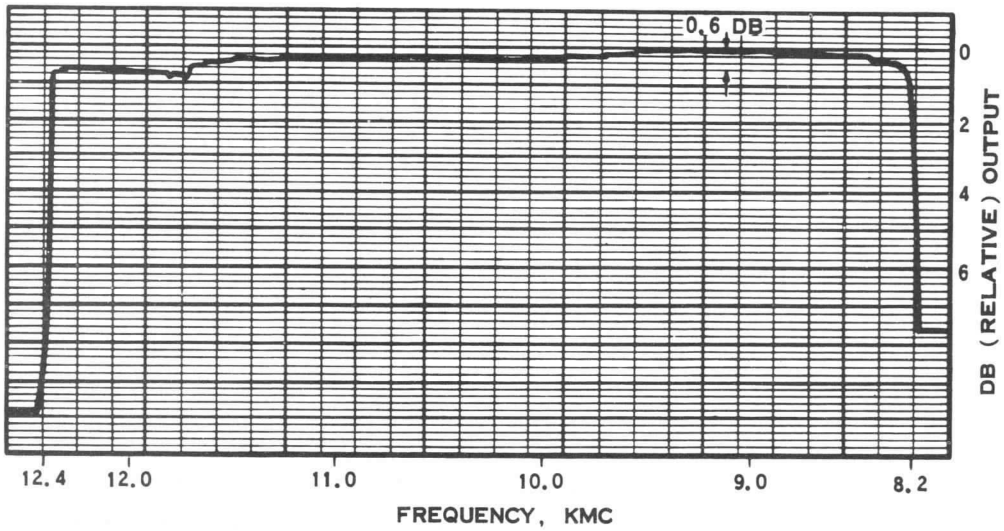

Figure 5. Output of Overall System of Figure 4 Minus TWT Under Test.

cially available as matched pairs, and the relative response is better than 1 db over the band of interest and over a wide range of power levels.

The logarithmic amplifier allows the gain of the TWT under test to be recorded in terms of db units. Zero db reference on the recorder plot can be simply obtained by bypassing the amplifier under test with a short length of coax cable. Calibration of the plot in absolute db units is also simply accomplished by properly changing the precision attenuator setting with the TWT under test in and out of the system.

In Figure 5 we see a recorded plot of output power vs frequency as observed at point B in Figure 4 when the TWT under test is bypassed. It will be noted that the signal is constant to within ± 0.3 db. Compared to the leveling achieved with the system of Figure 2 in which the quency dependence of the directional coupler and crystal detector produced a considerably wider variation in the output (See Figure 3), ± 0.3 db represents a great improvement.

Figure 5 also demonstrates the affect on leveling of using matched crystal pairs. Matched

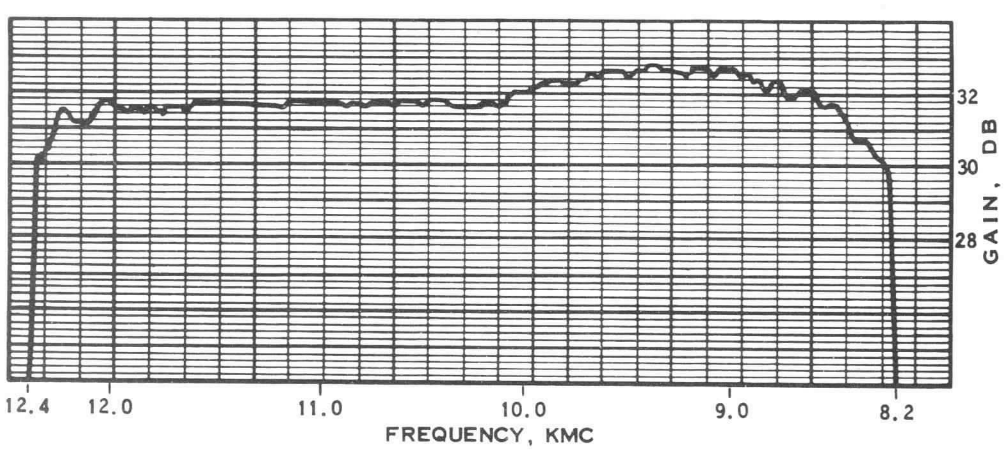

Figure 6. Recorded Plot of Small-Signal Gain vs Frequency of a Modified Huggins HA-4 TWT Amplifier.

crystal pairs are mandatory in systems in which a power ratio is desired to be accurately monitored as a function of frequency. Commercially available directional couplers are well matched across X-band, and no selection is necessary to obtain nearly identical characteristics.

The curve in Figure 6 is a recorded plot of the small-signal gain vs frequency characteristic of a modified Huggins HA–4 as measured with the system in Figure 4. In addition to showing an example of the results obtained by using a CP system to collect laboratory data, Figure 6 illustrates the extent to which the small-signal gain response of the TWT amplifier may be controlled during manufacture.

Here, modification of the standard X-band low level amplifier resulted in a gain characteristic which is within 0.75 db of an average value of 32 db over the 8.2 to 12.4 kmc band. Over the 8.2 to 12.0 kmc band the gain is flat to within 0.5 db. This type of response is possible with no operating adjustments; that is, the curve in Figure 6 was recorded with all TWT potentials and currents fixed.

Other types of gain responses are also possible. For example, TWT amplifiers can be manufactured to provide small-signal gain which varies at a fixed rate over certain specified frequency bands. A great deal of control is possible during manufacture, and TWTs having special gain vs frequency characteristics are commercially available in a wide range of frequency bands.

MEASUREMENT OF REFLECTION COEFFICIENT

Another example of the use of a constant power system is given in Figure 7. This system continuously monitors reflection coefficient as a function of frequency. The example chosen is again at X-band, but the measurements can be performed in other bands providing components are available.

The constant output from the HA-4 is used as the incident power in investigating the reflection coefficient vs frequency characteristic of the test load. With incident power constant, monitoring reflected power gives a measure of the magnitude of the load impedance. Depending upon the mon-

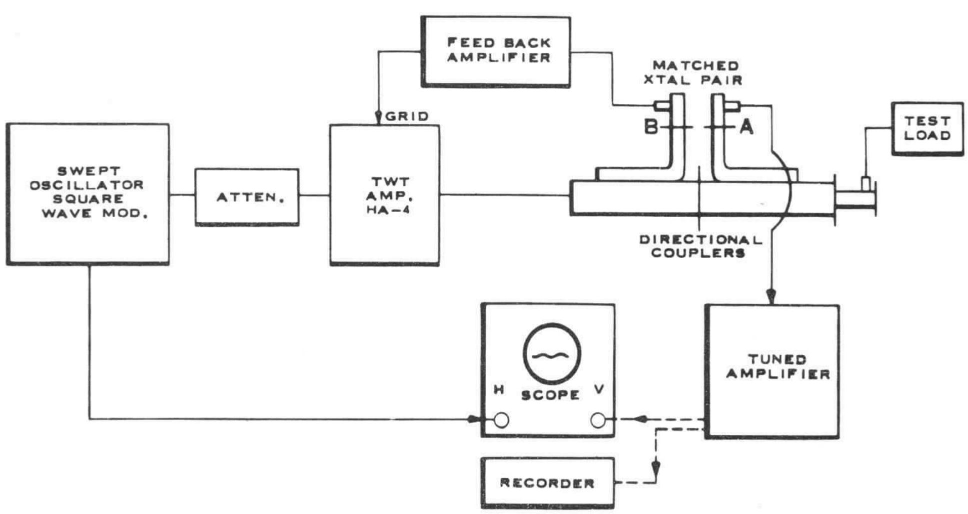

Figure 7. Use of Constant Power System for Displaying Reflection Coefficient.

itoring system used, the absolute load impedance can be measured by suitable calibration. The monitor may be a pen recorder, an oscilloscope, or a tuned indicating voltmeter consistent with the square wave modulating frequency used. Calibration, if required, can be accomplished with the use of short circuits and standard mismatches.

The advantages of a system which continuously monitors impedance magnitude as a function of frequency are obvious to anyone who has performed such measurements by slotted line techniques. This system is particularly advantageous in applications where broadband impedance matches are desired.

EXTENDED-RANGE CP SIGNAL SOURCE

Frequency extension of a given signal generator is frequently desirable to enable further wideband testing of system components. Such an extension becomes more attractive if a constant power output can be maintained.

Extending the frequency range of a microwave signal generator and holding a constant output is possible with the system described in Figure 2. We can attain this goal because the TWT amplifier is inherently capable of functioning as both a frequency multiplier[1] and a power leveler.

The Huggins HA-16 S-band to X-band frequency multiplier is an example of a commercially available TWT frequency multiplier. Operating on the 5th harmonic, the HA-16 provides a 9 kmc output with a 1.8 kmc input at a conversion gain between 0 and -10 db. Similarly, the Huggins HA-34 provides an output signal in the 2.0 to 4.0 kmc band with an input in the 400 to 1000 mc frequency range. Conversion gains of 10 db are achieved in the HA-34 when operating at the 5th harmonic.

To operate as a harmonic generator, the traveling-wave tube must be driven into saturation. This means that the variable attenuator between the signal source and the TWT leveling ampli-

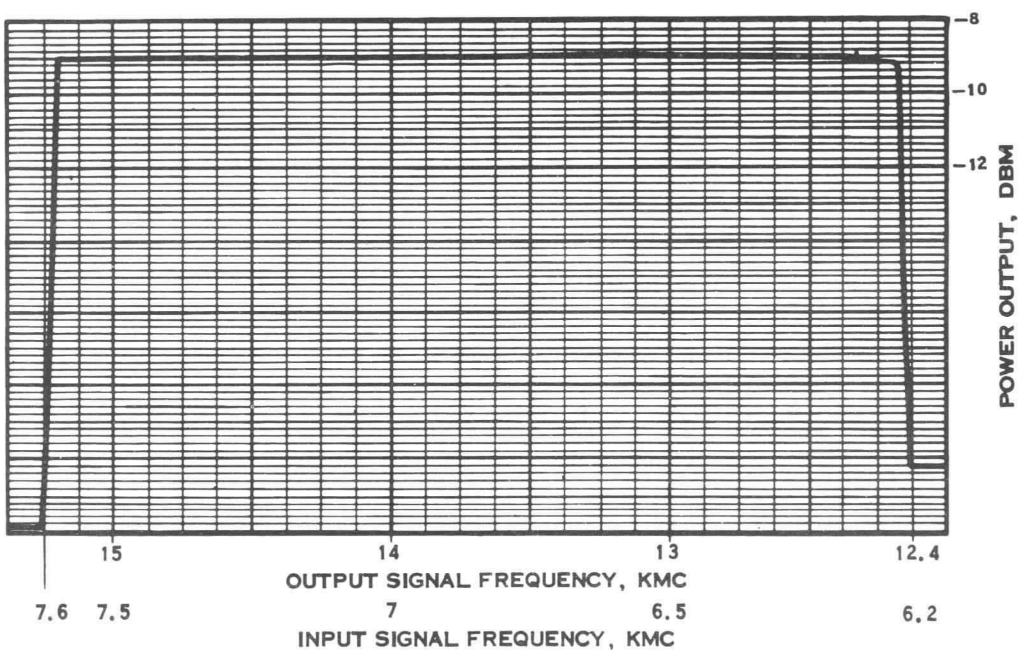

Figure 8. Response of second harmonic constant output system as observed with a matched crystal pair, with Huggins HS-24 TWT amplifier used as harmonic generator.

fier in Figure 2 must be set so as to provide saturation drive – approximately –5 to –20 dbm (See Figure 1).

Harmonic operation, of course, can be ensured only by proper filtering at the output of the leveling amplifier. This is fairly simple in waveguide systems since the filter must sufficiently attenuate the primary signal frequency and pass the harmonic of interest. Harmonic constant power systems, therefore, are most easily fabricated at frequency ranges in which the necessary waveguide components are available.

The recorded response of a constant output harmonic generator is presented in Figure 8. This curve was recorded using matched crystal detectors in a manner identical to that used to obtain the curve in Figure 5.

In the extended range system, as indicated in Figure 8, the input frequency varied from 6.2 to 7.6 kmc. The 2nd harmonic output covers the 12.4 to 15.2 kmc band at an essentially constant power level of –9 dbm. In this application, an 8.2 to 12.4 kmc directional coupler was used to ensure that the upper primary frequency of 7.6 kmc would be attenuated before reaching the feedback chain.

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[1]D. J. Bates and E. L. Ginzton, "A Traveling Wave Frequency Multiplier," Proceedings of the IRE, Vol. 45, July 1957, pp. 938-944.

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ENGINEERING NOTES

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POWER SUPPLY REQUIREMENTS FOR TRAVELING-WAVE TUBE AMPLIFIERS

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Satisfactory performance of traveling-wave tube amplifiers is largely dependent upon the power supplies used and the associated metering of currents necessary to insure proper operation.

Therefore, this Engineering Note deals with: 1) the general nature of a TWT amplifier power supply, 2) regulation requirements, 3) overload protection, and 4) the important power supply considerations for a few commonly used circuits.

BASIC POWER SUPPLY CONSIDERATIONS

A traveling-wave tube amplifier, such as that shown in Figure 1, generally requires three dc sources for supplying electrode potentials and an ac source for furnishing filament voltage. Also, some means of monitoring the helix voltage and the beam, helix and collector currents should be provided. The helix voltage measurement is used to insure synchronism, the helix current is used to indicate proper focusing and helix voltage, and the collector and beam currents are useful in determining TWT transmission. Because low and medium power tubes vary in their construction, however, power-supply circuitry needs will necessarily be modified.

Low Power Output (10 mw order of magnitude). Low power tubes normally use an aperture-type grid for beam-forming purposes, and these tubes can be made in either one of two ways: 1) the grid may be connected to the cathode internally, in which case the grid supply is eliminated, or 2) the grid may be brought out on a separate connection to permit beam control by operating the grid negative with respect to the cathode. Specified rf performance is attained when the grid and cathode are operated at the same potential.

The anode is also an aperture-type electrode which is used for beam controlling purposes in that its potential determines the amount of beam current. Except in unusual cases, the anode supply must be adjustable from zero volts for initial focusing purposes.

Normally the helix and collector are operated at the same potential, but the manufacturer may recommend that certain tube types be operated with the collector slightly more positive than the helix to improve performance (e.g., noise figure).

Medium and High Power Output (1 w and above). Medium and high power tubes are more effectively controlled with an interception-type grid which is operated positive with respect to the cathode to obtain specified rf performance.

Like the low power tubes, the anode supply for high power TWTs should be adjustable for initial focusing purposes.

Medium and high power tubes are normally operated with the collector potential above the helix.

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Tube Construction and Power Connection Considerations. The cathode of low level amplifiers (beam powers of three watts or less) can be operated at ground or capsule potential, for insulating the collector from the tube capsule is a relatively simple matter. Operation of the cathode at ground is often convenient for modulating purposes.

On the other hand, amplifiers providing 1 watt or more rf power—with their attendant high beam power—are not so easily manufactured with the collector insulated from the capsule. Consequently, these tubes are usually constructed with the collector and capsule in intimate contact for improved heat dissipation characteristics; especially is this applicable to high frequency tubes where the tube components are small.

REGULATION REQUIREMENTS

Regulation of the power supplies is used to improve both the phase (frequency) and amplitude stability of the TWTs output signal, whether it be an oscillator or amplifier. The degree of regulation to be employed is dictated by the function the tube is to serve and the tolerances allowed in performance.

Measurements made at Huggins Laboratories have resulted in the following data which may be used to obtain an order of magnitude approximation of the effect which a given amount of ripple on the supply voltages has on the amplitude and phase stability of a TWT amplifier's output signal.

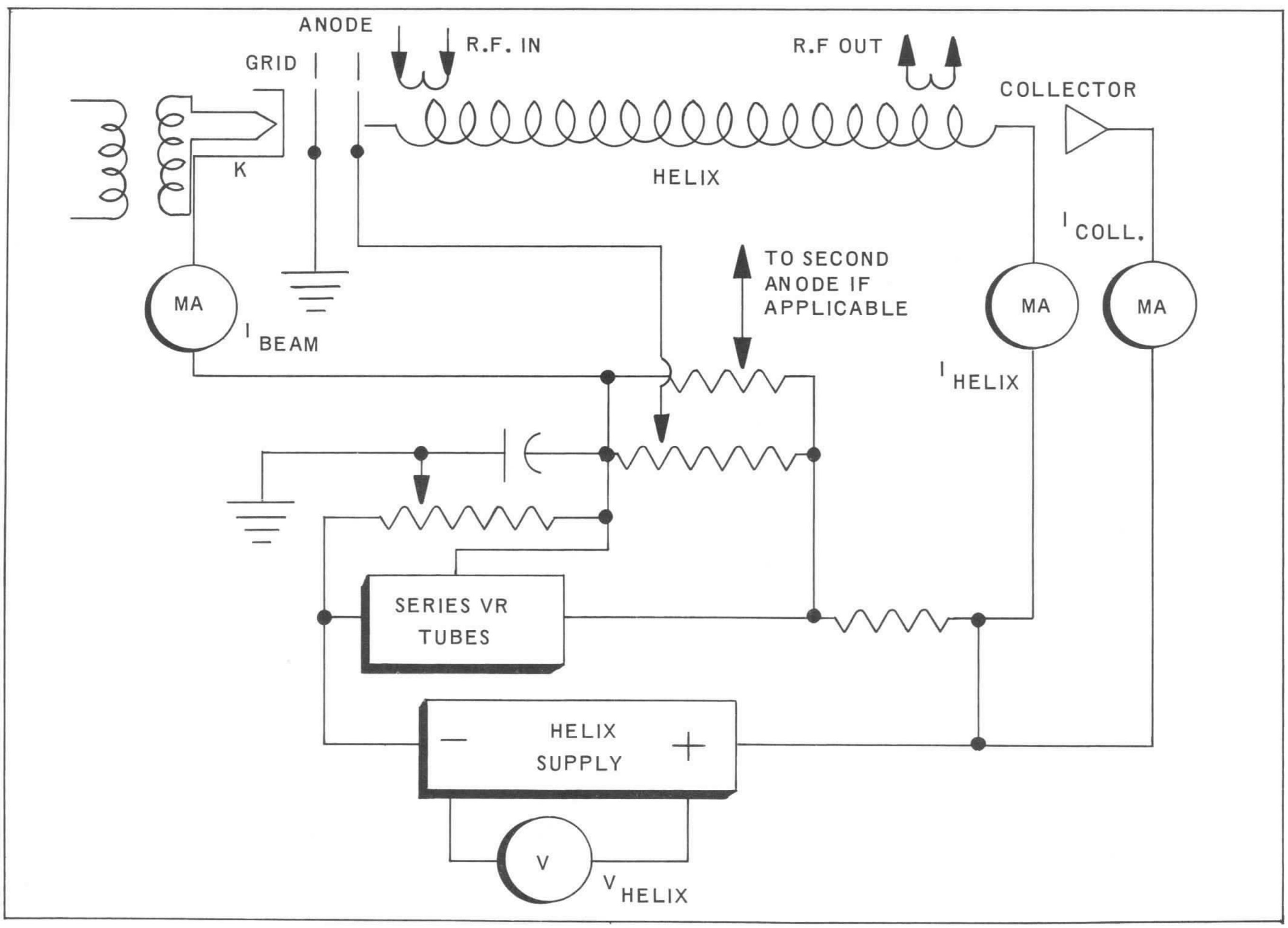

Figure 1. Basic supply and metering requirements for a TWT amplifier.

Helix Supply. Where regulation is concerned, the helix voltage is the most critical of all electrode potentials. A peak-to-peak ripple of less than 0.01% will result in less than 1° peak-to-peak phase excursion; a peak-to-peak ripple of less than 0.10% will produce amplitude modulation of less than 0.02 db; regulation of better than 0.50% will cause a change of less than 0.10 db in power output.

Anode Supply. A peak-to-peak ripple of less than 1% on the anode voltage will result in less than a 1° peak-to-peak phase excursion; a peak-to-peak ripple of less than 0.10% is required for less than 0.03 db AM; better than 1% regulation is necessary for a negligible change in power output.

Grid Supply. The grid supply should be regulated to at least 1% for negligible phase excursions, amplitude modulation, and power output changes. This is particularly true in the medium and high power tubes where the grid is operated positive with respect to the cathode.

Collector Supply. When operated from the regulated helix supply, the collector will have virtually no effect upon power output or the phase and amplitude of the output signals. In medium and high power TWTs, the collector is normally maintained a few hundred volts positive with respect to the helix. A ripple of 3% on the collector will not impair tube performance in such cases as long as the collector potential does not go below the helix during the ripple cycle.

Solenoid Supply (magnetic field). Providing the TWT is not misaligned in the field, or the focusing has not deteriorated to give appreciable helix current, ripple – within reason – in the solenoid current is of little importance. If helix interception takes place, however, phase excursions on the order of 50° per milliampere change in helix current will occur.

OVERLOAD PROTECTION

The primary cause of TWT failure is excessive helix current: Overheating of the helix and its support structure may result in a mechanical failure, and bombardment of the cathode by the positive ions released through out-gassing of the helix and its support structure may cause cathode "poisoning". Therefore, helix current is most conveniently used to indicate and control an overload.

Except for the HA-9, 21, 35, and the PA-1, removal of the high voltages is sufficient to prevent damage from an overload. One of the better ways to remove the high voltages is shown in Figure 2. Relay B, in series with the helix, functions as a sensing device; when helix current goes above a safe value, Relay B closes. In closing, Relay B actuates Relay C which in turn removes the primary voltage from the transformers. Relay C also acts as a holding relay: high voltages are applied only after the main power switch, S_1, is opened and closed again.

In the HA-9, 21, 35 and the PA-1 tubes, the amount of beam current and the helix size dictate that the beam current be cut off within one second to avoid tube damage. Because the magnitude of the beam current in a TWT is determined by the accelerating anode potential, Huggins Laboratories recommends that a system which will short the anode to the cathode during an overload be used for protecting the aforementioned tubes.

Such a system would use a sensing and control relay in series with the helix, and this relay would be connected in such a manner that it would, when activated, immediately equalize the anode and cathode potentials and maintain this condition until a manual reset is operated. When again applied, the anode voltage should be increased from the cathode potential to the proper voltage.

Use of an overload relay in the helix return will protect TWT's from failures caused by: 1) misalignment of the tube in the magnetic focusing field, 2) loss of the magnetic field, and 3) loss of the collector potential, for in each case the beam current is drawn to the helix.

Should the helix and anode supplies be separate, a helix relay will not offer protection against a loss of helix voltage because most of the beam current is drawn to the anode. When the anode potential is derived from the helix supply, a loss of helix voltage will result in no beam current. This, in itself, may not prevent tube damage because the voltage may not fall rapidly enough.

By-passing Stray A-C Components. When using a dc relay in series with the helix, stray ac components may be introduced in series with the helix supply. Therefore, steps should be taken to heavily bypass the stray ac voltages (ac power supply frequency) that may be coupled into the relay's high impedance winding. Furthermore, if relatively high helix current is normal, the high relay impedance can nullify the regulation of the helix supply against load variations.

Use of fuses. Use of fuses as protective elements may lead to TWT failures – especially if the tube employs a high helix voltage. With the exception of special high-voltage types, low cur-

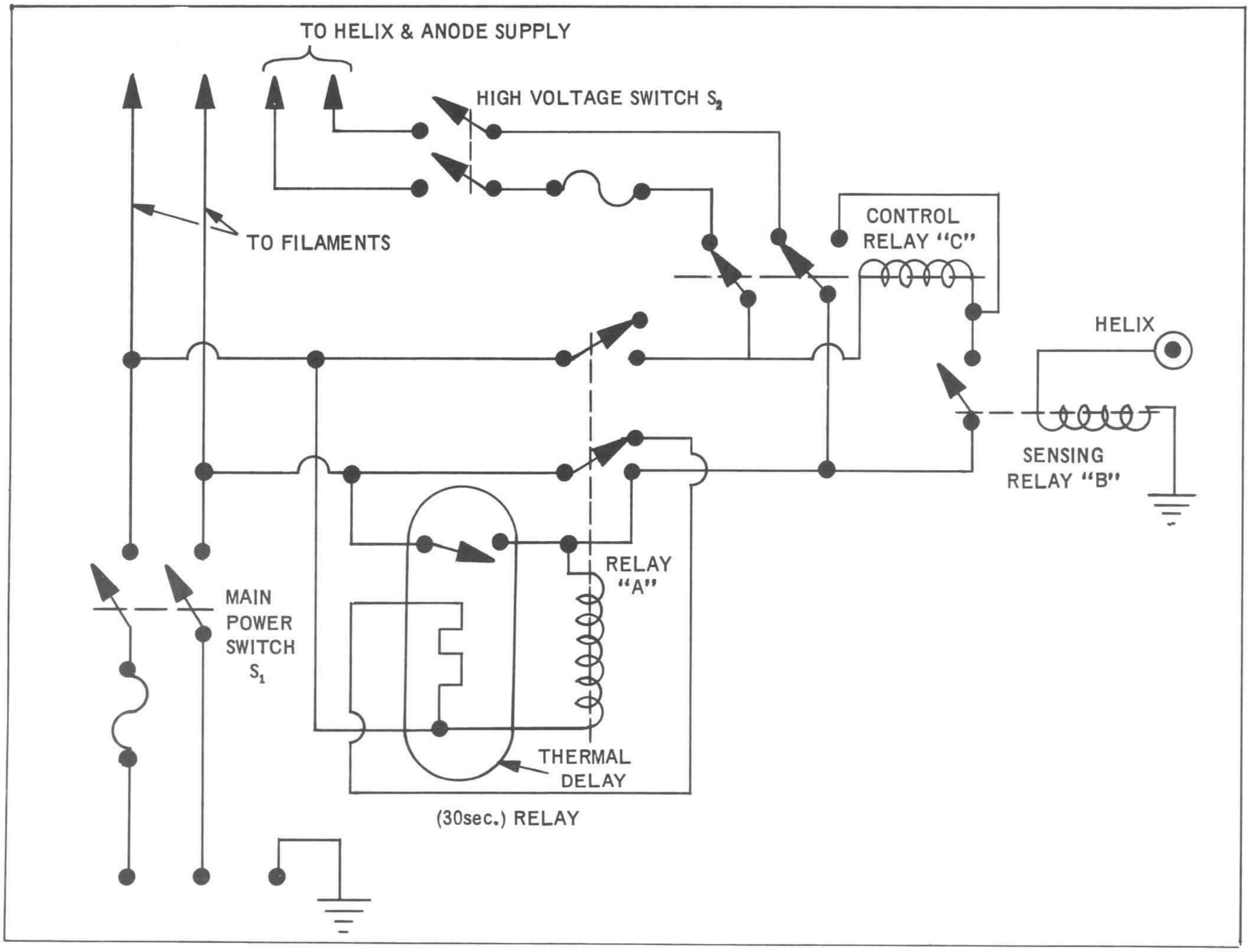

Figure 2. Overload protection system employing relays.

rent fuses which have opened will arc continuously at voltages in excess of 1500 volts, and the resultant voltage drop across the fuse is large enough to drastically change the helix voltage. This change will, of course, cause defocusing of the TWT.

Consequently, the SAFEST procedure is to provide means for reducing beam current to zero in the event of an overload. DO NOT USE FUSES.

SUPPLY AND METERING REQUIREMENTS FOR SPECIFIC APPLICATIONS

According to the TWT's circuit application, power supply considerations and metering requirements needed to insure proper operation will vary from the basic specifications discussed in the early parts of this Engineering Note. A few of the more important aspects of the modifications required for grounded collector, grounded grid, and phase modulation applications are discussed.

Grounded Collector Operation. Figure 3 describes a typical power supply and the metering arrangement for a traveling-wave tube that is operated with its collector grounded. Here, external grid control is assumed to be unnecessary, and the grid is shown directly connected to the cathode.

A single meter, I, can be used to monitor all currents necessary to maintain proper operation. In position 1 the meter is connected to read helix current only, and in position 2 the sum of the helix and collector currents is read. Because the gun electrodes draw negligible current, the read-

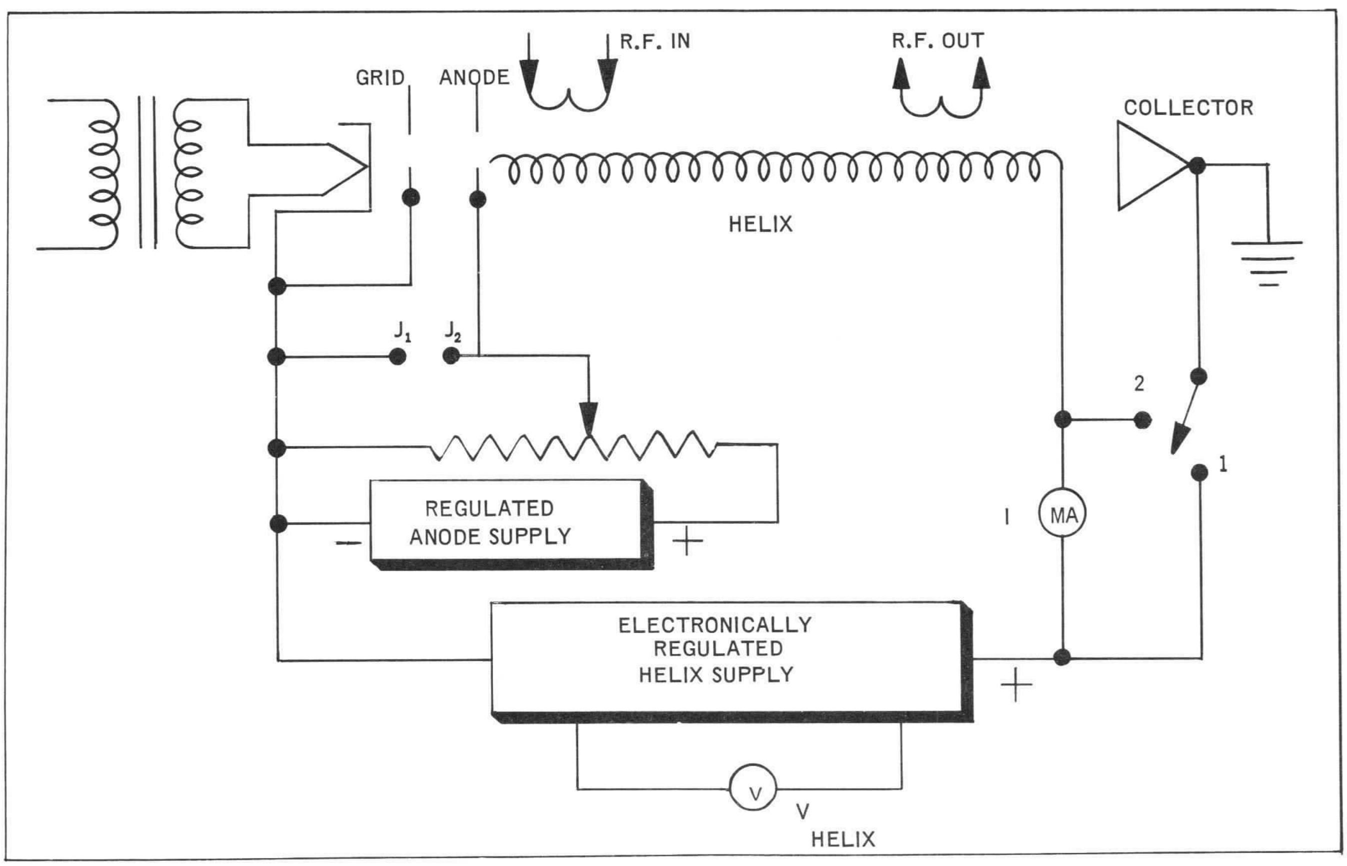

Figure 3. TWT Amplifier with grounded collector.

ing taken in position 2 also gives a sufficiently accurate indication of beam current.

This metering arrangement has two inherent disadvantages: The helix supply floats above ground by the amount of meter IR drop, and the meter switching is performed at a high potential off ground.

Monitoring the anode voltage is not important because its level is set to provide a specified beam current in the tube. Test jacks J_1 and J_2 are suggested, however, so that the anode potential can be checked with a high-impedance dc meter during the initial set-up procedure.

If the cathode and one side of the heater are internally connected, the precaution of connecting the dc return to the cathode should be taken in order to minimize the 60 cps component on the helix voltage caused by the common dc and ac junction.

Grounded Grid Operation. Grounded grid operation, Figure 4, is applicable in situations where rapid grid pulsing is important. Running the grid at ground potential permits connecting a terminating resistance directly across the grid driving source.

The main disadvantage of grounded grid operation is that the anode and helix supplies must float off ground by the amount of grid bias used, but where metering is concerned, this circuit has a definite advantage: the necessity of switching the meter at a high potential off ground is eliminated.

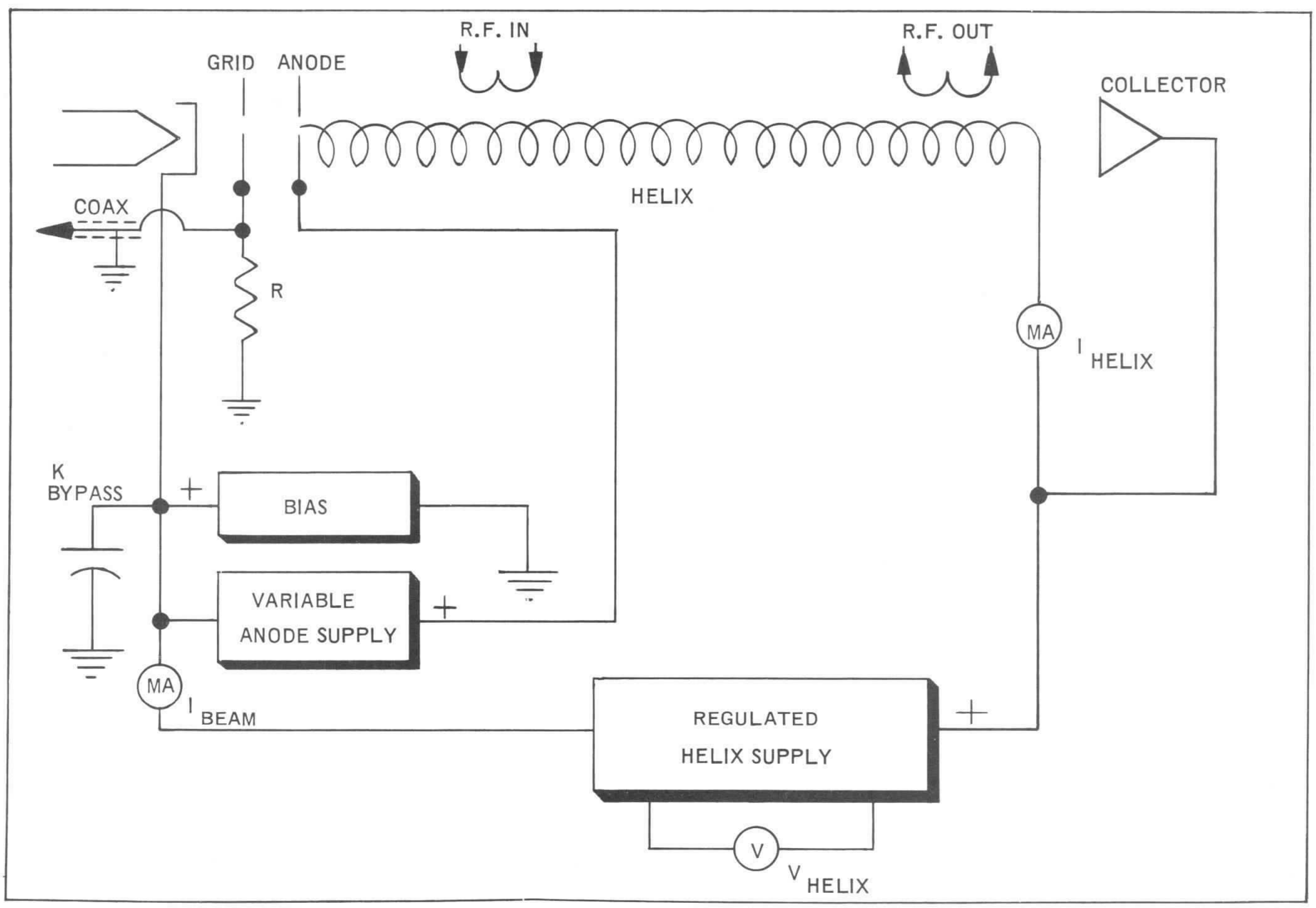

Figure 4. TWT Amplifier with grounded grid.

Anode Modulation. Because the grid has more sensitivity than the anode in controlling the electron beam, modulating the anode is not recommended for phase modulation.

Helix Modulation. Many applications of the TWT take advantage of the phase-modulation characteristics afforded by proper modulation of the helix (beam) voltage.[1] Phase modulation may be accomplished by modulating the beam current in some fashion (e.g., grid modulation), but the simplest method is to modulate the helix.

Modulating frequencies up to the order of 10 to 20 mc may be employed with relative simplicity. In these cases it is only necessary to insure that the collector potential never goes below the helix voltage during the modulation cycle. A possible method of obtaining helix modulation is shown schematically in Figure 5.

In high frequency serrodyne applications, where the helix length becomes a large enough fraction of the modulating wave length, helix modulation may cause standing waves to be set up on the helix. Consequently, it has been found most satisfactory to ground the helix and collector and to modulate the cathode by alternately charging and discharging the stray capacitance.[2]

SUMMARY

Quality in traveling-wave tube amplifier performance is primarily dependent upon the regulation characteristics of the power supplies used. Generally speaking, all voltages except the helix should have no more than 1% peak-to-peak ripple in order to maintain negligible amplitude modulation, phase excursions, and changes in power output. Ripple on the helix potential, however, should not exceed 0.01% to achieve stability between the input and output signals.

To maintain proper operation, some metering arrangement must be included to monitor the beam, helix, and collector currents. For protection, a helix relay system designed to remove all the high voltages, or short the anode to cathode in special cases, is recommended. Care must be taken to bypass stray ac voltages that may be coupled into the helix relay winding.

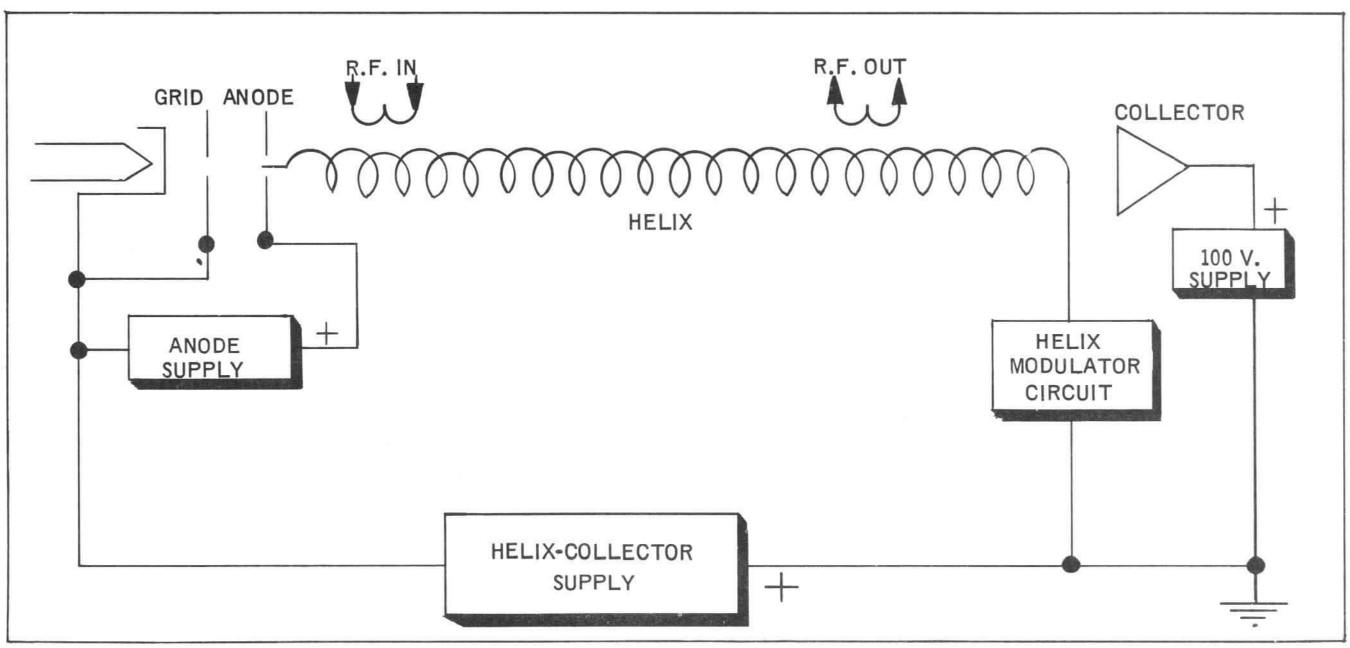

Figure 5. One method of modulating the helix.

In grounded collector operation, the helix supply must float above ground because of the metering arrangement, and grounded grid operation requires that the helix supply float above ground by the amount of bias employed. Helix modulation is relatively simple for modulating frequencies up to the order of 10 to 20 mc, the only precaution being to insure that the collector potential never goes below the helix during the modulation cycle. Above these frequencies it is best to operate the collector and helix at ground and modulate the cathode.

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[1]D. R. Bellis and R. A. Huggins, "Phase Modulation of Traveling-Wave Tubes," HUGGINS LABORATORIES Engineering Notes, Vol. 1 No. 7.

[2]Raymond C. Cumming, Frequency Translation by Modulation of Transit-Time Devices; Stanford University (Applied Electroncis Laboratory) Technical Report #39, August 1, 1955.

ADDENDUM

HUM CONSIDERATIONS

Should the heater and cathode not be connected internally, filament power supply modulation can be minimized by externally connecting the heater to the cathode. A convenient place to make this connection is at the dc mating connector.

Grounded Cathode Operation. Consistent with filament power supply modulation requirements, a heater-cathode connection is optional for grounded cathode operation.

Grounded Collector Operation. A typical grounded collector circuit is shown in Figure 6. Transformer T should have high voltage insulation, and it should also have an electrostatic shield to minimize capacitive coupling between the heater and ground. If a heater-cathode connection to minimize filament power supply modulation is necessary, caution should be taken to ensure that the cathode – not the heater – lead is returned to the negative terminal of the high voltage supply. This must be done to prevent modulation of the cathode by any IR drop across the filament lead. This latter effect is illustrated in Figure 7.

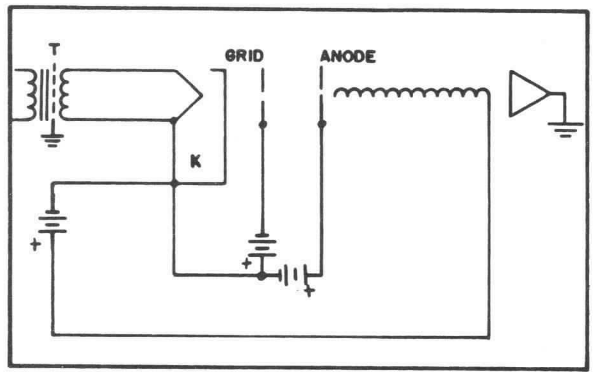

Figure 6. TWT amplifier with grounded collector.

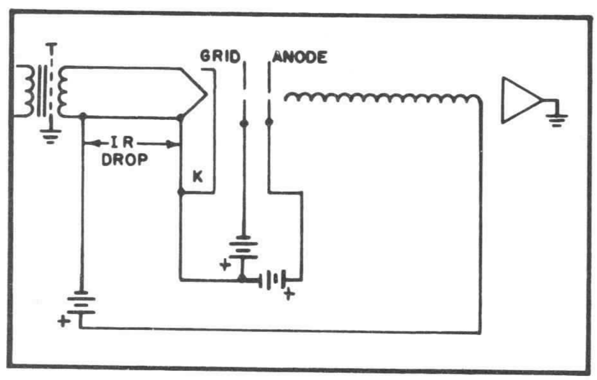

Figure 7. Cathode modulation by IR drop across filament lead.

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INTERMODULATION PHENOMENA IN TRAVELING-WAVE TUBES†

Edited By J. C. Stevenson*

The non-linearity of TWT gain characteristics can result in the generation of spurious signals when two or more legitimate signals are present. This Engineering Note describes the results of tests that were made on a Huggins' HA-2 TWT to obtain quantitative data concerning these effects. The solenoid-focused HA-2 TWT has specified performance of 1 watt minimum saturation output power over the 2.0 to 4.0 kmc band.

TEST PROCEDURE

The necessary data for determining the IM (intermodulation) characteristics of the HA-2 was obtained in the following way: Two S-band signal generators were connected through a 10 db coupler into the TWT. The TWT output was connected through a 30 db attenuator into the input of a panoramic receiver used as a spectrum analyzer. The detector output was directly coupled to a scope whose sweep was derived from the sweep circuits in the receiver.

The actual receiver frequency sweep was about 50 mc wide (sector sweep) centered on the IM component. A narrow sweep was used to prevent the two large input signals from appearing on the sweep and blocking the IF section of the receiver. The sensitivity of the receiver by itself was such that the TWT noise far exceeded the receiver front-end noise so that the TWT noise could be measured. Measurements made at various frequencies in the band showed the HA-2 to have a noise figure around 31 db.

The two signal generators were tuned about 40 mc apart and adjusted so as to present equal power levels into the receiver. These equal level signals were simultaneously increased until an IM product appeared in the receiver output. The level of detected IM was noted, and—using only one of the two signal generators—an equivalent output was obtained by suitably adjusting the level of the generator temporarily tuned to the IM frequency.

This process was continued, each time increasing the level of the two signal generators together. The amplitude of the detected IM product from the receiver was noted, and the equivalent input power level which produced the output was determined. This equivalent output power can be used to determine the corresponding level of input power if the gain is known. The gain of a TWT is not constant, however, for high level signals, but it is a function of the power level of these signals.

The gain of the TWT will be more in the case of substituting an equivalent input power representing the IM signal than results when the comparatively high level IM producing signals are actually present.

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†Manuscript submitted by William E. Budd, Melabs.

To determine the change in gain for different input power level signals, cross-modulation data was taken. This test was run by watching only the weaker of the two inputs while the stronger signal was varied in level. Each time the stronger signal level was raised, the level of the weaker signal had to be increased to maintain a constant receiver output. The amount of this increase, which actually represents a corresponding decrease in the weaker signal output from the TWT, was recorded as a function of the stronger signal level.

Figure 1 shows the results of these cross-modulation tests. The gain variation for different signal levels can now be found. This, as pointed out before, is necessary for calculating the equivalent IM input power when the IM output power is known. The corrected data can now be plotted in a way found desirable for many applications.

CORRECTED IM DATA

A compact representation of the way in which the IM products are related to the input power levels, as determined from the experimental data, can be constructed by normalizing the input power with respect to that power level which just causes saturation. This particular level of input power is defined as P_S. The IM power is conveniently normalized to the input level of the signals causing the IM so that a plot of (IM - P_i) vs (P_i - P_S) results. A description of these parameters follows:

P_i: The level of each of 2 equal power input signals separated slightly in frequency.

P_S: That level of input power which causes the TWT to produce maximum output power.

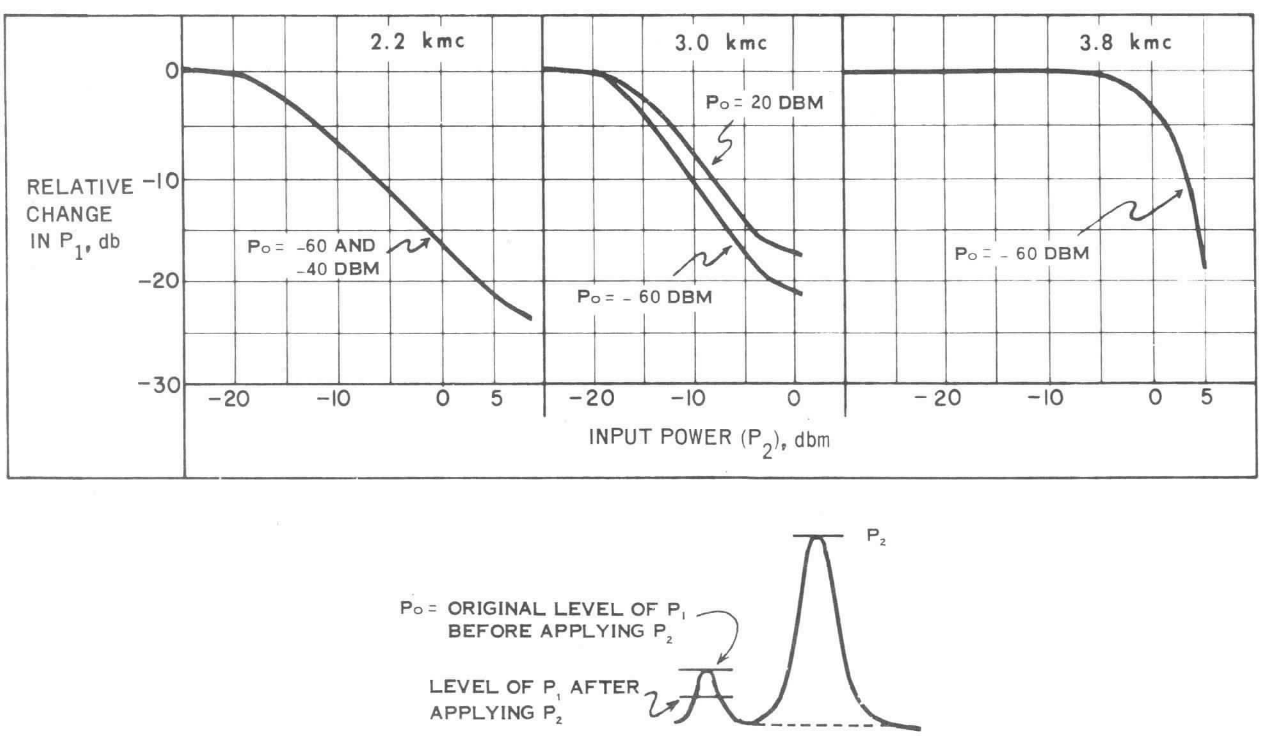

Figure 1. Cross-Modulation Characteristics.

IM: The power level of the resulting intermodulation.

Figure 2 shows such a curve based on the experimental data as well as a pictorial description of the parameters. [It has been experimentally determined at Huggins Laboratories that this curve is characteristic of all traveling-wave tubes – regardless of frequency or saturation power output. Figure 2 applies to only third-order modulation.] The curve is nearly linear for levels of $(P_i - P_s)$ less than - 10 db and can be expressed at lower levels as:

$$(IM - P_i) = 2(P_i - P_s)$$

[all units in db or dbm]

Since P_i is the level of each of the two equal power input signals which cause the IM products, the actual total power into the TWT is 3 db greater. In addition, unless the resulting IM products (both of them) are at least 10 db less than $(P_i + 3)$ db, they also contribute to the input power. This total input power is important in the accurate plotting of the IM data since it permits the calculation of the TWT gain from the cross-modulation curves in Figure 1.

USING FIGURE 2

Three examples will illustrate the use of the third-order IM curve in Figure 2.

Example 1. Suppose one wanted to know what level of input signal power would be allowable such that the IM products are at least 10.5 db below the signal. In this case $(IM - P_i)$ is -10.5 db. From the IM curve, $(P_i - P_s)$ is found to be about -3.5 db. Since P_s [at 2.2 and 3.0 kmc] is -8.5 dbm, P_i, the level of signal input, becomes -12 dbm.

It is interesting to examine the total input power for this case which was specifically set up in a particular way. Since both input signals are -12 dbm, then total power is -9 dbm. The IM products, taken together, probably raise the total power to -8.5 dbm which just equals P_s, the level of input power which causes saturation. Under these conditions then (total input power equal to P_s), either one of the first order IM products are about 10.5 db below the signal level causing them. The total output power is found from reference to the input-output curve in Figure 3. For -8.5 dbm input, the total output power is 32.5 dbm of which 32 dbm is due to the signals.

Example 2. It is desired to determine the level of third-order intermodulation products in a Huggins HA-31. The PPM-focused HA-31 has a specified minimum performance of 30 db small-signal gain and 10 dbm saturation power output over the 1.0 to 2.0 kmc band.

Assume that at 1.5 kmc the HA-31 exhibits 36 db small-signal gain (G) and 15 dbm saturation power output (P_o) with an input of -11 dbm (P_s). Further, an input of two signals which have a power level of -40 dbm (P_i) will be experienced.

Under these conditions $(P_i - P_s) = -29$ db, and from Figure 2, $(IM - P_i) = -59$ db. Using the latter equation we find that $IM = -59 + P_i$ which equals -99 dbm. That is, the intermodulation products are -99 db down referred to the input. At the output,

$$\begin{aligned} IM_{out} &= IM_{in} + G \\ &= -99 + 36 \\ &= -63 \text{ dbm} \end{aligned}$$

Since P_o at 1.5 kmc is 15 dbm, the intermodulation products at the output are 78 db below the fundamental.

Example 3. As our third example, assume that a two-stage preamplifier is to be designed. The first stage will be a TWT with 35 db small-signal gain (G_1), 10 dbm saturation power output (P_{o_1}), and 5 db noise figure. A 15 db gain (G_2),

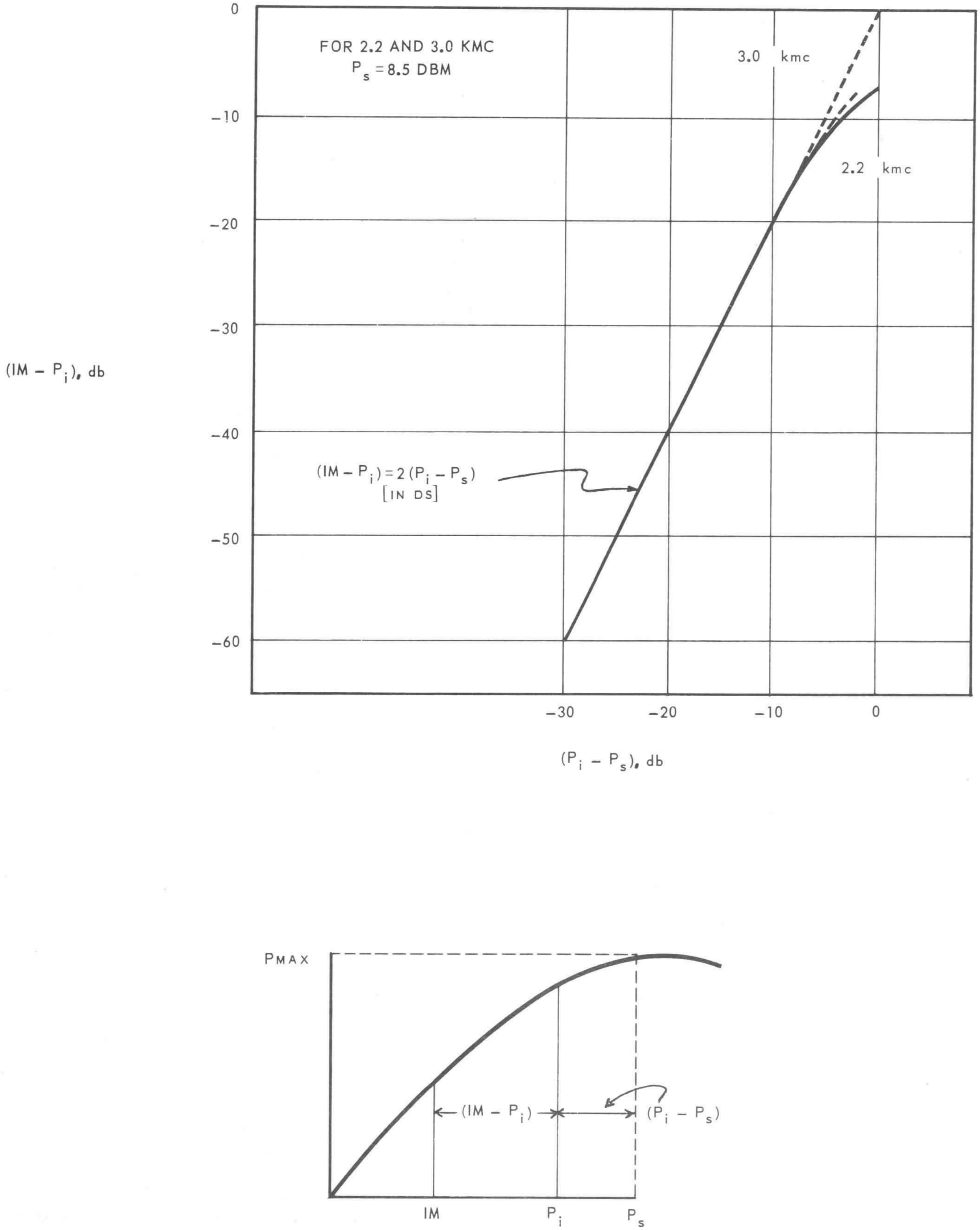

Figure 2. Third-order IM data for Traveling-Wave Tubes.

27 dbm output power (P_{o_2}) TWT with a 25 db noise figure will be used as the second stage. Intermodulation products are required to be at least -110 db down when referred to the input (i.e., IM_1 = -110 db) with an input of two signals which have a power level of -50 dbm (i.e., P_{i_1} = -50 dbm). We wish to determine if the proposed preamplifier will meet the IM requirement.

For the specified intermodulation requirements, $(IM_1 - P_{i_1}) = -110 + 50 = -60$ db, and from Figure 2 we see that $(P_{i_1} - P_{s_1}) = -30$ db. Hence,

$$\begin{aligned} P_{s_1} &= P_{i_1} + 30 \\ &= -50 + 30 \\ &= -20 \text{ db} \end{aligned}$$

Assuming that the first-stage saturation gain (G_{s_1}) is approximately 7 db lower than G_1, we have

$$P_{o_1} = G_{s_1} + P_{s_1} = 28 - 20 = 8 \text{ dbm}$$

This power output of 8 dbm is well within the capability of the low noise tube used in the first stage.

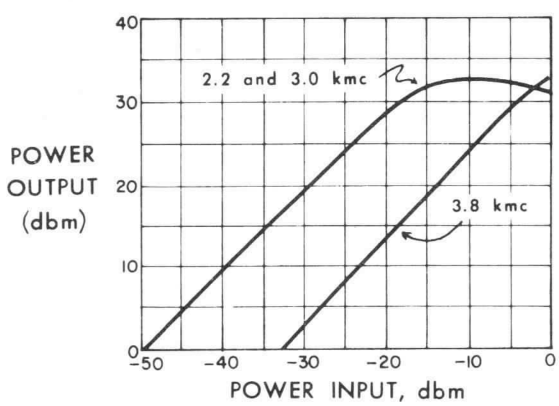

Figure 3. Input-Output Characteristics: Maximum Gain Conditions.

Since the input power to the second stage (P_{i_2}) equals the output power of the first stage,

$$\begin{aligned} P_{i_2} &= P_{i_1} + G_1 \\ &= -50 + 35 \\ &= -15 \text{ db} \end{aligned}$$

for the highest anticipated power level of two equal level input signals to the preamplifier.

Consistent with the IM specifications of -110 dbm, the intermodulation products at the output of the first stage are 60 db below P_{i_2} as determined from the equation $IM_{out_1} = IM_{in_1} + G_1$. To meet the IM specification, any intermodulation products generated in the second stage must be smaller than IM_{out_1}, say -81 db (i.e., IM_2 = -81 db). Thus, $IM_2 - P_{i_2} = -81 + 15 = -66$ db.

Following the same procedure as used for the first stage and the equation $(IM - P_i) = 2(P_i - P_s)$, we find

$$P_{i_2} - P_{s_2} = \frac{-66}{2} = -33 \text{ db}$$

Then,

$$P_{s_2} = P_{i_2} + 33 = -15 + 33 = 18 \text{ dbm}$$

and the saturation output power for the second stage is

$$\begin{aligned} P_{o_2} &= P_{s_2} + (G_2 - 7) \\ &= 18 + (15 - 7) \\ &= 26 \text{ dbm} \end{aligned}$$

Again, the power requirement of 26 dbm is within the capability of the TWT used in the second stage.

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PERIODIC PERMANENT MAGNET FOCUSED TRAVELING-WAVE TUBE AMPLIFIERS

Edited By J. C. Stevenson*

Over the past decade the versatile traveling-wave tube has enjoyed wide acceptance as a broadband microwave amplifier. And in recent years the TWT's versatility has been evidenced by its use in such applications as linear amplifiers, mixers, phase and amplitude modulators, frequency multipliers and dividers, limiters, and frequency-stable oscillators. But the advantage of versatility has often been offset by the method used to focus the electron beam.

With the perfection of periodic focusing techniques, however, TWT versatility has increased even more. This Engineering Note deals with focusing methods in general, and the more significant aspects of these methods are discussed.

After discussing general focusing methods, the periodic permanent magnet focusing structure is examined with respect to its advantages, field requirements, and gun and beam considerations. Construction of Huggins first PPM focused amplifiers is then described, and the characteristics of these tubes are compared to those of similar Huggins solenoid focused TWT amplifiers: power dissipation and weight, small-signal gain, saturation power output, and noise figure constitute this comparison. The effect of temperature on the focusing and gain of PPM amplifiers is then presented.

FOCUSING METHODS

Some method of focusing the electron beam in a traveling-wave tube is required because thermal velocities, rf fields, local stray fields, and induced image charges cause perturbations in the beam. More important, the ever-present space charge spreading of the beam must be counteracted.

Confined-Flow Focusing. "Confined-flow" focusing uses a strong axial magnetic field that threads the cathode and surrounds the electron beam over the length of the helix. The electrons enter parallel to the lines of the focusing field. Consequently, the electrons are forced to travel axially – with small deviations – around magnetic field lines.

The simplest means for generating an axial field is to use a solenoid, for solenoids can be built to provide, within limits, magnetic field regions of nearly any size and controlled uniformity. A plot of such a solenoid field is shown in Figure 1.

As can be observed in Figure 2, a straight-field permanent magnet could be used to generate an axial field. But a straight-field permanent magnet must also support external fields that are produced by the magnet; thus, for any increase in length there must be a corresponding increase in

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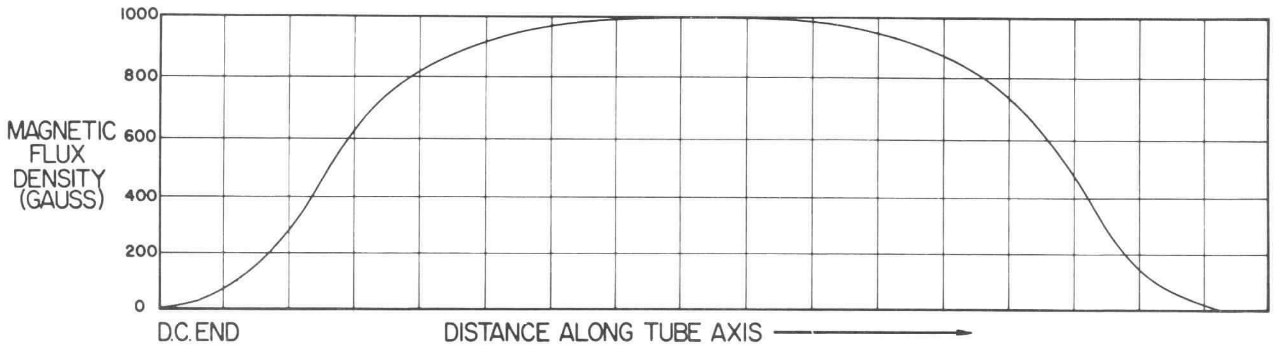

Figure 1. Plot of an Axial Field that is Generated by a Solenoid.

all transverse dimensions. This means that as tube length increases the volume and weight of the magnetic material needed to focus the beam increases as a function of the tube length cubed. Therefore, a straight-field permanent magnet is practical only when used with physically short tubes.

Though confined-flow focusing is a simple and effective method of controlling the path of the electron beam, it is relatively inefficient, the inefficiency being due to the requirement for an infinite field to keep electron deviation from the desired path at zero.[1]

In addition, the field producing device has the inherent disadvantage of weight, and in the case of solenoids, costly power is consumed and troublesome heat is generated. In trying to overcome these drawbacks, thought must be given to more efficient means of focusing, and such thought leads to the so-called "space-charge-balanced flow" method.

Space-Charge-Balanced Flow Focusing. The space-charge-balanced flow method is one in which the inward and outward forces that act on the electrons are balanced. One form of this type of focusing is known as Brillouin flow.

Brillouin flow utilizes an axial field that surrounds the electron beam over the length of the helix. Unlike confined-flow focusing, however, no flux threads the cathode. By using a convergent-flow electron gun, electrons are made to enter the focusing field at some angle to the field lines. The $\overline{B} \times \overline{V}$ force (where $\overline{B}$ = magnetic flux density and $\overline{V}$ = angular velocity of the electrons) causes the electrons to spin about a field line in a helical path toward the collector. The strength of the focusing field must be of such a value that the centrifugal forces of the spinning electrons and the $\overline{B} \times \overline{V}$ forces balance. Brillouin flow, as all space-charge-balanced systems, requires a smaller field than a confined-flow system.

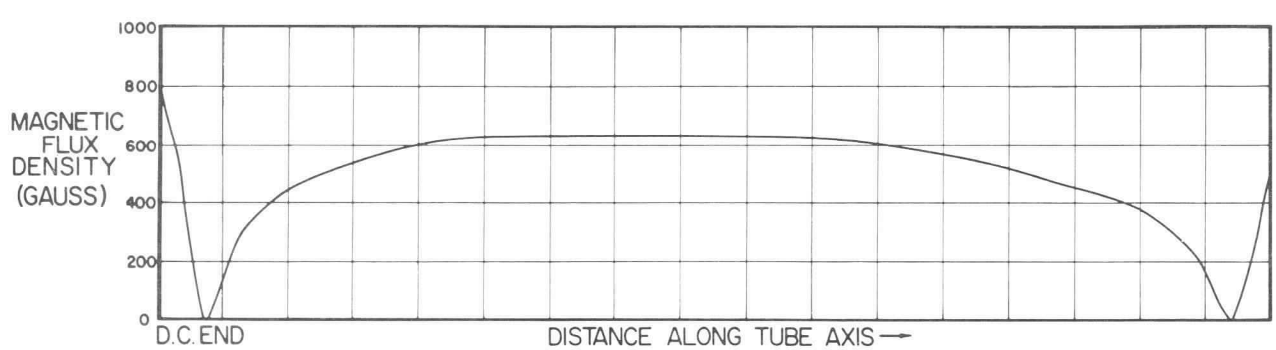

Figure 2. Plot of an Axial Field that is Generated by a Straight-Field Permanent Magnet.

PERIODIC AXIAL FIELD SYSTEMS

Periodic axial focusing systems employ a field that reverses polarity periodically as a function of distance along the electron beam. Because of the field's periodicity, the beam undergoes periodic changes of radii and velocities. These changes can, however, be kept small through proper design so that there is negligible effect upon tube operation. Periodic fields can be produced either electrostatically or magnetically.

Electrostatic Focusing. Electrostatic focusing is to be the subject of a future Engineering Note. Hence, suffice it to say that a periodic field can be generated by the difference potential obtained by operating a series of metal rings — or the windings of a bifilar helix — at different dc potentials.

Magnetic Focusing. A periodic axial magnetic field can be produced by using a series of magnets. Since weight and power dissipation preclude the use of electromagnets, a structure that consists of a series of permanent magnets is used.

THE PPM FOCUSING STRUCTURE

The particular PPM focusing unit discussed here utilizes ring magnets which are magnetized perpendicularly to their flat faces. Adjacent magnets are separated with an iron pole piece and are polarized in opposite directions. Compared to other periodic magnetic field systems, the PPM structure has many advantages.

Advantages. Where weight is concerned, periodic PM focusing has a decided advantage. Because the magnetic fields of alternate polarity tend to cancel, the field in the region external to the structure is nearly zero. Thus, very little field is found in the space outside of the tube. As a result, the amount of magnetic material required for focusing the beam is significantly reduced; hence, the weight is also reduced. Unlike a straight-field permanent magnet, the weight of a periodic structure is a linear function of tube length.

A second advantage lies in the area of field utilization. The focusing field of a periodic structure is used most efficiently because the main component of the magnetic field is confined to the electron-beam region.

Other advantages of the PPM focusing unit are low external magnetic fields and elimination of the adjustments normally needed to focus the tube. Furthermore, the structure is basically self-shielding; hence, external magnetic shielding can be added in close proximity to the magnetic unit without decreasing the field in the electron-beam region. And because the fields outside of the tube are very small, two tubes can be operated side by side with no change in performance.

Field Requirements. The rms value of a periodic magnetic field required to focus the beam in any given tube is theoretically equal to the Brillouin field needed to focus the same beam. It has been found experimentally, however, that the rms value of a periodic magnetic field must be approximately 0.5 to 2 times greater than the Brillouin field in order to attain nearly 100 % transmission.[2] Such strong field requirements dictate that the magnetic material used must have a large coercive force. Lightweight barrium ferrite ceramic magnets exhibit this property.

GUN AND BEAM CONSIDERATIONS

When designing any traveling-wave tube focusing device, consideration must be given to the type of electron gun that will be used. If the use of a magnetic lens is not considered, a parallel-flow gun, for example, cannot be used with a Brillouin field. Similarly, a confined-flow focusing system precludes the use of a convergent-flow gun.

The choice of focusing system is also determined by the type of beam. A hollow beam requires a minimum of two forces: an inward and an outward force. The former prevents the beam from spreading, and the latter keeps the beam from collapsing. A solid beam, on the otherhand, needs only an inward focusing force.

PPM FOCUSED AMPLIFIER CONSTRUCTION

The electron gun used in the tube described here is of the parallel-flow type. Proper beam entrance conditions into the periodic field in the helix area are obtained by using a magnetic lens instead of a convergent-flow gun: When noise figure is of no importance, convergent-flow guns are normally used in tubes with a power output in excess of 100 mw because of the higher beam current required.

Use of a parallel-flow gun simplifies tube construction. Simplicity of construction stems from the fact that a solenoid focused tube which employs a parallel-flow gun need not be redesigned for a convergent flow gun to operate in the periodic system. Therefore, a minimum amount of engineering effort is needed to make the transition from the solenoid unit to the periodic focused version of a given tube type.

Furthermore, the parallel-flow gun used in a PPM system will not generally increase the tube noise figure over that found in its solenoid version. Using a convergent-flow gun will increase the tube noise figure over that found in its solenoid focused counterpart because a convergent-flow gun introduces a greater thermal velocity spread in the electron beam.

The aforementioned magnetic lens is provided by a stack of magnetic rings that surround the electron gun. Proper entrance conditions (beam diameter and slope) of the beam into the helix area depends upon the transition region between the gun and periodic helix magnets. This transition region is actually a combination of the divergent electrostatic lens that exists between the gun's anode and the beginning of the helix and the external convergent magnetic lens. The magnetic lens must be designed so that it just compensates for the divergent electrostatic lens.

Figure 3 shows a plot of absolute value of the magnetic flux density measured along the axis of a Huggins HA-20. We see that the flux at the cathode is in opposition to the field of the first helix cell. A difference in the shape of the field between the cathode and helix and the remaining periods within the structure accounts for the magnetic lens action.

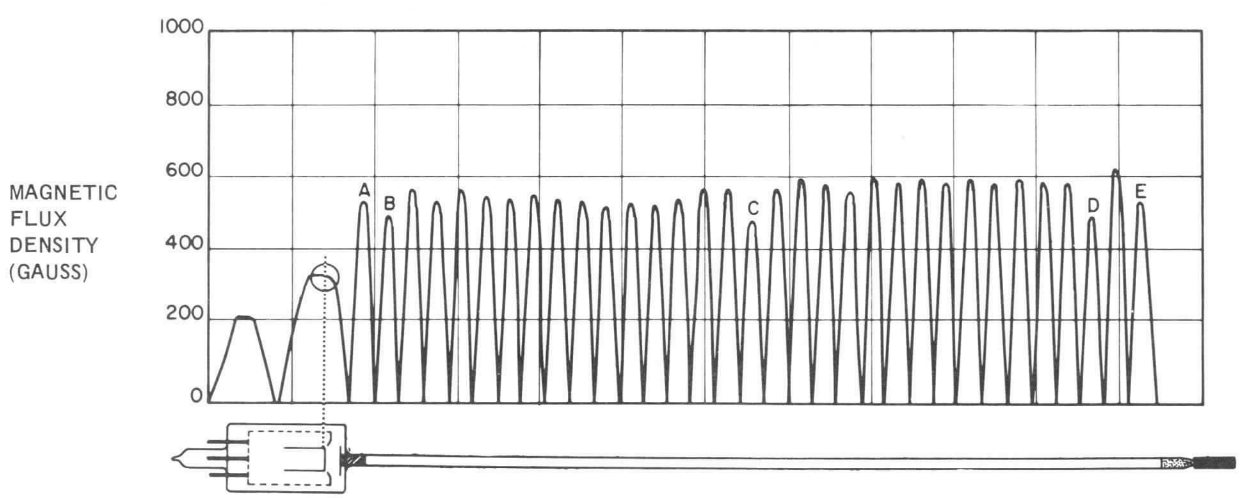

Figure 3. Plot of PPM Focusing Field used for Huggins HA-20 TWT Amplifier.

We also notice that the flux in each alternate cell is in the opposite direction and that the peak values of flux in the first few cells vary widely. The latter occurs at both ends of the helix stack (pulses A, B, D, and E) and is caused by a combination of 1) nonsymmetrical demagnetization due to leakage flux at both ends and 2) using slotted magnets (B and D). Despite this effect, PPM tubes will focus satisfactorily. In some instances an effort has been made to obtain a more uniform field such as that shown in Figure 4. The peak level of the end cells (F and J) can be increased by using pole pieces and smaller ID magnets at the ends of the helix stack. Also observed is the fact that the level of the slotted magnet pulses is increased too. The interesting point of placing adjacent cells in aiding is illustrated in Figure 5. Pulses K, L, M, and N show the resultant field of magnets so positioned.

A cutaway drawing of a typical tube and magnet assembly appears in Figure 6. The outer shell or capsule serves as a magnetic shield, and the inner capsule houses the helix magnet stack which consists of a series of ring magnets of alternate polarity separated by iron disc pole pieces. Though not shown, coupling helices are used for input and output helix to coax transitions, and thus, pole pieces with the smallest possible inside diameter can be used. The coax lines are brought out through slotted magnets to pass between the inner and outer capsules and terminate at rf connectors which are rigidly mounted in the cap on the helix end of the tube. DC connections, also not shown, are made to a plug which is mounted on the cap at the gun end of the tube.

TUBE CHARACTERISTICS

The operating characteristics of the PPM version of a given tube type are very close to those for its solenoid-focused counterpart. In the area of weight and power dissipation, however, the PPM tube has a very favorable edge.

Power Dissipation And Weight. Table I contains information on both 10 mw and 1 watt tubes and compares the weight and power dissipation between the PPM and solenoid-focused versions of several tube types. The weights of the PPM tubes lie within the range of 3.6 to 4.5 lbs, and the solenoid tube weights are reduced by a factor varying from approximately 5.25 to 7.

This reduction accounts only for the weight of the tube itself. The weight of a system using PPM tubes will usually be reduced even more by elimination of the solenoid power supply. Power dissipation in the PPM tubes ranges from 6 to 43 watts which represents a reduction factor varying from approximately 7 to 12.

Small-Signal Gain. A composite small-signal gain characteristic of typical 10 mw PPM amplifiers covering the frequency range from 1.0 to 11.0 kmc is shown in Figure 7. As can be seen, the gain is greater than 31 db, and the gain curves are asymmetrical across the bandwidth of an individual tube.

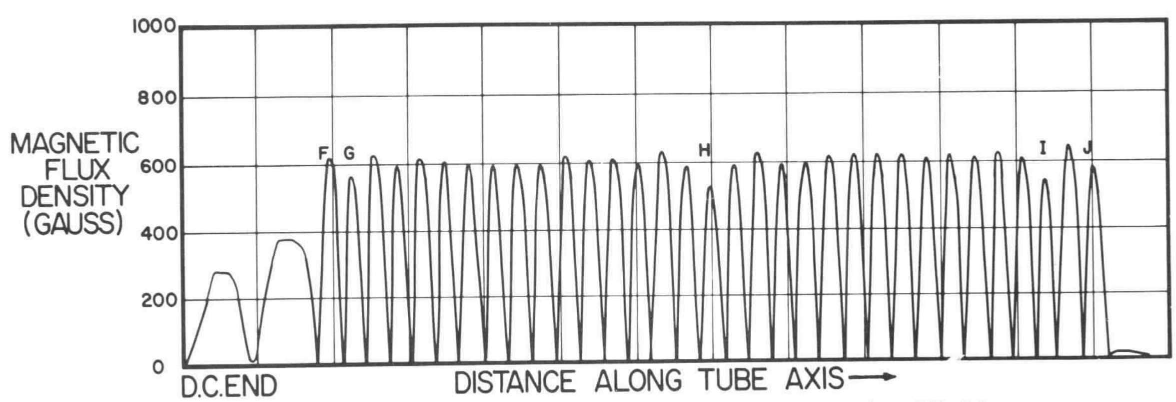

Figure 4. Plot of a Nearly Uniform Periodic Focusing Field.

TABLE 1 Comparison of Power Dissipation and Weight of PM Focused and Solenoid Focused TWT's

FREQUENCY RANGE (KMC)	POWER LEVEL	STANDARD TUBE PLUS SOLENOID		PM TUBE		WEIGHT REDUCTION FACTOR	SOLENOID PLUS TUBE BEAM AND HEATER POWER (WATTS)	PM TUBE BEAM AND HEATER POWER (WATTS)	POWER REDUCTION FACTOR
		TUBE	WEIGHT (LBS.)	TUBE	WEIGHT (LBS.)				
1.0—2.0	10 mw	HA-5	23	HA-31	4.0	5.75:1	78	8.0	9.75:1
2.0—4.0	10 mw	HA-1	23	HA-29	4.4	5.23:1	47	6.0	7.83:1
2.0—4.0	1 watt	HA-2	25.3	HA-30	4.5	5.57:1	292	26.0	11.2:1
4.0-8.0	10 mw	HA-3	25	HA-28	4.0	6.25:1	73	6.0	12.2:1
8.2-12.4	10 mw	HA-4	25	HA-20	3.6	6.94:1	73	11.0	6.65:1
8.2—12.4	1 watt	HA-9	28	HA-21	4.3	6.5:1	498	43.0	11.6:1

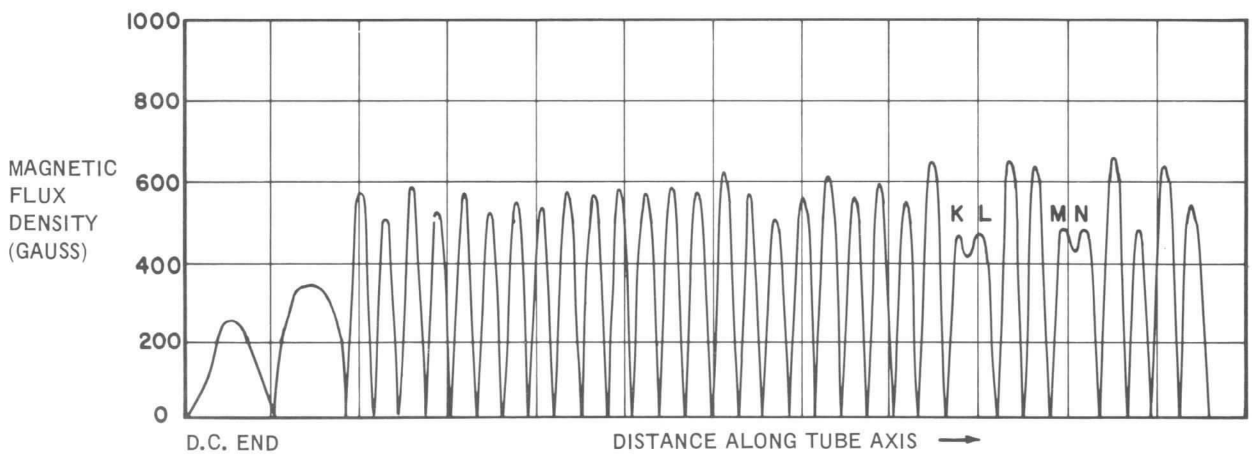

Figure 5. Plot of Periodic Stack in which Some Adjacent Magnets have been Placed in Aiding.

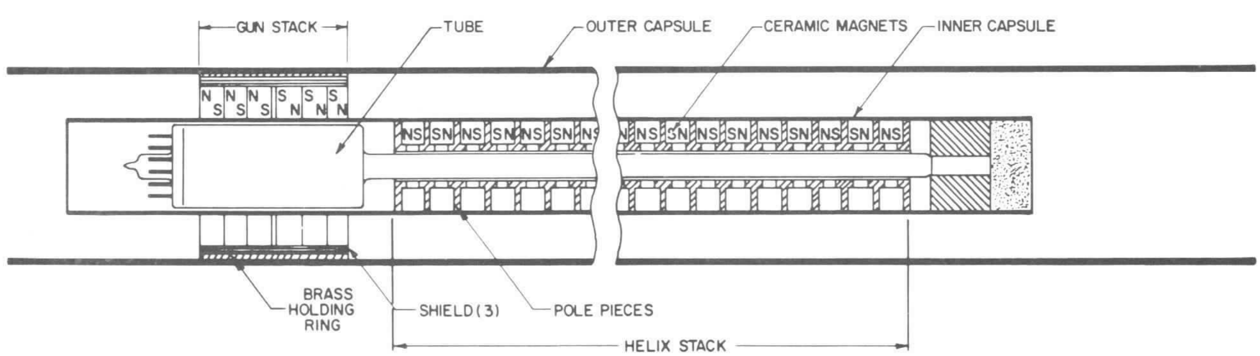

Figure 6. Cutaway Drawing of Typical PPM Focused TWT Amplifier.

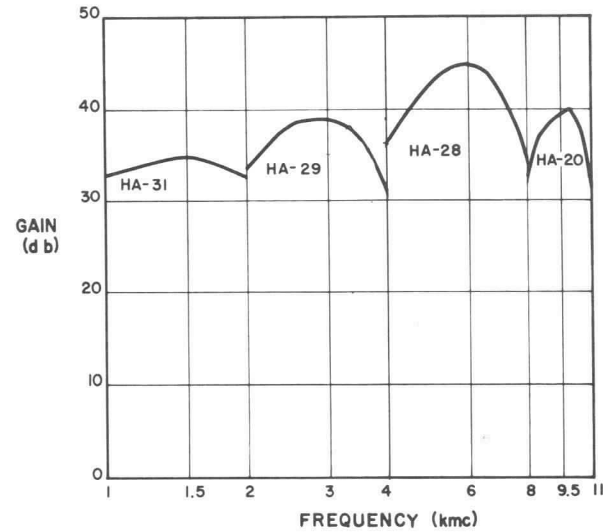

Figure 7. Composite Small-Signal Gain Characteristics of Typical 10 mw PPM Amplifiers.

Figure 8 compares the gain characteristics of a solenoid and PPM focused tube of the same type. The solenoid focused HA-1 curve represents a typical case. A slightly higher than average gain characteristic was used for the PPM focused HA-29, but it does illustrate the typically large difference in gain between maximum and high-frequency-end gain. By properly shaping the loss of the tube though, an octave bandwidth gain curve that is flat within ±2 db can be readily obtained.

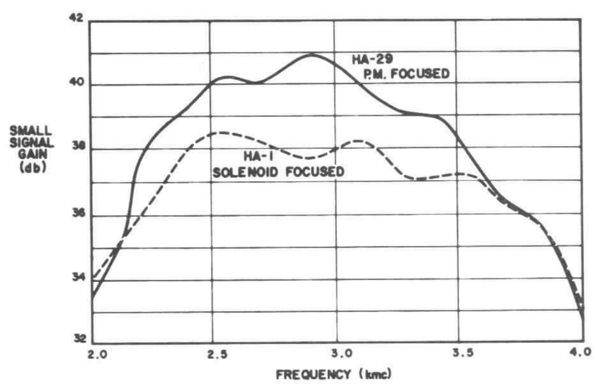

Figure 8. Gain Comparison of S-Band Solenoid and PPM Focused Amplifiers.

As in Figure 8, a gain comparison between a solenoid and periodic PM focused TWT is shown in Figure 9. Here, however, the tubes operate in the 8.0 to 11.0 kmc region of the frequency spectrum. The best and typical periodic-focused

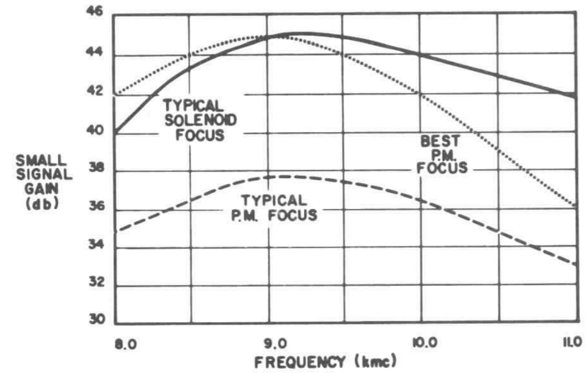

Figure 9. Gain Comparison of X-Band Solenoid and PPM Focused Amplifiers.

tubes are 6 and 9 db below the solenoid tube respectively at the high frequency end of the band.

In Figure 10 we see the results of data that was taken to indicate the total frequency range where appreciable gain can be realized. With the voltage and current conditions shown, useful gain can be obtained over approximately a 9.0 kmc bandwidth.

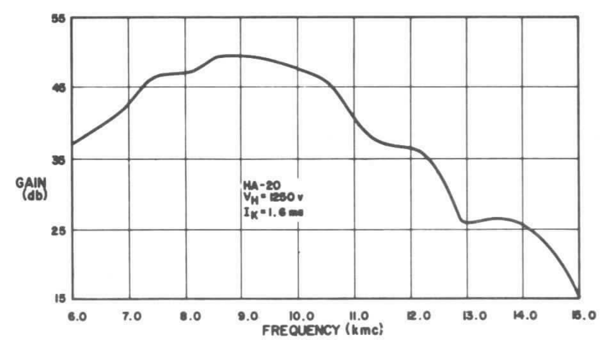

Figure 10. Curve Showing Wide-Band Gain Capabilities of an HA-20.

Saturation Power Output. A comparison of saturation power output for solenoid and periodic focused 10 mw amplifiers in two different frequency ranges is shown in Figure 11. We can see that the power output is at worst only 2.5 dbm low in either range. Figure 12 shows a similar comparison for 1 watt amplifiers in the same frequency ranges. Sufficient data has been taken to determine that the saturation power output curve for a PPM focused traveling-wave tube amplifier will generally be nearly the same as for its solenoid focused counterpart.

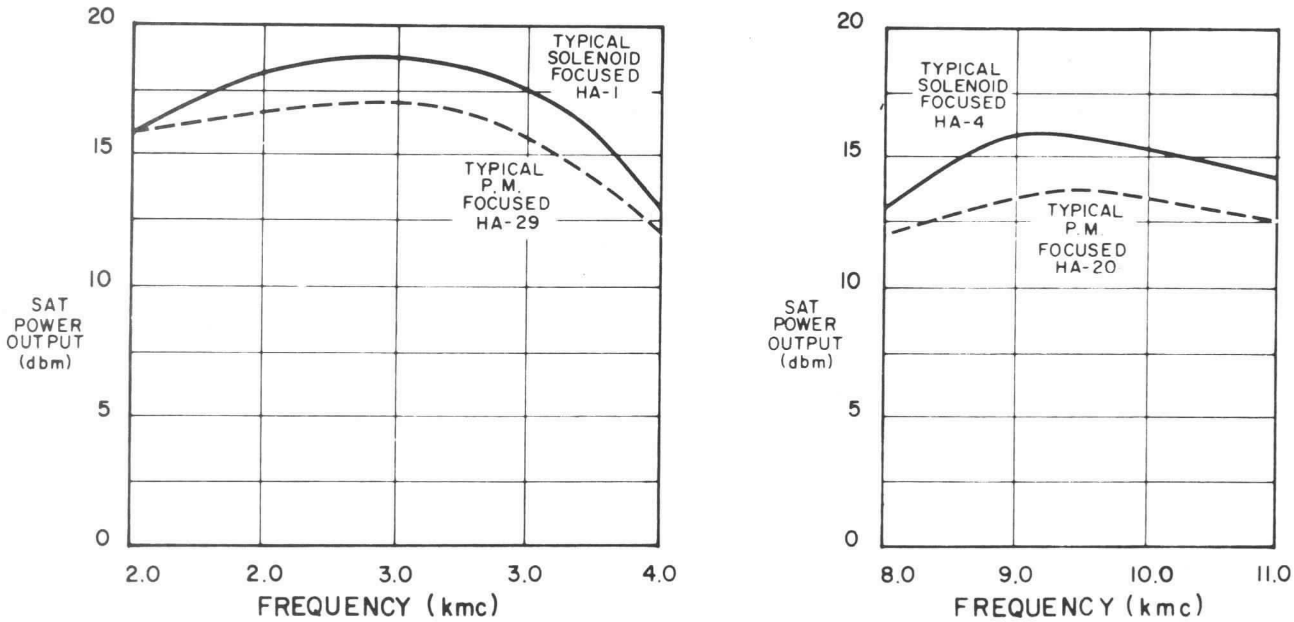

Figure 11. Output Power Comparison of Solenoid and PPM Focused 10 mw Amplifiers.

Noise Figure. Measurement of the noise figure in periodic PM tubes confirms the expectation that the noise figure should remain nearly the same as in the solenoid case because of the parallel-flow type electron gun used in the PPM tubes. Measured noise figures lie in the range between 20 and 28 db and average approximately 25 db.

Focusing. Figure 13 shows a typical focusing curve for PPM focused amplifiers. As cathode current is increased, the helix current at first rises to a maximum and then dips to a minimum value that is very near to the value of operating current. Beyond the operating point the intercepted helix current rises for all higher values of cathode current. The dip in the focusing curve

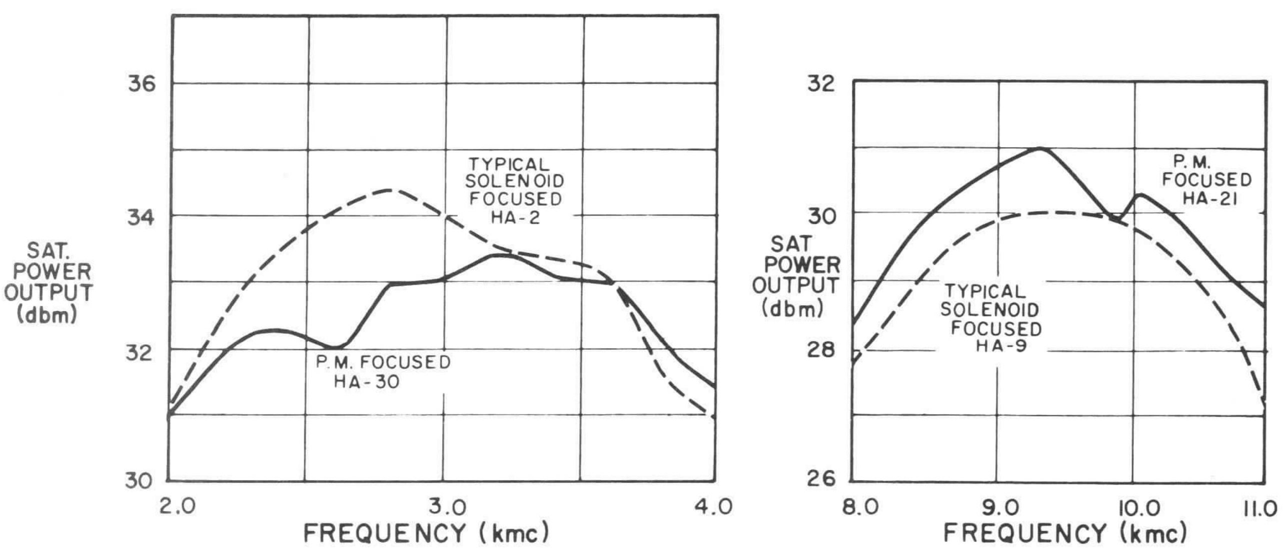

Figure 12. Output Power Comparison of Solenoid and PPM Focused 1 watt Amplifiers.

is called the focusing notch. In initially adjusting periodic PM tubes in their magnets at the factory, the focusing notch is made to correspond as closely as possible to the operating current. Tubes in which the focusing notch and operating current coincide exhibit the best wide temperature characteristics.

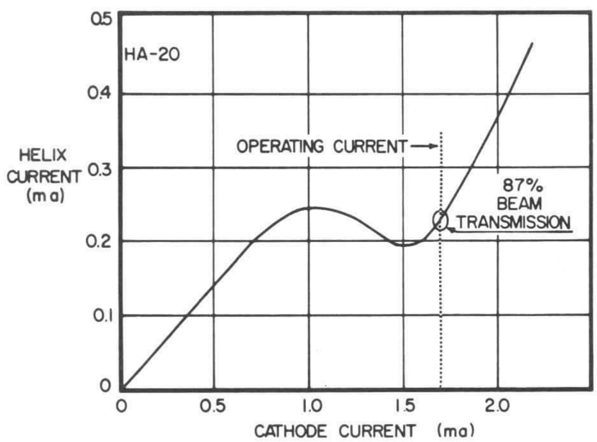

Figure 13. Typical Focusing Curve for PPM Focused TWT Amplifiers.

Temperature. The flux density which ceramic magnets can develop across the air gaps of the pole pieces is a linear function of temperature: flux in the air gaps increases toward lower temperatures and decreases toward higher temperatures. Because focusing of the electron beam is a critical function of the magnetic field, focusing will change with temperature.

The change in focusing as a function of the ambient temperature of the surrounding air is shown in Figure 14. Data for this curve was taken in the following manner: The tube was initially adjusted in the magnets at room temperature, and after each change the temperature was held constant long enough for the tube characteristics to reach equilibrium. As can be seen in Figure 14, the focusing notch occurs approximately at the operating current at room temperature. Tubes upon which the notch is correctly placed at room temperature have the smaller excursions of helix current as a function of temperature.

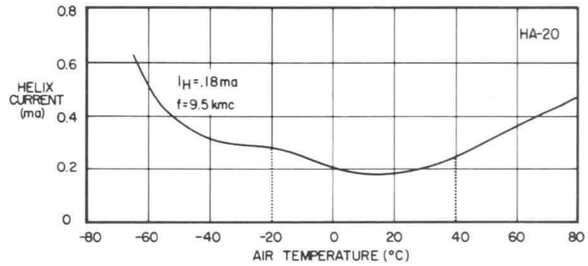

Figure 14. Change in Focusing as a function of Temperature.

At the extreme temperatures of -65 °C and 80 °C the intercepted helix current in uncompensated structures is high enough to appreciably affect the life of the tubes. This deterioration in life is caused by ion bombardment of the cathode which in time destroys its emitting properties. The ions are generated by the current which is intercepted on the helix. Probable allowable limits on intercepted helix current lie in the vicinity of 20 °C and 40 °C.

Figure 15 shows the small-signal gain for the same tube shown in the preceding figure (14) over the temperature range -55 °C to 80 °C. This particular tube shows no more than a 4.5 db gain variation. Though not indicated by the curve, gain generally decreases at the extreme temperatures.

Within the limits of acceptable focusing (20 °C to 40 °C), the gain variation – referenced to room temperature – is approximately 1.25 db at the low temperature end and 0.25 db at the high end.

Small-signal gain as a function of frequency over the temperature range -55 °C to 80 °C ap-

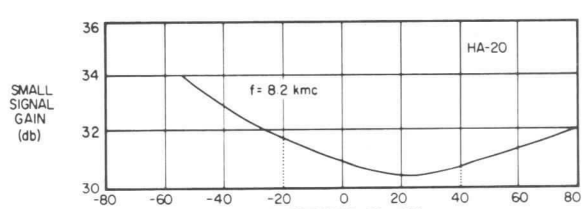

Figure 15. Typical Small-Signal Gain vs Temperature Curve for a PPM Focused Amplifier.

pears in Figure 16. Gain variations from the room temperature (20 °C) curve are approximately 0 to 3.5 db. As is normally expected, the greatest deviation exists between the room and lowest temperature gain curves. If TWTs are to be used over a very wide temperature range and gain stability is important, it is possible to use some means of temperature compensation to make TWT operation nearly independent of the temperature.

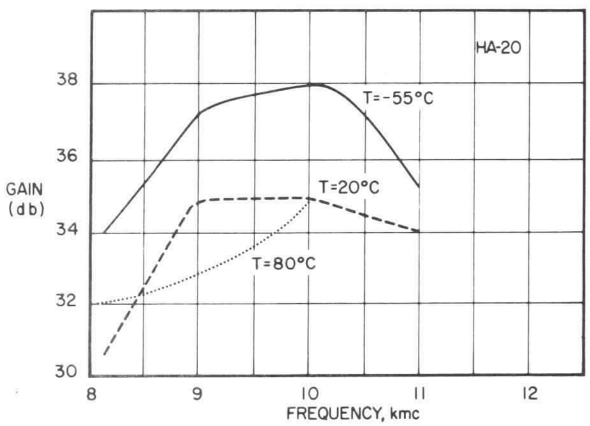

Figure 16. Small-Signal Gain Spread over Temperature Range –55°C to 80°C.

Extending the lower ambient temperature limit is not too difficult when thermostatically controlled heaters are used or temperature compensation in the stack is incorporated. The upper limit is somewhat more difficult, but Huggins Laboratories, Inc. has successfully extended the upper limit by using temperature compensating shields [see Huggins Laboratories, Inc. "Engineering News," Vol. I, No. 3]. And the once-remote possibility of optimizing the focusing at some higher temperature than room temperature by using a magnetic material that has no temperature coefficient is rapidly becoming more probable.

CONCLUSIONS

As we have seen, the gain, power output and noise figure characteristics of periodic PM focused tubes are close to those of similar tubes that use solenoid focusing.

Periodic PM focused TWTs are quite simple to use, and they give satisfactory performance over a reasonably wide temperature range. With improved techniques in temperature compensation, it is certain that the temperature limits for satisfactory operation of PPM focused TWTs will be extended. Furthermore, the great reduction in weight permits designing longer tubes; this means, of course, that higher gains can be achieved.

In the latter part of 1959, Huggins Laboratories, Inc. successfully developed a periodic permanent magnet focused backward wave oscillator. In view of this advance in the state of the art, it is apparent that PPM focused BWOs are as practical as PPM focused amplifiers.

ACKNOWLEDGEMENT

The bulk of this Engineering Note is based on work performed at Huggins Laboratories by L. A. Roberts.

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ELECTROSTATICALLY FOCUSED TRAVELING-WAVE TUBE AMPLIFIERS

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During the last few years there has been a great demand to reduce the over-all weight of traveling-wave tubes. This demand for weight reduction has been met by using periodic magnetic fields and periodic electrostatic fields as the focusing mechanism rather than a heavy solenoid or straight-field magnet.

The possibility of using periodic electrostatic fields of short periods to focus the electron beam in a TWT has been studied both theoretically and experimentally by many research laboratories (e.g., Huggins, RCA, Stanford, and Varian). In the middle of 1958, an electrostatically focused forward wave amplifier having a small-signal gain greater than 30 db and an output power of approximately 10 mw over an octave bandwidth was developed at Huggins Laboratories, Incorporated.

Further development at the Laboratories produced an electrostatically focused amplifier with 30 db small-signal gain and a power output of 1 watt over an octave bandwidth. The gross weight – including the capsule and connectors – is one pound for the 10 mw tube (the HA-27) and two pounds for the 1 watt tube (the HA-58).

SOLENOID FOCUSING

In solenoid focusing, parallel flow in a beam of circular cross section is maintained by a magnetic field that is positioned parallel to the path of the electrons. Throughout the length of the tube, electron flow is held in a tight pencil beam by the confining forces of the longitudinal magnetic field that surrounds the tube. Space-charge repulsion causes an initially parallel beam to spread; thus, the electrons will intersect magnetic field lines, and each electron will follow a helical path about a magnetic line whose distance from the axis corresponds to the radius of the electron's initial trajectory.

The radius of the helical path, or spiral, of the electron is inversely proportional to the magnetic field intensity. Consequently, high intensity magnetic fields are required to keep radial velocities of the electrons small. Unfortunately, the weight of the focusing system is proportional to field intensity.

ELECTROSTATIC FOCUSING

In electrostatic structures, however, the focusing system adds very little weight to the TWT because the bifilar helix acts as the focusing mechanism as well as the slow wave propagating circuit.

As in a conventional traveling-wave tube, the electron beam in an electrostatic tube is injected into the bifilar helix region by a parallel-flow or a convergent-flow electron gun.

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Focusing Forces. Figure 1 shows a bifilar helix and the forces of the electrostatic field that act on the beam. The potential difference that exists between the two helices provides a radial inward focusing force on the electrons. Figure 2 depicts the trajectory of the electron beam in the helix region. In region A the radial component of the field is directed inward, and the electrons will move more slowly in this region. In region B the radial component of the field is directed outward, and the electrons will move faster in this region. Because of this difference in velocity, the electrons spend more time in the vicinity of the low voltage helix (Region A). It is in this region that the electrons experience the inward focusing force that approximately balances the outward space-charge force of the electron beam.

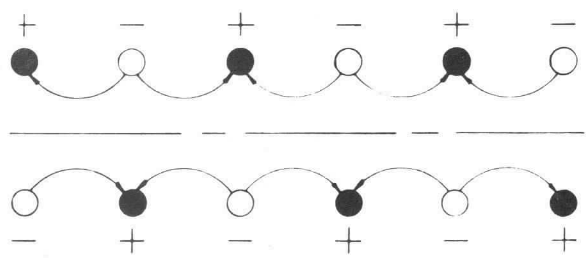

Figure 1. Periodic Electrostatic Focusing Field Forces that act on the Electron Beam in a TWT.

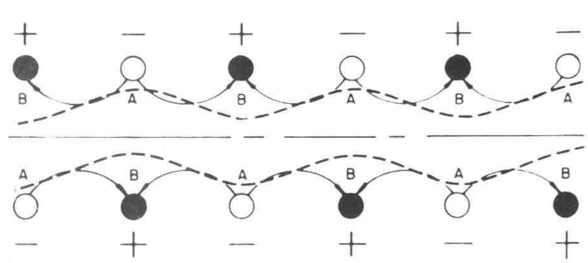

Figure 2. Electron Beam Trajectory Produced by a Periodic Electrostatic TWT Focusing Field.

Optimizing Performance. The average beam velocity, which is a direct function of the helix voltages, is varied until it nearly equals the phase velocity of the rf waves to obtain optimum interaction, and consequently energy transfer, between the electron beam and the rf wave. Because the phase velocity is a direct function of frequency, the helices are optimized at the high frequency end of the band to obtain octave bandwidth coverage.

Coupling. Single-filar helical couplers are used for coupling rf energy on and off the bifilar helix since the mode configurations of the bifilar and single-filar helices are essentially the same. Experimental coupling loss measurements indicate between 1.0 and 1.5 db loss per coupler across an octave bandwidth.

The Electron Gun. As mentioned before, the electron gun may be of the convergent-flow or the parallel-flow type. But experimental results indicate that more success is obtained with a hollow beam parallel-flow gun, especially in medium power tubes. This success stems from the fact that a hollow beam has less space-charge repulsion than a solid beam.

Rather than use a weight-adding magnetic structure over the gun section to obtain proper entrance conditions, a special gun arrangement is used to focus the beam into the helix region. A drift tube positioned near the helices provides the final focusing anode. This drift tube plays a vital part in providing optimum entrance conditions, and care must be taken to insure that the drift tube doesn't deform the beam in such a way as to cause the traveling-wave tube to oscillate.

The Bifilar Helix. Construction of the bifilar helix was one of the more difficult problems encountered in the development of an electrostatic TWT. Proposed solutions to the problem evolved into a successful method of manufacture: once the helix is wound, a glass shrinking technique is used in which the glass envelope is shrunk onto the helical structure, and the amount of glass that comes in contact with the helix is accurately controlled. This process locks the helices down at three points in every turn. In addition to providing an insulating medium, the glass prevents any relative motion between the helices. This, of course, allows the tube to be operated in any position without disturbing the rf performance.

ADVANTAGES OF ELECTROSTATIC FOCUSING

Because no magnetic structures (either solenoid or permanent magnet) are used on Huggins electrostatic tubes, the problems of temperature environment are greatly reduced. Another outstanding characteristic of this type of structure is the reduction in weight as compared to a solenoid or periodic permanent magnet focused TWT. For example, a weight reduction of 13:1 was obtained by building an electrostatic version of the Huggins HA-8. Furthermore, an electrostatic tube is free from ion oscillation because of the electrostatic field. When comparing an electrostatic tube with a solenoid tube, the elimination of the solenoid and its power supply is an obvious, but important, credit. The advantages of an electrostatically focused TWT can be summarized as follows:

1. Lightweight (lighter than a PPM tube)
2. Independence of environment
3. Ruggedized helix
4. Elimination of solenoid and its power supply

HUGGINS ELECTROSTATIC TUBES

The first success in the development of electrostatically focused traveling-wave tubes by Huggins Laboratories was realized with the L-band HA-27. One and one-half years later came the UHF-band HA-58. A few of the more important characteristics of these two tubes follow:

The HA-27

Frequency range .... 1.0 to 2.0 kmc
Saturation power output .... 10 mw min
Small-signal gain .... 30 db min
Weight .... 1.0 lb

The HA-58

Frequency range .... 0.5 to 1.0 kmc
Saturation power output .... 1 w min
Small-signal gain .... 30 db min
Weight .... 2.0 lbs