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In This Issue...

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Editorial

by John E. Remich
Manager, Technical Department

ELECTRONICS AND NUCLEAR RADIATION

In the past decade the words *nuclear energy* have become familiar household words. Thanks to motion pictures, television, and newspapers, practically everyone has at least a non-technical understanding of what nuclear energy is, how it may be used, and the effects of nuclear radiation on human beings and other forms of life. However, a comparatively recent development on which research has only barely begun is the determination of the effects of the various types of nuclear radiation on electronic components and equipment. This problem is especially important to the military establishment, since the acquisition and communication of information by electronic means is a vital factor in any military action.

Preliminary research on this problem has indicated that nuclear radiation has definite detrimental effects on at least some types of components. For example, it has been found that bombardment by neutrons can cause changes in the lattice structure of germanium, a metal widely used in transistors. These changes affect transistor operation to the point where the units can become unusable.

This is only a single example. Other possible problems such as the effects of nuclear radiation on the glass and metals used in vacuum tubes, the gas used in thyratron tubes, the alloys used in silicon transistors, and many others, have yet to be completely investigated. Determination of the effects on communications and radar at many frequencies and on the boxes in which equipments are packaged also constitutes a pressing requirement.

Since the field engineer is concerned with the operation and maintenance of electronic equipment under field conditions, it is becoming increasingly apparent that each engineer should become familiar with the reports and articles which deal with the effects of nuclear radiation, the methods devised to overcome any detrimental effects, and the electronic devices used to detect and measure the various types of radiation.

LETTERS TO THE EDITORS

"In reference to Mr. Hilary J. Burton's letter on page 2 of the September-October, 1957 BULLETIN requesting information on r-f radiation hazards, I should like to call to your attention an article on page 36 of the October, 1957 BUSHIPS JOURNAL (Navy Department), entitled 'Radar Radiation Hazards.'

"Practical safety precautions with quantitative data which might be quite useful to Mr. Burton are given in the cited article."

Howard Freiberger
Electronics Scientist

"After reading the Letters to the Editors column in your September-October, 1957 issue, I submit the following as additional information on the question of radiation hazards submitted by Hilary J. Burton. This information was extracted from BuShips Notice 9672, dated 22 July 1957.

"a. Based on an average radiation intensity of 0.01 watts/cm², the following minimum distances from the rotational center of radar antennas are tentatively established as limits beyond which constant exposure (one hour or more) is safe and within which there is a definite hazard from radiation:

RADAR	MINIMUM DISTANCE (FEET)	
	In Axis of Primary Beam of Non-rotating Antenna	In Horizontal Plane of Primary Beam of Rotating Antenna
AN/BPQ-2	45	5
AN/BPS-1	10	1-2/3*
AN/BPS-2	40	8*
AN/BPS-3	46	5*
AN/BPS-4	20	3-1/2
AN/SPN-6	186	21
AN/SPS-2	340	35
AN/SPS-6	40	9*
AN/SPS-8	175	25
AN/SPS-8A	245	34
AN/SPS-12	40	9*
AN/SPS-17	35	11
AN/SPS-26	320	37
AN/SPS-28	10	9*
CXRX	152	21
SA, SC series	7	8*
SRa, SK series	7	9*
SR-3, SR-6	40	9*
SS-1	10	1-1/3*
SS-2	11	1-1/3*
SV-3	22	4-1/2*

*These limits are the swing-circle safety radius.

"The intensity level of 0.01 watts/cm² has been tentatively accepted as a working tolerance subject to further review by the Bureau of Medicine and Surgery. On board ship, normal radar use practices (scanning and tracking) are such that the likelihood of long exposure in the main beam is remote. However, during target examination or testing the exposure could be an extended period.

"b. Do not make a direct visual examination of any microwave radiator, reflector, waveguide opening, or waveguide horn during periods of transmission.

"c. Photographic personnel should be cautioned regarding the dangers of exposing photoflash bulbs to radar beams even at considerable distances."

Darrell N. Carlson
Philco TechRep Field Engineer

"In answer to Robert L. Matthews' query in 'Letters to the Editors' in the July-August Tech-Rep BULLETIN, the following book should serve his purpose, whether or not it is the one he saw in the library: George Boole, AN INVESTIGATION OF THE LAWS OF THOUGHT, Dover Publications. All subsequent books on Boolean or logical algebra are based on this one."

Lewis F. Garber
327 North Madison Ave.
Pasadena 4, Calif.

(In addition to the reference given in the letter above, there are a number of other books which deal with symbolic logic, the first application of Boolean algebra. For those interested in applications, the Computer Issue of the *Proceedings of the Institute of Radio Engineers*, October, 1953 is also recommended reading. Ed.)

"According to my calculations, the most common type of MTI system, which uses i-f locking and i-f addition (Fig. 16.13(a), page 635, 'RADAR SYSTEM ENGINEERING'), differentiates between fixed and moving targets at the i-f frequency by beating together the target echoes and an i-f frequency from an oscillator which preserves the transmitter phase. Therefore, if the radial speed of a target with respect to the radar were one wavelength (10 meters at an i-f frequency of 3 mc) per interpulse interval (2.5 milliseconds), the target would be 'invisible.' This speed is equivalent to approximately 130 knots. Also, of course, targets having speeds of 2v, 3v, 4v, etc., where v equals 130 knots, would be invisible. Please confirm or correct my findings.

"Also, if a target, preferably an aircraft shaped like a dipole (an actual aircraft would probably invalidate the following), were to move at a radial speed equal to one-half wavelength (at the transmitter frequency) per interpulse interval, this target would be invisible to the radar, whether or not it was operating on MTI. Am I right? According to this conclusion, of course, the other blind speeds would be 1½ L/i, 2½ L/i, etc., where L is the wavelength of the rf in the transmitted pulse and i is the interpulse interval."

Gerald R. Brookman
Philco Site Engineer

(It is true that there are certain target radial velocities, with respect to a radar operating with MTI, at which the target will not appear on the scope. These radial velocities are known as "radar blind speeds," and are directly proportional to the r-f wavelength and PRF of the radar set. The expression for the first radar blind speed is:

$$V = \frac{\lambda \cdot PRF}{2}$$

This formula will give the target velocity in meters per second for the first blind speed. Any multiple of $\lambda/2$ will also be a blind speed. To obtain velocity in miles per hour, the following expression can be used.

$$V(mph) = \frac{\lambda\ (cm) \cdot PRF}{89.4}$$

The problem of radar blind speeds is discussed in section 16.9 of *Radar System Engineering.*

Radar blind speeds exist only because of cancellation circuits within the radar, which compare the return echoes from pulse to pulse. Since a radar set without MTI does not contain these cancellation circuits, each target is detected regardless of its radial velocity in relation to the radar. Ed.)

SUN-STROBE TECHNIQUES

By Bud M. Compton
Philco Site Engineer

This article describes the use of solar radiation in checking the calibration of both search and height-finder radars.

INTRODUCTION

Investigations in the field of radio astronomy, particularly those related to a study of the r-f energy radiated by the sun, are promising benefits to radar. It has been found that this energy is readily picked up by radars in the "L" and "S" bands, and the resulting signal displayed on the indicator is called the *sun strobe.* Proper use of this strobe offers many opportunities for the calibration of radar equipment and the evaluation of its performance.

As seen on a PPI scope, the sun strobe closely resembles the strobe produced by continuous-wave interference. In fact, there have been times when it was reported as such. Because of the correlation between its direction of reception and the sun's position, however, the sun strobe is easily distinguished from other types of interference.

Practical use of the sun strobe is becoming more common, as the reader may surmise. Probably the oldest use has been in orienting antennas. Related to this use is the alignment of azimuth servos by identical means.

Directed more toward research is the application of the sun strobe in probing deeper into refraction. However, this type of study is useful in determining the atmospheric peculiarities that exist at a radar site.

Quality control is an important potential use of the sun strobe. True, the results are not as objective, say, as the MDS measurement. On the other hand, there is much to be gained. For example, if the sun strobe should be received on several beams (lobes) and/or radars but observed only on part, it can be assumed with considerable certainty that the receiving equipment failing to "see" the strobe has impaired performance. Before this potential use can be fully realized, however, more information is needed concerning certain factors, particularly ionospheric influence and variations in the sun's radiation. These factors have an important effect on the strength of the sun strobe, as evidenced by the fact that the strobe is much stronger at sunrise than at sunset. It is likely that current research will provide the necessary information.

Finally, the sun strobe provides a means for checking height-finder accuracy under controlled conditions. For convenience and economy, the sun-strobe method surpasses all other methods. One qualification should be made, however. During winter at high latitudes little or no sun-up time is available, but during summer the condition is reversed. Actually, so few sites are affected in this manner that little consideration need be given to such a disadvantage.

HOW TO USE THE SUN STROBE

The basic problem in using the sun strobe for calibration purposes is to compare where the sun should be with where it is being received. Any difference that exists indicates the need for investigating such factors as antenna level, general alignment, and refraction. After the sun strobe has been used for several days, it becomes rather easy to pin-point causes of inaccuracy.

Since the radar supplies "where the sun is received" information, it remains necessary to determine where the sun should be. To accomplish this, the navigators' method of determining his position by sighting the sun is reversed. Through the use of a current Air Almanac and Sight Reduction Tables practically all the calculations involved are eliminated.

To illustrate the method of using the sun strobe to check height-finder accuracy, a hypothetical site near Las Vegas, Nevada, will be used as the example. Assume that the exact site location is latitude 37° 22′ 12″ north and longitude 116° 54′ 0″ west. The antenna is situated 4000 ft above sea level and the screening angle is 1/4° positive.

It is desired to use the sun strobe shortly after sunrise at 0540 local time, or 1340 Z (Greenwich) time, 1 Sept. 1957. The first step is to determine the sun's elevation and azimuth relative to the site for this time. From the Air Almanac (1 Sept. page) the sun's Greenwich Hour Angle (GHA) 25° 00′ west and declination 8° 16′ north are obtained. This pin-points the spot on the surface of the earth where the sun is directly overhead at this time. Since GHA is the sun's meridian relative to Greenwich, it must be converted to Local Hour Angle (LHA), which is relative to the site. The LHA is obtained by subtracting the site's west longitude from GHA. This gives 268° 06′. The sun's elevation (Hc) and azimuth (Z) relative to the site can now be obtained from the Sight Reduction Table for 0° to 39° Latitude. Interpolation will be necessary for the minutes and seconds in site location and for minutes in the sun's declination. For interpolating declination, the tables provide a "d" column for use with an interpolating table in the back of the book. Accordingly, it is found that elevation (Hc) is 03° 28′ 36″ and azimuth (Z) is 82° 06′ true. Notice that elevation changes more rapidly than azimuth at this time of day. For checking antenna azimuth where magnetic values are used to orient the antenna, determine the local magnetic variation and subtract east; add west to the 82° 06′ true as found above.

In applying the above information it should be obvious that azimuth errors are found by simple comparison of the computed values with those indicated by the radar.

For checking height-finder accuracy and refraction the matter becomes a little more complicated. It is necessary to convert the angular position of the sun in elevation to the units of altitude used on the height finder. Most height finders indicate altitudes in feet above sea level. By referring to figure 1 it may be seen that the RHI altitude for a given antenna angle is:

RHI ALT. = (Tan α) Range + Mod. Earth's Curv. + antenna elevation

In this example:

RHI ALT. = (Tan 03° 28′ 36″) 6.08 x 10^5 + 8040 + 4000

RHI ALT. = .06075 x 6.08 x 10^5 + 12,040

RHI ALT. = 48,976 ft

The sun's angular position in units of altitude vs time for two or more times are plotted to provide a sun track chart (figure 2). The sun strobe is then tracked during sunrise and the indicated height at which the sun strobe crosses the 100 nautical mile range is plotted on the chart. It is emphasized that the time at which these observed altitude readings are taken should be as accurate as possible. Radio time signals from WWV or WWVH are usually available on site and provide the necessary accuracy.

There is usually an initial ambiguity because a difference between radar and computed altitude may result from either alignment or refraction, or both. Therefore, it is convenient to apply

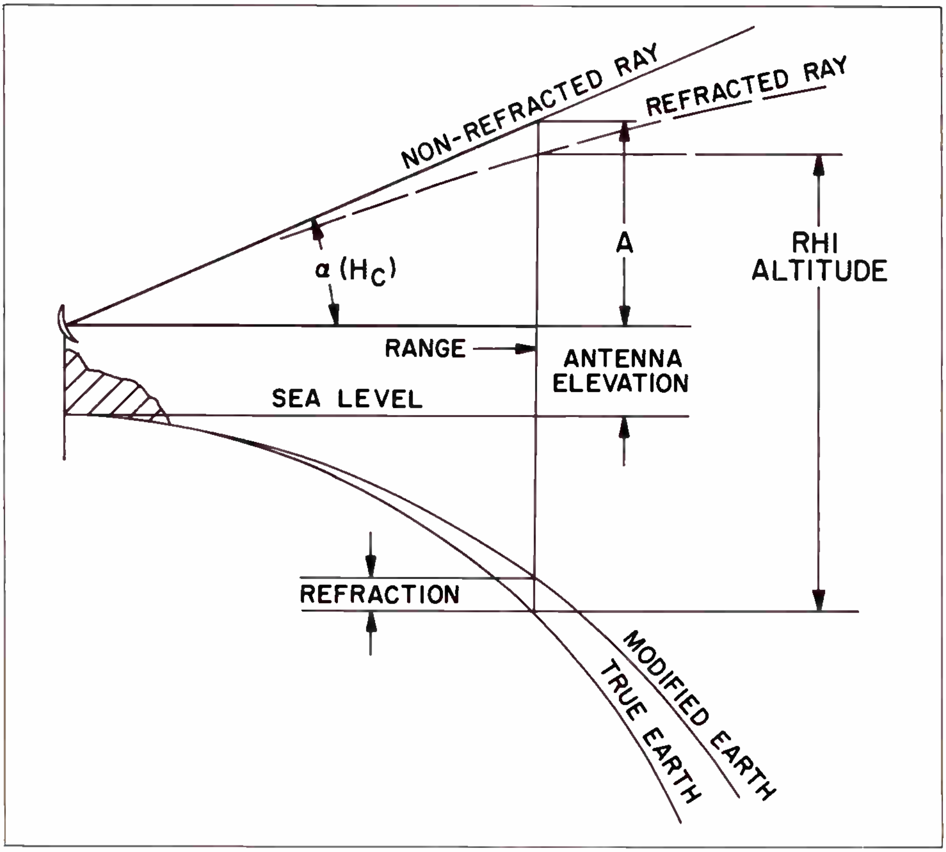

Figure 1. RHI Altitude Factors (Grossly Exaggerated)

sun-strobe information obtained from the search radar or a back-up height finder. In figure 2 it is assumed that the search radar was used to back up the height finder. The tilt of the search antenna is 1-3/4° and is converted to feet of altitude at 100 nautical miles and then drawn on the chart. The sun strobe is observed on the search radar's A scope, and the exact time of maximum signal is recorded on the 1-3/4° line. In this case the maximum occurred at 1329.8 hr Z time. The slight difference between the search radar plot and the height-finder track is insignificant in this case because of the difficulties in determining the exact time of maximum signal on the A scope. However, a large difference plotted over several sun tracks would indicate antenna misalignment in either the height finder or search radar, or both.

Returning to the height-finder track and sun-altitude track, let the altitude difference be ΔR. If this difference is consistent with recent plots and both radars, it may be decided from such evidence that the difference is primarily due to refraction. If desired, ΔR values can be converted to earth's curvature correction factors for easy comparison with the standard curvature factor of 4/3. This standard factor is known as the "K" factor; it is given by:

$$K = \frac{1}{1 - \Delta R \dfrac{2a}{D^2}}$$

where: a = true earth's radius
D = range distance

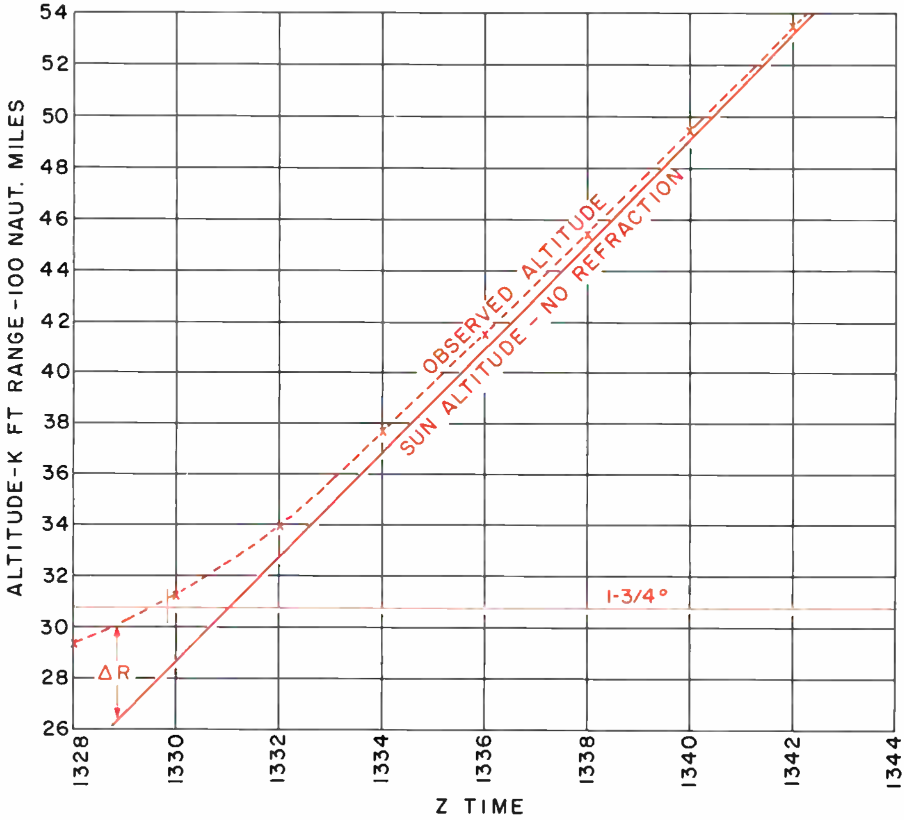

Figure 2. Typical Plot of Computed and Observed Sun-Strobe Altitudes

It is more convenient to give this equation for a range of 100 nautical miles:

$$K = \frac{1}{1 - 1.13\ \Delta R 10^{-4}}$$

Perhaps by now it would seem that all this involves a sizeable amount of labor, especially for daily use. However, this is far from the truth if standardized procedure is used and the sun track curves are plotted a week apart. Individual days of the week on curves so plotted are found simply by proportional division.

In trying the sun strobe method, remember these few pointers. Only the basic principles are given — leaving plenty of room for individual ingenuity. Time accuracy is of utmost importance and any procedure used should be reliable[3]. Ducting and wave trapping may cause unusual sun strobe presentation, but refraction abnormalities are usually limited to small antenna tilt angles. Because of this, it is comparatively simple to check height alignment at higher tilts, and refraction at lower ones.

Also, results of actual tests can be studied[4].

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4. "Sun Strobe as Aid to Radar", Bud M. Compton, (Unclassified) *Air Defense Command C&E Digest*, November, 1957.
5. Field Memo #296, General Electric Co. (Pertaining to the FPS-6 Radar Height Finder), 1 February, 1956.
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CLOSER FREQUENCY TOLERANCES FOR COMMERCIAL MOBILE RADIO

A new F.C.C. ruling stipulates that transmitters with an output of over 3 watts operating in the land-mobile services, installed after November 1, 1958, must meet closer frequency tolerances.

Transmitters operating in the frequency range between 25 and 50 mc are presently required to maintain a frequency tolerance of 0.01%. Transmitters of over 3 watts output operating in these services installed after November 1, 1958, must maintain frequency within 0.002% for frequencies between 25 and 50 mc. The range of transmitting frequencies above 50 mc has been extended in these services from the previous upper limit of 220 mc to include bands of frequencies up to 1000 mc. The frequency tolerance for transmitters of over 3 watts output operating above 50 mc is presently 0.005%. Transmitters installed after November 1, 1958, in these services, operating above 50 mc, must maintain an accuracy of 0.0005%.

The new ruling does not apply to transmitters installed before November 1, 1958, until November 1, 1963, at which time all transmitters in these services are affected.

Since much of the frequency-measuring and monitoring equipment in present use does not meet the required tolerances (0.00025% for frequencies above 50 mc) for transmitter checks under the new ruling, it will be advisable to ascertain that the proper equipment, meeting the required tolerances, will be available.

TechRep Field Engineers and other engineers and technicians who may be working within these services should become familiar with these and other important changes in the F.C.C. regulations pertaining to the land-mobile services.

A pamphlet, "Part 60 — Land Transportation and General Radio Services," which describes these changes, is available from the Superintendent of Documents, Washington 25, D.C., at 10 cents per copy, or from the nearest field office of the U. S. Department of Commerce.

DEPENDABLE OPERATION OF A-F-C CIRCUITS IN THE FPS-3

by Fred F. Thomas
Philco Site Engineer

In the past there has been considerable trouble in maintaining continuous and reliable operation of the a-f-c circuits in the C-708/FPS-3. This article is intended to point out some of the necessary performance checks ond adjustments that must be made before successful and dependable operation can be maintained. Included also is an alignment procedure that has been used by the author for some time now with very good results.

RELIABLE OPERATION of the C-708/FPS-3 a-f-c circuits can be maintained continuously with a minimum of attention if the alignment procedure and performance checks given below are closely followed.

PRELIMINARY ALIGNMENT CHECKS AND ADJUSTMENTS

Before going into the alignment procedure, it is necessary to point out several items that must be correct before a-f-c operation can be obtained.

The first and most important of these items is the magnetron spectrum of both transmitting systems. The 5J26 magnetron used in the FPS-3 has a major side mode whose peak normally falls approximately 1 megacycle lower in frequency than the main mode. The amplitude of this major side is not only very important to good a-f-c operation, but also affects the capability of the FPS-3 transmitting and receiving equipment. For best results, the maximum amplitude of this mode should not exceed 25% of the main mode. There is a tendency at quite a few AC&W sites to put too much emphasis on the performance figure of a radar set and pay very little attention to the magnetron spectrum. The radar performance figure (ratio of receiver sensitivity to transmitted power, or the sum of the transmitted power in dbm and receiver MDS in dbm) does not give a true picture of radar capability if the transmitter spectrum and the bandpass of the receiving system are not taken into account.

Transmitted power levels of 500 kw to 750 kw peak power from the FPS-3 cannot be considered good performance if the greater portion of this power does not fall within the bandpass of the receiving system. For good a-f-c action and system performance, it is essential that the bandwidth of the spectrum not exceed 0.6 to 0.7 megacycle at the half-power points. A bandwidth of 0.6 to 0.7 megacycle in magnetron spectrum is not too difficult to maintain on the AN/FPS-3. The increase in magnetron consumption in maintaining this performance is not excessive — in fact, as compared to the crystal and TR tube consumption required to maintain a high MDS, the increase in magnetron consumption is negligible. A high MDS is of importance only when all other factors are also considered. Long magnetron life cannot be compared to good magnetron performance.

The second item of importance is the adjustment of the mechanical brake in Stalo O-195/FPS-3.

The O-195/FPS-3 has two braking systems on the tuning motor—a dynamic brake and a mechanical brake. The dynamic brake, which consists of a d-c voltage applied to the tuning motor when the manual tune switch is returned to neutral, operates in conjunction with the mechanical brake when manual tuning of the Stalo is employed. Only the mechanical brake and the lead voltage from the a-f-c unit itself are used when the a-f-c is in operation. The dynamic brake is disconnected from the Stalo unit when the manual-auto-track switch is in either the AUTO or the TRACK position.

Proper adjustment of the mechanical brake must be made when the Stalo is operating in manual tuning, because proper a-f-c operation is impossible until the mechanical brake has been properly adjusted.

Adjustment of the mechanical brake is complicated by the presence of the dynamic braking system. With both brakes operating at the same time, proper operation of the mechanical brake cannot be accurately determined. One method of adjusting the mechanical brake when bench alignment of Stalo Unit OA-195 is being accomplished is to temporarily disable the dynamic braking circuit.

At this site, a single-pole, single-throw toggle switch has been mounted in the individual Stalo units. During bench alignment, the switch is thrown to the OFF position and the brake adjusted to allow the motor to turn two or three complete revolutions after the manual tuning control is returned to neutral. The toggle switch intercepts the power to CR-1701, the dynamic brake rectifier. Proper adjustment of the mechanical brake is possible without interruption of the dynamic brake, but requires more time and experimentation.

Another item which affects a-f-c operation and which should be checked is erratic contacts on K-1308, the 0.5-second time-delay relay in the high-voltage power supply. This relay temporarily interrupts power to the Stalo scan motor transformer primary, thus preventing the Stalo from running away when the automatic reclosure circuit is operated because of misfiring, or when the high voltage is removed from the transmitter. Either of these conditions results in loss of the a-f-c or coho lock pulse. Many sites have reported the failure of this relay in the *Site Engineers Digest*. It has been necessary to replace K-1308 twice in one year at this site. The first replacement was necessitated by complete failure of the relay and the second by erratic operation.

With regard to the a-f-c unit itself (C-708/FPS), before any alignment of the a-f-c circuits is attempted, a thorough check of the upper and under sides of the main chassis should be made, and all loose nuts and screws should be tightened. Many of the tube-socket mounting screws, as well as other mounting screws, are used for mounting grounding lugs in the various circuits. It is necessary that all screws be drawn up as tightly as possible in order to ensure good grounds in the various circuits, particularly the i-f amplifier and discriminator circuits.

All tubes should be checked for transconductance and microphonics. Although it may be a local condition, considerable trouble has been encountered with microphonic 12AT7 and 12AU7 tubes.

All voltages from the a-f-c power supply should be checked.

In damp climates, it is necessary to burnish the pins on K-5201 at least once every six months. Relay F-5201 is a sensitive plate-operated relay and good contact must be ensured. Trouble has been encountered with poor contacts at the socket of this relay even in dry climates such as exist in New Mexico. A little cleaning goes a long way.

A-F-C ADJUSTMENT AND ALIGNMENT PROCEDURE

The only equipment required for alignment of the a-f-c is a VTVM, a small screwdriver, and an 8-inch lead with an alligator clip on each end.

Most of the following alignment and

adjustment procedure can be performed while the equipment is operating. However, final adjustment must be made while the antenna is not rotating.

1. To adjust the Centering and Balance controls, proceed as follows:

a. Disconnect J-5201 and remove V-5210.

b. Connect short lead with alligator clips between TP-5201 and TP-5202. This prevents any unbalance of the balanced modulator due to d-c unbalance of the discriminator and pulse stretcher.

c. Set MANUAL - AUTO - TRACK switch to TRACK position, and turn Servo Gain control, R-5241, fully clockwise (maximum gain).

d. Set the VTVM to the 150 volt, a-c range, and read voltage at pin 1 or 7 of V-5210.

e. Adjust the Centering control, R-5240, for minimum reading on the VTVM (reduce range of VTVM as necessary so that minimum indication occurs near mid-scale). Minimum voltage should be between 5 and 15 volts, a-c.

NOTE: More accurate adjustment of the Centering control can be accomplished by replacing V-5210, thus allowing the Stalo motor to run, and adjusting the Centering control to stop the motor while observing the Stalo frequency dial. TP-5201 and TP-5202 must remain shorted for this procedure. Also, this adjustment must be made during a preventive-maintenance period because the Stalo will be out of tune.

f. Remove the jumper between TP-5201 and TP-5202. An increase of voltage at pin 1 or 7 of V-5210 may be noted. Adjust the Discriminator Balance control, R-5257, for minimum voltage at pin 1 or 7 of V-5210, increasing or decreasing the VTVM range as required. (This step results in an approximate adjustment of the discriminator balance.)

g. Set the VTVM to the +20 volt, d-c range.

h. Connect the high side of the VTVM to TP-5201, connect the low side to ground, and, as accurately as possible, record the voltage. Then, using the same procedure, record the voltage at TP-5202.

NOTE: It is absolutely necessary that the readings at TP-5201 and TP-5202 be taken with the high side of the VTVM connected to these points and the low side connected to ground. Tests made with TS-375, TS-505, and RCA Volt-Ohmyst show erroneous readings if the meter case, common or neutral lead is connected to the test points.

i. Adjust the Pulse Stretcher Balance Control, R-5257, until the two voltages obtained in step h are as close to the same value as possible. Both voltage readings should be between 12 and 15 volts positive.

j. Return the MANUAL - AUTO - TRACK switch to MANUAL. Replace V-5210 and reconnect P-5201.

2. Before aligning the i-f amplifiers and discriminator, accurately peak the Stalo for maximum echo. Proceed as directed below:

NOTE: The i-f strip and discriminator circuits can be aligned more accurately if the antenna is stopped and frequent manual tuning of the Stalo made during alignment of the i-f and discriminator. However, the circuits can be satisfactorily aligned with only one peaking of the Stalo. Very slight readjustment of the 29-mc coil may be necessary if the Stalo is peaked only once.

a. Set the VTVM to the +25 volt, d-c range.

b. Connect the high side of the VTVM to TP-5201 and TP-5202, in turn. Adjust L-5207 and C-5217 for maximum voltage at TP-5201, and adjust L-5208 and C-5218 for maximum voltage at TP-5202. Repeat adjustments until no further increase is noticed. The voltage should read between 18 and 25 volts.

NOTE: While making these adjustments, it may be necessary to readjust the Pre-amp Gain control, R-5202, to a point between limiting and no signal values several times to accomplish accurate sharp alignment.

3. With both sections of the discriminator aligned to the exact i-f frequency, the i-f amplifiers may be aligned. Proceed as follows:

a. Set the VTVM to the +25 volt, d-c range.

b. Connect the high side of the VTVM to TP-5201 or TP-5202.

NOTE: Test jack J-5202 has been provided for alignment of the i-f strip. However, it has been found that much sharper tuning of the i-f coils can be obtained with the meter connected to TP-5201 or TP-5202. Even at these points the tuning of L-5201 is rather broad.

c. Adjust L-5201, L-5203, and L-5206 for maximum voltage. Repeat until no further increase is noted.

d. Repeat step 2b.

4. Before proceeding with the alignment of the discriminator (step 5), it is necessary to explain the function of capacitors C-5217 and C-5218. These capacitors are not intended for the purpose of obtaining resonance of the 29- and 31-megacycle sections of the discriminator. They are strictly coupling capacitors, and are to be adjusted only to obtain dynamic balance of the discriminator.

The Discriminator Balance control, R-5257, is used to obtain static, or quiescent balance, and should not be readjusted after step 1h is completed. Dynamic balance of the discriminator is accomplished with C-5217 and C-5218 as will be indicated later. After steps 2b and 3d are completed, the screwdriver slots in C-5217 and C-5218 should be parallel to the sides of the a-f-c chassis. (A check may be made on the correct setting of these two capacitors by rotating them 180 degrees. The screwdriver slots of the capacitors should be parallel to the sides of the chassis at the point giving maximum voltage at TP-5201 and TP-5202 before proceeding with the alignment.) When further adjustment of these capacitors is necessary, it must be done with a plastic or fiber alignment screwdriver.

5. To align the discriminator, proceed as follows:

a. With the VTVM set to the +25 volt, d-c range, connect the high side of the VTVM to TP-5201. Adjust the pre-amp gain for a reading half way between no signal and limiting. Tune L-5207 for maximum voltage, and then reduce the voltage 2.5 volts by turning the coil slug clockwise.

NOTE: The 2.5-volt reduction detunes the discriminator by 1 mc from the center frequency. The exact value of this center frequency, although theoretically 30 mc, will actually depend upon the individual set.

b. Connect the high side of the VTVM to TP-5202. Adjust L-5208 for maximum voltage, and then reduce the voltage 2.5 volts by tuning the coil slug counterclockwise. Note that L-5208 requires a considerably greater number of turns to reduce the voltage at TP-5202 than was necessary with L-5207 to reduce the voltage at TP-5201. This condition is normal.

c. Turn Pre-amp Gain control R-5202 and Relay Adjustment R-5223 fully clockwise. With fingers of one hand touching the case of K-5201, slowly turn Relay Adjustment R-5223 counterclockwise until relay clicks, and continue counterclockwise a small fraction of a turn. When properly adjusted, Relay Adjustment R-5223 should be from one half to three quarters counterclockwise.

d. Slowly turn Pre-amp Gain control R-5202 counterclockwise until relay clicks again. Then back-up clockwise until relay again energizes, and continue clockwise a small fraction of a turn.

NOTE: When adjusting the Pre-amp Gain control, it is important that both sections of relay K-5201 energize, as normally indicated by a good solid click of K-5201. If the click is light, it may be necessary to continue clockwise rotation of the Pre-amp Gain control until another click is felt. This adjusts the approximate operating point of the Pre-amp Gain and makes possible more accurate alignment of the discriminator.

e. Repeat steps 5a and 5b until interaction of L-5207 and L-5208 is eliminated.

f. After completing step e, record the voltage at TP-5201 and TP-5202 to the nearest fraction of a volt. A difference of as much as 0.5 volt may be indicated at these points.

g. It is now necessary to adjust C-5217 or C-5218 to balance these voltages. If the voltage at TP-5201 was the higher of the two, reduce to the reading at TP-5202 by adjusting C-5217. If voltage at TP-5202 was the higher of the two, reduce to the reading at TP-5201 by adjusting C-5218.

h. Repeat steps 5a, b, e, f, and g, as necessary to eliminate all interaction and exactly balance the voltages at TP-5201 and TP-5202.

i. After step h is completed, repeat steps 5c and d.

6. If alignment was made while the antenna was not rotating and frequent checks of Stalo tuning were possible, turning the MANUAL-AUTO-TRACK switch to the AUTO or TRACK position should result in immediate and accurate locking of the afc. If it was necessary to make the alignment while the antenna was rotating and frequent checks of Stalo tuning were not possible, then slight adjustment of L-5207 may be necessary. This is accomplished as follows: Stop the antenna, and turn the MANUAL-AUTO-TRACK switch to the TRACK position. If the Stalo hunts, turn the Servo Gain Tuning-Servo Gain control, R-5241, counterclockwise. Tune L-5207 for maximum echo signal on "A" scope. Then lock down all coils.

7. A daily check of the Pre-amp Gain control, R-5202, and the Relay Control, R-5223, will ensure continuous operation of the afc. To make this check, use steps 5c and d. Remember that in AUTO position, the a-f-c will lock-in only on an increasing i-f. This occurs when the Stalo is tuning away from the transmitter frequency. If the Stalo is tuned just slightly farther away from the locking point and the switch is thrown to the AUTO position, the Stalo motor will run, hit one or both reversing switches, and again lock-in when the Stalo is tuning to an increasing i-f. In TRACK position, the afc will return directly to the locking point from up to 2 megacycles either side of the locking point. The exact range depends on the accuracy of a-f-c alignment and on the a-f-c sensitivity. If the Stalo is tuned beyond the sensitivity point of the afc, the Stalo motor should not run at all in the TRACK position. Loss of the lock pulse while in AUTO position will result in continuous running of the Stalo motor.

SYMPTOMS OF IMPROPER A-F-C ADJUSTMENT

The following is a list of common indications of improper adjustment and their cause.

SYMPTOM	CAUSE
Stalo lock-in over too broad a range.	Pre-amp gain too high. Mechanical brake too tight. Servo gain too low.
Stalo hunts.	Servo gain too high. Mechanical brake too loose.
Stalo pauses at locking point and then continues to run.	Relay adjust R-5223 not quite far enough counterclockwise. Servo gain too high. Mechanical brake too loose.
Stalo locks in at several places or on wrong beam.	Pre-amp gain too high.
AFC not locking exactly on frequency.	L-5207 (29-mc coil) not properly tuned, or discriminator unbalanced.
Stalo tunes across entire band on track position.	Balanced modulator not centered.

The above procedure has been in use for approximately 2-1/2 years with very good results. The AFC's have required very little attention except for compensating for aging tubes. Alignment of the i-f strip is required only when some component burns out in the i-f strip itself, or once every six months or a year. Slight readjustment of the discriminator, especially L-5207 (the 29-mc coil), is usually necessary when a signal pre-amp is changed.

At this site the AFC is operated in the short scan position because of the variation in amplitude of the lock pulse over wide scan. The MANUAL-AUTO-TRACK should be left in TRACK position except for initial lock-on.

ERRATA

In the September-October, 1957 issue, equation (12) on page 14 should read as follows:

$$\frac{E_o}{E_i} = - \frac{A}{1 + A\beta}$$

In the November-December, 1957 issue, the letter β was inadvertently omitted in two places on page 14. The open loop gain of a system is not referred to as μ or A, but rather as $\mu\beta$ or $A\beta$. Our apologies to the author for this omission.

CONTROL-GRID EMISSION

By T/Sgt Lee R. Bishop
Airways and Air Communications Service
United States Air Force

(Editor's Note: This article deals with a subject about which little information is commonly available. This subject was the basis of the "What's Your Answer?" problem in the May-June, 1956 issue of the TechRep Division BULLETIN, with the explanation given in the July-August issue.)

ALTHOUGH MANY PEOPLE feel that control-grid emission is a problem which lies in the realm of the design engineer, it is actually a fairly common maintenance problem without even being recognized as such. Since control-grid emission is dependent upon the vacuum tube itself, its effects are usually cured by tube replacement without a thought ever being given to the actual type of tube defect. As will be shown, control-grid emission can be a source of hard-to-solve trouble that tube replacement will not always cure. It is for this reason that some time spent in reviewing its cause and effect can prove to be time well spent.

The two types of grid emission likely to present a service problem are primary (thermionic) emission and secondary emission. Primary emission of electrons takes place when the grid is heated, and secondary emission is caused by ion or electron bombardment of the control grid. Secondary emission is the most troublesome type, and is, therefore, of greatest interest to the serviceman. Considering any type of control-grid emission, it should be remembered that, as a rule, the grid is the most negative element in the tube. Therefore, it can and will act as an emitter unless steps are taken to guard against such action. The effect of any type of control-grid emission is to subtract from the operating bias of the tube by an amount equal to $I_g \times R_g$, as shown in figure 1.

Primary, or thermionic, emission is encountered mainly with small oxide-coated-cathode receiving tubes and is caused by evaporation of some of the cathode material and resultant condensation onto the grid while the tube is operating. In addition, the elements

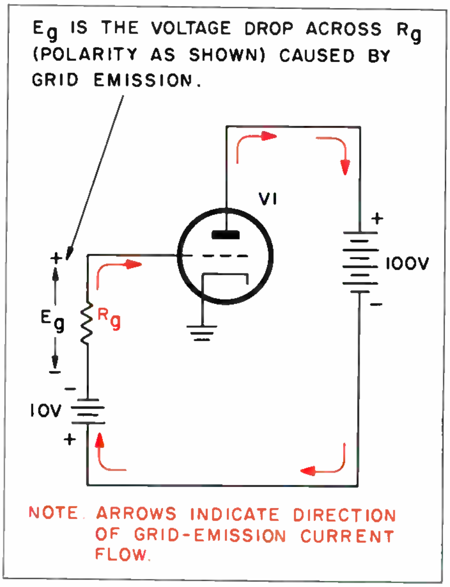

Figure 1. Loss of Bias Caused by Control-Grid Emission

of these tubes operate at high temperatures because of close grid-cathode spacing and high electrode dissipation in relation to element size. Primary emission in itself is seldom of sufficient magnitude to cause trouble, because steps are taken in the design of the tube to make the grid a poor emitter and to assist in heat dissipation. These steps consist of providing adequate heat dissipation, careful selection of materials used in grid construction, and coating the grid with a poor emitting substance. These same steps also serve to reduce secondary emission. Gold plating is very effective in making the grid a poor emitter, and is frequently used when the requirement for low grid emission is severe enough to warrant the cost. For example, the 6J4, a low noise tube designed for r-f amplifier service, has a gold-plated grid. Coating the control grid with powdered boron-carbide is another method used to reduce grid emission in less severe applications.

Secondary emission, as previously mentioned, is of primary interest to the serviceman and is generally caused by ion bombardment of the control grid. In cases where the grid swings positive on a signal alternation, emission can also be caused by bombardment of the momentarily positive grid by the negative electrons.

In the tube manufacturing process, the materials used are carefully degassed by means of a heating process just before sealing of the tube envelope. The bulb is then evacuated and sealed, and the getter blown. In spite of all such precautions, a certain amount of gas remains in the envelope. As the tube ages, more gas is driven out of the cathode and plate materials, the actual amount being determined by the quality of materials, the materials themselves, and the techniques used in the manufacturing process.

Molecules of this gas are struck by electrons on their way to the plate from the cathode. In striking the gas molecules, these electrons dislodge an electron, leaving a positive nucleus, or ion. By nature of its charge, this ion is then attracted to the negatively charged grid. When the ion collides with the grid, one electron must flow from the power supply to neutralize this ion. Actually, when an ion strikes the grid it may dislodge several electrons, depending upon its velocity and how good an emitter the grid is. This increases the current through R_g, and hence increases E_g. It should be realized that all of the electrons so dislodged will not reach the plate, but will, in a random fashion, collide with and neutralize some of the positive ions formed by earlier collisions. Since these collisions cause a random current flow through R_g and hence a random voltage (i.e., noise at the grid), the reason for gold plating the grid of the 6J4 to reduce grid emission becomes clear. It can also be seen that gassy tubes can be the cause of a poor signal-to-noise ratio in a receiver.

An actual case of grid emission due to gas content was encountered in connection with the coaxial-line driving circuit shown in figure 2. It proved to be an extremely trying trouble, the cure of which necessitated circuit modification.

The 6Y6 output tubes had been in service about one year when loss of regulation occurred in the —150 volt power supply which furnished the operating potential for these tubes. The 6Y6 tubes seemed unusually hot, so a plate-current check was made to see whether they were overloaded. The check showed that the tubes were drawing excessive plate current. Bias was subsequently checked at point A in the circuit, but it proved to be correct. New 6Y6 tubes were substituted, but the trouble still existed. A more thorough check of the circuit revealed that the actual bias between grid and cathode was —2 volts instead of the —20 volts set by the divider.

Finally, the voltage across R_g was checked and found to be 18 volts, with the same polarity as the voltage drop

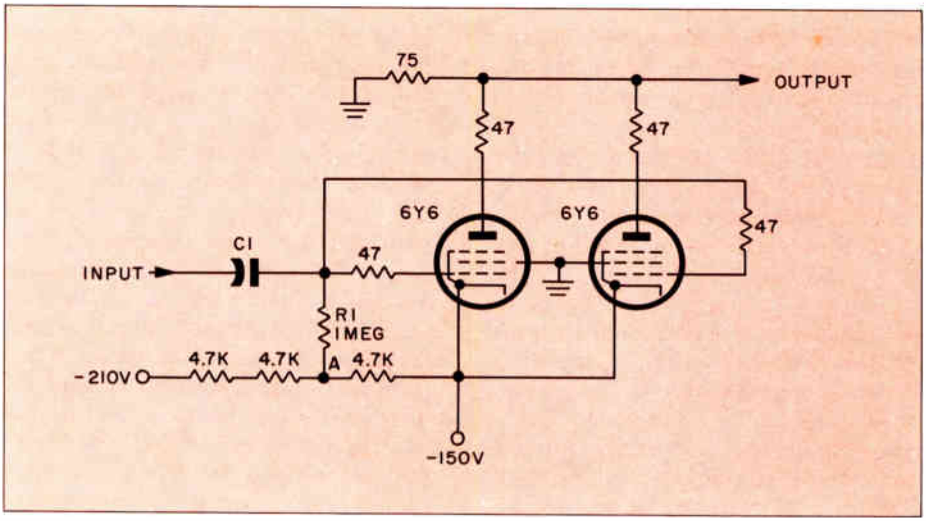

Figure 2. Coaxial-Line Driving Circuit

across R_g in figure 1. It was then realized that 18 microamperes of grid-emission current was flowing through the 1-megohm grid-leak resistor, R1, and creating this drop. Inasmuch as a replacement tube could not be found to work, it was reasoned that the manufacturer had changed techniques or materials on the replacement tubes, thereby permitting a higher gas content.

Changing R1 from 1 megohm to 100K reduced the drop across R1 from 18 to 1.8 volts. This modification proved to be the ultimate solution, allowing the circuit to operate with all makes of 6Y6 tubes.

"What's Your Answer?"

The problem given below illustrates a situation which often occurs in electronics—that of the loading effect of a voltmeter.

Two resistors of unknown value are connected in series across a 120-volt battery which has negligible internal resistance. A voltmeter having a resistance of 20,000 ohms is used to measure the voltage across first one resistor and then the other. The first reading is 30 volts and the second is 50 volts. What are the values of the two unknown resistors?

POWER SUPPLY FOR THE GPS-T-2

By Albert C. Picarel
Philco TechRep Field Engineer

The high-voltage power supply described in this article was constructed to replace the original power supply for the GPS-T-2 trainer and overcome the problems of unreliability and instability inherent in the original. Results obtained over a period of approximately one year indicate that the replacement power supply provides substantial improvement in the operation of the equipment.

It has been known for some time that the high-voltage power supply furnished as a part of the GPS-T-2 trainer does not have the reliability and stability required for prolonged periods of operation. Numerous simulated training program (STP) missions have been aborted as a result of the failure of this particular power supply.

The major cause of the large number of failures occurring in the original power supply appears to be an unfortunate tendency towards overheating. The metal top plate becomes too hot to touch, attesting to the high temperature within the unit. In addition, when a breakdown does occur, the inadequate provisions for maintenance require that the unit be sent to a depot repair shop. A delay of up to several weeks may then result before a replacement unit is made available. During this time the equipment stands idle, and hours that could be utilized for the simulated training program are effectively wasted insofar as training is concerned.

In an effort to overcome some of the problems caused by the failure of the original power supply, it was decided that the best approach would be to construct a replacement power supply which would embody the principles utilized in the normal television receiver. The construction, of course, was subject to the restrictions that the unit fit in the same space and provide the same output voltages as the original power supply. The unit which resulted from the project incorporates a standard television horizontal multivibrator circuit and other basic circuits which are used throughout the television industry.

Although certain minor changes were required, the circuits and their operation should be familiar to all television technicians.

The unit is built in an adequately ventilated rectangular housing of the same outer dimensions as the original unit, thus meeting the requirement that it fit in the same space as the original power supply. Also, the output voltages are identical to those of the original power supply. The top plate is hinged to allow ready access to tubes, test point jacks, and components located on the upper chassis base. Figure 1 shows the unit installed in the equipment, and figure 2 is a top view showing the hinged cover in the open position. This hinged cover consists of a solid aluminum radiation shield mounted on a one-half inch thick bakelite sheet used for high-voltage insulation. The supply is readily accessible for maintenance, and also can be inspected during normal preventive maintenance periods. As a safety measure in protecting personnel who are not

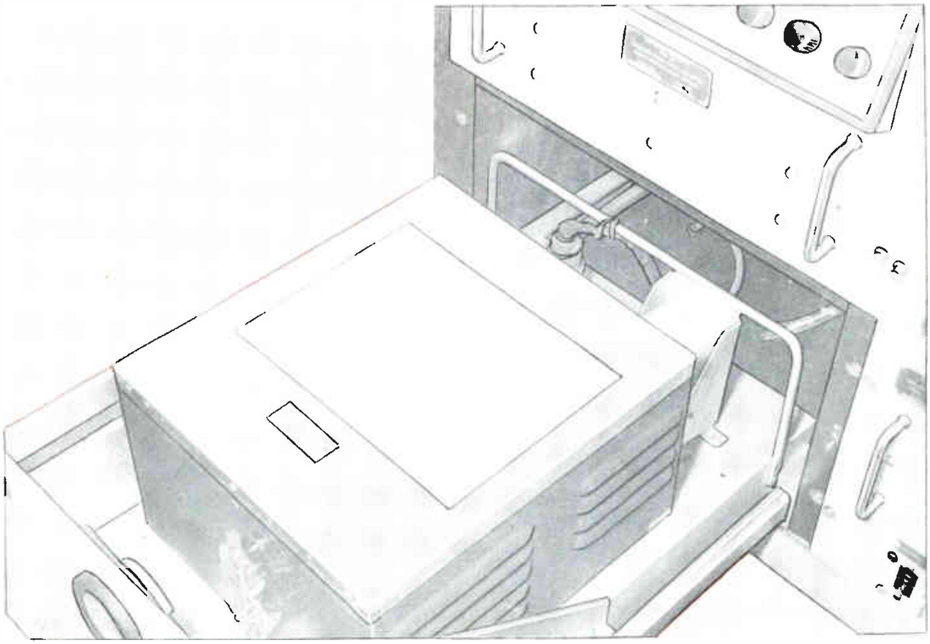

Figure 1. High-Voltage Power Supply Installed in Cabinet (Drawer in Open Position)

familiar with the voltages which are found in this type of power supply, an interlock switch effectively kills the high voltage and associated circuits when the cover is opened.

Figures 3 and 4 show the wiring and component placement at the bottom of the unit. A terminal strip, located on the front housing plate and visible in figure 3, was incorporated to eliminate the necessity of soldering the external connections (a definite drawback on the original unit). A pocket screwdriver is the only tool needed to remove or install the unit.

The circuit design centers around a flyback transformer of the type generally found in color television receivers. This transformer should be capable of delivering the 25 kv required for the high-voltage output. A Philco transformer, part number 32-8817, may be employed in the circuit as shown. Other transformers can be used, but may require minor modifications in circuitry and component values.

The flyback transformer is driven by a 6CB5 horizontal output tube. The 6CB5 is a newer and improved version of the 6BG6, and pin connections are identical. Since the current drain of the 6BG6 is less than that of the 6CB5, the two tubes can be directly interchanged, and both types have been used in this unit. There is no appreciable difference in performance, but the maximum high-voltage output is somewhat more stable when the 6CB5 is used.

A schematic diagram of the power supply is given in figure 5. The multivibrator uses a 6SN7 dual-triode tube, and is free-running at a frequency of approximately 15,750 cps. However, since the frequency is not critical in this case, and since the frequency may be audible and annoying to some persons with a keen sense of hearing, provision is made to vary the bias on the oscillator-discharge section by means of the 1-megohm potentiometer. By adjusting this potentiometer, the frequency can be raised above the audible

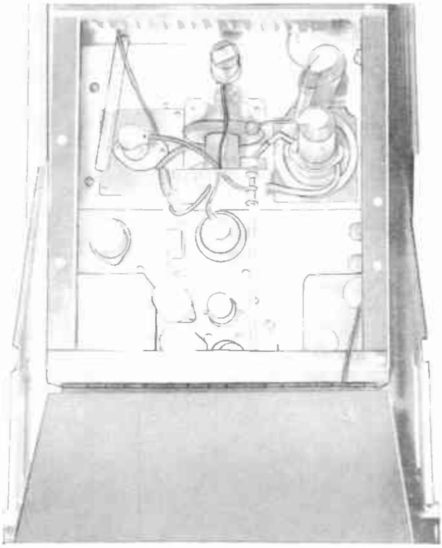

Figure 2. Supply with Top Lid Open to Expose Components and Tubes on Upper Chassis. Note the locally fabricated tube bases.

range. Although increasing the frequency decreases the maximum high voltage somewhat, there is no appreciable effect on performance.

The r-f potential developed by the flyback transformer is rectified by the 3B2 high-voltage rectifier tube. This tube is a heavy-duty version of the 1B3-GT, which should be familiar to television technicians.

The particular transformer used also features a provision for controlled focus take-off. The r-f voltage is tapped off at a lower potential and fed through a control potentiometer to the 1B3-GT focus rectifier stage, where a positive voltage of 5500 to 7500 volts is developed. The added feature of being able to adjust the focus potential in this power supply provides a means of controlling, if necessary, the sharpness of the sweep trace on the flying-spot scanner tubes without interruption during the course of an STP mission. It should be noted that this focus voltage control should be kept near the low end of its range to avoid exceeding the ratings

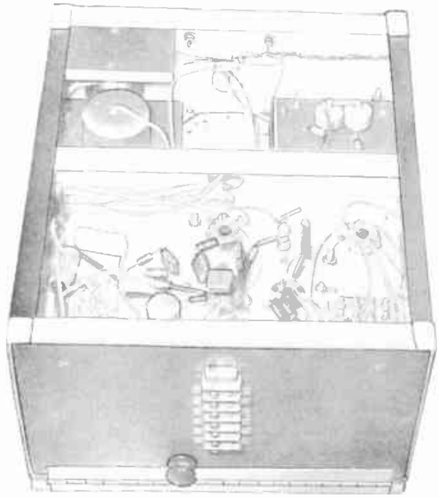

Figure 3. Underside View of Chassis. Note the terminal strip and focus control knob.

of the 6CB5. Proper adjustment of the focus is achieved by watching the target echoes on the monitor scope while adjusting the focus potentiometer.

A beam-power 6BK4 regulator tube is used to control the 25-kv anode voltage. It is shunted across the output, and effectively stabilizes the potential at the desired level. Varying the bias on the tube provides the necessary controlling action for this stage.

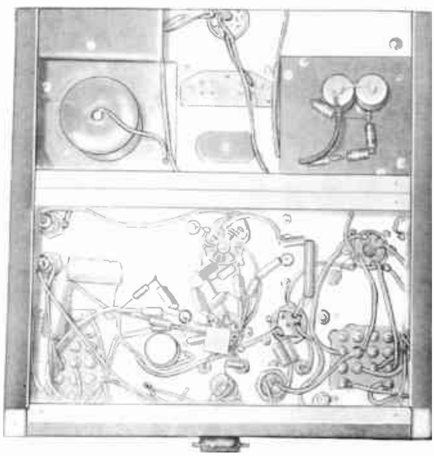

Figure 4. Bottom View Illustrating Location of Components

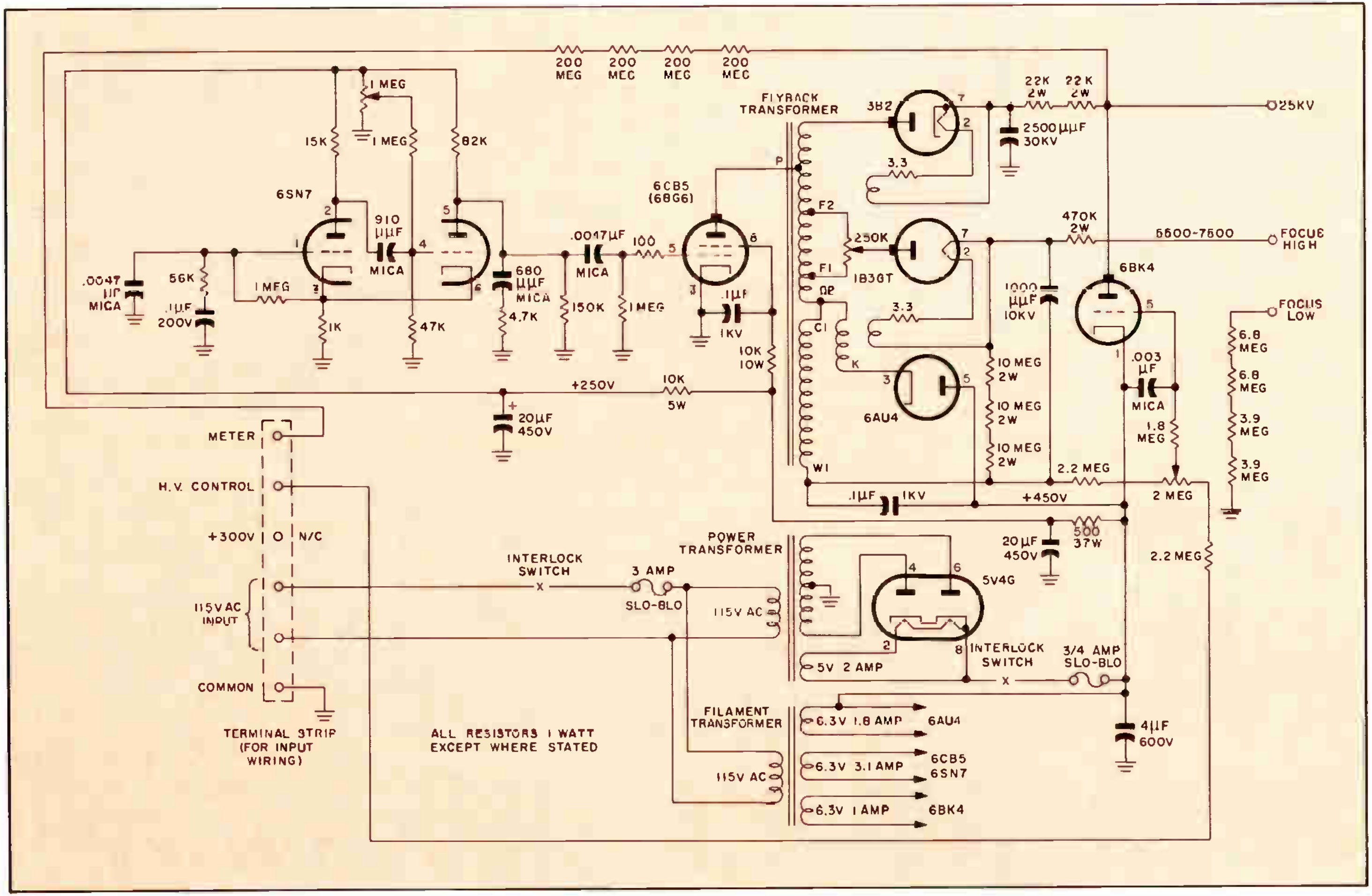

Figure 5. Schematic Diagram of High-Voltage Power Supply

A word of caution to anyone who contemplates building this power supply — the high-voltage section must be thoroughly insulated to guard against arcing and corona effect. This is especially true with regard to the high-voltage-rectifier tube socket. It is recommended that a commercial type socket be used if at all possible. If not, extreme care should be exercised in prefabricating a tube-socket base. The insulating material should have a puncture voltage rating of at least 54,000 volts. Also, the power supply should have adequate shielding to eliminate r-f radiation.

The low-voltage portion of this power supply employs a conventional full-wave rectifier circuit. A 5V4-G rectifier tube was used because of its cathode take-off arrangement, which allows a gradual build up of the low B plus potential. This eliminates the possibility that any sudden surge will be thrown on the circuitry when the power is initially applied.

A separate filament transformer was used to provide isolation for the filament circuits. It should be noted that the shunt regulator and damper tubes have their own independent filament windings. This feature prevents interaction of circuitry in the event of a tube short. The damper heater circuit is placed at a +450-volt potential to minimize the possibility of heater-to-cathode breakdown. The input labelled "H.V. CONTROL" serves to complete the return to ground, and the H.V. CONTROL potentiometer (2 megohms) allows adjustment of the high voltage. This potentiometer, which is located near the external connection terminal strip, is adjusted for a correct reading on the meter located on the front panel of the high-voltage drawer.

Most of the parts needed for this power supply were obtained from service stock, and the remainder were acquired through authorization of local purchase.

The power supply has been tested by the squadron and found to perform very satisfactorily. All STP missions conducted with this power supply have evidenced a degree of stability in performance that has not been encountered in some time.

PRINCIPLE OF DUALITY

By H. W. Merrihew
Headquarters Technical Staff

In GEOMETRY a new theorem may often be derived from a known theorem simply by interchanging two key words, or terms, in the known theorem. For example, consider the theorem, "Any two points have one and only one straight line in common." If the words *straight lines* are substituted for the word *points,* and the word *point* for the words *straight line,* the given theorem becomes: "Any two straight lines have one and only one point in common." Both theorems are obviously true. The two words, or terms, that are interchanged (*point* and *straight line* in this case) are said to be *duals* of each other. Therefore, the principle on which this method of derivation is based is called the *principle of duality.*

The principle of duality is frequently employed in mathematics to create new equations. For example, given an equation or group of equations, the quantities in the given equation are replaced by duals to produce new equations which are analogous to the given equations but represent the relationship between the dual quantities. Duality may be used in geometry to create new theorems based on formulations which bear a dual, or reciprocal, relationship to a given theorem, and because the original theorem is true the dual theorem is also true.

The principle of duality was first stated by Gergonne in the period 1825 to 1827 and by Poncelet in 1829. However, hints at duality appear in writings of Pappus (about 300 A.D.). The theorem of Pappus states: Given two straight lines A and B, in a plane, choose any three points (p_1, p_2, p_3) on A and any three points (p_4, p_5, p_6) on B. If these points are joined as shown in figure 1, the three points of intersection (p_7, p_8, p_9) are also on a straight line C.

It should be observed in the figure that the points, in groups of three, are associated with three pairs of straight lines. Thus the hint at duality is present; however, it went unrecognized as a principle until the nineteenth century when the following dual theorem was stated: Given any two points, draw any three straight lines through the first point and any three straight lines through the second point. The three straight lines joining certain pairs of points all pass through the second given point.

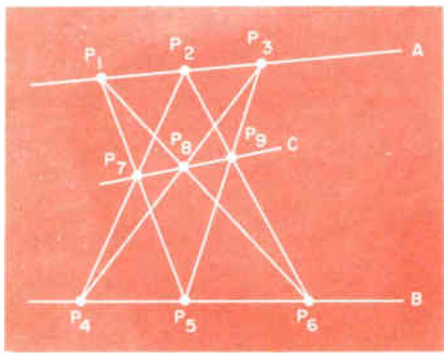

Figure 1. Theorem of Pappus

The fact that the principle of duality can be extended to mathematical equations is exemplified by the pairs of equations, given in Table 1. The pairs of equations designated a, b, c, and f represent relationships between electrical quantities, whereas the pairs marked d and e represent relationships between mechanical quantities. Equations are said to be the duals of one another when, by a change of symbols, the equations can be made identical. For example, the equations given under

TABLE 1

Pair Desig.	*Given Equation*		*Dual Equation*	
a	$E = RI$	(1)	$I = GE$	(2)
b	$I_T = I_1 + I_2 + I_3$	(3)	$V_T = V_1 + V_2 + V_3$	(4)
c	$Z = R + j\omega L - \frac{j}{\omega C}$	(5)	$X = G + j\omega c - \frac{j}{\omega L}$	(6)
d	$F = DV$	(7)	$V = \frac{F}{D}$	(8)
e	$F = M \frac{dv}{dt}$	(9)	$V = \frac{1}{K} \frac{df}{dt}$	(10)
f	$e_L = L \frac{di}{dt}$	(11)	$i_C = C \frac{de}{dt}$	(12)

c are of the same form, and the given equation

$$Z = R + j \quad L - \frac{j}{\omega C}$$

can be transformed to its dual by making the following substitutions:

Y for Z
G for R
C for L
L for C

Hence, by a mere change of symbols, equation (5) is converted to equation (6). The networks represented by equations (5) and (6) can also be considered as duals of one another. Figure 2A illustrates the network represented by equation (5), and figure 2B the network associated with equation (6). Networks are the duals of each other when the node equations of one network are of the same form as the loop equations of the other network.

The loop equation for the network of figure 2A is:

$$e = v_L + v_R + v_C$$

and the node equation for the network of figure 2B is:

$$i_T = i_L + i_R + i_C$$

Therefore, the networks are duals of each other. It must be noted that the networks are not equivalent circuits; that is, the dual relationship holds at any frequency.

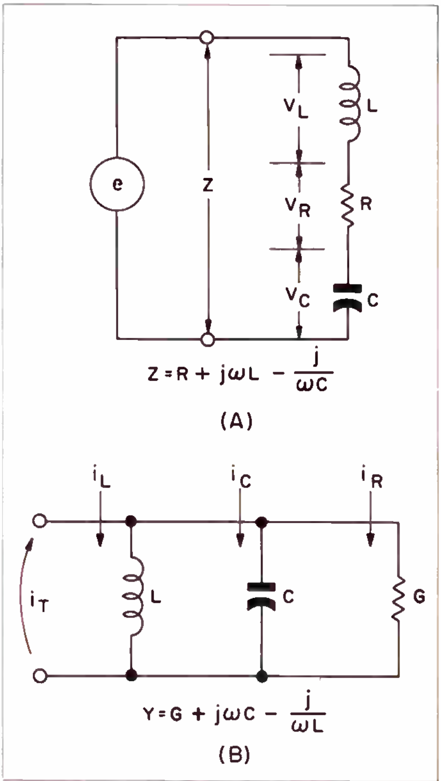

Figure 2. Dual Networks

Another feature of duality arises when components or devices are considered. For example, components or devices are said to be duals if the current in a given component or device reacts in the same way as the voltage in another component or device. Therefore, it can be seen, from equations (11) and (12), that the current in the inductor behaves in the same way as the voltage across the capacitor. Hence, inductance and capacitance are duals of each other. The reader can, with a little thought, supply many more instances of duality in electrical networks or devices.

In Table 1 equations (7) and (8) are given as the duals of one another. Equations (7) and (8) relate to mechanical quantities wherein:

F = damping force
D = damping constant
V = velocity

Hence, it can be seen that dual relationships exist between mechanical quantities. For example:

V is the dual of F.

$\frac{1}{D}$ is the dual of D.

F is the dual of V.

In equations (9) and (10) another dual relationship between mechanical quantities is given in these equations:

V is the dual of F.

$\frac{1}{K}$ is the dual of M.

F is the dual of V.

In equations (9) and (10):

M = mass
K = spring constant

Therefore, it can be said that duality is a form of analogy based upon physical systems of the same type; that is, duality exists in electrical systems or mechanical systems. Between mechanical and electrical systems there exist analogies that are frequently employed; however, these are not considered dual relationships.

To see how duality comes about, consider the following electrical equations:

$$e = iR \tag{13}$$

$$e = L\frac{di}{dt} \tag{14}$$

$$e = \frac{1}{C}\int i\,dt \tag{15}$$

Equations (13), (14) and (15) are all solved in terms of e; equation (13) is the expression for the voltage across a resistor, equation (14) is the expression for the voltage across an inductor, and equation (15) is the expression for the voltage across a capacitor. If each of the three equations is solved for current, a set of dual equations result, for example:

$$i = \frac{e}{R} \tag{16}$$

$$i = \frac{1}{L}\int e\,dt \tag{17}$$

$$i = C\frac{de}{dt} \tag{18}$$

Observe that equations (13) and (16), equations (14) and (18), and equations (15) and (17) have the same form; hence it can be seen that:

i is the dual of e.

C is the dual of L.

$\frac{1}{R}$ is the dual of R.

The advantage gained by the use of the principle of duality is that for a given network or device the equations and behavior of its dual can be obtained by appropriate substitution of dual quantities in the data available about the given network or device. For example, consider the following statement relating to a series resonant circuit: "In a *series* circuit at resonance, the *current* is maximum, the *impedance* is minimum, and the *voltage across* the reactive elements is maximum." The insertion of dual relationships in the above statement modifies it to read: "In a *parallel* circuit at resonance, the *voltage across* the circuit is maximum, the *admittance* is minimum, and the *current through* the reactive elements is maximum." It can be seen that duality is especially useful when the operation of a given device or circuit is known and it is desired to determine how a dual device or circuit operates. The dual quantities can be inserted in the statement of operation for the original circuit, and the result is a statement of operation for the dual circuit.

Table 2 contains a list of electrical quantities, networks, and devices which meet the requirements of duality. This table will serve to summarize the electrical items, presented in this article, which are related in a dual position.

TABLE 2

Item	*Dual*
Current (I)	Voltage (E or V)
Resistance (R)	Conductance (G)
Impedance (Z)	Admittance (Y)
Reactance (X)	Susceptance (B)
Inductance (L)	Capacitance (C)
Series	Parallel
Loop	Node
Tube	Transistor

LOGARITHMIC I-F AMPLIFIERS

PART 1

by Donald R. Taylor, Jr.
Philco Research Division

This first of two articles provides an introduction to the subject and treats the theory of operation. Receivers having an output which is a logarithmic function of the input have now been known and used for several years. Much of the early work in this field was done at Philco, one of the major contributions being the use of the same tubes for amplifiers and detectors, thus obtaining the desired characteristics with little added complication.

INTRODUCTION

In this series of two articles, a logarithmic-type receiver for radar use will be discussed. The attention is specifically directed toward that form of receiver which utilizes detection in the i-f stage itself. A discussion of the methods of obtaining the logarithmic performance, the theory of operation, and the associated practical considerations will be given. This discussion should be useful to those interested in the application and construction of this form of receiver.

For many radar applications it is desirable to have a receiver with a non-saturating response that is capable of handling input signals of widely varying amplitudes. A conventional linear receiver has a rapidly rising, early saturating characteristic which, though yielding good contrast over a limited range of input, gives a saturated output for stronger signals. All tonal graduations in intensity due to relative size, distance, and character of targets are lost in this overload region. Land mass, clutter, and other background return obscure targets by saturating the receiver. Response of the logarithmic type can be obtained to handle a very large range of input signals and present the information usually lost in the overload region of conventional i-f strips.

A logarithmic response may be obtained in several ways. The earliest type of logarithmic receiver provided a non-saturating i-f input signal to the second detector. To prevent overload at the i-f stages by large signals, crystals suitably biased were placed across the tuned circuits to increase the damping with increased signal strength. To obtain a good input-output characteristic, it was necessary to establish rigid specifications for the characteristics of the crystal damping units. It was sometimes found necessary to incorporate the crystals in temperature-controlled packages to ensure uniform performance. For those cases where the output i-f signal is to be unsaturated, this method is instantaneous and suitable; but for cases where interest is essentially in the detected video signal, less complicated ways are available. The successive detection method is the easiest to handle and has been used before. Either crystal detectors or vacuum diodes are used for detection of signal at each i-f stage. As the signal strength increases earlier detectors contribute. The summation of outputs then does not saturate over wide ranges.

The method of building a lin-log receiver described here makes use of the successive detection idea, but eliminates the need for additional detection devices. No tubes are required except those normally used in the linear amplifier; the i-f stages themselves supply the desired signal. This results in considerable circuit simplification and conveniently allows both linear and logarithmic outputs to be provided.

The potentialities of the logarithmic receiver have never been fully realized because of the serious restrictions placed on it by the indicator. Since the indicator is able to handle only a small dynamic range, the log receiver output must be compressed. The result is a great reduction in output contrast between adjacent signal strengths at the input.

THEORY OF OPERATION

The range and shape of the transfer characteristic of a linear receiver is determined principally by the saturation of the last i-f stage. Beyond the limiting level of this stage no increase in output results from an increase in input signal. It is always possible, however, to obtain an unsaturated signal by going back to some previous stage where the signal has not received sufficient amplification to overdrive the stage. By arranging to pick off the signal at some unsaturated point back down the line, the limiting of later stages becomes unimportant. To accomplish this, detectors can be stationed at each stage along the i-f strip, and the outputs can be added so that for any level of input signal there is always an unsaturated component of output.

The detectors used may be ordinary diodes. It is much more convenient and efficient, however, to use some nonlinearity in the stage itself to provide the detected voltage. The nonlinear nature of the amplifier-tube transfer characteristic produces demodulation in the plate current on large signals. When the driving signal becomes sufficiently large, the grid-cathode circuit goes into operation as a diode.

Grid-Cathode Rectification

The grid-cathode circuit of the i-f amplifier stage constitutes a convenient diode for detection purposes. Operation is similar to that of a biased diode, since the potential developed across the cathode resistance by the tube current prevents conduction from cathode to grid until the swing of voltage is sufficient to override it. When overriding occurs, a detected signal appears across any video impedance located in the cathode and grid circuits. When this

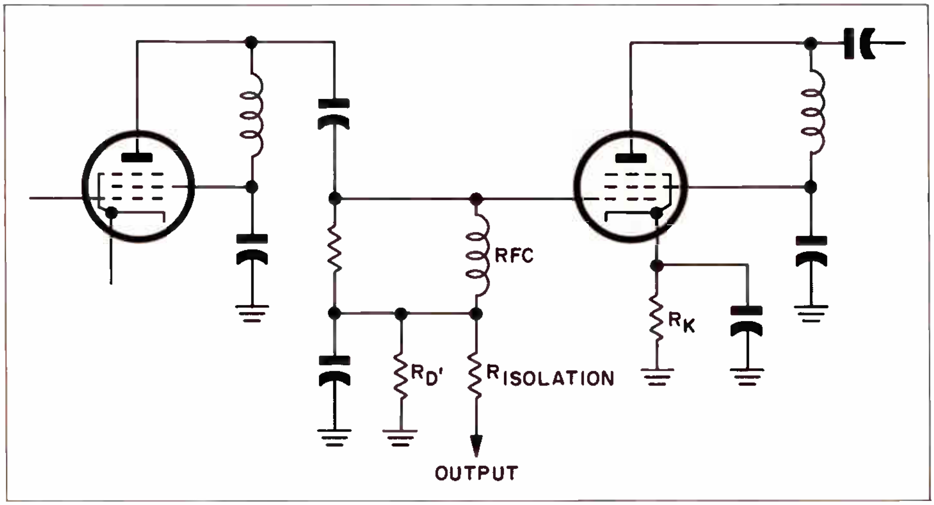

Figure 1. Typical Grid Detection Stage

rectification action is employed, the video impedance in the cathode circuit is usually made as low as possible, and the signal is picked off across a video load placed in series with the i-f impedance in the grid circuit. A typical stage in which grid-cathode rectification is used is shown in figure 1. In order to prevent attenuation of the detected signal, while still affording r-f filtering, an r-f coil is usually placed between the grid and the video diode load. This coil offers a high impedance to the rf and a low impedance to the video voltage. A plot of the detection characteristic of this circuit is shown in figure 2. Over a portion of the lower range the curve is approximately linear, but over most of the range it is logarithmic in nature. The saturation of the curve is dependent on the preceding stage, as is the shape of the curve to some extent.

grid swings. At larger signal input levels grid-cathode conduction begins and contributes even more output of intelligence. At this point the demodulating action is a combination of nonlinearity detection and grid rectification.

With a 100-percent-modulated signal, the amplitude of detected voltage at the cathode is shown in figure 3 plotted against the log of the amplifier input voltage. The curve is substantially logarithmic over a considerable range of input. After rising to a maximum, it then levels off to a constant amplitude. Beyond this point no increase is obtained since the intelligence between cutoff and saturation remains the same for a 100-percent-modulated signal. A similar curve for the signal at the plate is obtained, but the signal is of larger amplitude.

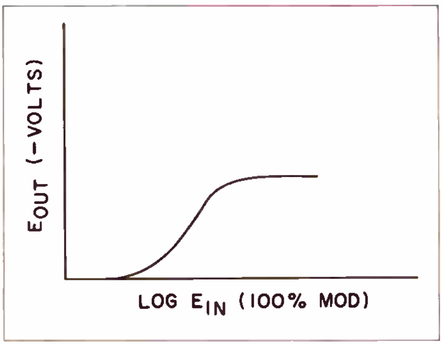

Figure 2. Grid-Detection Characteristic

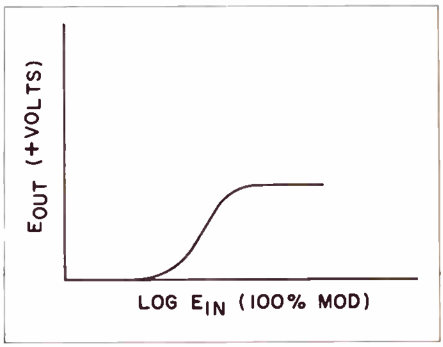

Figure 3. Cathode-Detection Characteristic

Plate Current Demodulation

Because the e_g-i_p curve of the tube is not linear, but curved to some extent, some demodulation of the i-f signal takes place with signals of appreciable magnitude. The demodulated component of signal appears in the plate current so that in both the cathode and the plate circuits a detected voltage is found. As the driving signal increases, the detected signal increases because the e_g-i_p curve is even less linear for larger

Detection Nomenclature

As an i-f stage is progressively overdriven, a signal appears first in the cathode and plate circuits and later in the grid circuit when the signal input has increased even more. At high levels of signal, then, a detected voltage appears at all these tube elements. When a logarithmic receiver is said to function by grid, cathode, or plate detection, the nomenclature serves only to indicate from which tube element the signal is removed. Grid detection functions by virtue of grid-cathode rectification, while plate detection and cathode detection function by virtue of both grid-cathode rectification and demodulation caused by tube nonlinearity. The output polarity is positive for cathode detection and negative for plate and grid detection.

Formation of the Complete Curve

If several i-f stages with similar characteristics are cascaded, and their individual detection characteristics are referred to the input of the amplifier, they appear as in figure 4. The lateral displacement of one characteristic from another is due to the gain between stages. In logical order, the last stage develops output and saturates first, with the others following in their order from output end to input.

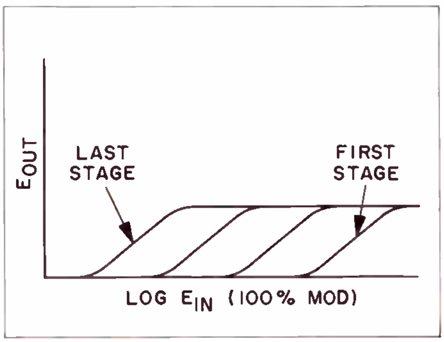

Figure 4. Characteristics of Stages in Cascade

These outputs of the individual stages can be removed from the circuit element at which they appear and added together arithmetically to form a total curve possessing the same logarithmic character. Removal of signal is effected through isolation resistors, and addition is accomplished by applying the signals to a common load. In order to compensate for the envelope delay in the amplifier, a delay is inserted between the stage outputs so that the video signals will coincide in phase and thus add directly in the common load at the end of the line. Assuming correct overlap of the individual curves, the addition forms the over-all curve shown in figure 5. The over-all characteristic has the same slope as that of the individual curves with saturation extended to a high level of input. By correctly overlapping the individual detection characteristics, the over-all curve can be made to approach a straight line.

Log Range and Dynamic Range

It is advisable at this point to distinguish between log range and dynamic range.

The range of input, expressed in db, over which the input output characteristic is approximately logarithmic is termed the log range. The degree to

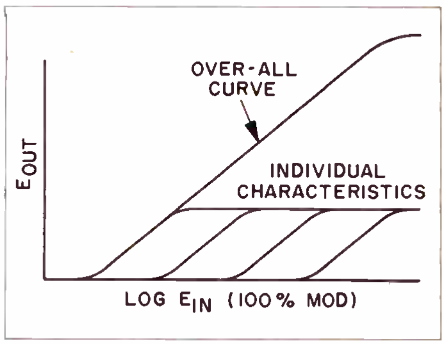

Figure 5. Addition of Characteristics to Obtain Over-all Curve

which the curve approaches logarithmicy should be specified. For this purpose it may be given in terms of percent deviation in slope. The maximum limit of the range is that from noise to saturation.

The output dynamic range is the range, in db, of the output voltage from minimum perceptible signal to saturation. In this case, lower practical limit is the noise level. This range from noise to saturation is important when designing the receiver because of the restrictions on the range that the indicator imposes.

Effect of the Gain Control

The effect of the i-f gain control on the receiver transfer characteristic is similar to that for a linear receiver. Varying the gain — which is done ahead of the log stages — moves the curve laterally along the input axis. Since the noise at the input remains at a constant abscissa and the saturation level of input moves along the axis, increasing the i-f gain causes the log and dynamic range to contract, and decreasing the i-f gain causes the log and dynamic range to increase. See figure 6. It thus becomes necessary to specify the gain setting for measurement of log range. The gain control setting is usually specified in terms of a certain noise level at the receiver output.

Effect of Percent Modulation of the Signal

For signals having a low percentage of modulation, the individual detection characteristic goes through a maximum and then falls toward the axis again, as shown in figure 7. The reason for failure to maintain output in the usu-

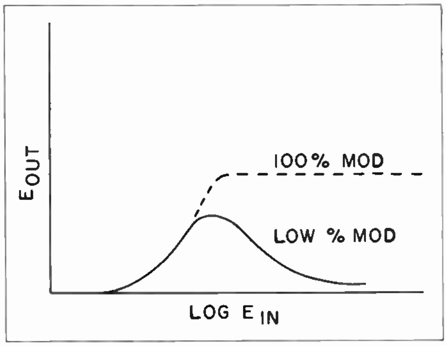

Figure 7. Detection Curve for Low Percentage of Modulation

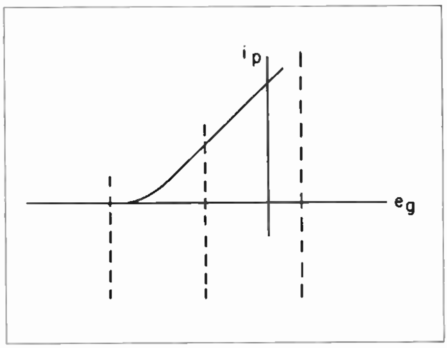

Figure 8. Conditions Existing for High-Amplitude 100-Percent-Modulated Signal

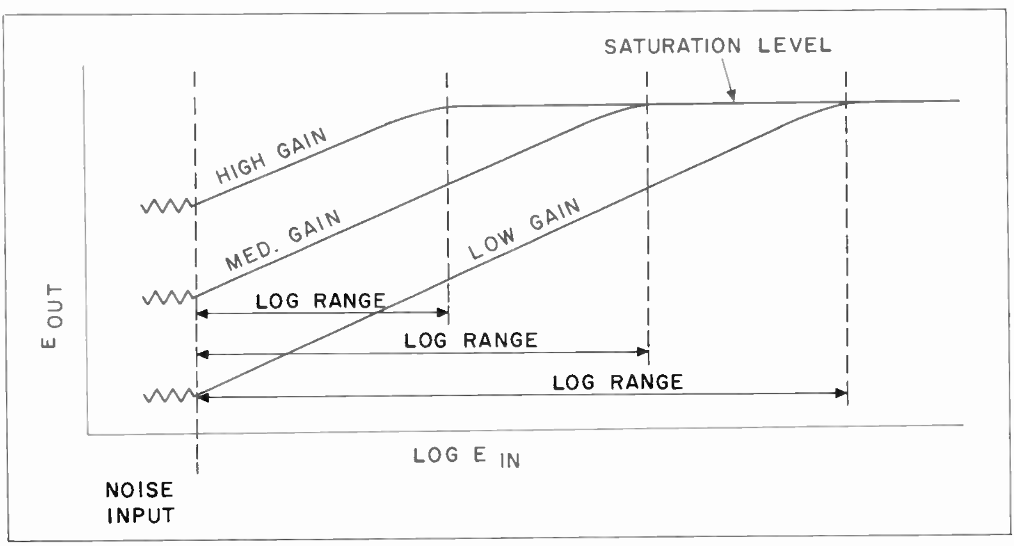

Figure 6. Effect of Change in Gain on Log Range

ally flat portion of the curve may be seen by considering the signal presented to the grid of the stage as viewed on the e_g-i_p curve. In figure 8, a 100-percent-modulated signal of saturation level is shown. It can be seen that as the signal swings further and further beyond cutoff and positive saturation, no decrease in signal content occurs. When the signal swings beyond the limiting regions, the amount of intelligence contained in the useful grid range remains constant.

Figure 9 shows a low-percent-modulated signal of high level on the e_g-i_p curve. As the signal increases from small levels, the output increases to the point where the peaks of the signal enter cutoff and positive saturation. Beyond this point the intelligence contained in the envelope begins to be lost. Less and less intelligence is contained in the useful grid swing region until finally nothing but a c-w component remains there. At this ultimate point, the output has dropped to nothing.

The effect of the falling individual characteristics on the over-all curve is shown in figure 10. For the lower-percentage modulations the curve displays a rising and falling characteristic after passing the first peak of signal. As the percent modulation increases the variations become smaller in amplitude, and the point at which the curve stops rising initially moves to higher inputs.

The variation in log range and dynamic range with percent modulation is evident. For circuit test purposes, the

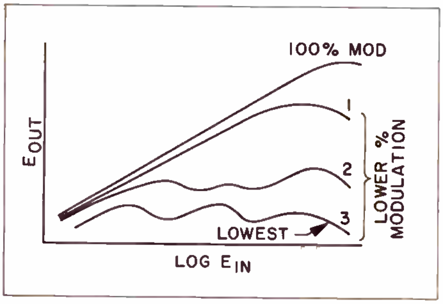

Figure 10. Log Curves for Various Percentages of Modulation (Complete I-F Amplifier)

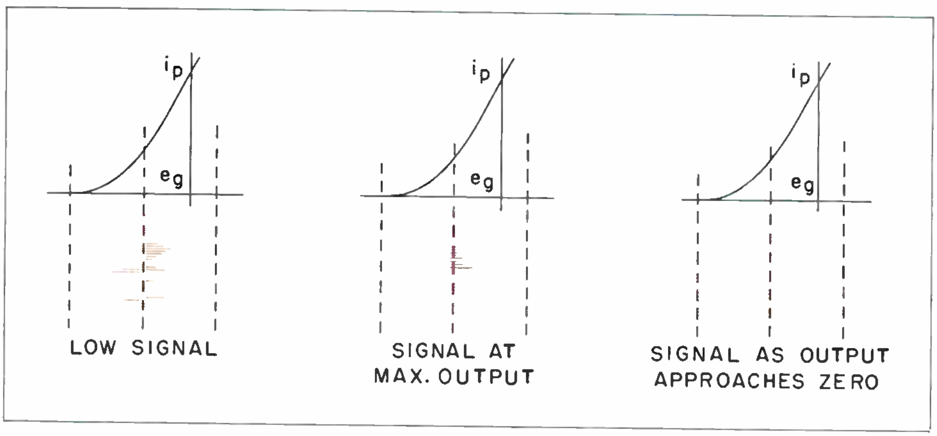

Figure 9. Conditions Existing for Various Signal Levels with Less Than 100-Percent Modulation

importance of having a fully modulated signal can be seen. Misleading conclusions can be drawn from curves plotted unknowingly with signals less than completely modulated.

Output Pulse Shape

If the input pulse to the receiver has poor rise and fall times, the logarithmic output pulse will not have a leading edge that is continuous. The effect of an input pulse of finite rise time is to cause the signal in later i-f stages to appear broadened when large signal amplitudes are introduced at the input. A cascaded amplifying and clipping action as a result of overloading causes the increased width. The pulse obtained at the last stage is the widest, with the width diminishing in each stage toward the front end of the receiver. In a linear receiver the widening is observed in the output pulse as a stretching of the pulse width. Since the log i-f amplifier collects a part of the signal from each stage, the addition of these portions of different widths produces a net pulse whose shape is like that shown in figure 11. The degree of non-coincidence of the leading edges is greatest in the upper regions of the pulse, causing the pulse to resemble the shape of the input but emphasizing the slopes.

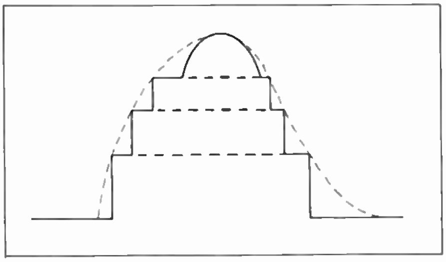

Figure 11. Output Pulse for Poor Input Pulse (Exaggerated)

Differentiation of Log Signal Output

Because of the compressed output range demanded by the indicator, adjacent targets show only a small difference in intensity on the PPI indicator. To offset this somewhat, a fast-time-constant circuit is frequently desired to differentiate the log output signal and provide better discrimination between proximate signals, and between signal and background. The noncontinuous character of the leading edge of the log pulse is disconcerting when an attempt is made at differentiation. Instead of a single spike, a multiple spike is obtained which changes its character with input. A sketch of such a pulse appears in figure 12. The effects of this phenomenon are also felt in the log curve. Curves with and without differentiation are shown in figure 13.

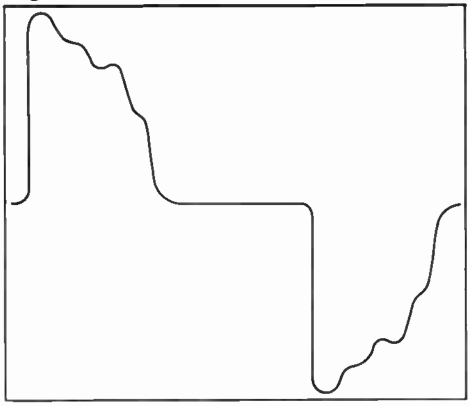

Figure 12. Differentiated Log Output with Poor Pulse Input

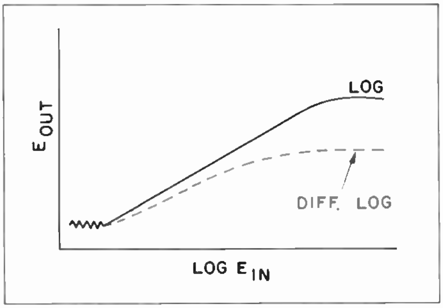

Figure 13. Effect of Differentiation on Log Curve

PPI Limitations

Unfortunately, present cathode-ray tubes used as PPI's can handle only a relatively small range of input signals. The range from minimum perceptible signal to blooming is only about 12 db. This range is not particularly bothersome for linear receivers since the usual output dynamic range is not much greater. The PPI, however, definitely limits the performance of which logarithmic receivers are capable. Many more tonal gradations are obtained from the log receiver than the indicator can handle. The log i-f amplifier, therefore, is usually designed with an output dynamic range equal to that which the PPI can handle to prevent the log range from being cut down by blooming of the scope. This compression of the range of output reduces the contrast between targets differing by a small amount in strength. Design factors affecting the dynamic range of the log receiver for compression will be discussed in Part 2, the concluding installment of this series.

Solution to November-December "What's Your Answer?"

Let an orange band and an adjacent white band constitute one pair. Then there are x pairs plus one orange band on the total tower height. Since each white band must be between 6 and 9% of the total tower height, then each orange band must be between 12 and 18% of the total tower height. From this it follows that each pair must be between 18 and 27% of the total tower height. Thus, if the tower height is k,

$x(.18k) + .12k = k$ for the lower limit, and

$x(.27k) + .18k = k$ for the upper limit

Solving these for x:

$x = 4+$ for the lower limit

$x = 3+$ for the upper limit

Since x must be an integer, it must be equal to 4, or there must be 4 pairs plus an extra orange band, making a total of 5 orange bands.

If the height of each white band is y, the height of each orange band is 2y; therefore:

$$4 \times 3y + 2y = k$$

$$14y = k$$

$$y = \frac{k}{14},$$

But k is 322; hence:

$$y = \frac{322}{14} = 23 \text{ ft}$$

$$2y = 46 \text{ ft}$$

Therefore, each orange band is 46 feet high.

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