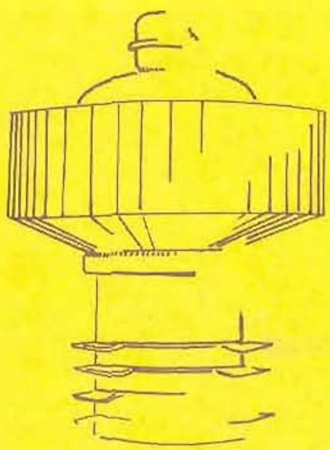
amateur service newsletters

WILLIAM I. ORR

W6SAI



EIMAC

DIVISION OF VARIAN

301 Industrial Way
San Carlos, California

amateur service newsletters

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4167

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Varian, EIMAC Division, 301 Industrial Way, San Carlos, CA 94070

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USE OF TRIODE CONNECTED TETRODES AS GROUNDED GRID AMPLIFIERS

The tetrode tube may be connected for hi-mu triode operation by placing the grid and screen elements at the same d.c. and signal potential (figure 1). Low-mu triode operation may be approximated with the same tube by connecting the screen to the plate (figure 2). This connection is not recommended for grounded-grid operation, as the tube amplification factor is extremely low.

Hi-mu triode connection, however, offers several advantages for sideband operation. First, no grid bias or screen power supplies are needed. In addition, the drive level of the grounded grid stage is compatable with the power output level of the modem sideband exciters. Finally, neutralization is not required in a properly designed amplifier employing modem tubes.

Certain tetrodes do not perform well when connected in the grounded grid configuration of figure 1. These tubes are characterized by high perveance, together with extremely small spacing between the grid bars, and between the grid structure and the cathode. Thus, while performing in excellent fashion as high gain tetrodes, this family of tubes are unsuited for grounded grid operation. Tubes of the 4-65A, 4X150A/7034, 4CX250B and 4CX1000A type are in this class.

For proper operation of the tetrode the screen requires much larger voltages than the control grid. When these electrodes of these high perveance tubes are tied together the control grid tends to draw tremendous currents and there is grave risk of destroying it. For example, in the following table, the control grid current of the 4X150A is 1.3 amperes at the positive peak of the driving cycle and the screen current is about 0.5 amperes. At the same instant, the plate current is only about 0.8 amperes. In other words, the plate is getting only a third of the current emitted by the cathode instead of nearly all the current! By any standards, such a triode is unsatisfactory. Observe that the grid dissipation is one thousand times as great for the high-mu connected tetrode as it is for the tetrode-biased tube.

4X150A Tetrode Comparison of Tetrode Biased and High-mu Triode Operation of Driven Cathode Amplifiers

	Hi-mu (Not recommended)	Tetrode Biased	
D-C Plate Voltage	2000	2000	volts
D-C Screen Voltage	0	250	volts
D-C Grid Voltage	0	-50	volts
D-C Plate Current	250	250	ma
D-C Screen Current	105	20	ma
D-C Grid Current	305	3	ma
Plate Dissipation	145	145	watts
Screen Dissipation	5.7	6.3	watts
Grid Dissipation	18	0.02	watts
Plate Power Output	355	355	watts
Plate Power Input	500	500	watts
Driving Power	38	13.0	watts
Stage Gain	10	28	
Cathode Impedance	86	120	ohms

By far the best way to operate tetrodes such as the 4X150A, 4X250B or 4CX300A in a cathode driven linear amplifier is to ground the grid and screen through bypass capacitors and to operate them at their rated d.c. voltages, as shown in figure 3. The grid dissipation reduces to little or nothing when this is done and the stage gain is greatly increased. The screen dissipation is nearly the same as in the tetrode connection. Greater stage gain can be obtained with this circuit because the driver does not have to supply large screen and grid losses. If it is desired to dissipate some excess of driving power, it should be expended in a resistive load (figure 4).

Tetrodes such as the 4-125A, 4-250A, 4-400A and the 4-1000A are suitable for connection as grounded grid tetrodes because of their more favorable current division characteristic. In the case of the smaller tubes, the maximum power capability is limited by maximum grid dissipation.

The following ratings apply to these tubes for triode connected, grounded grid service.

OPERATING CHARACTERISTICS, EIMAC TETRODES, GROUNDED GRID CONFIGURATION

4-125A

D-C Plate Voltage	2000	2500	3000	volts
0-signal D-C Plate Current	10	15	20	ma
Single-Tone D-C Plate Current	105	110	115	ma
Single-Tone D-C Screen Current	30	30	30	ma
Single-Tone D-C Grid Current	55	55	55	ma
Single-Tone Driving Power	16	16	16	watts
Driving Impedance	340	340	340	ohms
Load Impedance	10,500	13,500	15,700	ohms
Plate Input Power	210	275	345	watts
Plate Output Power	145	190	240	watts

4-400A

(ratings apply to 4-250A, within plate dissipation rating of 4-250A)

D-C Plate Voltage	2000	2500	3000	volts
Zero-Signal d.c. Plate Current	60	65	70	ma
Single-Tone d.c. Plate Current	265	270	330	ma
Single-Tone d.c. Screen Current	55	55	55	ma
Single-Tone d.c. Grid Current	100	100	100	ma
Single-Tone Driving Power	38	39	40	watts
Driving Impedance	160	150	140	ohms
Load Impedance	3950	4500	5000	ohms
Plate Input Power	530	675	990	watts
Plate Output Power	325	435	600	watts

4-1000A

D-C Plate Voltage	3000	4000	5000	volts
Zero-Signal d.c. Plate Current	100	120	150	ma
Single-Tone d.c. Plate Current	700	675	540	ma
Single-Tone d.c. Screen Current	105	80	55	ma
Single-Tone d.c. Grid Current	170	150	115	ma
Single-Tone Driving Power	130	105	70	watts
Driving Impedance	104	106	110	ohms
Load Impedance	2450	3450	5550	ohms
Plate Input Power	2100	2700	2700	watts
Plate Output Power	1475	1870	1900	watts

In all cases, grid current should be monitored. This may be accomplished by grounding the control grid through a 1-ohm composition resistor, bypassed by a .01 µfd disc ceramic capacitor (figure 5). The R-C combination serves to hold the control grid very near to ground potential. Grid current is monitored by measuring the voltage drop across the resistor. The indicating meter is calibrated in terms of grid current. For example, to have a meter range of 100 milliamperes, the series resistor plus the internal meter resistance should equal 100 ohms.

For voice operation, the plate or grid current (as read on the meter) will reach some peak value less than the single tone meter reading. Under average conditions, the "voice" current peaks should be approximately one-half the indicated single tone current. For example, a single tone plate current of 300 ma is approximated by voice meter peaks of 150 ma. Driving the indicated voice meter peaks to equal the value of single tone current will result in severe overload distortion.

Use of a high-C tuned cathode circuit in any grounded grid amplifier is mandatory if maximum efficiency and lowest intermodulation products are desired. The circuit need have only a Q of 2 or so.

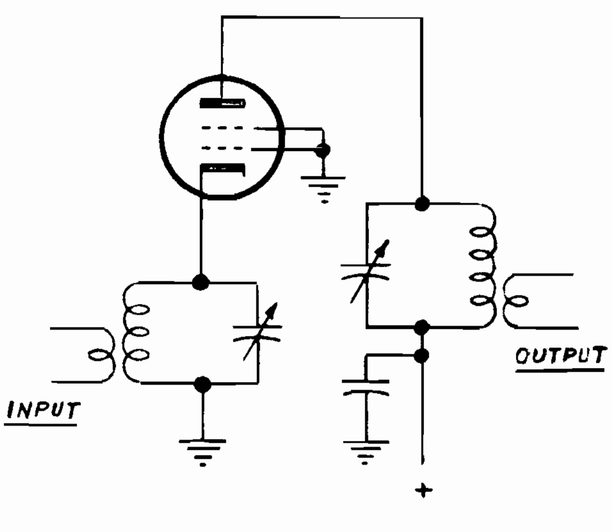

FIGURE 1

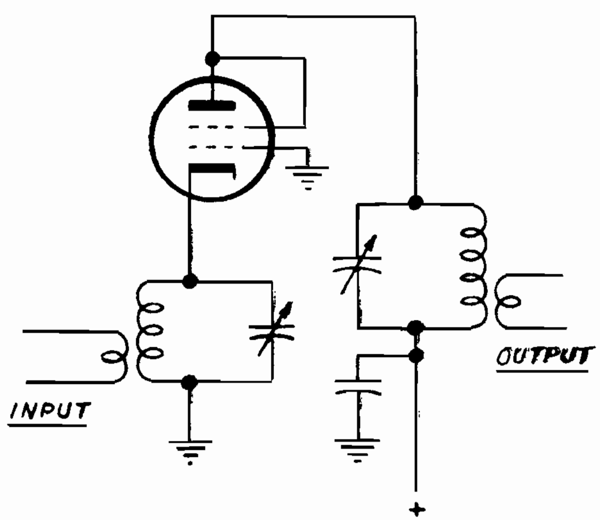

FIGURE 2

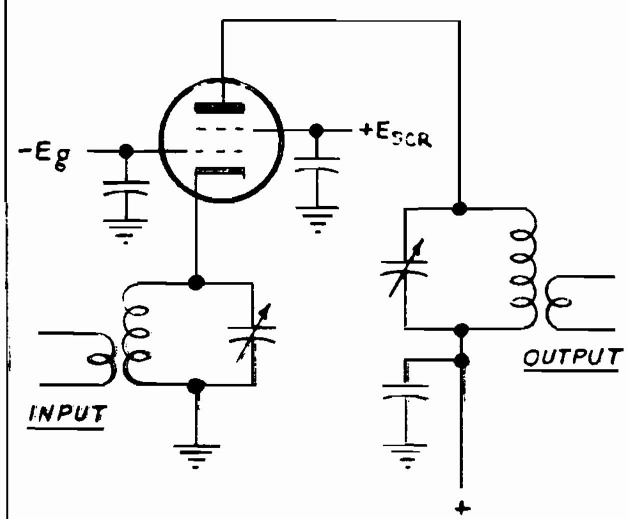

FIGURE 3

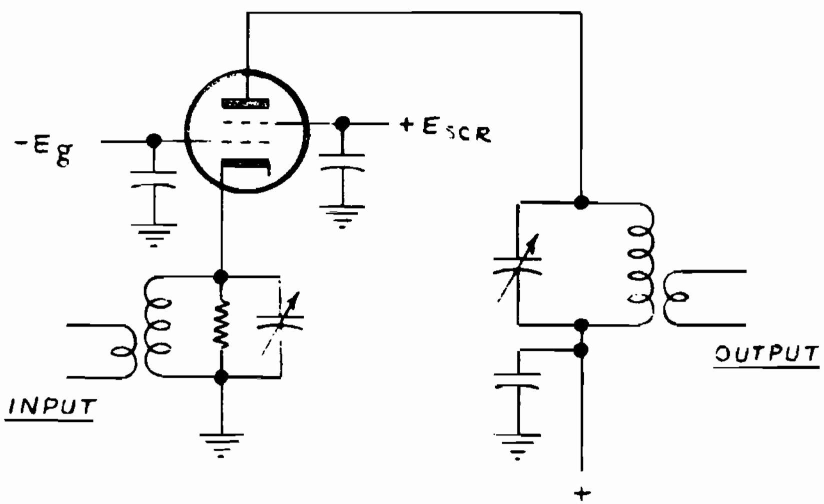

FIGURE 4

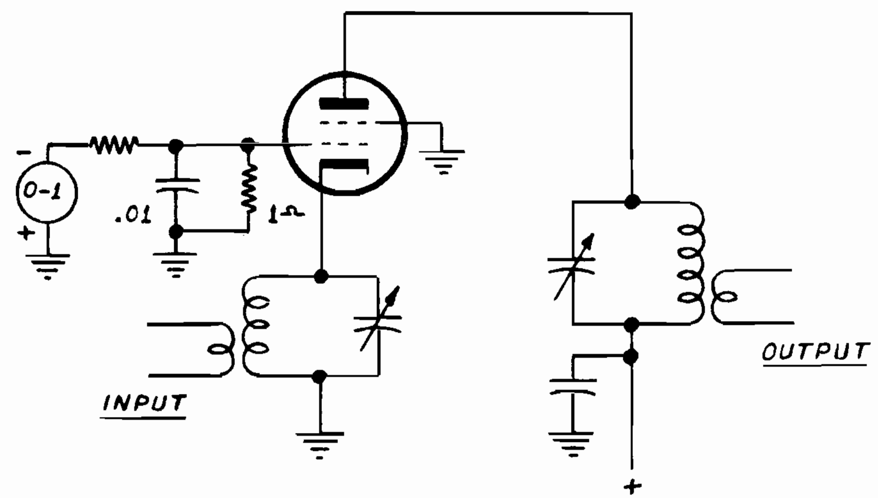

FIGURE 5

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GROUNDED SCREEN OPERATION OF TETRODE AMPLIFIERS

One of the design problems encountered in tetrode amplifiers is ensuring that the screen element of the tube is at rf ground even though a dc potential is applied to it. The problem arises because the perfect screen bypass capacitor has not yet been invented. Even the best capacitor contains residual inductance which (when added to the inductance of the screen lead of the tube) inhibits neutralization and encourages VHF parasitics. One solution to this problem is to eliminate the screen bypass capacitor by rearranging the circuit.

Figure 1 shows the conventional dc return paths wherein all power supplies are returned to the grounded cathode of the stage. Meters are placed in the common return leads, and each meter reads only the current flowing in the particular circuit.

The dc ground connection is removed from the cathode in figure 2 and placed at the screen terminal of the tube. Circuit operation is still the same as all power supply returns are still made to the cathode of the tube. The only change is that the screen rf ground and dc ground are now the same. The cathode circuit is at a negative potential with respect to the chassis by an amount equal to the screen voltage. Also, the return of the high voltage plate supply and the grid bias supply are now negative with respect to ground by the screen potential. The cathode is bypassed to ground, as shown.

A practical version of this circuit is shown in figure 3. The grid bias and screen supplies are incorporated in the amplifier and terminals are provided for positive and negative connections to the high voltage supply. A "tune-operate" switch is added which removes the screen potential for tune-up purposes. The negative of the plate supply "floats" below ground by the value of the screen voltage. A resistor is placed across the plate milliammeter as a safety device to prevent the circuit from being accidentally opened by chance failure of the meter. Note that the cathode bypass capacitor must be rated to withstand the full screen voltage.

Operation of this circuit is normal in all respects and it may be applied to any form of tetrode amplifier with good results.

Another version of this circuit is shown in figure 4. The screen supply is placed between the cathode and ground, and the supply must be capable of passing the dc plate current. The screen current is thus "swamped" by the plate current. Aside from the extra complication of high current screen supply, the advantage of an extremely stable screen voltage source is gained, plus the fact that the screen potential is added to that of the plate power supply. The plate voltage, therefore, is the sum of the plate and screen supply voltages.

Screen current and plate current should be monitored separately, because screen current is still the best indicator of loading in a tetrode amplifier stage, and also because the screen dissipation rating must be observed.

The resistors "R" serve to tie the plate and screen supplies to ground in the event of meter burn-out. The resistors are just large enough so that they do not materially affect the accuracy of the meters. The grid meter and power supply are "off ground" by the amount of the dc screen voltage and must be adequately insulated.

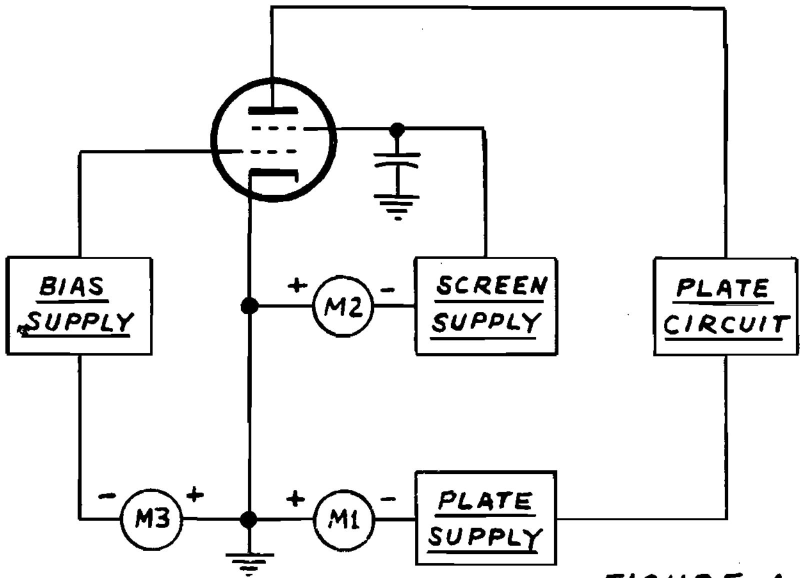

FIGURE 1

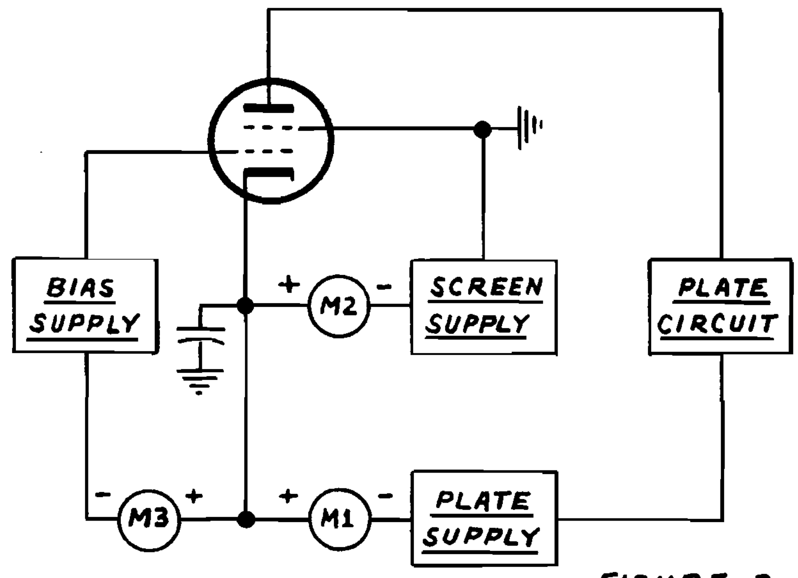

FIGURE 2

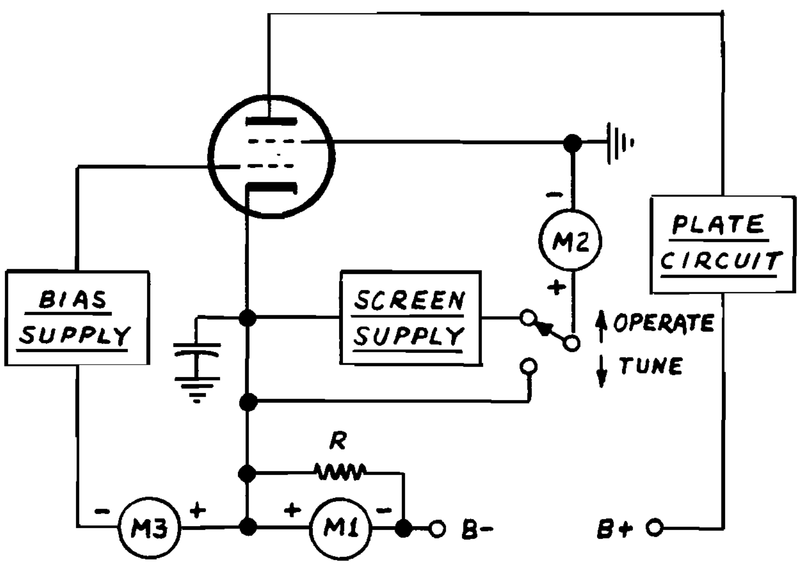

FIGURE 3

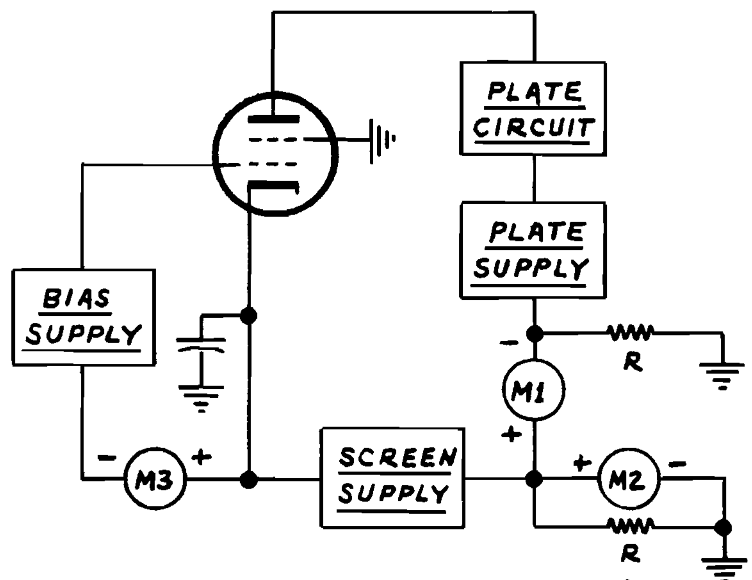

FIGURE 4

AS-2

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THE 3-400Z and 3-1000Z FOR AMATEUR SERVICE

by William I. Orr, W6SAI

The EIMAC 3-400Z and 3-1000Z are zero bias triodes designed for grounded grid service in the high frequency spectrum. The tubes are rated at 400 and 1000 watts plate dissipation, respectively. No external bias supply is required over the plate potential operating range of 2000 to 3000 volts.

These tubes are especially suited for single sideband operation in the amateur service. Costly and bulky screen and bias supplies are not required. The tubes are small and rugged, and are designed to fit into modern, compact transmitter design. Best of all, the 3-400Z and the 3-1000Z provide **improved linearity** and a reduction of bothersome intermodulation products when operated in an approved circuit.

The 3-400Z is rated to 1000 watts PEP input, and the 3-1000Z is rated to 2000 watts PEP input. These ratings are established at moderate plate potentials, and result in third-and high-order product distortion figures better than -35 decibels below maximum output!

Preliminary operating data for these tubes is given in figure 1, and suggested circuits are shown in figures 2 and 3.

Circuitry for the 3-400Z

A simple operational circuit for the 3-400Z is shown in figure 2. The input circuit comprises a high-C tank (L1-C1), with excitation applied at a point which matches a 52 ohm driving source. The coil is bifilar wound, with the filament voltage applied to the tube via the coil. The grid of the tube is at ground potential for both d-c and r-f The plate circuit consists of a pi-network (C5-L2-C6) with the output voltage monitored by a simple diode voltmeter.

The Cathode Circuit

Capacitors C3 and C4 form part of the cathode tuned circuit and comparatively high valves of r-f current flow through them. The specified capacitors are satisfactory for the 3-400Z tube in continuous service, and will serve for the 3-1000Z in intermittent duty. These two capacitors should be grounded to a common point at the rotor of capacitor C1. Capacitor C2 carried the full excitation current and should be a transmitting type, as specified.

Although specifically designed for class B service, the 3-400Z may be operated as a class C power amplifier or oscillator or as a plate-modulated radio-frequency power amplifier. One can take advantage of the zero bias characteristic of the 3-400Z in class C amplifiers operating at plate voltages below 3000 volts by employing only grid-leak bias. If driving power fails, plate dissipation is then kept to a low value because the tube will be operating at the normal static zero bias conditions. Operating conditions are listed below:

MAXIMUM RATINGS	Class C Amp. or Osc.	Class C Plate-Modulated
D C PLATE VOLTAGE	4000 VOLTS	3000 VOLTS
D C PLATE CURRENT	350 MA	275 MA
PLATE DISSIPATION	400 WATTS	270 WATTS
GRID DISSIPATION	20 WATTS	20 WATTS
TYPICAL OPERATION		
D C Plate Voltage	3000 volts	3000 volts
D C Plate Current	333 m A	245 m A
D C Grid Voltage	-75 volts	-90 volts
D C Grid Current	130 m A	100 m A
Grid Driving Power	25 watts	18 watts
Plate Output Power	730 watts	550 watts

Air Sockets and Chimneys for the 3-400Z and 3-1000Z

In order to properly cool the seals and envelope of the 3-400Z and 3-1000Z, use of the EIMAC Air System socket is recommended. The SK-400 and SK-500 sockets are satisfactory; however, the new series SK-410 and SK-510 sockets are recommended for new equipments. These modernized sockets feature low cost, lighter weight design and simplified mounting. Low lead inductance insures proper tube operation.

The SK-410 socket may be used with the 3-400Z, 4-250A, 4-400A series tubes, in conjunction with the SK-416 chimney (for the 3-400Z) or the SK-406 chimney (for the 4-250A or 4-400A). The SK-510 socket may be used with the 3-1000Z (with the SK-516 chimney) or the 4-1000A (with the SK-506 chimney). See tube data sheet for air flow and full cooling information.

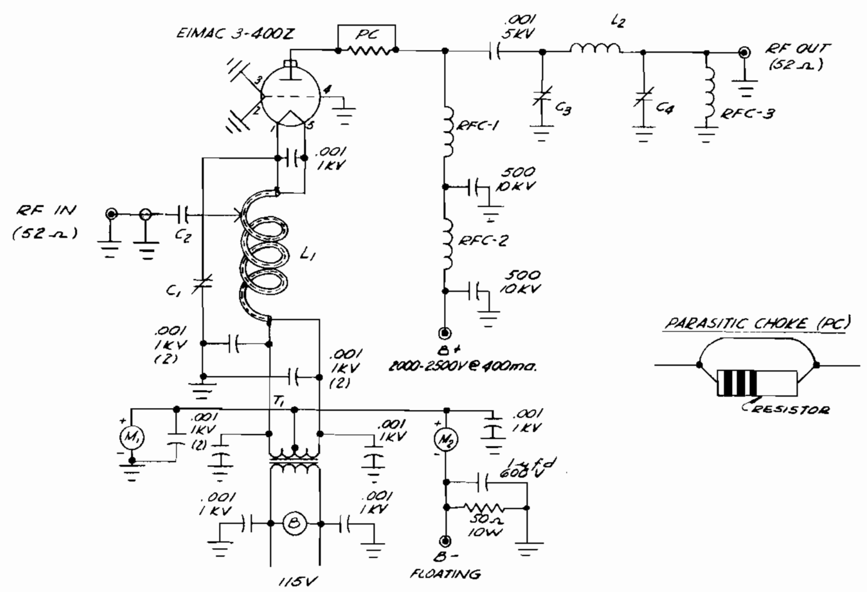


FIGURE 6

Using the 3-400Z on 50 MHz

The 3-400Z is rated for maximum service to over 100 MHz and makes an excellent grounded grid. linear tube for 6 meter band. A suitable circuit for this service is shown in Figure 6. Neutralization is not required and the amplifier is stable and free of parasitics. A standard tuned cathode input and pi-network output circuit are used, with the grid and plate current meters placed in the d-c ground return. The power supply shown in Figure 3 can be used with this amplifier.

The amplifier may be built on a 10" x 14" x 3" aluminum chassis-box. Cathode circuit components are mounted within the box, with the plate circuit components placed atop the box. A complete cabinet, or enclosure should be built atop the chassis using perforated aluminum sheet. This will contain the various harmonics and reduce coupling between input and output circuits. The three grid pins of the 3-400Z socket (EIMAC type SK-410) are grounded directly to the chassis by means of wide copper straps passing through the adjacent socket slots. The straps (cut from flashing copper) are soldered to the socket pins, then bolted to the chassis directly next to the socket. Tuned cathode circuit L1/C1 is mounted close to the socket to insure short leads. The ''cold'' end of the bifilar filament coil (L1) is bypassed with two ceramic capacitors in parallel on each lead to bypass any r-f current flowing at this point. Input tuning capacitor C1 is mounted to the chassis in close proximity to the tube socket. All leads should be short and direct. Coupling capacitor C2 is placed near the tuned circuit and the lead from C2 to the input connector may be a length of 52 ohm coaxial cable.

The blower is mounted to the chassis, forcing air into the box which escapes via the tube socket and chimney.

FIGURE 1
Operating Data For
EIMAC 3-400Z and 3-1000Z

3-400Z

Filament: 5 volts @ 14.5 amperes

Socket: EIMAC SK-400 Air System Socket

Cooling: Radiation and forced air

Maximum Operating Temperatures: Base, 200° C; Plate seal, 225° C

Typical Operation for minimum distortion products with 1 kw PEP input

DC Plate Voltage	2500 volts
Zero Signal Plate Current	73 mA
Single Tone d-c Plate Current	400 mA
Single Tone d-c Grid Current	142 mA
Two Tone d-c Plate Current	274 mA
Two Tone d-c Grid Current	82 mA
Useful Power Output (PEP)	560 watts
Resonant Load Resistance	3450 ohms
Intermodulation Products	Typically more than -35 decibels below PEP level

Filament: 7.5 volts @ 21.3 amperes

Socket: EIMAC SK-500 Air System Socket

Cooling: Radiation and forced air

Maximum Operating Temperatures: Base, 200° C; Plate Seal, 225° C

Typical Operation for minimum distortion products with 2 kw PEP input

DC Plate Voltage	2500 volts
Zero Signal d-c Plate Current	162 mA
Single Tone d-c Plate Current	800 mA
Single Tone d-c Grid Current	254 mA
Two Tone d-c Plate Current	550 mA
Two Tone d-c Grid Current	147 mA
Useful Power Output (PEP)	1050 watts
Resonant Load Resistance	1700 ohms
Intermodulation Products	Typically more than -35 decibels below PEP level

Figure 2: Grounded grid circuitry eliminates expensive bias and screen supplies required with grid driven circuitry. Good tube linearity, plus use of tuned cathode circuit results in low distortion, high power sideband amplifier.

Parts List:

C1 -- 1000 pF (Three gang b-c variable, with sections connected in parallel. J.W. Miller #2113).

C2 -- .01 µF, mica. 1200 volt. Aerovox type 1446.

C3 -- .01 µF, mica. 500 volt. Aerovox type CM-30B-103.

C4 -- same as C3.

C5 -- 3500 volt rating. Effective tuning capacity; 2.5 pF per meter.

C6 -- 500 volt rating. Effective tuning capacity; 25 pF per meter.

L1 -- See text. Resonates to operating frequency with C1 setting of approximately 13 pF per meter. Approximate dimensions are: 80 meters, 10 turns, 1-5/8" i.d., 3-1/4" long, tap 6 turns from ground. 40 meters, 6 turns, as above, 2" long, tap 3-1/2 turns from ground. 20 meters, 4 turns, as above, 1-1/4" long, tap 2 turns from ground. 15 meters, 3 turns, as above, 1" long, tap 2 turns from ground. 10 meters, 1 turn, as above, tap 1/2 turn from ground. Make of 1/4-inch copper tubing, threaded with #12 insulated wire.

L2 -- Make of 1/4-inch copper tubing, 3" i.d. To resonate to frequency with settings of C5 and C6 as specified above.

M1 -- 0-750 mA M2 -- 0-1 mA M3 -- 0-1 mA

R1 -- Internal resistance of meter M3 plus R1 totals 550 ohms. Meter reads 0-500 mA, full scale.

T1 -- 5 volts at 14.5 amperes. Chicago type F-516.

RFC1 -- HF choke. B & W type 800.

RFC2 -- VHF choke. Ohmite Z-144.

PC -- Three 100 ohm, 2 watt composition resistors in parallel; shunt coil is 3 turns, 1" diameter, length of resistors.

B -- 115 volt blower. Minimum of 15 cubic feet per minute. Ripley #82.

FIGURE 3

T1 - 2900-0-2900 volts @ 500 mA CCS. 115-230 volt primary. 1600 VA capacity. Chicago #P-2126.

T2 - For 866A tubes: 2.5 volts @ 10 amp. 9 kV insulation. Chicago #F-210H.
For 872A tubes: 5 volts @ 15 amp. 10 kV insulation. Chicago #F-520HB.

CH1 - 10 Henries @ 500 mA. Resistance 40 ohms. Chicago #R-105

RYL - DPST relay. Potter & Brumfield PR7AY, with 115 volt a-c coil.

Note: 866A rectifier tubes may be used with 3-400Z. Use 872A rectifier tubes for 3-1000Z. Xenon type 3B28 may be substituted for 866A, and 4B32 for 872A.

FIGURE 4

Pi-network circuit for 3-1000Z

C1 - 1250 volt, transmitting-type mica capacitor. Aerovox type 1446 or 1651-L. Two capacitors may be connected in parallel to obtain odd values of capacitance.

80 meters:	1000 pF	15 meters:	150 pF
40 meters:	450 pF	10 meters:	100 pF
20 meters:	220 pF		

C2 - Same as C1. Capacitance as follows:

80 meters:	900 pF	15 meters:	90 pF
40 meters:	375 pF	10 meters:	30 pF
20 meters:	150 pF		

C3 - Same as C1. 0.02 μF

L1 - May be made of B & W Miniductor coil stock, as follows:

80 meters:	2.5 μH 15 turns, 1" diameter, 8 turns per inch. (B & W #3014).
40 meters:	1.1 μH 8 turns, same as above.
20 meters:	0.55 μH 7 turns, 3/4" diameter, 8 turns per inch. (B & W #3010).
15 meters:	0.37 μH 6 turns, same as above.
10 meters:	0.3 μH 6 turns, 1/2" diameter, 8 turns per inch. (B & W #3002).

Network may be adjusted by shorting capacitor C2 and trimming coil L1 until L1-C1 resonates to center of band of operation, or by making C2 variable and adjusting for maximum grid drive.

B - 115 volt, 60 cycle blower. Ripley #81. 45 cubic feet per minute.

RFC1 - 30 ampere bifilar filament choke. Barker & Williamson FC-30

Note: An alternative arrangement is to eliminate the bifilar filament choke and substitute a low-capacity filament transformer. (7.5 volts at 21.3 amperes). A suitable transformer is made by Transformer Technicians, Inc., 2608 No. Cicero Avenue, Chicago, Ill. (Type #TTI-4173).

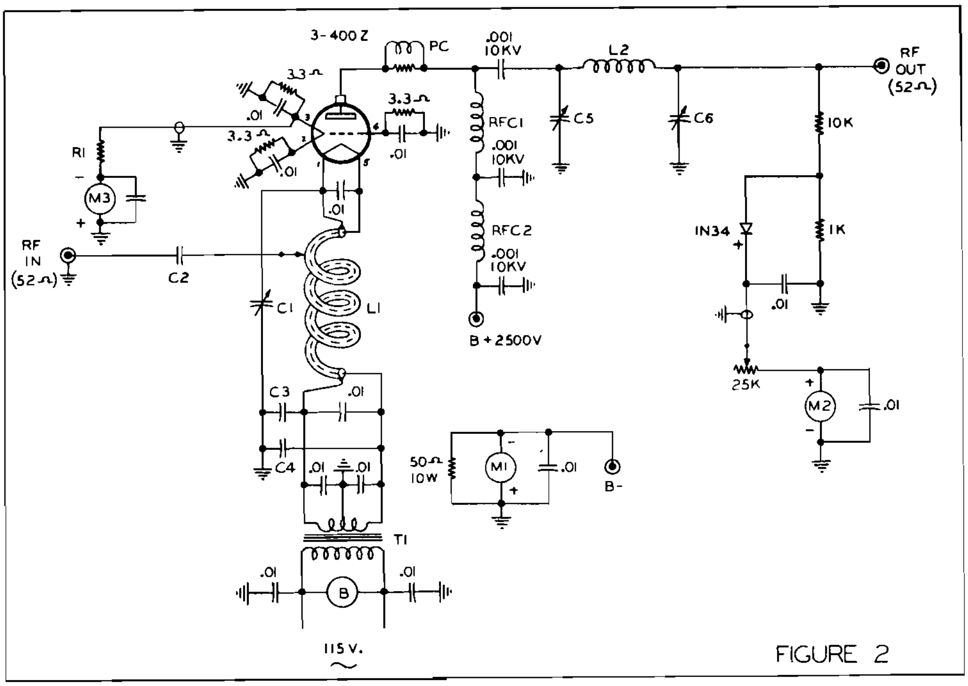


FIGURE 2

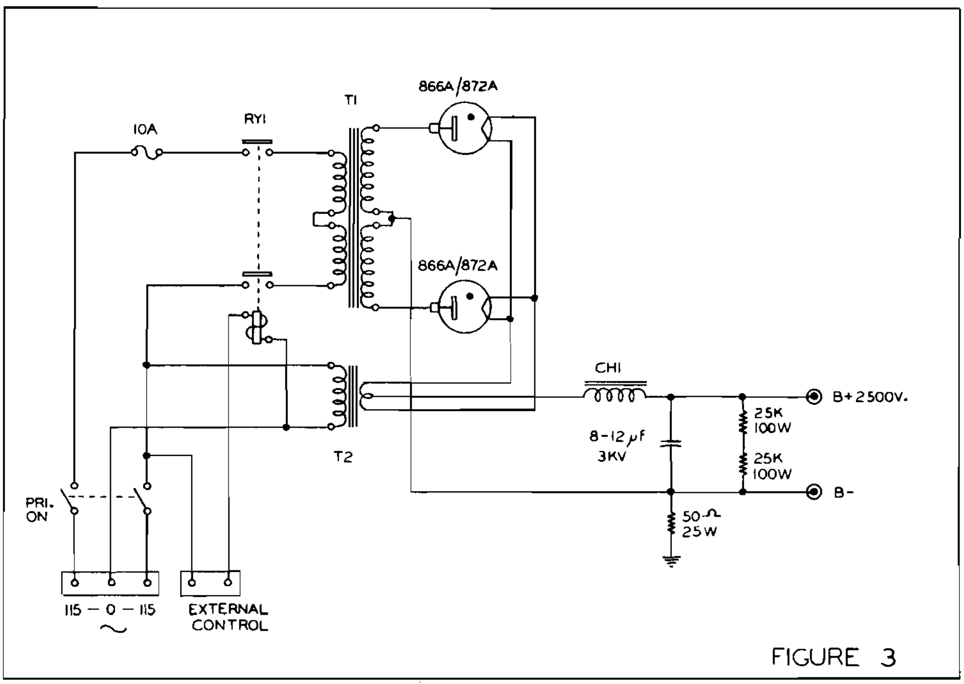


FIGURE 3

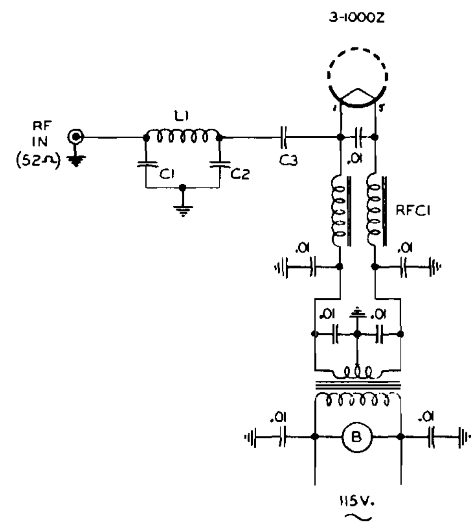
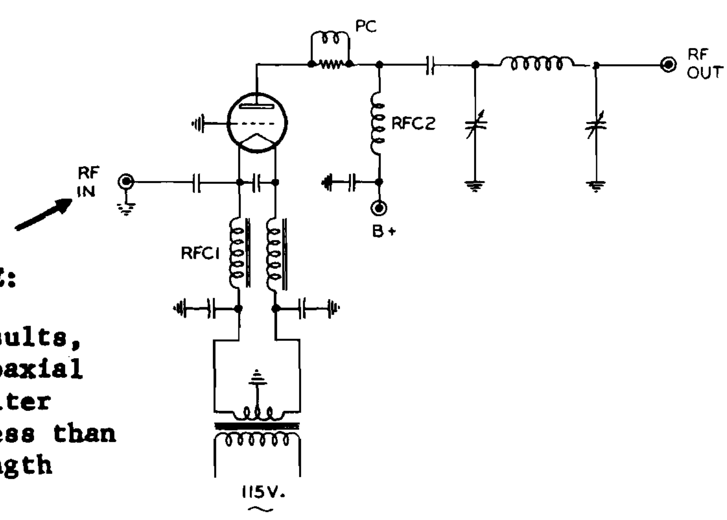


FIGURE 4



NOTE:

For best results, length of coaxial line to exciter should be less than 1/10 wavelength

FIGURE 5

The Untuned Cathode Circuit

Shown in figure 5 is a simple grounded grid amplifier circuit employing r-f chokes in the cathode in place of a tuned circuit. This circuit is very popular, and may be employed with the 3-400Z or 3-1000Z provided the limitations of the circuit are understood by the user.

Any single-ended Class B stage (regardless of the tube used) draws grid and plate current over only a portion of the operating cycle (approximately 180°). The input impedance of such a stage, therefore, does not represent a constant load. The waveform delivered by the exciter to the grounded grid stage is greatly distorted over the portion of the cycle that the amplifier draws grid and plate current. Although published "input impedance" values may look attractive, they actually represent only the fundamental component of input impedance (useful for tank circuit "Q" calculation). Since the input load impedance of the class B grounded grid stage is not a constant value, it is necessary to transform it to a **constant impedance** which will resemble 50 ohms over the complete operating cycle. This is best done by a high C tuned circuit placed directly at the cathode of the grounded grid stage.

The 3-400Z and 3-1000Z are not unique in requiring a tuned cathode circuit. Good engineering practice requires it for **any** grounded grid amplifier to achieve best linearity unless the amplifier is operating class A. In short, grounded grid amplifiers may be built **without** the cathode tank circuit, but the fullest capabilities of the equipment will not be realized. Power output will be sacrificed, and poorer values of odd-order product distortion will be observed when the cathode tank circuit is omitted. The degree of degradation will vary with the particular driver/amplifier combination used, and with the length of interconnecting coaxial line. Typical measurements made with a class A driver stage (well swamped) and a short length of interconnecting coaxial line between the driver and a grounded grid amplifier showed that when the amplifier cathode tank was removed, the power output of the amplifier stage dropped about 5%, the third order products increased approximately 3-5 decibels, and the fifth order products increased 5-7 decibels. The higher order products also rose accordingly. It is possible that with an exciter having poorer output circuit Q, and with other lengths of interconnecting coaxial line that a greater degradation might have been noticed in performance when the tuned tank was removed.

The tuned cathode tank can be thought of as a refinement necessary for best performance of a grounded grid class B linear amplifier. Its use is highly recommended.

Filament coil L1 is made of a section of ¼-inch diameter copper tubing with a length of #12 insulated wire passed through the center hole. Heavy conductors are soldered to each end of the coil to serve as filament circuit leads. After the correct position of the excitation tap has been found, it may be soldered in position.

It is necessary to use some form of tuned circuit at this point and to resist the temptation to substitute untuned filament chokes. Use of the latter by themselves will result in degradation of linearity to a marked degree, and will make the amplifier more difficult to drive properly.

Circuit Details

Plate meter M1 is placed in the B-minus lead to the amplifier in order to register plate current instead of cathode current (a combination of grid and plate current). The negative return of the power supply, therefore, should be "floating" as shown in figure 3.

Since a great deal of power is produced in a small package, an EIMAC air system socket and blower are recommended for use with these tubes in order to maintain the envelope and seals at a low temperature. An alternative arrangement may be two separate blowers, one directed at the envelope of the tube, and the other at the bottom of the tube socket.

Amplifier Construction and Adjustment

In order to complement the excellent internal shielding of the grounded grid tubes it is necessary to isolate the input and output circuits. The cathode circuit should be placed below the metal chassis, enclosed in a metal box. Plate circuit components should be mounted above the chassis. As a TVI preventive measure, the complete amplifier should be enclosed within a metal shield made of perforated material. The grid terminals are bypassed to ground by means of three .01 μF ceramic capacitors mounted (one at each socket terminal) by the shortest possible leads. Also placed at each grid terminal is a 3.3 ohm, 1-watt composition resistor. These resistors "de-Q" the capacitors and provide a short, direct ground return for the r-f and d-c currents in the grid circuit. Meter M3 measures the slight voltage drop across the resistors and is calibrated in terms of grid current for tuning purposes.

Before the amplifier is adjusted, the filament voltage at the tube socket should be checked to ensure that excessive voltage drop does not exist in the tuned circuit. Approximate setting for the tap of L1 is given in figure 2. The exact point may be found by loading the amplifier to full input and varying tap placement until maximum grid current is obtained at the same setting of capacitor C1 that provides minimum SWR on the exciter coaxial line. At incorrect tap settings, minimum SWR figure and maximum grid current settings of capacitor C1 do not coincide.

When properly loaded and tuned, grid current runs about 1/3 the value of plate current. Plate loading and tuning and the excitation level adjustments should be conducted to adhere closely to this ratio of grid/plate current. After the amplifier is loaded to maximum input in this fashion, the pi-network should then be overcoupled (C6 reduced in capacitance) slightly until the r-f output measured on M2 drops about 3%. This will approximate a condition of maximum linearity. Do not apply full excitation to the tube without plate voltage and proper loading as grid dissipation will be exceeded.

In sideband service, for voice waveforms, the indicated plate current measured on M1 will be approximately one-half the peak d-c plate current. One kilowatt PEP input to the 3-400Z, for example, may be achieved on voice by plate current meter peaks of 200 milliamperes at a plate potential of 2500 volts.

Notes on the 3-1000Z tube

The circuit of figure 2 may be used for the 3-1000Z. However, the higher filament current of this tube requires that a heavier bifilar coil be used having a length of #10 insulated wire for the center conductor. Also, as the input impedance of this tube is close to 52 ohms, the filament tap point for C2 occurs near to the top (filament end) of the coil. Adjustment of the tap point may not be required.

It may be somewhat easier from a mechanical point of view to substitute a pi-network input circuit for the parallel L-C circuit, as shown in figure 4. Untuned filament chokes are used to supply filament voltage to the tube, and a simple fixed tuned pi-network circuit is employed to couple the exciter to the filament circuit of the 3-1000Z. The transformation ratio of the network is 1:1, matching the line to the tube, yet providing the "Q" necessary for proper circuit operation. The pi-network circuit and filament choke may be used for the 3-400Z tube, if desired.

To assist in proper adjustment of the amplifier, an output tuning device should be placed in the coaxial lead to the antenna. A SWR power meter or r-f ammeter may be used. The amplifier is adjusted for maximum power output at the proper plate and grid current values. Carrier injection may be used for the tuning process, and antenna loading and grid drive adjusted to provide the current data given in figure 1 for 3-400Z.

FIGURE 6

50 MHz Linear Amplifier for 3-400Z

C1 - 100 pF. Bud MC-1855 (double bearing)

C2 - .001 mica capacitor

C3 - 15 pF. 3 kV (.075" spacing). Johnson 155-8 capacitor may be used, with every other rotor plate removed.

C4 - 200 pF. Bud MC-1858 (double bearing)

L1 - 2 turns of 1/4-inch copper tubing, 1 inch inside diameter, 1-1/2 inches long. Pass #12 insulated wire through center before winding coil. Tap for C2 placed at center of coil. Connections between L1 and C1 should be made with 1/8-inch copper strap.

L2 - 3-1/2 turns of 1/4 inch copper tubing, 2 inch inside diameter, 3 inches long. Mount between stator terminals of C3 and C4 with short leads.

PC - Plate lead to 3-400Z is made of 1/4-inch copper strap. The parasitic choke (PC) is made of a section of this lead with three 47 ohm, 2 watt resistors in parallel placed across the lead, as shown in Figure 6. Resistors are mounted side by side.

RFC1 - 8.3 µH. Space-wound on 1/2-inch diameter ceramic insulator, 2-1/2 inches long (Birnbach 447). Wind #28 double silk covered wire spacewound about wire diameter. Winding 1-3/4 inches long. Teflon rod may be used. (Avoid polystyrene as it may melt or deform from heat of tube).

RFC-2,3 - VHF choke. Ohmite Z-50.

M1 - 0-200 mA(grid current)

M2 - 0-500 mA(plate current)

B - 115 volt blower. Minimum of 15 cubic feet per minute. Ripley #82 or Dayton 2C782

T1 - 5 volts at 14.5 amperes. Chicago-Standard F-516

Socket: EIMAC SK-410 Chimney: EIMAC SK-616

Note: 500 pF, 10 kV capacitors are TV "door-knob" type. Aluminum chassis box 10" x 14" x 3" is used.

amateur service newsletter
W6SAI

THE KW-2. AN ECONOMY GROUNDED GRID LINEAR AMPLIFIER

by William I. Orr, W6SAI

The KW-2 sideband amplifier is designed for use with 4-400A, 4-250A, or 4-125A tubes, and will operate on the 80, 40, 20, 15 and 10 meter amateur bands. A pi-network output circuit is used, capable of matching 52 ohm or 75 ohm coaxial antenna circuits. Maximum power input is 2 kilowatts (p.e.p.), or 1 kilowatt, c.w. The amplifier may be driven by any of the popular SSB exciters having 70 to 100 watts output.

Full input may be achieved with the use of 4-400A tubes, but the unit may be run at reduced power rating with 4-250A or 4-125A tubes. No circuit alterations are necessary when tube types are changed.

The amplifier employs a passive (untuned) input circuit, and an adjustable pi-network output circuit. Air tuning capacitors are used in the network in the interest of economy and with no sacrifice in performance. The complete amplifier is housed in a TVI-suppressed perforated metal cabinet measuring 17¼" x 12" x 12½" - small enough to be placed on the operating table next to your receiver.

Amplifier Circuit

The schematic of the amplifier is shown in figure 2. Two tetrode tubes are operated in parallel, cathode driven, with grid and screen elements grounded. The sideband exciting signal is applied to the filament circuit of the tubes, which is isolated from ground by an r.f. choke. The resistance of the windings of the choke must be limited to .01 ohms or less, as filament current is 30 amperes for two 4-250A or 4-400A tubes. Neutralization is not required because of the excellent circuit isolation afforded by the grounded elements of the tubes.

The Input Circuit

The input signal is fed in a balanced manner to the filament circuit of the two tubes. Ceramic capacitors are placed between the filament pins of each tube socket, and excitation is applied to each tube through two 1250 volt, mica capacitors. The latter are employed because of the relatively high value of excitation current which may cause capacitor heating if ceramic units are employed at this point.

The filament circuit is wired with #10 stranded insulated wire to hold voltage drop to a minimum. The leads from the choke to the filament transformer are run in shielded loom which is grounded to the chassis at each end of the wire. The use of shielded leads for all low voltage d.c. and a.c. power wiring does much to reduce TVI-producing harmonics.

The Grid Circuit

The grid circuit of this amplifier is simplicity itself. Screen terminals of both sockets are grounded to the chassis of the amplifier. The best and easiest way to accomplish this is to bend the terminal lead of the socket down so that it touches the chassis. Chassis and lead are then drilled simultaneously for a 4-40 machine screw. Low inductance ground paths are necessary for the high order of stability required in grounded grid service.

It is helpful to monitor the control grid current for tuning purposes, and also to hold the maximum current within the limits given in the data chart. Maximum grid current for the 4-400A is 100 milliamperes. Under normal voice conditions this will approximate a peak meter reading of 50 milliamperes.

Grid current can be observed by grounding the control grid of each tube through a 1-ohm composition resistor, bypassed by a .01 µfd disc capacitor. The voltage drop across the resistor is measured by a simple voltmeter calibrated to read full scale when 100 milliamperes of grid current are flowing through the resistor monitoring grid current of either tube. With incorrect antenna loading, it is possible to exceed maximum grid current rating with some of the larger size SSB exciters. No circuit instability is introduced by this metering technique.

The Plate Circuit

Power is applied to the plate circuit via a heavy duty r.f. choke bypassed at the "cold" end by a 500 µµfd, 10 KV "TV-type" ceramic capacitor. In addition, a VHF choke and capacitor are used to suppress high frequency harmonics that might pass down the plate lead and be radiated through the power supply wiring. Two .001 µfd, 5 KV ceramic capacitors in parallel are used for the high voltage plate blocking capacitor, and are mounted atop the plate choke.

The pi-network coil is an Air-Dux #195-2S inductance, designed for service at a kilowatt level, and silver plated for minimum circuit loss. Use of the cheaper model having tinned wire is not recommended for continuous service at maximum power. The bandswitch is a Communications Products #88 high voltage, ceramic switch.

A circuit Q of 15 was chosen to permit a reasonable value of capacitance to be used at 80 meters. In this case, a 150 µµfd variable air capacitor is employed for operation above 80 meters, and an additional 50 µµfd parallel capacitance is switched in the circuit for 80 meter operation. The 50 µµfd padding capacitor is the small vacuum capacitor found in the "Command" set antenna relay boxes. These capacitors seem to be plentiful and inexpensive. A satisfactory substitute would be a 50 µµfd, 5 KV mica capacitor, also available on the surplus market.

The pi-network output capacitor is a 1500 µµfd unit. It is sufficiently large to permit operation at 80 meters into reasonable antenna loads. For operation into very low impedance antenna systems that are common on this band, the loading capacitor should be paralleled with a 1000 µµfd, 1250 volt mica capacitor. This capacitor may be connected to the unused 80 meter position of the bandswitch.

The Metering Circuits

It is always handy to have an output meter on any linear amplifier. A simple r.f. voltmeter can be made up of a germanium diode and a 0-1 d.c. milliammeter. The scale range is arbitrary, and may be set to any convenient value by adjusting the potentiometer mounted on the rear apron of the chassis. Once adjusted to provide a convenient reading at maximum output level of the amplifier, the control is left alone. Under proper operating conditions, maximum output meter reading will concur with resonant plate current dip.

It is dangerous practice to place the plate current meter in the B-plus lead to the amplifier unless the meter is insulated from ground, and is placed behind a protective panel so that the operator cannot accidentally touch it. If the meter is placed in the cathode return the meter will read the cathode current which is a combination of plate, screen and grid current. This is poor practice, as the reading is confusing and does not indicate the true plate current of the stage. A better idea is to place the meter in the B-minus lead between the amplifier chassis ground and the power supply. The negative of the power supply thus has to be "undergrounded", or the meter will not read properly (figure 5). A protective resistor is placed across the meter to ensure that the negative side of the power supply remains close to ground potential. Make sure that the negative lead between the power supply and the amplifier is connected at all times.

The Cooling System

It is necessary to provide a current of cool air about the base seals and plate seal of the 4-250A and 4-400A tubes. If small blowers are mounted beneath each tube socket it is possible to dispense with the special air sockets and chimneys, and use the inexpensive "garden variety" of socket. A **Barber Coleman type DYAB** motor and impeller is mounted in a vertical position centered on the socket, and about an inch below it. Cooling air is forced up through the socket and around the envelope of the tube. The perforated metal enclosure provides maximum ventilation, yet effectively "bottles up" the r.f. field about the amplifier. In order to permit air to be drawn into the bottom of the amplifier chassis, small rubber "feet" are placed at each corner of the amplifier cabinet, raising it about ½-inch above the surface upon which it sits.

Amplifier Construction

The amplifier is built upon an aluminum chassis measuring 13" x 17" x 3". Input circuit components, power circuits, and the blower motors are mounted below the chassis, and the plate circuit components are mounted above the deck. Placement of parts is not critical, except that the leads between the bandswitch and the plate coil must be short, heavy and direct. One-half inch, silver plated copper strap is used. The straps are bolted to the bandswitch with 4-40 nuts and bolts. Each lead is tinned and wrapped around the proper coil turn and soldered in place with a large iron. The operation should be done quickly to prevent softening of the insulating coil material. Low resistance joints are imperative at this point of the circuit. To play safe, you can submerge the coil in a can of water, with just the top of the turns showing above the surface. This will prevent the body of the coil from overheating during the soldering process. It is also helpful to depress a turn on each side of the tap in order to provide sufficient clearance for the soldering iron. This may be done by placing the blade of the screw driver on the wire, and hitting it with a smart tap.

The coil assembly is supported on four ceramic pillars, and placed immediately behind the band change switch, which is mounted on a sturdy aluminum bracket. The coil is positioned so that the taps come off on the side nearest the switch.

A set of auxiliary contacts are required to switch the padding capacitor into the circuit when the bandswitch is thrown to the 80 meter position. A simple switch may be made up from the metal portions of an insulated coupling and a block of insulating material, such as teflon, lucite, or micarta (figure 4). The insulated disc of the coupling is removed, and an oval of insulating material is substituted. This assembly is placed on the shaft of the bandswitch. A set of spring contacts are mounted on small standoff insulators attached to the side of the tuning capacitor and positioned so that the oval rotates between the contacts as the switch is turned. A hole is drilled in the oval, and a flat-head 8-32 brass machine screw is passed through it. A nut is run onto the screw, and screw end and nut head are filed flat. When the switch is rotated to the 80 meter position, contact is made between the two spring arms through the body of the screw, which completes the circuit between the switch contacts.

Amplifier Adjustment

Typical operating conditions for various tubes are tabulated in figure 8. For initial adjustment, four or five hundred volts plate potential is applied to the amplifier, and sufficient grid drive is supplied (five watts or so) to provide an indication on the plate meter. The loading capacitor is set at maximum capacitance, and the tuning capacitor is adjusted for resonance, which is indicated by the customary dip in plate current. After resonance is found full plate voltage should be applied to the amplifier, and resting plate current compared with the value shown in the table. If all is well, a carrier is applied to the amplifier for adjustment purposes. The signal may be generated by carrier injection, or by tone modulation of a sideband exciter.

Caution! Do not apply full excitation to any grounded grid amplifier without plate voltage on the stage, or with the stage improperly loaded. Under improper conditions, driving power normally fed to the output circuit becomes available to heat the control grid of the tube to excessive temperature, and such action can destroy the tube in short time. Adjustable control of the excitation level is mandatory.

The amplifier is now loaded to full, single tone input. (In the case of two 4-400A's this will be 3000 volts at 333 ma, 2500 volts at 400 ma, or 2000 volts at 500 ma). Driving power will be approximately 30 watts per tube. Under these conditions, power input will be 1000 watts p.e.p. for sideband operation.

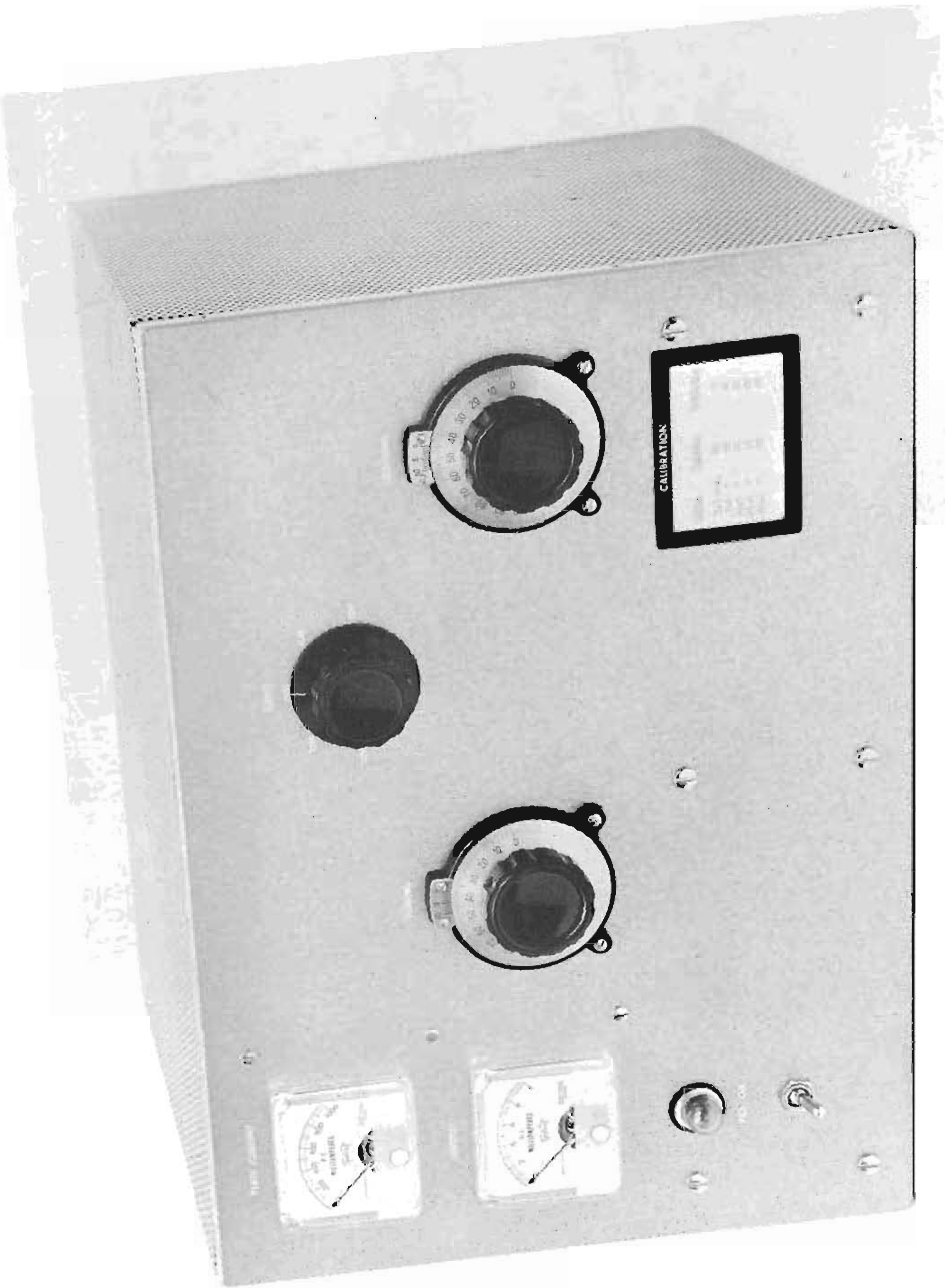
To properly load the amplifier for 2 kw p.e.p. operation it is necessary to have a special test signal. Tuning of this (or any other linear amplifier) is greatly facilitated by the use of an oscilloscope and envelope detectors. Even with two-tone or carrier input signal, however, it is difficult to establish the proper ratio of grid drive to output loading. In general, antenna coupling should be quite heavy: to the point where the power output of the amplifier has dropped about two percent. This point may be found by experiment for power levels up to 1 kw p.e.p. However, since neither this amplifier, nor most power supplies, are designed for continuous carrier service at two kilowatts and since this average power level is illegal, some means must be devised to tune and adjust a "legal" two kilowatt p.e.p. lineal amplifier without exceeding the limitations of the amplifier, and without breaking the law. A proper test signal having high peak to average power ratio will do the job, permitting the amplifier to run at less than a kilowatt d.c. input while allowing the 2 kw peak power level to be reached. This type of signal can be developed by an audio pulser, such as is described in QST magazine, August, 1947 (figure 9). The duty cycle of this simple pulser is about 0.44. This means when the amplifier is tuned up for a d.c. indicated meter reading 800 watts, using the pulser and single tone audio injection, the peak envelope power will just reach the 2 kw level. An oscilloscope and audio oscillator are necessary for this test, but these are required items in any well-equipped sideband station. Loading and drive adjustments for optimum linearity consistent with maximum power output may be conducted by this method.

A Tuned Cathode Network

Use of a tuned cathode network is recommended for optimum linearity and ease of drive (figure 10). As the input impedance of the amplifier is in excess of 50 ohms, a moderate value of s.w.r. will exist on the coaxial line coupling the amplifier to the exciter. With certain lengths of line, it may be found difficult to properly load the amplifier to maximum input because of a lack of excitation. This is a common difficulty encountered in amplifiers employing an untuned cathode circuit. A slight change in line length will alleviate this difficulty somewhat. Unfortunately, some SSB exciters have a fixed pi-network output circuit designed for operation into 50 ohm loads and no adjustment of exciter loading is possible. A tuned cathode network in the amplifier is required in this case to achieve a proper match between the exciter and the amplifier.

The cathode network should be placed directly at the filament circuit terminals of the amplifier, keeping the lead between capacitor C3 of the network (figure 10) and coupling capacitors C3 and C4 (figure 2) very short. The amplifier plate circuit return to the cathode now passes through capacitor C3 of the network instead of returning via the coaxial line and exciter tank circuit. In any grounded grid amplifier, the condition of maximum linearity may only be achieved by the use of such a network.

The network may be adjusted by placing an SWR bridge in the coaxial line to the exciter. When properly tuned, maximum grid current to the amplifier will coincide with minimum SWR on the line.


CALIBRATION

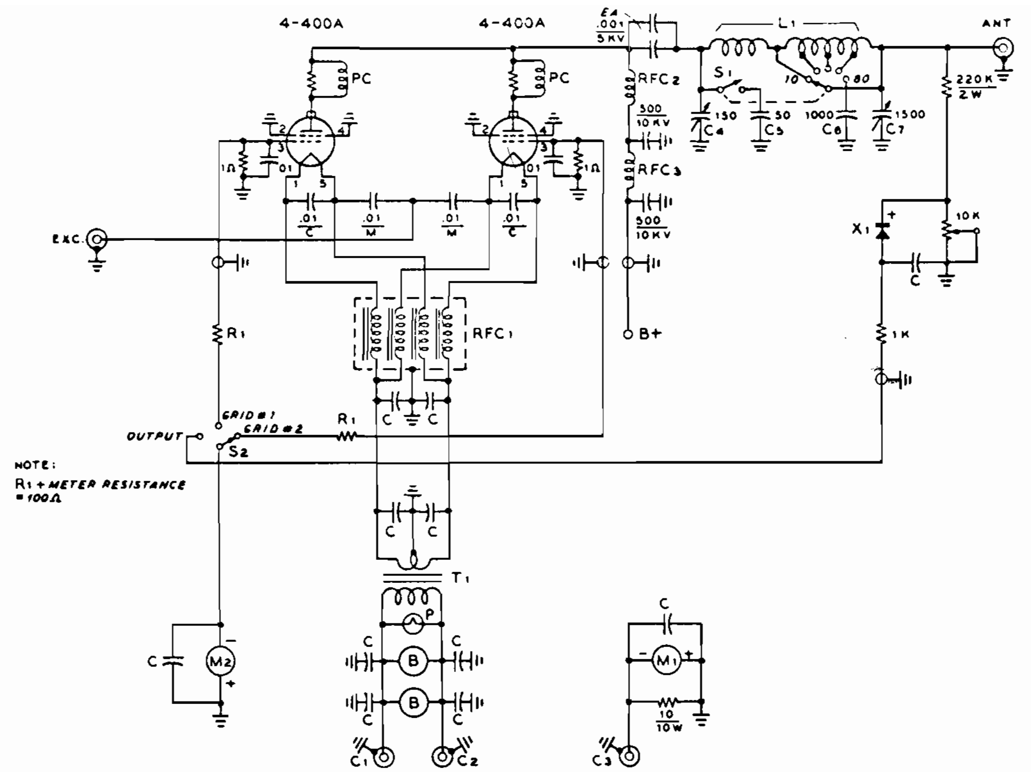


Fig. 2 Schematic, KW2

C--.001 μfd, 600 volt disc ceramic

C1-3, C8--0.1 μfd, 600 volt coaxial capacitor Sprague "Hypass" #80P3

C-4--150 μμfd, 4500 volt. Johnson #150D45

C-5--50 μμfd surplus vacuum capacitor (see text)

C-6--1000 μμfd, 1250 volt mica capacitor (see text)

C-7--1500 μμfd. Obtainable on special order from Barker & Williamson

Note: In drawing, C=ceramic, m=mica capacitor

L-1--Kilowatt pi-network coil. Air-dux #195-2S (silver plated). Modify as follows: Strap coil, 3 turns 1 3/4" diameter. Wire coil, remove turns from free end, leaving 11½ turns, counting from junction with tubing coil.

Tap placements:

10 meters: 1 3/4 turns from junction of tubing coil and strap coil.

15 meters: 3¼ as above.

20 meters: 1½ turns of wire coil, counting from junction with tubing coil.

40 meters: 5¼ as above.

80 meters: Complete coil in use.

RFC-1--30 ampere filament choke. B&W #FC-30

RFC-2--Kilowatt r.f. choke. Raypar, or B&W #800

RFC-3--VHF choke. Ohmite #Z-50

T-1--5 volts at 30 amperes. Stancor P-6468

PC--3½ turns #12e, 7/8" diam. 2" long. Wound around three 220 ohm, 2 watt composition resistors connected in parallel.

R-1--10 ohms, 10 watt, wirewound.

M-1--0-1000 ma. Triplett

M-2--0-1 ma. Triplett

X-1--Diode, type 1N34

B-1, B-2--Blower motor and fan. Barber Coleman #DYAB

P-1, P-2--Coaxial receptacle, SO-239

Chassis & Cabinet--California Co., Type "LTC" 17¼" x 13¼" x 12"

Dials--"Cal-Rad" 3 inch

Plate Blocking Capacitors--.001 mf, 5KV Centralab #858-S (two used).

Plate Bypass Capacitors--500 μμfd, 20 KV "TV-type". Mallory HV-20035B

Tubes--4-400A by EIMAC.

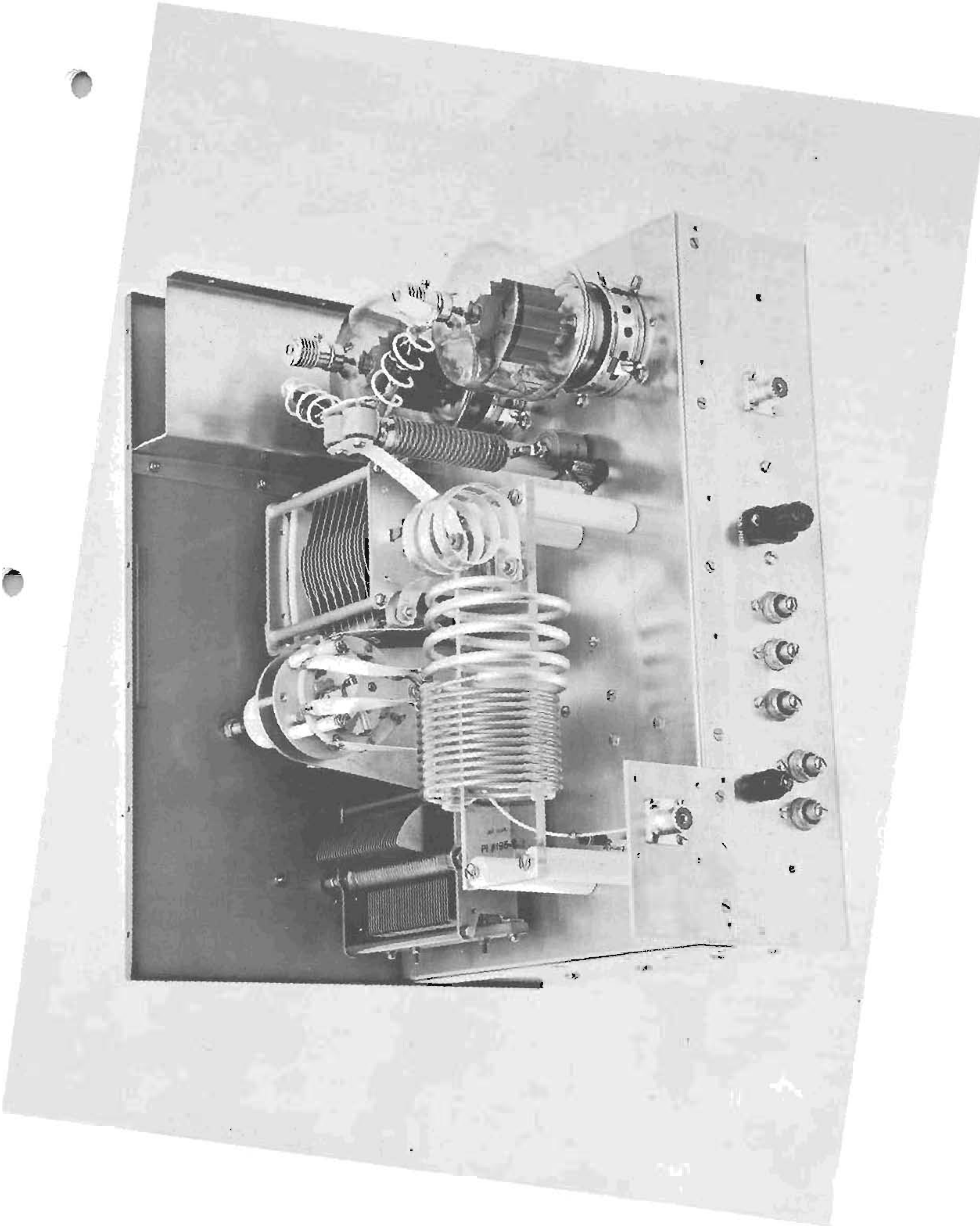


Fig. 8

OPERATING CHARACTERISTICS, EIMAC TETRODES, GROUNDED-GRID CONFIGURATION

4-125A

D-C Plate Voltage	2000	2500	3000	volts
Zero-Signal Plate Current	10	15	20	ma
Single-Tone Plate Current	105	110	115	ma
Single-Tone Screen Current	30	30	30	ma
Single-Tone Grid Current	55	55	55	ma
Single-Tone Driving Power	16	16	16	watts
Driving Impedance	340	340	340	ohms
Load Impedance	10,500	13,500	15,700	ohms
Plate Input Power	210	275	345	watts
Plate Output Power	145	190	240	watts

4-400A

(Ratings apply to 4-250A, within plate dissipation rating of 4-250A) (PER TUBE)

D-C Plate Voltage	2000	2500	3000	volts
Zero-Signal Plate Current	70	80	90	ma
Single-Tone Plate Current	265	270	330	ma
Single-Tone Screen Current	55	55	55	ma
Single-Tone Grid Current	100	100	100	ma
Single-Tone Driving Power	38	39	40	watts
Driving Impedance	160	150	140	ohms
Load Impedance	3950	4500	5000	ohms
Plate Input Power	530	675	990	watts
Plate Output Power	325	435	600	watts

4-1000A

D-C Plate Voltage	3000	4000	5000	volts
Zero-Signal Plate Current	100	120	150	ma
Signal-Tone Plate Current	700	675	540	ma
Signal-Tone Screen Current	105	80	55	ma
Single-Tone Grid Current	170	150	115	ma
Single-Tone Driving Power	130	105	70	watts
Driving Impedance	104	106	110	ohms
Load Impedance	2450	3450	5550	ohms
Plate Input Power	2100	2700	2700	watts
Plate Output Power	1475	1870	1900	watts

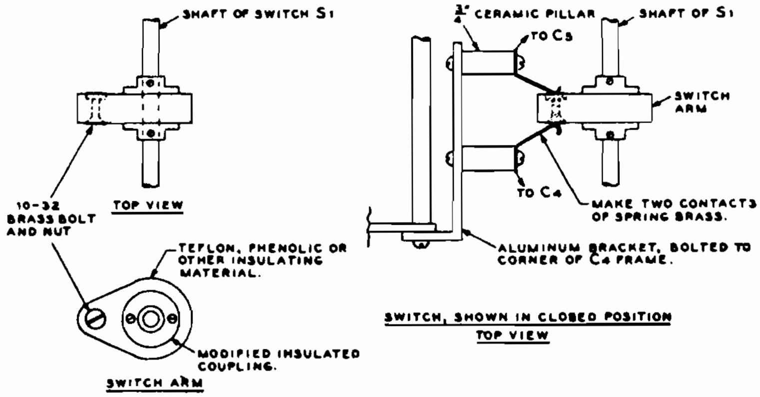


Fig. 4

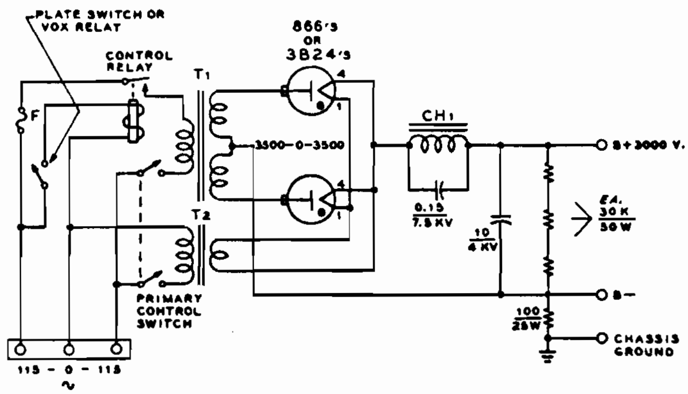


Fig. 5

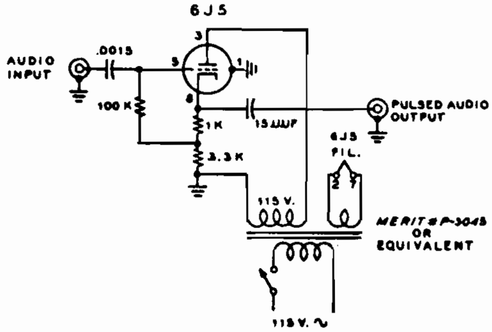
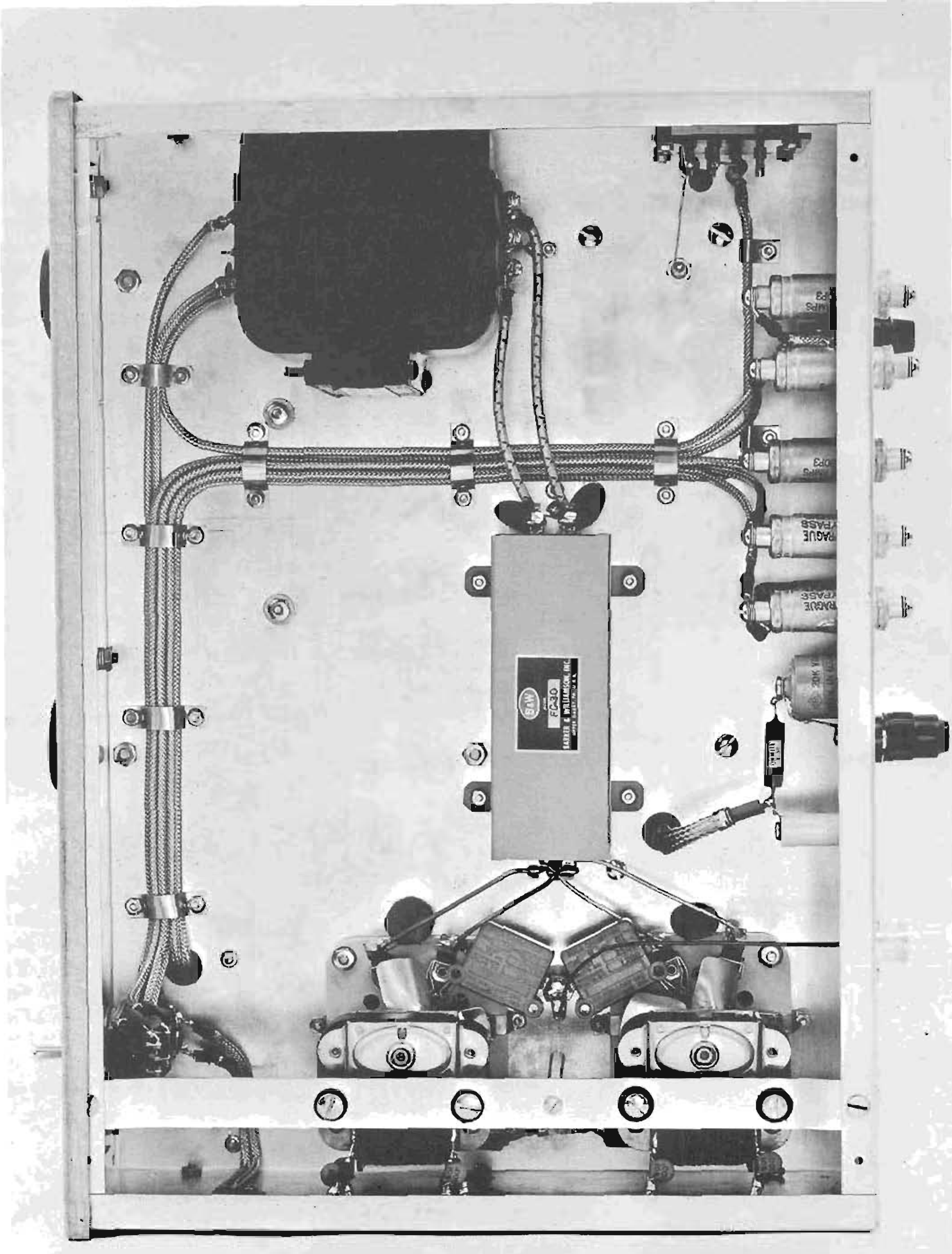


Fig. 9


B&W
FC-30

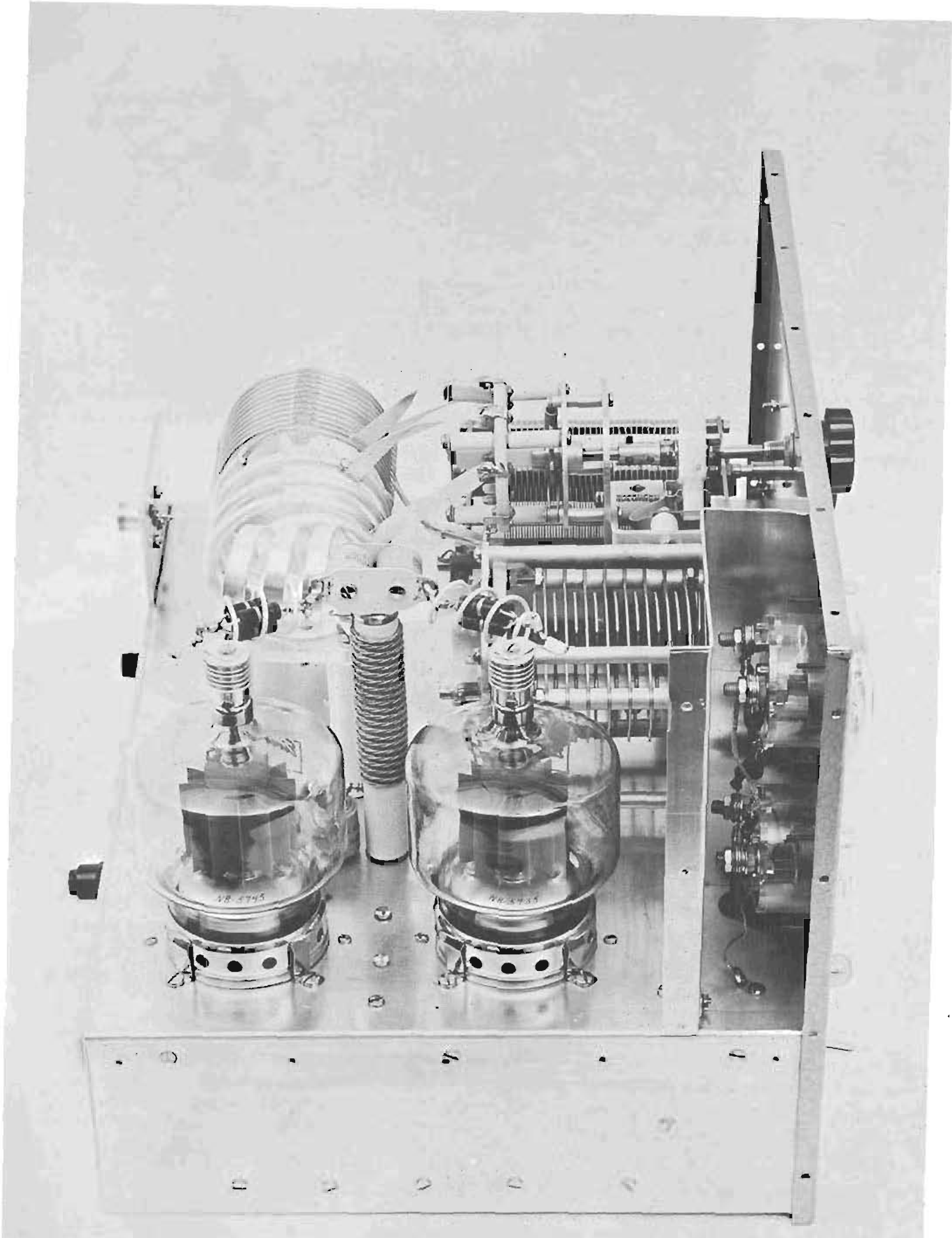
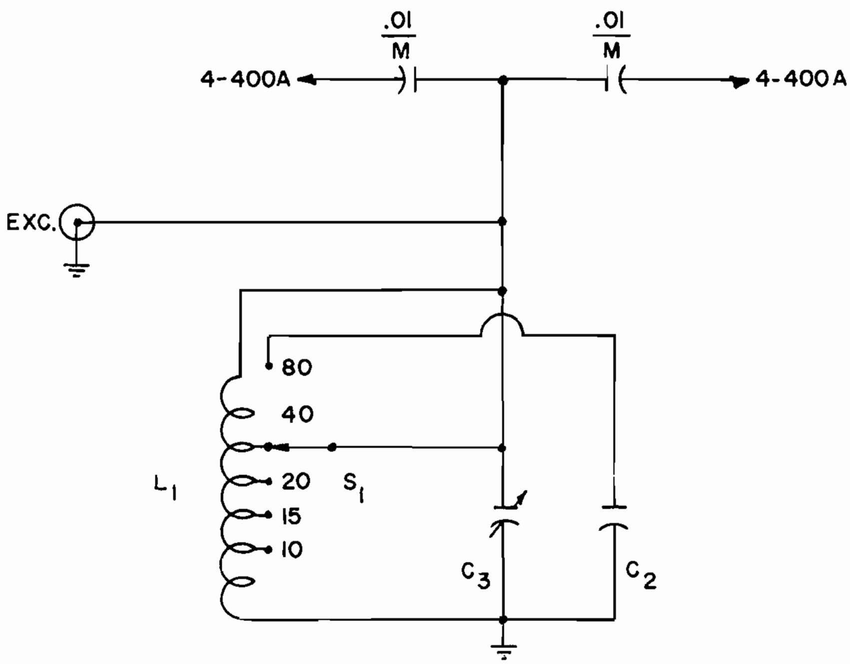

NB-5745
NB-5745

FIGURE 10



C_2 - 800 mmf mica, 1200v test

C_3 - Three gang b.c. capacitor. 1100 mmf. J.W. Miller #2113

L_1 - 9 turns #10e, 1" diam., 1-½" long. 40 meter tap 4-½ turns, 20 meter tap 2-½ turns, 15 meter tap 1-¾ turns, 10 meter tap 1/2 turn plus 2" lead, (all taps measured from the ground end).

Taps adjusted so circuit resonates with tubes in socket with following tuning capacitance:

80 meters -	1660 mmf
40 meters -	830 mmf
20 meters -	415 mmf
15 meters -	240 mmf
10 meters -	210 mmf

SEE HR-10 FOR ...

linear amplifier design

The designer of a linear amplifier should be concerned with the proper potentials required to make the power tube operate in a linear manner. The word *linear* implies that the output signal of the amplifier is an amplified replica of the input signal. There's no such thing as a perfect linear amplifier, and the designer's problem is to make the practical amplifier (*i.e.*, the amplifier that can be built) as linear as possible.

When a linear amplifier is driven by a complex signal, such as the human voice, nonlinearity results in intermodulation distortion. This unpleasant form of distortion creates a broad, raspy signal that throws annoying "buckshot" into adjacent channels. Proper design and operation of a linear amplifier reduces this distortion to a minimum.

amplifier circuit and mode

There's a lot of confusion with regard to the so-called "grounded-grid" amplifier. Rf power amplifiers are classified according to circuitry and mode of operation. The two classifications should not be confused with one another. For Amateur service, the two most popular circuits are the grid-driven circuit and the cathode-driven circuit. As shown in **fig. 1**, the circuits are remarkably similar, the most obvious difference being the placement of the ground point in relation to the input and output circuits.

The mode of operation refers to the dynamic operating characteristics of the tube (class AB_1, class B, or class C). Characteristics of the classes are given in reference material listed at the end of this article. For linear service, the power tube amplifier is commonly run in either class AB_1 or class B service. Thus, modern equipment may have an intermix of circuitry and mode — the cathode-driven amplifier may be operated in a class AB_1 mode, for example, or the grid-driven amplifier may be operated in the class B mode.

So far, I've not discussed the popular grounded-grid amplifier. This is a sloppy term which usually refers to a cathode-driven amplifier, working in the class B mode. "Grounded grid" implies cathode drive, but in such a circuit the grid may not necessarily be at dc ground potential, especially with respect to screen voltage (see **fig. 2**). Rf ground and dc ground are not always the same in a linear amplifier, and most circuit engineers shudder at the use of the term.

amplifier plate circuit

While this series of articles concerns itself with linear, cathode-driven-amplifier design, the remarks about the plate circuit apply equally well to grid-driven amplifiers. It is desirable to operate any linear amplifier with a very minimum of intermodulation distortion, with high-plate efficiency, and with high power gain. The latter is especially important, as it affords maximum power output with a given amount of drive power. The class B mode of operation meets these requirements.

Shown in **fig. 3** is a graphical representation of a class B amplifier, showing the operating cycle of the tube. This is the portion of the electrical cycle over which the tube grid is driven positive (approaching $+e$) with respect to the cathode (or the cathode driven negative with respect to the grid). When the grid potential is highly negative with respect to the cathode (approaching $-e$), the tube is cut off and is inoperative. In the class B amplifier, the operating cycle is about one-half the electrical cycle, or approximately 180 degrees. The transfer curve plot shown indicates that the tube delivers power only over one-half of the electrical cycle and is cut-off during the other half of the cycle. Does this mean that the output signal consists of half-sine waves as shown, and is therefore highly distorted? Not at all.

The amplifier plate circuit (often called the tank circuit) saves the day, since the energy storage ability (Q) of the circuit balances the energy between the halves of the cycle, much as the flywheel stores energy during the operating cycles of a gasoline engine. The plate circuit must, therefore, be designed to have sufficient Q, or energy storage, for good operation. A Q value of 12 is commonly used for linear amplifier service, as it provides ample energy storage and at the same time provides reasonable reduction of harmonics generated in the amplifier.

By William I. Orr, W6SAI, 48 Campbell Lane, Menlo Park, California 94025

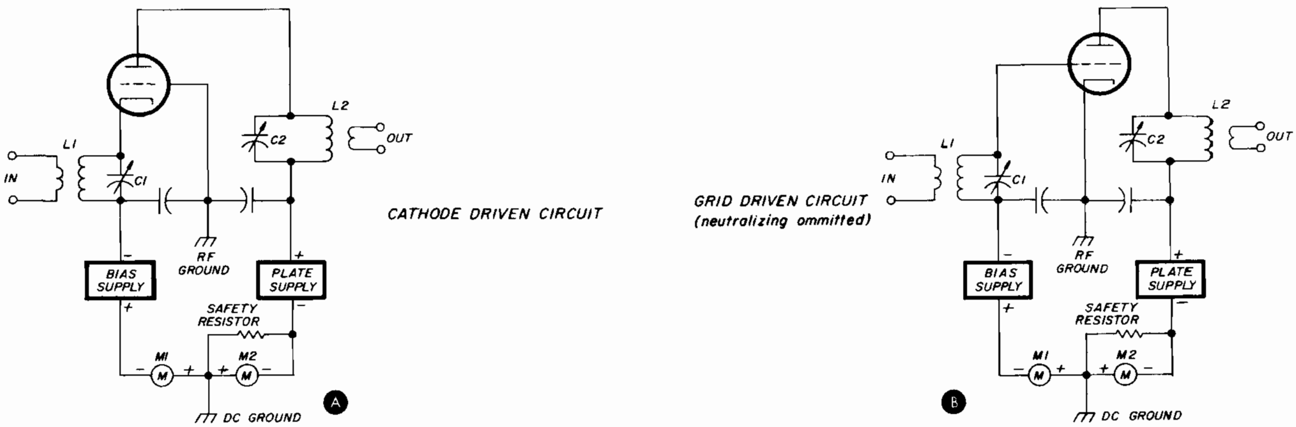


fig. 1. A comparison between grid-driven and cathode-driven amplifiers. Rf and dc circuits have been simplified for clarity. In both cases, the grid- and plate-current meters are placed in the ground return circuits to remove any dangerous voltage from the meter movement. This, however, places the plate supply above dc ground by virtue of the voltage across the plate meter. If the meter coil should open, the negative lead of the supply rises to the value of the plate voltage. As a safety factor, a wirewound resistor is usually placed across the plate meter, and often the grid meter. The circuit configuration determines the difference between cathode- and grid-driven service. The applied voltages determine the mode of operation.

A rigorous design of the plate circuit calls for manipulation of the plate voltage and current to determine the operating parameters of the tube. The results of these tedious calculations can be summed up in simple formulas that provide the designer with circuit data in everyday terms.

A network is required that matches the plate load impedance of the power tube to the characteristic impedance of the transmission line, while at the same time maintaining a Q value of 12. The popular pi network can do the job. The plate load impedance (Z_L) for a class B rf amplifier can be closely approximated by:

$$\text{load impedance (ohms)} = \frac{\text{plate voltage}}{2 \times \text{peak dc plate current (amperes)}}$$

As an example, a pi network is to be used to match a pair of 3-500Z tubes to a 50-ohm transmission line. The tubes operate with 2500 volts plate potential with a peak dc plate current of 800 mA (0.8 amp) for a PEP input of 2 kW.

$$\text{load impedance} = \frac{2500}{2 \times 0.8} = 1560 \text{ ohms}$$

Thus, the pi network plate circuit has to match a load impedance of 1560 ohms to a 50-ohm termination.

designing the plate circuit network

The approximate values of the pi network can be determined from three simple graphs. The plate inductance from **fig. 4**, the tuning capacitance (C1) from **fig. 5**, and the loading capacitance (C2) from **fig. 6**. The graphs are entered at the x axis and read up until the sloping line denoting a particular Amateur band is intersected. The value of the component is then read horizontally off the y axis. For example, the required inductance for a plate load of 1560 ohms for the 15 meter band is about one microhenry — as close as the graph can be read. Note that capacitor C1 is commonly referred to as the tuning capacitor and C2 the loading capacitor.

The graph for C2 tells us that the pi network cannot cope with impedance transformation values much greater than 100-to-1 at this value of Q. Note how the curves bunch together and "fall-off the graph" at plate impedances much higher than 5000 ohms.

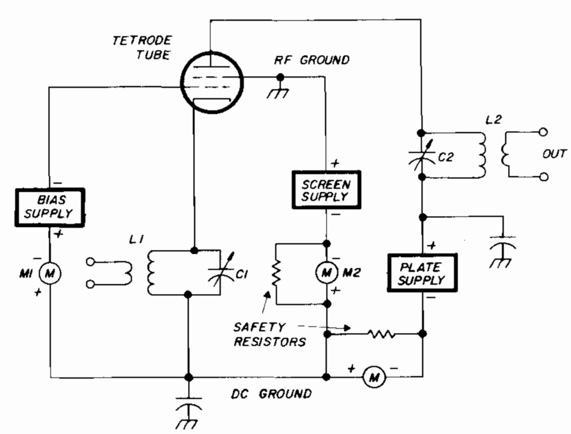


fig. 2. Diagram of the so-called "grounded-grid" amplifier. The grid and screen elements are bypassed to ground as far as rf is concerned, but each element has normal operating voltages applied and are "above ground" as far as dc is concerned. Metering is inserted in the supply return leads to dc ground. Rf ground is placed at the positive screen voltage level. This eliminates the screen bypass capacitor, a tricky component that often causes circuit instability at the higher frequencies.

A more accurate, computer-derived summary of pi network values is given in **table 1**. Note that, for a given plate impedance, when the operating frequency is doubled the capacitance and inductance values are halved. (Fifteen- and forty-meter constants are related by a factor of three as 21 MHz is the third harmonic of 7 MHz.)

coil winding

Winding plate coil L1 to a given value of inductance takes an inductance meter, or a degree of exper-

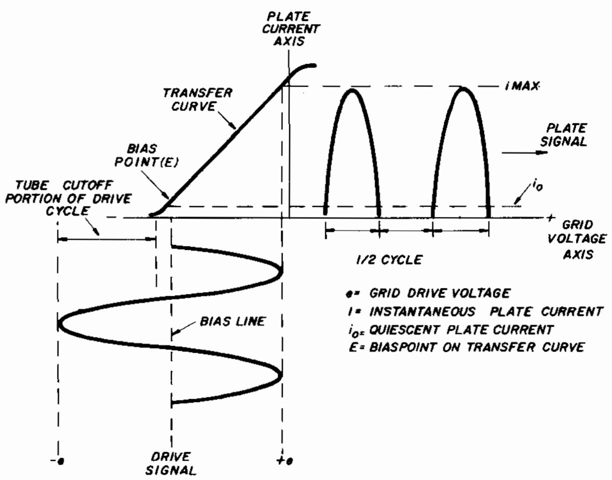


fig. 3. Transfer curve and operating cycle for a class B amplifier. The transfer curve is determined by a static test of the tube where plate current is plotted against grid bias. Once the transfer curve is established, the operating cycle may be determined. The sine wave drive signal (e) is drawn about the bias line, determining both the zero-signal plate current (i_o) and the peak plate current (i_{max}). Note that when the grid driving signal swings negative, no plate current is drawn and the tube is cut-off for one-half cycle. Pulses of plate current only appear when the drive signal is positive with respect to the bias voltage. Thus, the output waveform of a class B rf amplifier consists of a series of half-cycles, much in the manner of a half-wave rectifier. The distorted waveform is restored to a sine wave by the plate tank circuit which, by virtue of its *Q*, or flywheel effect, stores energy on the active half of the cycle and releases it on the inactive half. Circuit engineers, working from a transfer curve, can determine actual dc operating potentials for a linear amplifier.

tise and a dip-meter. A simple formula for calculating inductance when the coil dimensions are known is:

$$\text{Inductance } (\mu H) = \frac{R^2N^2}{9R + 10S}$$

where R is the radius of the coil in inches
S is the length of the coil winding in inches
N is the number of turns

These calculations have been simplified in the ARRL type-A "Lightning Calculator," which is a sim-

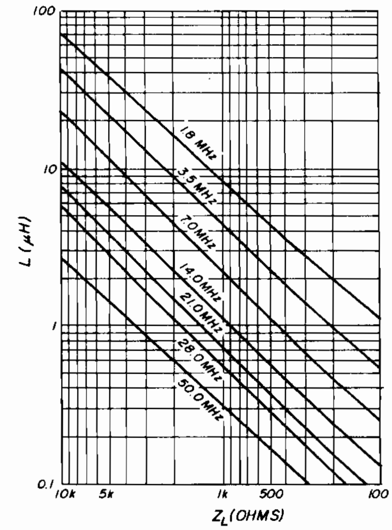


fig. 4. Plot of the plate inductance vs. plate load impedance for the high frequency Amateur bands (Q = 12).

ple slide rule providing direct read-out of the coil dimensions if the inductance is known. It takes the hard work out of designing coils.

Once the plate circuit has been designed and built, it is a good idea to "breadboard" it up and check it out with a dip-meter before the connections are finally soldered. Coil taps may have to be moved a bit to compensate for capacitance of the components to the chassis and adjacent parts.

amplifier-cathode circuit

The cathode-input circuit provides an impedance match between the 50-ohm coaxial output circuit of the driver/exciter and the input impedance of the cathode-driven amplifier (see **table 2**). The input im-

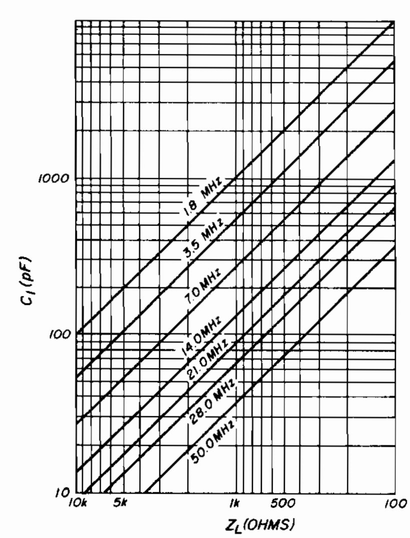


fig. 5. Plot of the tuning capacitance (C1) vs. plate load impedance (Q = 12).

pedance (Z_t) of a cathode-driven tube is related to the ratio of the peak cathode signal voltage to the peak cathode current (sum of grid and plate currents), and is commonly given in the tube data sheet. For the 3-500Z at 2500 volts, it is about 110 ohms. And for two tubes in parallel, it is about 55 ohms, but *only* over the operating cycle.

It is tempting to jump to the conclusion that if the amplifier input impedance is about 55 ohms and the coaxial line impedance driving it is 50 ohms, that no cathode impedance matching circuit is required. In fact, many commercially manufactured amplifiers leave it out for economy's sake. This omission is poor engineering practice, as the circuit Q is required in the cathode circuit as well as in the plate circuit. Omission of the cathode-tuned circuit can lead to distortion of the driving signal, increased intermodulation distortion, reduced amplifier efficiency, and driver loading problems. A circuit Q of 2 is adequate, and a simple rule of thumb is that the network circuit capacitances at resonance should be about 20 pF per meter of wavelength for one-to-one impedance transformation.

practical amplifier circuit

Armed with the information discussed so far, it is possible to draw up a schematic for a cathode driven, 2-kW PEP linear amplifier using two 3-500Z tubes in parallel (see **fig. 7**). This is a true "grounded-grid" circuit, as the grids are at both dc and rf ground potential.

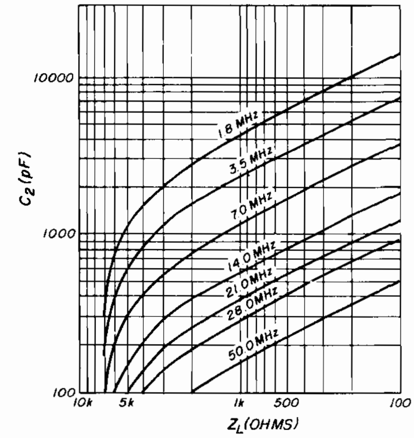


fig. 6. Plot of the loading capacitance (C2) vs. plate load impedance (Q = 12).

Note that plate and grid currents are measured in the cathode return circuit. This requires the amplifier plate power supply to "float" a little above ground potential in order to insert a meter in the negative lead to measure plate current. This removes the lethal plate voltage from the meter. The grid meter is out of the critical rf ground return path, which simplifies the metering circuit. A filament voltmeter is included. Filament voltage should be held to within

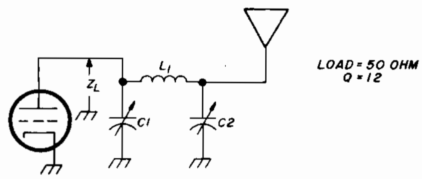


table 1. Computer-derived values for a pi network having a Q of 12 and working into a 50-ohm load. Values for C1 include the output capacitance of the tubes. These values are taken from a computer program derived by Bob Sutherland, W6PO.

component		Z_L plate load impedance (ohms)							
	band	1000	1500	2000	2500	3000	3500	4000	5000
C1	160	1060	690	531	430	354	309	265	212
	80	546	364	273	220	182	159	136	109
	40	273	182	136	110	91	80	68	55
	20	136	91	68	55	45	40	34	27
	15	91	61	45	37	30	26	23	18
	10	68	45	34	30	23	20	17	14
C2	160	4421	3487	2865	2440	2105	1849	1594	1186
	80	2274	1784	1473	1263	1082	951	820	610
	40	1137	892	737	632	541	475	410	305
	20	568	446	368	316	271	237	205	153
	15	379	297	246	211	180	158	137	102
	10	284	223	184	158	135	118	102	76
L1	160	8.84	13.26	16.61	20.10	24.13	27.80	31.47	38.63
	80	4.55	6.57	8.54	10.90	12.41	14.29	16.18	19.87
	40	2.27	3.28	4.27	5.50	6.20	7.15	8.09	9.93
	20	1.14	1.64	2.14	2.70	3.10	3.57	4.05	4.97
	15	0.76	1.09	1.42	1.82	2.07	2.38	2.70	3.31
	10	0.57	0.82	1.07	1.36	1.55	1.78	2.02	2.48

±5 per cent of 5 volts, and it is prudent to monitor this voltage when expensive tubes are used. A plate voltmeter may be included in the amplifier, but it is easier to place it in the power supply.

Amplifier standby plate current is reduced by means of a 10-kilohm, 25-watt cathode resistor which is shorted out by the VOX relay of the exciter, causing the tubes to operate at the proper resting plate current when the amplifier is on the air. A zener diode is placed in series with the cathode dc return path to reduce the quiescent plate current during amplifier operation.

A 50-ohm wirewound resistor from the negative side of the plate supply to ground makes certain that the negative supply terminal does not rise to the value of the plate voltage if the positive side of the supply is accidentally shorted to ground.

Two reverse-connected diodes are shunted across the safety resistor to limit any transient surges under a shorted condition which might cause wiring insulation breakdown. In addition, the diodes protect the meters from transient currents. A resistor across the zener diode provides a constant load for it and prevents cathode voltage from soaring if the zener safety fuse opens.

Note that a 10-ohm, 50-watt wirewound resistor is placed in series with the B-plus lead to the plate rf choke. This resistor serves as a vhf choke to suppress harmonic currents in the power lead and also protects the tube and associated circuitry in case of a flash-over in the tube or plate circuit. The tremendous amount of energy stored in the power supply is instantaneously "dumped" into the amplifier when a

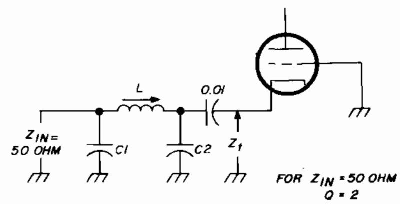


table 2. The pi-network circuit for a cathode-driven amplifier. This chart provides approximate values for the components of the cathode circuit. Capacitors should be 1-kV silver mica or equivalent. The inductor can be wound on a slug-tuned form. Value of C2 should take into account the cathode-grid capacitance of the tube which appears in parallel with C2 (information is from a computer program by W6PO).

cathode Z_t (Ω)	band	C1(pF)	C2(pF)	L(μH)	cathode Z_t(Ω)	band	C1(pF)	C2(pF)	L1(μH)
	160	3300	4100	2.50		160	3300	2870	3.81
	80	1700	2120	1.34		80	1700	1540	2.05
20	40	900	1120	0.68	75	40	900	770	1.03
	20	440	560	0.33		20	440	380	0.51
	15	300	370	0.22		15	300	250	0.34
	10	220	275	0.16		10	220	180	0.25
	160	3300	3900	2.84		160	3300	2520	4.20
	80	1700	2100	1.52		80	1700	1350	2.26
30	40	900	1050	0.77	100	40	900	680	1.14
	20	440	520	0.38		20	440	330	0.56
	15	300	350	0.25		15	300	220	0.38
	10	220	258	0.19		10	220	160	0.28
	160	3300	3360	3.01		160	3300	2100	4.81
	80	1700	1800	1.62		80	1700	1130	2.59
40	40	900	910	0.82	150	40	900	570	1.30
	20	440	450	0.40		20	440	280	0.66
	15	300	300	0.27		15	300	180	0.43
	10	220	220	0.20		10	220	138	0.32
	160	3300	3300	3.33		160	3300	1800	5.32
	80	1700	1700	1.79		80	1700	980	2.86
50	40	900	900	0.90	200	40	900	490	1.44
	20	440	440	0.45		20	440	245	0.71
	15	300	300	0.30		15	300	164	0.48
	10	220	220	0.22		10	220	120	0.35
	160	3300	3100	3.53		160	3300	1640	5.78
	80	1700	1670	1.90		80	1700	880	3.11
60	40	900	840	0.96	250	40	900	440	1.57
	20	440	417	0.47		20	440	220	0.78
	15	300	275	0.32		15	300	140	0.52
	10	220	205	0.23		10	220	100	0.38

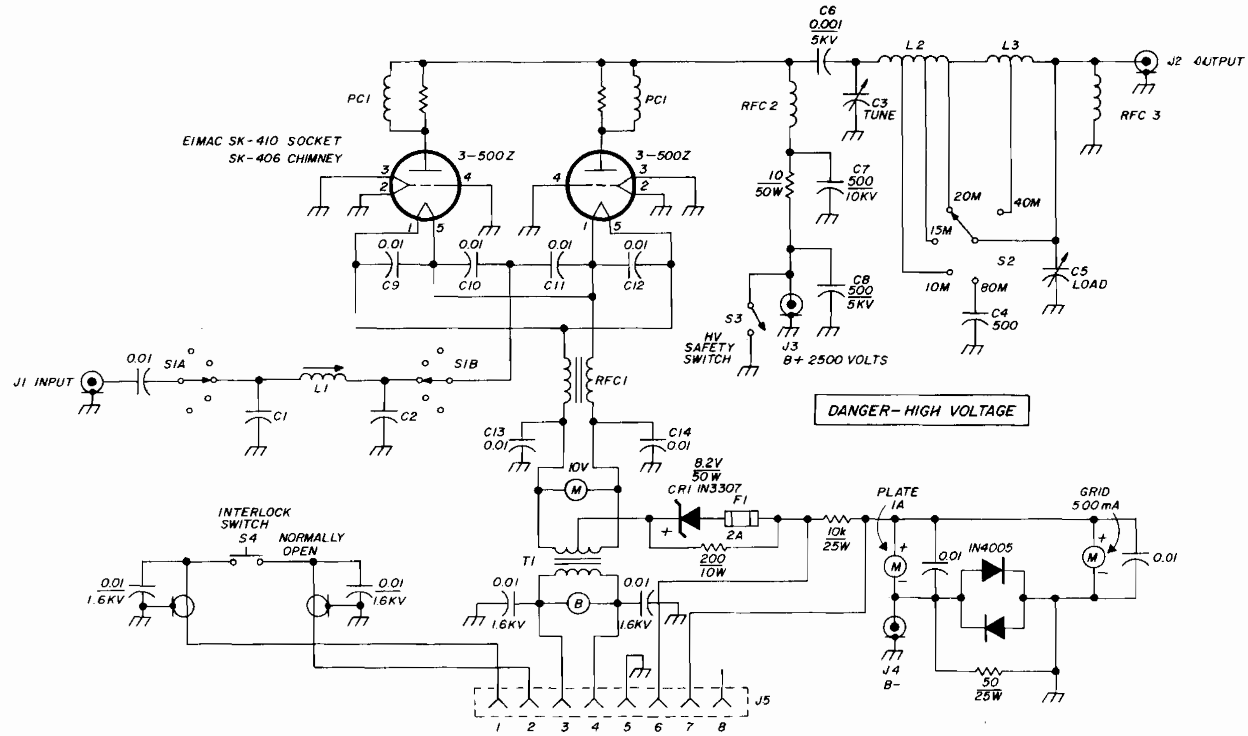


fig. 7. Schematic diagram of the 3-500Z linear amplifier.

C3	250 pF, 4.5 kV plate spacing — Johnson 154-16
C4	500 pF, 4.5 kV
C5	1000 pF, 500 volt plate spacing
C6	0.001 μF, 5 kV — Centralab 858S-1000
C7, C8	500 pF, 10 kV TV-type "door knob"
C9-C14	0.01 μF, 500 volt mica capacitor. Ceramic disc is a suitable substitute if rated 1 kV.
PC 1	Three 100-ohm, 2-watt resistors in parallel
PC 2	Three turns of no. 14 AWG (1.6 mm) wound with 12.5-mm (0.5-inch) diameter and 19-mm (0.75-inch) length connected in parallel with the resistors. The coil may be wound around one of the resistors.
RFC 1	50 μH; 14 bifilar turns of no. 10 AWG (2.6 mm) enameled wire wound on ferrite core 12.5 cm (5 inches) long and 12.5 cm (0.5 inch) in diameter (Indiana General CF-503 or equivalent).
RFC 2	100 μH, 1 ampere dc; 112 turns no. 26 AWG (0.4 mm) spacewound wire diameter on 2.5 cm (1 inch) ceramic form 15 cm (6 inches) long (Centralab X-3022H insulator). Series resonant at 24.5 MHz with terminals shorted (B&W 800).
RFC 3	2.5 mH, 100 mA
T1	5 volts at 30 amps (Chicago-Standard P-4648)
Blower	13 cu. ft./min. Use a no. 3 impeller at 3100 rpm (Ripley 8472, Dayton 1C-180, or Redmond AK-2H-01AX)

flash-over occurs, and much of this destructive energy is dissipated in the resistor.

Many modern-generation Amateurs have never worked with equipment operating at voltages higher than 12 volts. This amplifier, with the high-voltage plate supply, is positively lethal and the operator can be killed if his hands are inside the unit when the high voltage is on. It is *imperative*, therefore, that safety switches be incorporated in the amplifier design. It is poor engineering practice to leave these devices out! S4 is a normally open, pushbutton device that is closed only when the lid is placed on the amplifier enclosure. S3 is a shorting switch that shorts the high voltage to ground when the lid is removed. Construction of this special switch will be covered in a future article. *Always remember — high voltage kills*! Take necessary precautions.

Although not shown on the schematic, it is a good idea to use a filament transformer having a primary winding tapped for 105, 115, and 125 volts. This provides a plus or minus ten per cent adjustment from a normal line voltage of 115 volts. If a closer filament adjustment is desirable, the transformer can be run on the 105 volt tap with a rheostat in series with the primary winding to place the filament voltage "on the nose."

The plus and minus leads to the high voltage supply should be run through high-voltage connectors and high-voltage cable. Test prod wire having a 10-kV breakdown is satisfactory. As an alternative, RG-58/U coaxial cable can be used for high-voltage leads along with PL-259 plugs and reducers and SO-239 receptacles. The shield of the coaxial line is grounded by the connectors. **ham radio**

amateur service newsletter
W6SAI

PUTTING THE 3-400Z TUBE TO WORK

Using A Zero Bias Tube in a Grounded Grid 1 Kilowatt Linear Amplifier

The "ideal" linear amplifier package would contain no more than a tube, a filament transformer, a plate supply and a tuned circuit. It would be simple to build and cost but a few pennies. Unfortunately, such a perfect device does not yet exist, and is not foreseeable in the near future. On the contrary, the relatively simple linear amplifier has "grown" to become an object of astounding complexity, requiring grid bias supplies, regulated screen supplies, power dissipating grid resistors and other awesome and complicated devices that add to the cost and weight of the linear, but do nothing to make the signal louder or clearer at the receiver. Indeed, some linear amplifier designs have been almost lost in the maze and complexity of expensive, regulated power supplies required to make the beast "tick".

A large quantity of auxiliary equipment can be swept aside and junked if a zero bias tube is employed in a simple grounded grid configuration, such as shown in Figure 2. Various types of transmitting tubes (originally designed for grid driven service) such as the 813, 811A and 4-400A have been used with success as "zero bias" grounded grid amplifiers, but no true zero bias triode of large power capability has been at hand for this class of service. The amplifier described in this article is designed around the Eimac 3-400Z, a member of a family of zero bias triode tubes now available to the amateur.

The 3-400Z Zero Bias Tube

The new 3-400Z tube is a high μ triode having a plate dissipation of 400 watts. It is rated to 1 kilowatt d c input for linear amplifier service (figure 2). Within the maximum plate voltage rating of 3000 volts, the 3-400Z has the very desirable characteristic of having no need for either a bias or a screen power supply. The Old Timers will remember with nostalgia the old '46 tube. (Remember the 160 meter pre-war transmitter using a flock of these bottles)? When excitation was removed from the '46 it would simply relax and stop working. The 3-400Z will do this trick, too.

The seated height of the 3-400Z is only 4½" to the top of the plate radiator cap, making it extremely attractive for the new, modern concept of linear amplifier design. Because of the small tube size, and because no one has yet been able to miniaturize a watt, it is necessary to cool the tube seals, envelope and plate lead with an auxiliary blower.

Elimination of the bias and screen supplies allows a large savings in cash normally spent for these items, and also saves the builder the labor (and skinned knuckles) required to drill the holes, mount the parts, and do the necessary wiring on these electronic nuisances. A large bonus in the forms of simplicity and low cost accrues to the user of a zero bias tube!

The Amplifier Circuit

The 3-400Z grounded amplifier circuit is shown in Figure 2. It is designed for an input of 1 kilowatt PEP sideband, or 1 kilowatt CW operation. In addition, it may be run as an AM linear amplifier at an input level of 600 watts. Bandswitching circuits are ganged, and cover the amateur bands between 3.5 and 29.7 MHz with generous overlaps. A high-C pi-network output circuit is used to enhance a high order of linearity. It is necessary to monitor the output level of any linear stage, and a simple semiconductor voltmeter is incorporated in the output portion of the network. The voltmeter range is variable, as absolute readings are not necessary.

Proper operation of the amplifier may be established by maintaining a given ratio between grid and plate current. The grounded grid, therefore, is "ungrounded" sufficiently to permit insertion of a simple metering circuit. Done properly, the stability and operation of the grounded grid circuit remain unchanged. To achieve this, each grid pin of the 3-400Z socket is grounded by a low impedance resistor-capacitor combination. The resistors act as a shunt across the milliammeter, but have a value sufficiently high so as not to disturb the calibration of the meter to any great degree.

Plate current is measured in the negative lead of the power supply rather than in the filament return circuit, as the latter current is a combination of grid and plate current. The negative of the power supply is "above ground" by the voltage drop across R1 so it is necessary to "float" the power supply above the chassis as shown in Figure 2.

The driving impedance of the 3-400Z is a nominal 122 ohms. Since this figure varies widely over the operating cycle, a high-C tuned cathode circuit is employed to present a constant load impedance to the exciter. Filament voltage is applied to the tube via a bifilar coil, and excitation is applied to taps on the coil which are set for minimum standing wave ratio on the coaxial line from the exciter. The usual driving difficulties experienced with grounded grid amplifiers are entirely absent and no coupling problems are noted when switching from band to band. Increased power output, reduced intermodulation distortion, and ease of drive are gained when a tuned cathode circuit is used in preference to the old-fashioned untuned r f choke input circuit.

Construction of the Tuned Cathode Circuit

The tuned cathode circuit was built as a complete sub-assembly in a manner similar to a conventional grid-driven type coil turret. The unit consists of a bifilar coil, a suitable tuning capacitor, a bandswitch, and appropriate bypass capacitors. The photographs show various views of this sub-assembly.

The bifilar coil is wound from a 61" length of standard 3/16" diameter soft copper tubing, available at auto parts houses, refrigerator repair departments, and large hardware stores. Before the coil is wound, a length of #12 Formvar insulated copper wire is passed through the tubing, leaving about three inches protruding from each end. Be sure you sand the ends of the tubing to a smooth, rounded surface to prevent the insulation of the wire from being scraped or marred during this operation. Copper wire with enamel insulation should not be used as the enamel is too soft and may be easily damaged. Next, the coil is wound around a 1-5/8" diameter form (a section of water pipe may be used), and the turns are spread apart as required.

Soft 3/16" copper strap is used for the bandswitch leads, and the taps should now be soldered in position. The 50 ohm driving points are tapped with #18 enameled wire.

The completed coil is mounted on a piece of ¼-inch bakelite or phenolic sheet measuring 4" x 1-5/8". The sheet is drilled and tapped to mount vertically on small ceramic standoff insulators bolted to the sub-assembly chassis. The chassis measures 6" x 4", with a 2-3/8" lip on the front end.

When mounting the bandswitch, keep in mind that the plate inductor and the cathode inductor will be switched simultaneously by means of a chain and sprocket drive. Therefore, the cathode turret must have the 80 meter setting fall in the full clockwise position corresponding to the tap sequence of the plate turret.

The capacitor in series with the exciter input (C23) carries the full excitation current and must be a transmitting-type mica unit. Filament capacitors C8 and C9 are paralleled ceramic units chosen to conserve space and yet provide sufficient capacity to insure that the secondary of transformer T1 is at r f ground potential. These capacitors are mounted directly at the "cold" terminals of the bifilar filament coil. The plate-cathode r f return circuit is via the cathode tuning capacitor, C12. The lead from the stator terminals of C12 to the filament circuit and the bifilar coil is made of ¼-inch copper strap.

The series input capacitor is wired directly to the arm of the bandswitch with copper strap. The center conductor of the coaxial line from the exciter input receptacle is soldered to the capacitor terminal and the shield is grounded directly to the frame of the cathode tuning capacitor. The impedance of this tuned circuit is extremely low, and care must be taken in the design and assembly to make sure that the impedance is in the tuned circuit, and not in the various interconnecting leads and switches.

Amplifier Construction

This little powerhouse measures only 8-¾" high, 14" wide, and 15" deep - small enough to sit on the desk beside your sideband exciter or receiver. Construction is unique in that no chassis is used. The cabinet serves as the chassis! The TVI-proof enclosure is fabricated from 0.063" aluminum sheet and ½ inch aluminum angle stock. The front panel is cut from 1/8-inch dural and measures 8-¾" high by 14" wide. The sub-panel and rear panel are cut of the thinner aluminum to the same dimensions. All three pieces are framed with the corner stock as shown in the illustrations. Spacing between the panel and the sub-panel is 2½".

The bottom of the enclosure is formed in the shape of a "U", wrapping around the sides of the unit. This piece measures 14" wide and 15" deep. The sides turn up 3-5/8". The front panel is set back ½ inch from the edge, and is held to the sub-panel by means of four corner posts cut from ½-inch square aluminum stock.

The top cover is also "U" shaped, and is made of perforated aluminum to allow the exhaust air to escape from the main compartment. The cover measures 14" wide, 15" deep and 5-1/8" high. Top and bottom pieces are attached to the frame by means of sheet metal screws.

The input circuit of the amplifier is contained within an "L" shaped box, as shown in the under-chassis photograph (figure 5). The compartment is approximately 12" deep (this depth is determined by the finish dimension between the sub-panel and the rear panel) and 3-¾" high. It has two ½-inch lips, one along the side and the other along the bottom. Together with the bottom cover and the panels, it makes an r f tight and air tight compartment for the cathode input circuit and blower.

The plate circuit components require no chassis. The two pi-network capacitors are mounted to the sub-panel by means of 6-32 machine screws and spacers. The plate coil is affixed in a similar fashion (figure 3).

Component Layout and Assembly

General component placement may be seen from the photographs. The panel meters are isolated from the r f circuits by virtue of the sub-panel. In the meter area is also located the chain drive for the cathode tank circuit (figure 6). The plate bandswitching inductor and the cathode circuit switch are ganged for ease of operation. A two-to-one drive ratio is needed as the plate inductor has 60-degree indexing and the cathode switch has 30-degree indexing.

The filament transformer is placed at the front of the chassis-box. Although slightly under rated, this unit has operated for hours with no evidence of overheating. The tube socket and chimney are centered on the chassis-box 5½" behind the sub-panel, and the remaining space is occupied by the centrifugal blower and motor (figure 3). A **Johnson** ceramic socket was used for the tube, but the new **EIMAC SK-410** air socket and **SK-416** chimney are recommended as an inexpensive substitute.

The bandswitching plate inductor is at the opposite side of the main compartment. The unit is rated at 500 watts input. However, it was disassembled, silver-plated, and modified for one kilowatt sideband and CW operation. A new 10 meter coil section was wound, and the turret taps altered to provide the proper L/C ratio for optimum amplifier linearity (see parts list).

Amplifier Wiring

Shielded wire is employed for all low voltage circuits and small "feedthrough" capacitors pass the leads from the amplifier compartment into the meter compartment. Coaxial capacitors are employed for the low voltage terminals on the rear apron of the chassis. Silver-plated, ½-inch wide copper strap is used for the output wiring of the pi-network circuit. The four stator sections of the output capacitor of the network are paralleled by a short length of strap. All wiring is short, and direct.

Testing the Amplifier

The amplifier is entirely free from unwanted regenerations or parasites, and operation is simple and straightforward. It is designed to operate with a 2500 volt, 400 milliampere power supply of good regulation. Preliminary adjustments should be made at reduced plate voltage and a minimum value of excitation. Excitation should never be applied without plate voltage being on the stage. Once resonance is established, the tube should be loaded up to approximately 400 mA plate current. The grid current at this particular operating point should be about 140 mA. The ratio of about three plate milliamperes to one grid milliampere should be maintained for all operating conditions. If the grid current is excessive, it indicates that the plate circuit loading is too light. Low grid current indicates that plate loading is too heavy. As a final check, it should be observed that the power output of the stage (as observed on the output voltmeter) should increase in direct proportion to the excitation level. Finally, to achieve a condition of maximum linearity, the plate output circuit should be **overcoupled** (by decreasing the value of the pi-network output capacitor) until power output drops about 3%. With a two-tone test signal, the maximum signal plate current read on the meter will be 275 mA, and the grid current will be about 80 mA. With an average voice, plate current as read on the meter should kick to about 180 to 200 milliamperes, with grid current peaks of about 60 to 70 milliamperes. PEP input under these conditions will be one kilowatt, and all spurious distortion products will be reduced better than -35 decibels below peak signal level. Under proper operating conditions, signal-to-distortion ratios better than -42 decibels with a two-tone test signal have been achieved with this tube in this circuit. Distortion ratios of this order can only be obtained otherwise with conventional amateur tubes employed in feedback circuits.

The cost of all parts, including the tube, air socket, and chimney is under two hundred dollars. Amateurs owning a good junk box, or who are "surplus hounds" can cut this cost figure considerably. Considered both on a watts-per-dollar basis, and on a linearity basis, this little powerhouse is hard to beat for maximum performance!

Harold C. Barber, W6GQK and
Robert I. Sutherland, W6UOV

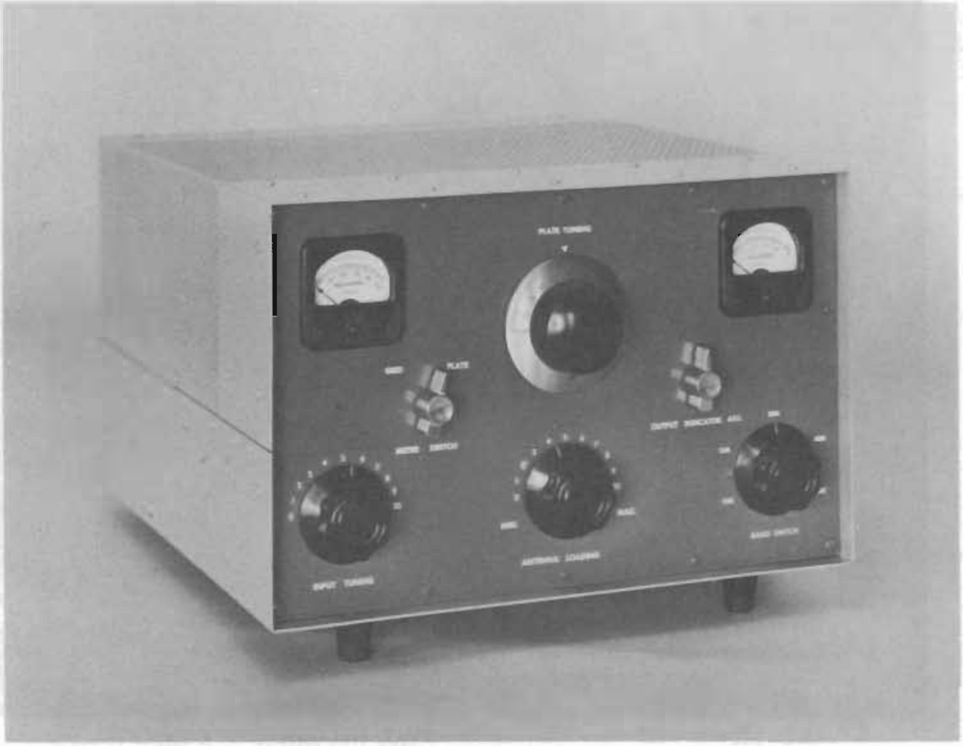


Figure 1. 3-400Z grounded grid amplifier runs 1 kw PEP input on all amateur bands between 10 a
80 meters, with a distortion figure better than -35 decibels below maximum output. Amplifier is enclos
in TVI-proof cabinet.

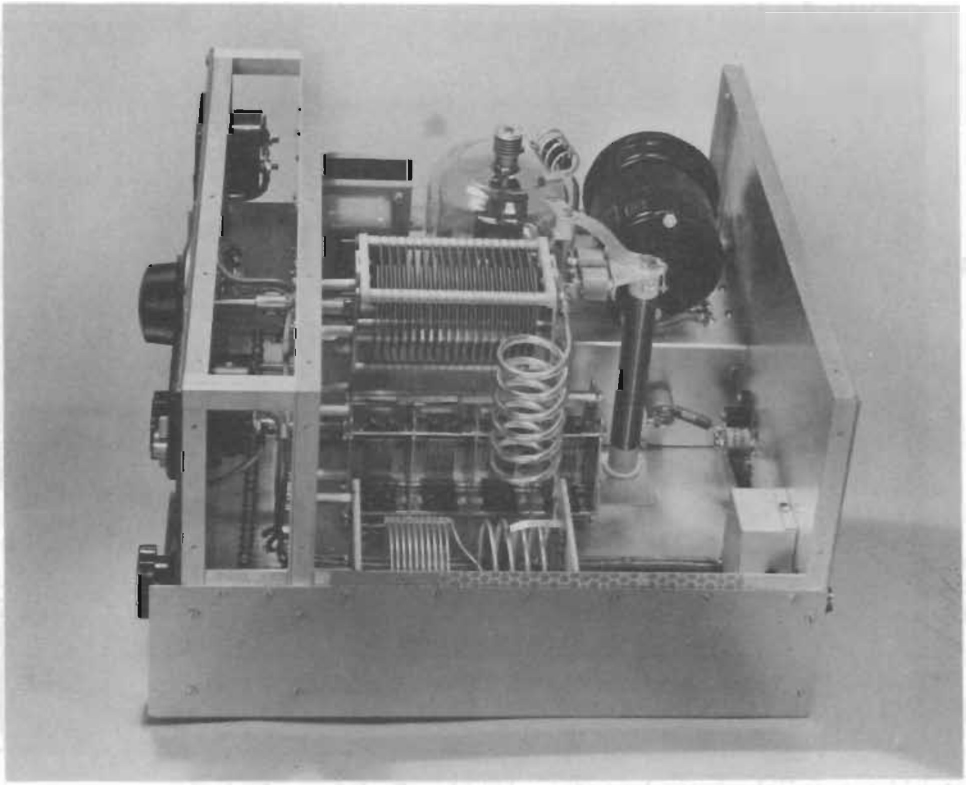


Figure 3. Plate tuning capacitor and loading capacitor are supported from sub-panel by three me
pillars, each 1-3/8" long. Loading capacitor sections paralleled by copper strap, and angle plate on r
of capacitor mounts plate choke. Ten meter inductor is placed in vertical position, in foreground.

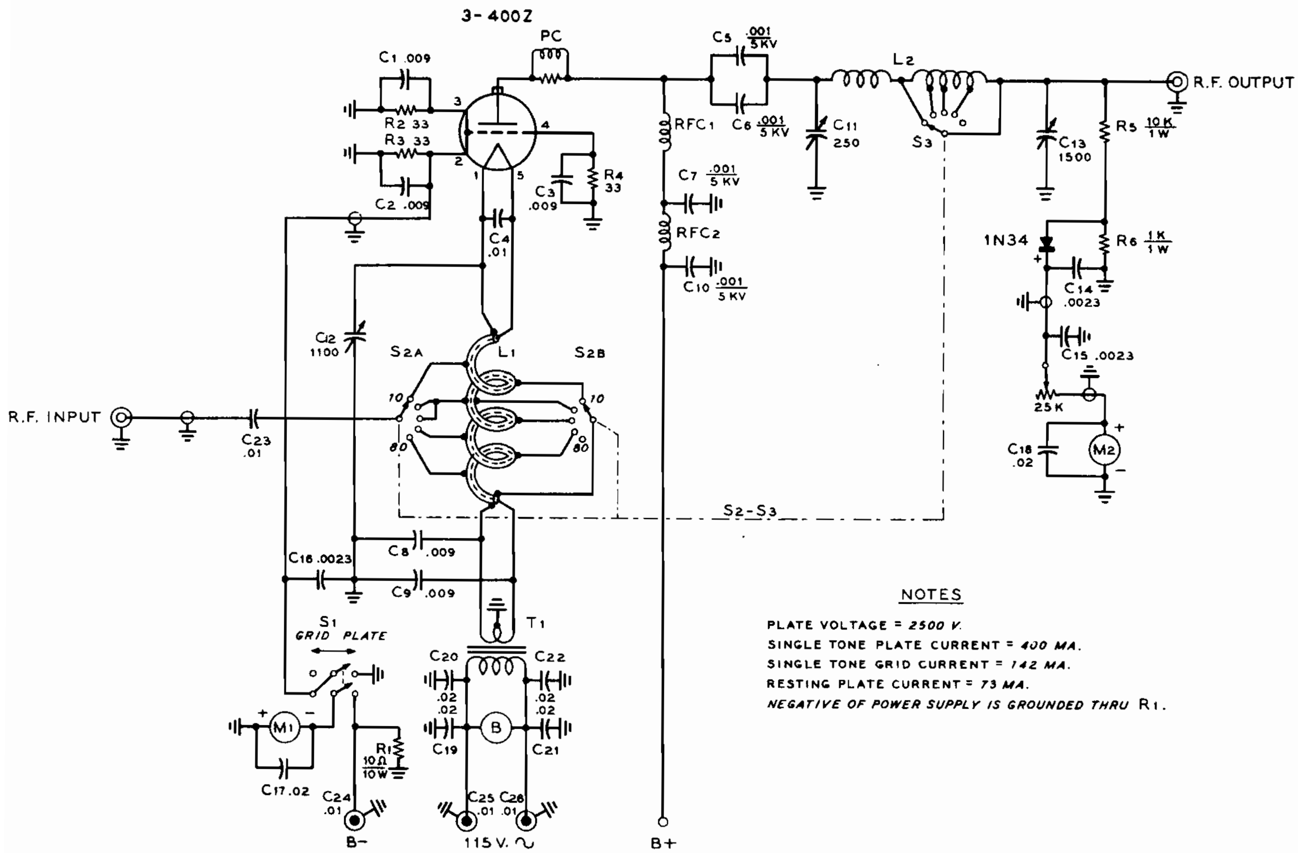


Figure 2

C1-C3,C8,C9 -- Each: Two 4700 pF, 3 kV ceramic disc capacitors in parallel (10 req.)

C4 -- .01 µf, 1 kV ceramic disc capacitor

C5,C6,C7,C10 -- .001 µf, 5kV. Centralab type 858S-1000

C11 -- 250 pF, 3kV. E.F. Johnson 250E30 (154-9)

C13 -- 1500 pF. Four section b.c.-type capacitor. J.W. Miller #2104

C12 -- 1100 pF. Three section b.c.-type capacitor. J.W. Miller #2113

C14,C16 -- 2300 pF "feedthrough" capacitor. Centralab FT-2300

C17,C22 -- .02 µf, 600 V ceramic disc capacitor

C23 -- .01 µf, 1.2 kV mica capacitor. Aerovox #1446

C24,C26 -- 0.1 µf, 600 V. Sprague "Hypass" capacitor. Type 70P8

M1 -- 0-500 d c milliammeter, 2" diam.

M2 -- 0-1 d c milliammeter, 2" diam.

PC -- Three 120 ohm, 2 watt composition resistors in parallel. Place inside of coil made of 4 turns #14, 5/8-inch diam., 1" long

R1 -- 10 ohms, 10 watt wire wound resistor

R2,R4 -- 33 ohm, 2 watt composition resistor

R5 -- 10,000 ohm, 1 watt

R6 -- 1,000 ohm, 1 watt

R7 -- 25,000 ohm, 1 watt potentiometer

S1 -- 2 pole 5 position rotary switch

S2A,B -- Make up of two Centralab "RR" decks (2 pole, 5 position, 30 degree index) and Centralab P-122 Index Assembly

S3 -- Part of L2 assembly

T1 -- 5 V @ 13 A Triad F-9A

Blower -- Fasco Industries #50745-IN, 115 V

Coil data:

L1 -- Wound of 3/16-inch copper tubing. 10½ turns, 1-5/8-inch diam. Spread turns apart 1/16-inch. Top two turns (tube end) are spread apart 5/16-inch. S2A taps coil (from "cold" end) as follows: 10 meters, 9½ turns; 15 meters, 9 turns; 20 meters, 9 turns; 40 meters, 7½ turns; 80 meters, 6 turns. S2B taps coil (from "cold" end) as follows: 10 meters, 9 turns: 15 meters, 8 turns; 20 meters, 7 turns; 40 meters, 4 turns; 80 meters, tap at "cold" end.

L2 -- Barker & Williamson #851, modified as follows:

A- Remove turns from main coil from "cold" end until 11½ turns remain.

B- Tap as follows: 80 meters, entire coil; 40 meters, tap 7½ turns from "cold" end; 20 meters, tap at junction of #12 wire and 1/8" wire coil; 15 meters, tap 1-3/4 turns toward "hot" end of coil from junction.

C- 10 meter coil consists of 6 turns, 3/16-inch copper tubing, 1-3/8-inch inside diam., 4" long.

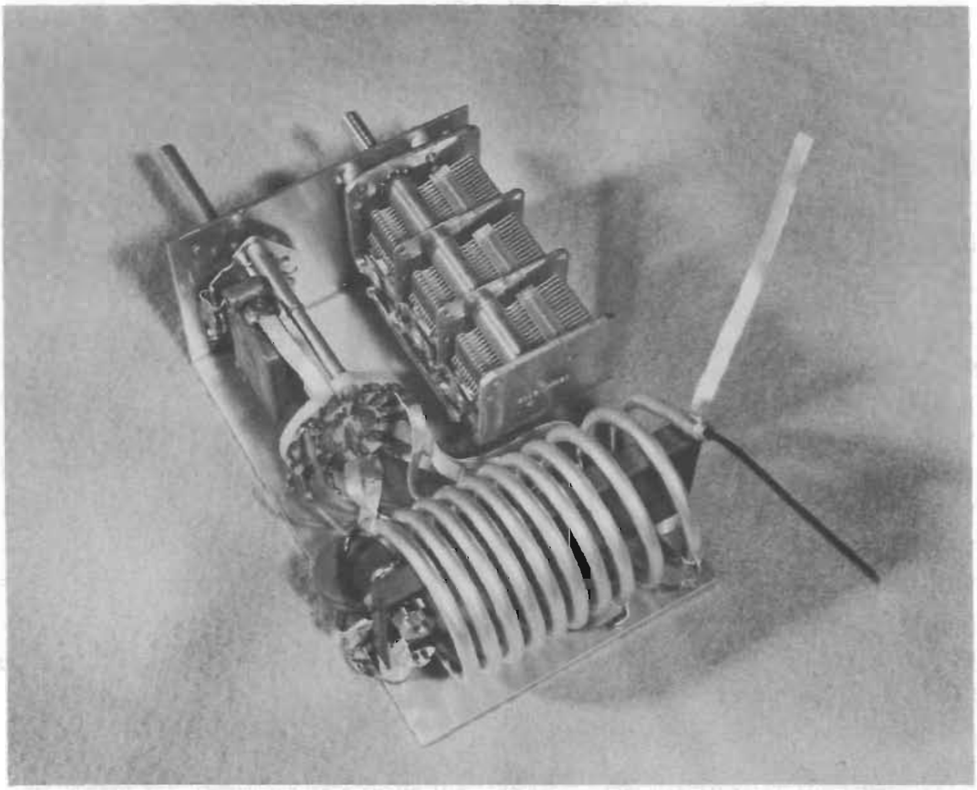


Figure 4. High-C cathode tank employs bifilar-wound coil to carry filament voltage to 3-400Z. Switch leads are made of copper strap, and coil is supported on phenolic strip, mounted to the support plate by two ½-inch ceramic insulators.

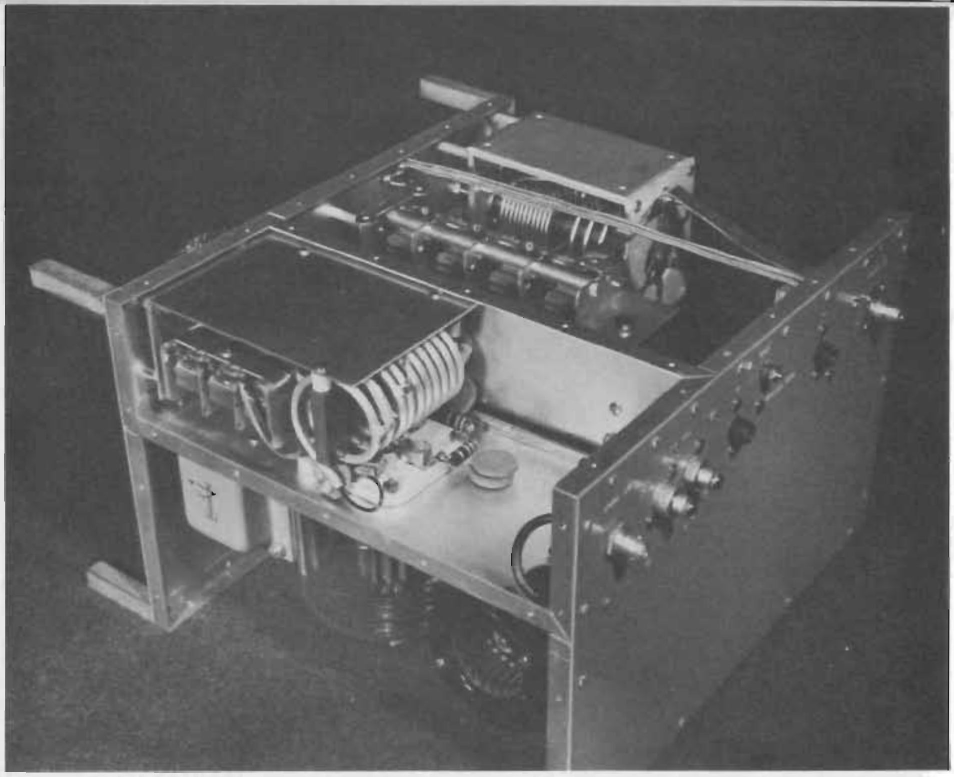


Figure 5. Under-view of chassis showing input compartment with tuned cathode circuit in place. Filament pins of socket face side of chassis. Each grid pin is grounded by R-C termination. All power leads run in shielded braid.

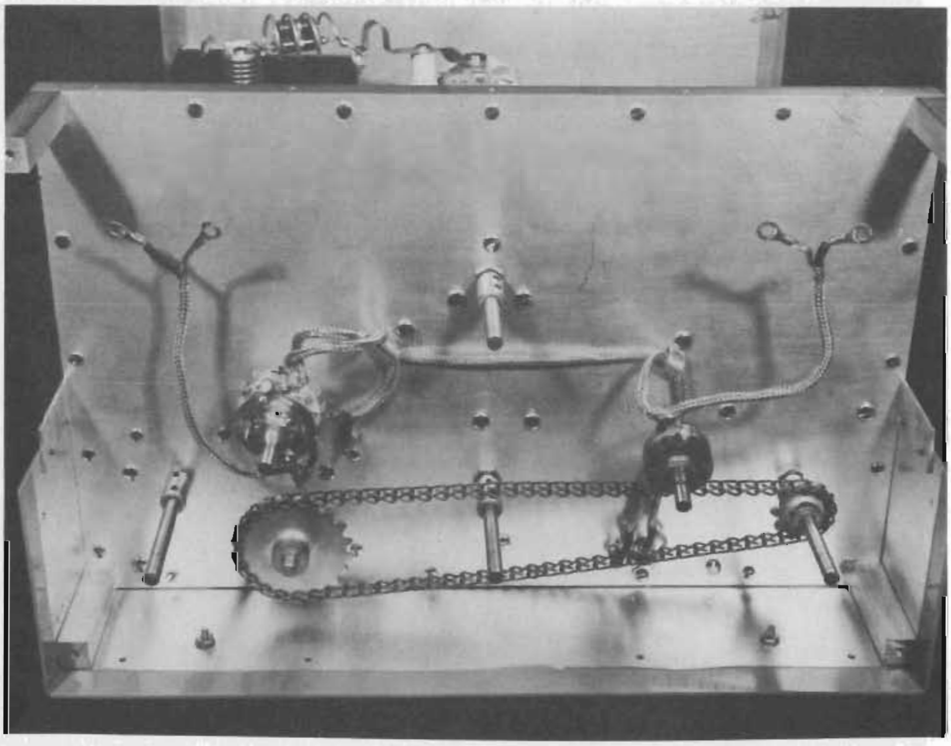


Figure 6. Meter wiring and chain drive behind panel. Cathode tank (left) and plate turret (right) are ganged, using American Stock Gear #C-10 sprocket (10 teeth, 1.125" diam.) left, and American Stock Gear #C-20 sprocket (20 teeth, 2.030" diam.) right. "Ladder" chain drive is American Stock Gear #18/42.

Figure 7

General characteristics, 3-400Z tube

Filament: 5.0 volts at 14.5 amperes

Interelectrode capacitances:

- Grid-filament: 7.4 pF
- Grid-plate: 4.1 pF
- Plate-filament: 0.07 pF

Typical operation, 2500 volts plate potential, grounded grid circuitry.

Zero signal plate current:	75 mA *
Single tone d c plate current:	400 mA
Single tone d c grid current:	140 mA
Two tone d c plate current:	275 mA
Two tone d c grid current:	82 mA
P E P input:	1000 watts
P E P output	560 watts **
Resonant load impedance:	3450 ohms
Intermodulation products:	-35 db or more below P E P signal level
Driving power (approx.):	32 watts, P E P

* Approximate value

** Includes circuit losses

The new EIMAC 3-400Z is a zero bias triode specifically designed for grounded grid r f linear amplifier service. It is rated to one kilowatt PEP input for linear or CW service, and is also rated for modulator and class C operation. Only 4½-inches high, this little "powerhouse" delivers a big signal in a small package! The big brother of this tube is the 3-1000Z, rated to 2-kilowatt PEP input in sideband or CW service.

The Grounded Grid Linear Amplifier

In the "good old days" of ham radio, linear amplifiers were used by a few amateur phone stations as a (relatively) inexpensive way of obtaining high power. Class B modulators were as yet unknown, and the cost of glassware necessary to generate two or three hundred watts of Class A audio power was exhorbitantly expensive for all but the "well-heeled" hams.

Loy Barton's classic QST article (Cira 1929) describing the inexpensive Class B modulator system sounded the death-knell of the linear amplifier for amateur service until the advent of single sideband, which recently blew the dust from this ancient mode of operation and modernized it to fit today's operating conditions.

What is a Linear Amplifier?

To the hi-fi enthusiast, the linear amplifier is a high fidelity music amplifier. To the SSB enthusiast, the linear amplifier package, when placed on the end of a sideband exciter, will make the exciter sound bigger, louder, and more commanding to other amateurs. The fact of the matter is that the SSB linear amplifier is a high fidelity amplifier in the true sense of the word. The "hi-fi" man thinks in terms of fidelity, and the "sidebander" thinks in terms of linearity. They are both talking the same language.

It is interesting to note that a good "hi-fi" audio amplifier can be theoretically converted to a low distortion linear amplifier for sideband service by replacing the audio circuits with suitable r.f. tank circuits. Indeed, for r.f. work, push-pull circuitry is not even required as it is in audio service, because the flywheel action of the r.f. tank circuits will supply the missing half-cycle. Finally, the operating parameters for a particular tube (plate, screen and grid voltage; driving voltage; and load resistance) are easily calculated for audio work, and apply equally well for r.f. service. For example, the 811A tube is rated for Class B audio service, as a high-μ triode (figure 1A). Compare these ratings with the Class B r.f. linear data listed in figure 1B.

Why Linearity?

For sideband service, the r.f. power amplifier must be truly linear. It must be capable of high fidelity reproduction, That is, the signal existing in the plate circuit must be an exact replica of the exciting signal impressed upon the input circuit. This statement is a good definition of a linear amplifier. It implies that the power gain of the stage must be constant regardless of the signal level. Any deviation from this happy state creates distortion products that appear in the signal passband and adjacent to it.

Unfortunately, many amateurs judge the excellence of the sideband signal by the quality of the signal; that is the pleasing aspect of the voice being transmitted. Many times one hears the report "Your quality is excellent, Old Man. You have a fine signal", yet the listener observes that the recipient of this flattering observation has a signal as broad as a barn door, complete with "whiskers" and "splatter" that obliterate half the phone band! Obviously, the criteria of quality of a sideband system is what you don't hear, not what you do! The place to examine a sideband signal for linearity and quality is in an adjacent channel, not in the frequency band of the signal itself!

How Good is "Good Quality"?

The excellence of a sideband signal is an ethereal concept and usually is judged by the amount of (or lack of) sideband splatter in nearby channels. Theoretically, a sideband signal should be about three or four kilocycles wide: just as wide as the voice passband of the equipment. However, the poor sideband operator's "tin ear" has been brutally deafened by so many "rotten signals" that he often accepts any SSB signal as "good quality" as long as it does not blanket the dial of his receiver!

Over the years a nice, easy, vague figure of "30 decibels down" for distortion products has become a password for good quality, low distortion, amateur sideband equipment. Since the measurement technique is usually undefined, and practically no amateurs have equipment sufficiently sophisticated to measure the intermodulation products of a sideband signal, this figure has become a byword for most commercial and home-made amateur equipment on the air. Valid or not, this magic number seems to be the socially correct distortion figure applicable in all cases to all equipment!

Distortion - What it Means

If the output signal of a linear amplifier stage is a replica of the exciting signal, there will be no distortion products. However, as vacuum tubes and circuit components are not perfect, this situation is as yet unreachable. As shown in figure 2, the transfer characteristic of a typical tube is approximately linear. This tube suffers no pain when amplifying a single signal (such as a carrier or a single tone), but has the interesting property of mixing when a multiple signal source is applied to it. This means that a voice signal (made up of a multiplicity of tones) will become distorted and blurred by the inherent mixing action of a so-called linear or "high fidelity" amplifier. A standard test to determine the degree of mixing for a given circuit or tube is the two-tone test, wherein two radio frequencies of equal amplitude are applied to the amplifier, and the output signal is examined for spurious products (figure 3). These products, or "garbage" fall in the fundamental signal region and atop the various harmonics. The tuned circuits of the amplifier filter out the spurious signals falling in the harmonic regions, which are termed even-order products. The odd-order products, unfortunately, fall close to the fundamental output frequency of the amplifier, and cannot be removed by simple tuned circuits. These are the spurious frequencies that cause a poorly designed or incorrectly adjusted linear amplifier to cover the dial with splatter. Shown in the illustration are two frequencies that make up a typical two-tone test signal. In this example, they are 2000 kc. and 2002 kc. Now, if the amplifier is perfect, these two signals will be the only ones appearing in the output circuit. An imperfect (but practical) amplifier, however, will have various combinations of sums and differences of the signals and their harmonics which are generated by the nonlinearity transfer characteristic of the tube. All of these unwanted products fall within the passband of the tuned circuits of the amplifier and are radiated, along with the two test tones.

If the odd-order products are sufficiently attenuated, they will be of minor importance and can be ignored. The sixty-four dollar question is: of what magnitude can these spurious products be without becoming annoying? How much "garbage" can be permitted before the signal becomes intolerable to the operator trying to maintain a QSO in an adjacent channel?

The answer to these questions depends upon the type of information being transmitted, and the degree of interference that can be tolerated in the adjacent channel. Certain forms of information (not voice) require an extremely low value of spurious products within and adjacent to the pass band; otherwise, the information will be seriously degraded. Odd-order products greater than .001% of the wanted signal may be damaging to the intelligence. Translated into terms of decibels, this means the unwanted odd-order products must be -50 decibels below the wanted signal! This takes some doing, and is orders of magnitude more strict than the level of intermodulation products than cat be tolerated in amateur voice communications.

In actual practice, it would seem that odd-order products less than 0.1% of the peak signal level are sufficiently attenuated so as to cause a tolerable level of adjacent channel QRM in everyday amateur communications. This indicates a distortion product magnitude of -30 decibels below the peak output power level of the transmitter. Such a state of affairs can be obtained by modern techniques without too much trouble provided attenuation is given to circuit design and operating parameters of the equipment. Of course, if distortion levels exceeding this arbitrary level can be reached, so much the better! Unfortunately, some equipments presently operating in the amateur bands and masquerading as "linear" amplifiers exhibit distortion levels of -20 decibels or less below peak power output! Use of equipment of this dubious quality quickly reduces the popularity of the operator to zero, and will probably lead to a brick through the shack window if continued!

The Grounded Grid Linear Amplifier

For amateur service the grounded grid circuit professes to be the answer to many of the ills besetting the linear amplifier. It requires a level of drive that is compatible with the great majority of sideband exciters (70 to 100 watts). With proper choice of tubes, it may be operated in a zero bias condition, eliminating the need of expensive and heavy grid (and screen) power supplies. Neutralization is not usually required. In addition, claims are made that the "inherent" feedback of the grounded grid amplifier improves the stage linearity and drops the magnitude of the distortion products. This all sounds too good to be true, and an examination of the grounded grid amplifier may be in order to see if it is "the answer to the sidebander's prayers".

The "classic" grounded grid amplifier is shown in figure 4A. The control grid is at r.f. ground potential, and the driving signal is applied to the cathode via a tuned circuit. The control grid serves as a shield between the cathode and the plate, making neutralization unnecessary at medium and high frequencies.

The input and output circuits of the grounded grid amplifier may be considered to be in series and a certain portion of the input power appears in the output circuit. This feedthrough power acts to somewhat stabilize the load the amplifier presents to the exciter, and also provides the user with some "free" output power he would not otherwise obtain from a more conventional circuit. The driver stage for the grounded grid amplifier must be capable of supplying the normal level of excitation power required by the amplifier plus the feedthrough power. Stage power gains of 5 to 25 can be achieved in a grounded grid amplifier.

Measurements made on various tubes in the Power Grid Tube Laboratory of EIMAC showed that an improvement of 5 to 10 decibels in odd-order distortion products may be gained by operating various tubes in the grounded grid configuration of figure 4A, in contrast to the same tubes operating in the grid driven mode. The improvement in distortion figure varied from tube type to tube type, but all tubes tested showed some order of improvement when cathode driven. (See Amateur Service Bulletin #1 for information regarding cathode driven service).

The tuned cathode circuit consisted of a bifilar coil which carried the filament current and a large capacity variable capacitor. The circuit was high-C, with the excitation tap placed to provide a low value of SWR on the coaxial cable to the exciter.

The Untuned Cathode Circuit

After sufficient measurements had been made with the circuit of figure 4A, the apparatus was modified to simulate the popular untuned cathode input circuit of figure 4B. It was immediately noted that the tubes tested in the previous circuit provided noticeably poorer results when used with an untuned cathode circuit. Power output dropped by 5% or so, greater grid driving power was required, and linearity suffered to a degree. Specifically, the third-order products rose approximately 3 to 4 decibels over the values produced by the circuit of figure 4A, and the fifth-order products rose 5 to 6 decibels over those figures recorded with the tuned cathode circuit. The higher order distortion products also rose accordingly. Observing the input waveform at the cathode of the grounded grid amplifier showed a pronounced distortion of the r.f. waveform caused by the loading effect over one-half cycle caused by a single-ended class B amplifier. Plate and grid currents drawn over the portion of the cycle loaded the input circuit. The exciter thus "sees" a very low load impedance over a portion of the cycle, and an extremely high impedance over the remaining part of the cycle. Unless the output regulation of the exciter is very good, the portion of the wave on the loaded part of the cycle will be seriously degraded, as shown in figure 5. Under normal circumstances, degradation of the input waveform may reach a more serious degree, as the exciter used for these tests was operating Class A and was well swamped to improve regulation. Obviously, the circuit Q of the exciter output tank at the end of a random length of interconnecting coaxial line is not sufficient to prevent this form of wave distortion. In addition to degrading the intermodulation figure, this waveform distortion also can cause mysterious TVI troubles as a result of the high harmonic content of the wave.

It was also noted that the degree of intermodulation distortion could be changed by varying the length of coaxial line between the driver and the linear amplifier! This pointed out a problem that had gone unnoticed until now (figure 6). When the untuned cathode circuit is used, the r.f. current path from the plate of the amplifier tube back to the cathode must follow the path of the dotted line in the drawing. Since the cathode choke offers a high impedance to this path, the alternative circuit is via the outer shield of the coaxial line, through the output capacitor of the exciter plate tank circuit, back to the linear amplifier via the center conductor of the coaxial line, and through the coupling capacitor to the cathode of the amplifier tube!

This alternative path presents several severe hazards. First of all, it is random, and varies with the length of interconnecting coaxial line. Second, the outer shield of the coaxial line is "hot" to the plate circuit ground return, and all sorts of weird intercoupling between the amplifier and the exciter may result. Third, the plate return current passes through the output pi-network capacitor of the exciter. There is a real danger that this capacitor may not be large enough (in a physical sense) to carry the current, and may be damaged when subjected to this form of abuse.

It is possible to employ either a high-C tuned circuit of the form shown in figure 7A, or untuned filament chokes in conjunction with a simple pi-network may be employed as shown in figure 7B. Either arrangement will supply the necessary "flywheel" effect to retain good r.f. waveform at the cathode of the linear stage, and both will provide a short, direct ground return path for the plate r.f. circuit.

Adjustment of the Cathode Circuit

The cathode circuit is resonated to the operating frequency by means of the variable capacitor. Resonance is indicated by maximum grid current of the amplifier. A low value of SWR on the coaxial line to the exciter is established by adjusting the tap on the tuned circuit, or by varying the "input" capacitor of the pi-network. SWR correction should be made with the amplifier running at maximum input. When the tap is correctly set, maximum grid current and minimum SWR will coincide at one setting of the capacitor. No cutting and trimming of the coaxial line is required, and the exciter will be properly loaded. This is a boon, indeed, to the owners of SSB exciters that have a fixed pi-network output circuit.

Summary

The use of the tuned cathode circuit in a grounded grid linear amplifier stage improves linearity, increases the power output, makes the stage easier to drive, and reduces the burden placed on the sideband exciter. The advantages of this circuit are well worth the added cost of parts and the extra controls. It is, of course, possible to dispense with the tuned cathode circuit, provided the user understands the handicaps he must assume by omission of this important circuit element.

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NOTE: The majority of the above material appeared in QST magazine, August 1961 issue, pages 16-21. Thanks are given to the editors of QST for permission to reprint this material.

Figure 1

811A OPERATING DATA			
(A) Class B Audio Service (Two Tubes)		(B) Class B r.f. Service (One Tube)	
		Grid Driven	Grounded Grid
Plate Voltage	1250	1250	1250
Grid Bias	0	0	0
Peak Grid Voltage	175	88	88
Zero Signal Plate Current (ma.)	54	27	27
Max. Signal Plate Current (ma.)	350	175	175
Load Resistance (ohms)	9200	4600	4600
Max. Signal Grid Current (ma.) (1)	26	13	13
Power Output (watts)	310 (2)	155 (2)	141 (3)

(1) Varies from tube to tube (2) Computed power output (3) Measured output (including circuit losses)

The operating parameters of a class B amplifier stage remain the same regardless of whether the tube functions in audio or r.f. service. Grounded grid operation is similar, except that the exciter must supply additional feed-through power required by this configuration. Since class B audio service requires two tubes, all currents and plate load resistance must be halved for single tube r.f. service. Class B audio data is readily available for most tubes and can be used for r.f. service, as shown above.

Figure 2

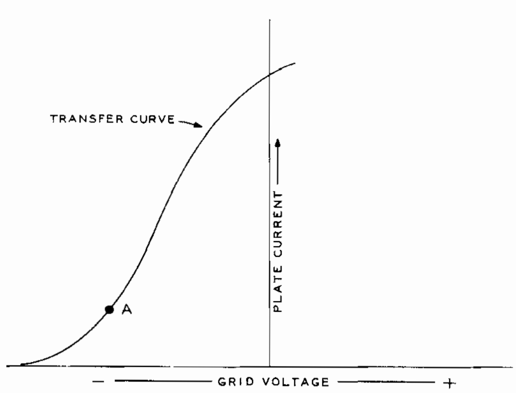
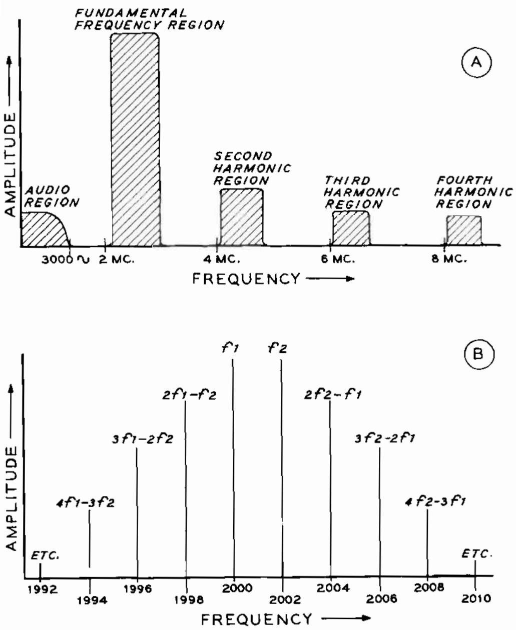


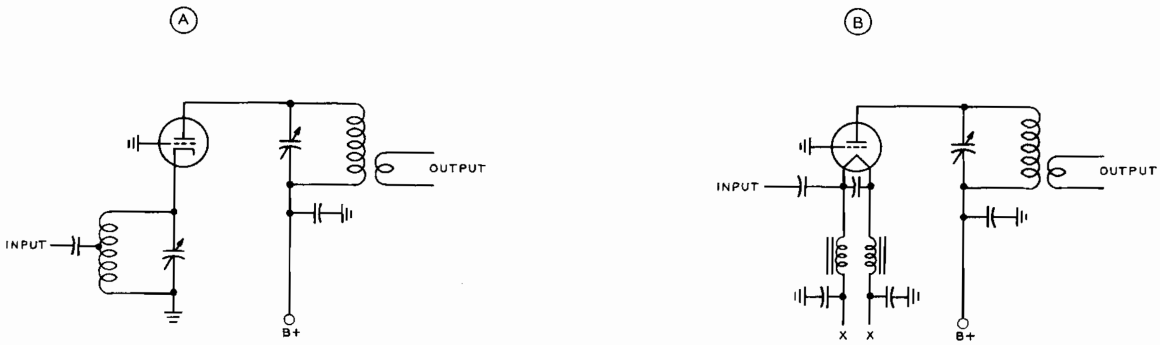
Plate current vs. grid voltage curve (dynamic characteristic) of a vacuum tube. This curve is linear in the center portion and exhibits deviations at either extremity. The shape of the curve and the choice of the zero-signal operating point (A) will determine the distortion produced by the tube. Mixing action caused by nonlinearity produces distortion products which cannot be eliminated by the tuned circuits of the amplifier.

Figure 3



Intermodulation (mixing) distortion caused by nonlinearity is illustrated by two-tone test signal (f_1 and f_2). Even-order products (A) are substantially eliminated by the tuned circuits of the amplifier, but odd-order products (B) fall within the passband of the tuned circuits and are not removed. (B) shows the mixture of spurious signals that make up distortion products falling within the fundamental range.

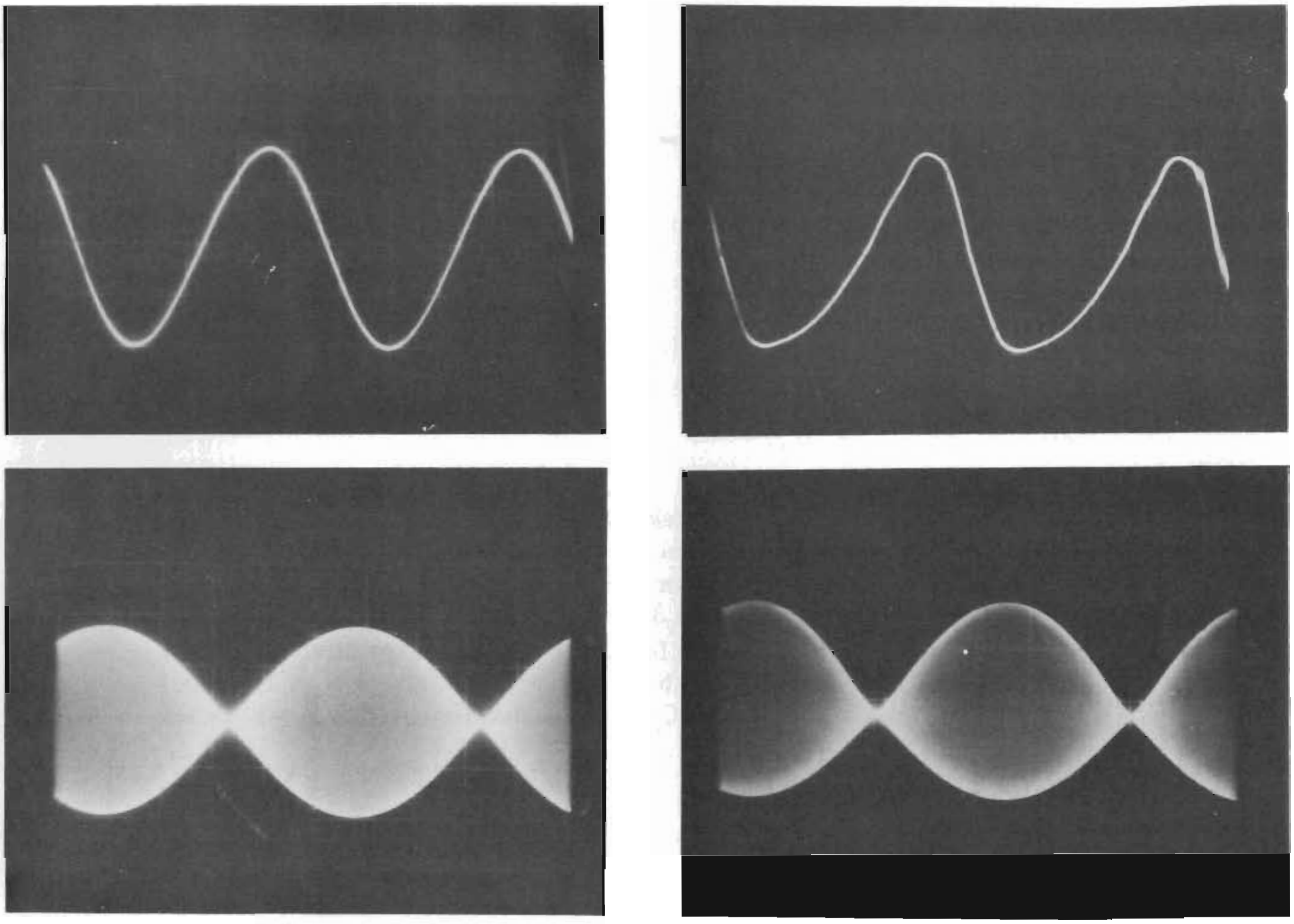
Figure 4



4(A) The grounded-grid amplifier has the input circuit between cathode and ground. The control grid acts as a screen between the plate and the cathode, making neutralization unnecessary in most circuits. The input and output circuits are in series and a portion of the input power appears in the output circuit. The driver stage for the grounded-grid amplifier must be capable of supplying normal excitation power plus the required feed-through power. High-C cathode tank preserves waveform of input signal and prevents distortion.

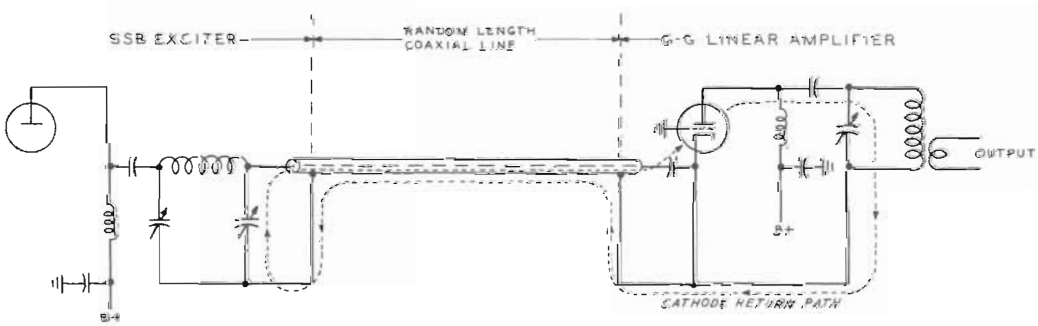
4(B) Popular amateur-style grounded-grid amplifier uses untuned filament choke in place of cathode tuned circuit. Laboratory tests showed that this simplified configuration produced higher intermodulation distortion products and had less power output than the "classic" circuit of Fig. 4(A), regardless of the type of tube used. In addition, the untuned input circuit proved hard to match and drive with pi-network sideband exciter.

Figure 5

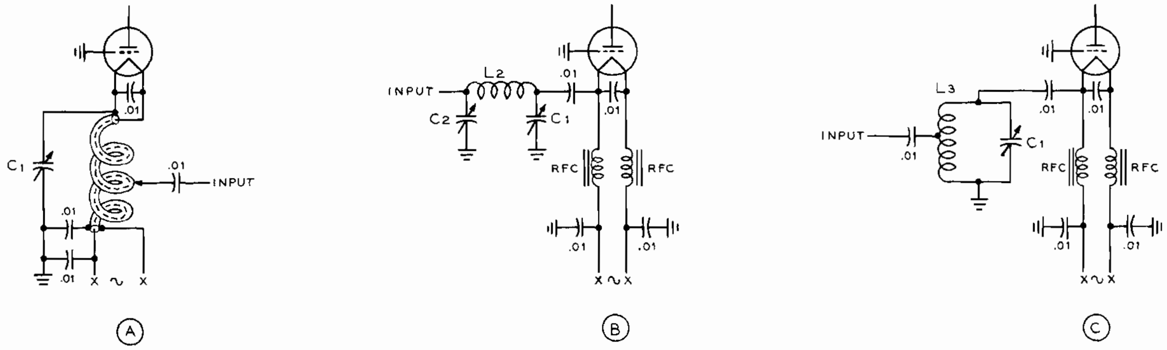


Waveform distortion caused by half-cycle loading at cathode of grounded-grid amplifier can be observed in oscilloscope studies. Lower left: Two-tone test signal when tuned cathode circuit is used. Upper left: 3.5-Mc. waveform (single tone) from sideband exciter as seen at cathode tank. Lower right: Two-tone test signal when untuned cathode circuit is used. Upper right: 3.5-Mc. waveform (single tone) from sideband exciter, showing severe distortion of waveform when untuned cathode circuit is used.

Figure 6



Untuned filament circuit of grounded grid amplifier offers a high impedance to the r.f. current path from plate to cathode of the amplifier tube. The alternative path is via the interconnecting coaxial line and tank circuit of the exciter. Lack of tuned circuit at the cathode of the g-g stage permits waveform distortion of the driving signal resulting in a higher degree of intermodulation distortion and reduced power output.



Tuned cathode network for zero-bias tube may take the form of bifilar circuit (A), pi-network (B), or a shunt LC circuit (C). A Q of 5 is recommended for optimum results. However, as this leads to rather bulky circuits at the lower frequencies, the Q may be decreased to 2 or 3 without serious effects. Capacitor C_1 is a 3-gang broadcast-type unit. Coils L_1, L_2, and L_3 are adjusted to resonate to the operating frequency with C_1 set to about 13 $\mu\mu$f per meter of wavelength. Capacitor C_2 is approximately 1.5 times the value of C_1. The input tap on coils L_1 and L_3, or the capacitance of C_2 are adjusted for minimum s.w.r. on the coaxial line to the exciter.

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Grid Current Measurement in G-G Amplifiers

Measuring the grid current of a cathode driven amplifier can be a delicate and exasperating task as it is a ticklish job to "unground" the grid sufficiently to permit a metering circuit to be used, yet still hold the grid at r.f. ground potential. The inherent inductance of most bypass capacitors permits the grid circuit to "float" above ground at some high frequency, and as a result the amplifier exhibits instability and parasites. This problem can be avoided with the measuring circuit of figure 1A. The control grid is grounded through a 1-ohm composition resistor, bypassed by a .01 μfd disc capacitor. The voltage drop generated by the flow of grid current across the resistor can easily be measured by a milli-voltmeter which is calibrated to read in terms of grid current. Individual grid current for each of a parallel pair of tubes may be measured by the circuit of figure 1B.

The internal resistance of the 0-1 d.c. milliammeter plus the series resistor R1 determines the maximum current that can be measured. Suppose it is desired to read grid current in the order of 60 milliamperes. The meter should therefore read 0-100 milliamperes. This is very convenient, as the reading of the meter scale can easily be multiplied by 100 to obtain the actual value of current. Now, when 100 milliamperes flow through 1-ohm, there exists a potential of 0.1 volt across the resistor. The meter should therefore read 0.1 volt full scale to correspond to a grid current of 100 ma. Assume the meter is Triplett #221-T, which has an internal resistance of 55 ohms. The voltage drop across the meter itself is 0.055 volts when one milliampere flows through it. To convert the milliammeter to a voltmeter reading 0.1 volt full scale a series multiplier must be added. A voltage drop of 0.1 volt exists across a 100 ohm resistor when one milliampere of current flows through it. The difference between 100 ohms and 55 ohms, or 45 ohms, must therefore be added in series with the meter to convert it to read 0.1 volt, full scale. On the other hand, placing the meter itself across the 1-ohm resistor without the series multiplier will result in a full scale reading corresponding to 55 milliamperes. Thus, if maximum grid current is below this latter figure, no series resistor is required for the meter. Conversely, high values of grid current produce greater voltage drop across the 1-ohm resistor, and larger values of series multiplier resistance are needed.

FIGURE 1A

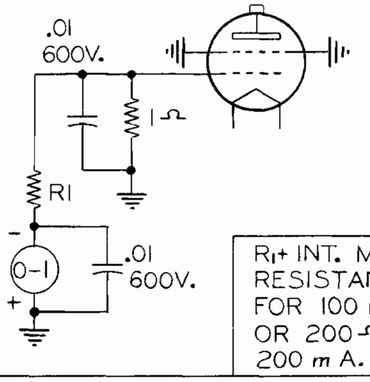
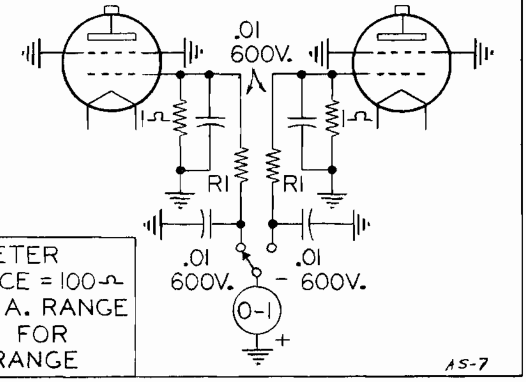


FIGURE 1B



R1+ INT. METER RESISTANCE = 100Ω FOR 100 mA. RANGE OR 200Ω FOR 200 mA. RANGE

TRANSMITTING TUBES - How To Use and Abuse Them

The story has been told at hamfests that the data sheet enclosed in a vacuum tube box was included with the intention that it should be thrown away with the box and packing material. Contrary to this little story, the data sheet has been placed therein with good reason: to inform the user of the tube of the capabilities and limitations of the tube. The data sheet is a summation of the functions of the tube and covers the **electrical and mechanical characteristics**, the **maximum ratings**, and the **typical operating conditions.**

The rugged individualist usually ignores the data sheet and runs his tube at a temperature just below that at which the plate will start to melt. This may be fun, and may even evoke "oh's" and "ah's" from the visiting hams, but it violates the old "watts-per-dollar" evaluation of the vacuum tube!

After all, let's face it: There is an economic point beyond which it is impractical on a "watts-per-dollar" basis to push a tube. In general, the harder the tube is pushed, the shorter will be its life. This is analogous to the story of the cowboy who wandered into a western saloon and saw an old, grizzled prospector at the bar, drinking whiskey neat from a bottle, and smoking a huge, black cigar taken from a box of stogies at his elbow. Striking up a conversation, the cowboy learned the old gent drank three quarts of booze a day, and smoked five boxes of cigars a day, too. Said he, "You are amazing! All this hard living at your age! You look like you must be ninety years old!"

"Ninety!" screeched the prospector slamming his drink on the bar and reaching for his gun, "I'm only twenty-two!"

So it is with the vacuum tube. Moderation is the secret to a happy, long tube life. The tube manufacturer sets maximum ratings on a basis of expected tube life. Each rating has been determined as the maximum value which will permit a reasonable life expectancy for the tube. The enemy of unlimited tube life is **heat**, but unfortunately heat is a natural consequence of making the tube work. A compromise of some kind must thus be made, and this is the purpose of the data sheet. Let's look into the compromise and see what establishes the various ratings given for transmitting tubes.

Plate Dissipation

The **plate dissipation** (rated in **watts**) of all radiation (air) cooled tubes is limited by the maximum safe temperature of the plate, and the effects of this temperature on parts of the tube other than the plate. In general, the plate of radiation-cooled tubes will withstand several times its maximum rated plate dissipation **for a short period of time**. Other parts of the tube, however, are affected greatly by excessive heat radiated by the plate. High levels of plate temperature cause the grid, filament, and glass envelope to become over heated, while the heat conducted away from the plate by the plate lead contributes to the heating of the plate seal.

These effects are not instantaneous, and short overloads do not usually overheat the adjoining tube structure to a damaging extent. However, the user has no way of telling to what degree he can safely exceed the plate dissipation, or over what length of time this abuse can take place. The maximum plate dissipation rating is intended to set a point at which **continuous** operation may be carried out without damage to any part of the tube, even though the other tube elements may at the same time be operating at their maximum ratings.

Regardless of other conditions, the maximum plate dissipation should not be exceeded in continuous operation.

Maximum Plate Voltage

Voltage limitations are set at a point above which the internal insulators of the tube may arc over, or above which the glass envelope will become damaged from dielectric losses. In addition, a plate voltage ceiling tends to set a limit to the r.f. **charging current** flowing in the plate and filament leads. The charging current is a function of the r.f. plate voltage, which in turn is a function of the d.c. plate voltage; this makes it possible to set a limit on r.f. charging current without the difficult task of determining the current directly.

Tube envelopes having grid and plate leads in close proximity are subject to a greater degree of glass stress than those having widely separated electrode terminations. In general, however, most glass tubes have maximum plate voltage ratings that fall in the r.f. charging current limit category.

Minimum Plate Voltage

Each tube has a particular plate voltage below which it is uneconomical to operate the tube. That is to say, the filament power consumed by the tube (and the initial cost of the tube) are so high that the cost of power developed by the tube is high in comparison to the same power generated by a cheaper tube. Of course, if the tube is purchased "surplus" at low cost, the economic picture changes so that the initial cost is of secondary importance. Even so, tube efficiency tends to drop when extremely low values of plate voltage are employed. In addition, multi-element tubes, such as the tetrode (the 813 or 4-250A, for example) have a definite **minimum** plate potential below which it is not wise to operate the tube. As the plate potential is lowered, the average screen current tends to rise and the screen dissipation increases accordingly. It can be possible to thereby damage a tube by excessive screen dissipation by operating it at a low plate potential.

Lowering the screen voltage to decrease the screen dissipation is but a make-shift cure, as the power gain and efficiency of the tetrode tube drops sharply as the screen voltage is lowered beyond the normal operating range.

Maximum Plate Current

Maximum plate current is based upon the available supply of electrons emitted by the filament of the tube. **Filament emission** is therefore the controlling factor determining maximum allowable plate current. The maximum plate current figure is intended to set a value which may be easily realized throughout the life of the tube. If operating conditions are chosen which require the maximum plate current limitation to be exceeded at the start of tube life, it may become increasingly difficult to maintain the desired plate current as the tube ages. To have ample filament reserve, it is important to make sure that filament voltage is "up to snuff" at all times.

Filament Voltage

Proper **filament voltage** and the allowable departures therefrom are usually specified in the tube data sheet. In general, quick-heating thoriated tungsten filaments used in the larger power tubes may be operated over a range of plus or minus 5-percent of the recommended voltage. Slower heating cathode-type filaments used in small power tubes usually have a filament operating range of plus or minus 10-percent of the recommended voltage. External anode tubes have a filament voltage range of plus or minus 5-percent. Some variation in power output must be expected as the filament voltage is varied in this range. Lower than normal filament voltage will impair the power output of the tube, and higher than normal voltage will cause critical parts of the tube to run at an excessive temperature, and may even cause damage to the grid structure in extreme cases. In passing, it should be noted that an inexpensive a.c. type meter of plus or minus five-percent accuracy can tell the operator little about filament voltage, when the voltage must be held to the same value of accuracy. Use a **good** filament voltmeter of known accuracy.

Element dissipation sets the grid and screen power limits. Excessive dissipation can result in electron emission from the element (termed **primary emission**), or can cause deformation or melting of the structure through overheating. In addition, the grid and screen structures can be overheated by excessive radiation from the plate.

A common type of screen damage results when the tube is operated with full screen voltage and low or nonexistent plate voltage. The screen then tends to act like the plate, and excessive screen current quickly boosts the screen dissipation to the point where the structure is permanently damaged. Thus, the tetrode should be protected against loss of plate voltage. Either the screen and plate voltage should be taken from a common supply, or some form of overload relay or safety device should be used that will break the screen voltage lead when the screen current exceeds a predetermined value. When such a device is used, a ground return should exist between the screen and the negative of the plate supply. Otherwise, when the screen circuit is broken, the screen will "float" above ground at some high potential and the screen bypass capacitor may fail due to excessive voltage (figure 1).

Bulb Temperature

The glass envelope and lead seals of the transmitting tube must be maintained below a temperature at which the glass will soften, or the seals "leak" air. Tubes and components tend to become smaller year by year, but nobody has yet been able to miniaturize the watt, and assemblies run hotter as they are reduced in size. Also the tube's glass envelope will act as a conductor when it is too hot. Adequate ventilation is very important if maximum tube life is to be achieved. Heat is the great enemy of the vacuum tube and pains should be taken to conduct the heat away from the tube as efficiently as possible.

The most popular tubes used in amateur service are air cooled. The smaller tubes (and the larger, old ones having long element support stems) may be cooled by **convection**, the heat rising from the envelope creating sufficient air movement to ensure that excessive element and seal temperatures are not reached. Compact, higher power tubes that have to dissipate large quantities of heat in a small area must have assistance in the form of air blown across the envelope, seals, and pins by an auxiliary fan or blower.

Short, squat tubes may require more cooling air than long, thin tubes as the lead seals of the "shorties" are nearer the elements and are thus exposed to higher temperatures.

For most tubes the flow of cooling air is upward, consistent with the normal flow of convection currents. Large transmitting tubes have an open base structure and a matching socket which permits cooling air to enter the base end of the tube. The grid circuit area under the chassis, therefore, may be pressurized and the air introduced into this chamber by means of an external blower. The plate circuit area may have a mesh cover which permits the air to vent out readily, yet which provides a degree of circuit shielding. No holes in the chassis should be provided for the air to pass from the lower to the upper compartment other than by passage through the socket and tube base (figure 2).

Do not sub-mount a tube with a metal base shell so that the chassis comes above the vent holes of the base. **Do not** mount above the chassis a tube with a metal base shell or the proper circulation of air will be impaired (figure 3).

In the case of the external anode-style tubes (4X150A, for example) complete air system sockets are available that permit air to be blown axially on the base of the tube, past the base to the envelope, and then over the plate cooler. Use of other than such a special socket is "bad medicine" for the external anode tube, as tube temperatures cannot be adequately controlled.

Use of a receiving-type loctal socket with external anode tubes is not recommended, as dangerously high stem temperatures will be generated from the heat of the filament unless the base structure is cooled by an air blast, and the solid construction of the loctal socket blocks the normal flow of air above the tube stem.

Construction Techniques for Forced-Air Cooled Tubes

In general, the under-chassis area should be made air-tight, and a suitable fan or blower used to pressurize the compartment. The intake air vent should have a large area to provide a minimum resistance to the flow of air. Air holes may be screened as a TVI-preventive measure, but such impediment reduces the passage of the air by a large degree.

As it is difficult to measure the air flow to a tube, and even more difficult to measure the envelope and seal temperatures of the tube, the following "rules of thumb" may be observed in order to achieve optimum cooling and longest tube life.

1. Use the maximum amount of forced air possible. It is wise to employ a blower delivering at least twice the recommended volume of air. Turn the air blast on at the same time the filament is turned on, and leave it on as long as the filament is lit.

2. Inexpensive squirrel-cage blowers often do not work properly when delivering air into a back pressure "load" created by the socket, tube and chimney. A large quantity of air escapes through the sides of the blower. Make sure the air **enters** the socket and **escapes** via the tube chimney, and does not "windmill" in the blower cavity.

   See that the rotatable cage of the blower makes a close fit with the housing; otherwise, air will spill out of the unit when it is operated under back-pressure. In general, low speed blowers will lose pressure badly when subjected to back-pressure. A blower speed of at least 3,100 r.p.m. is recommended. Most transmitting tube data sheets reveal the required air pressure (in cubic feet per minute) and the back-pressure (in inches of water) that must be developed as pressure drop across the socket. Most blower manufacturers provide data sheets which show the blower output (in cubic feet per minute) that the unit develops for various values of back pressure (in inches of water, or static pressure). You can be reasonably sure your tube is adequately cooled if you choose a blower that develops about **twice** the required number of cubic feet you need at a specified back pressure. Beware of "midget" and "surplus", unmarked blowers, or blowers with loose-fitting housings!

3. Make sure the cooling air reaches the socket and make sure the exhaust air leaves the vicinity of the tube. It does no good to pump cooling air to a tube and then have no path for the warmed air to excape.

4. A large amount of heat escapes from glass tubes by radiation of energy from the tube plate. Placing the tube near polished metal surfaces that reflect radiant energy back to the elements of the tube is a sure way to raise the internal temperature of the tube. It is a good idea to space the tube away from such surfaces by at least the diameter of the tube envelope.

No simple rules can be given to accommodate all tube installations in all possible equipments. Tubes can be damaged by lack of air, but never by too much air, unless the blast is strong enough to lift the tube out of the socket and smash it against your ceiling! Use the largest blower you can afford. A great deal can be learned about air flow by puffing cigarette smoke into the blower and observing the path it takes in quitting the amplifier.

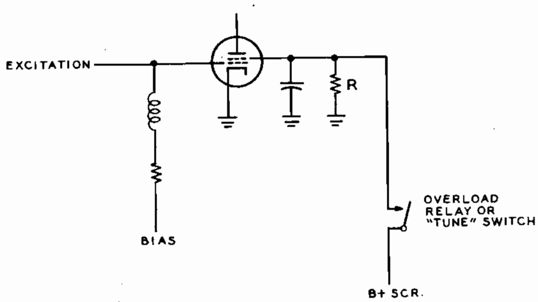
Connections to the Tube

Connections to the plate cap of transmitting tubes should be made with a section of flexible strap or braid to prevent any lateral strain from being placed on the tube electrode. Those tubes having a rod-type plate lead (such as the 304-TL) are prone to damage if the plate connector is forced on the rod until the connector touches the glass envelope. Under heat, the expansion rate of the glass, the plate rod, and the connector are all different, and it is possible for the connector to press against the envelope and cause a fracture of the glass at the point of contact.

There Is Hope!

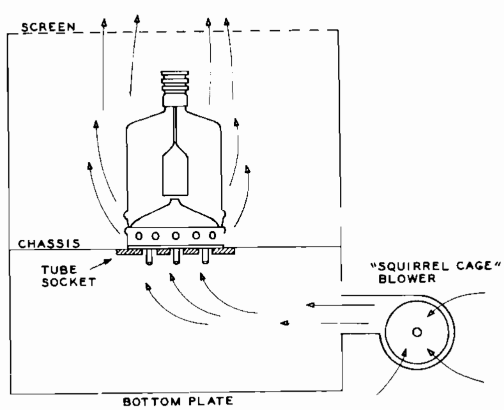
Don't let these warnings discourage you from building equipment and using modern high power transmitting tubes! It is merely that "forewarned" is "forearmed." By anticipating minor difficulties such as outlined in this article and eliminating the sources of trouble, the equipment in question can provide a long and happy life for the vacuum tube. This will make you (the owner and operator) happy, and - believe it or not! - make the vacuum tube manufacturer equally happy!

Figure 1



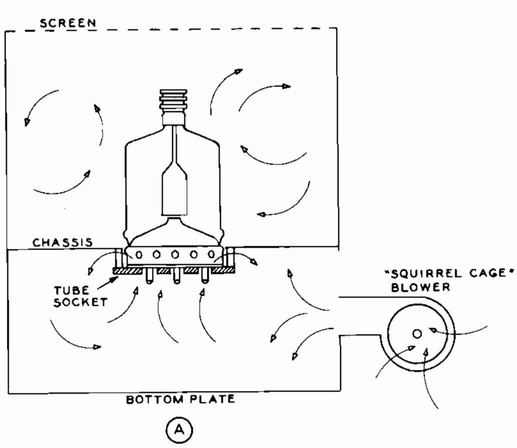
If the d.c. screen circuit is broken by an overload relay or a "tune-operate" switch, it is necessary to provide a d.c. return path from screen to ground. Otherwise, it is possible for the screen to assume the d.c. potential of the plate with the result that the screen bypass capacitor may be damaged. The screen resistor (R) should be chosen so as to "bleed" about 15 milliamperes per tube for the 4CX250B/4CX300A series tubes and approximately 70 milliamperes for the 4CX1000A.

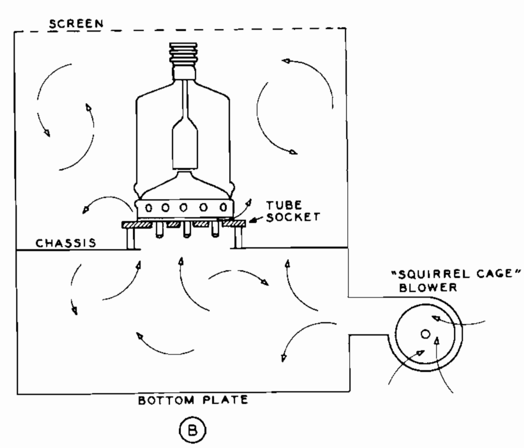
Figure 2



Forced air enters the grid compartment below the socket through a screened opening, passes through the socket cooling the base seals of the tube, sweeps upward cooling the the glass envelope. The plate circuit compartment has a screened cover which permits the air to vent out readily. This arrangement applies whether the tube is cooled by forced air or convection circulated air.

Figure 3





Do not sub-mount a tube with a metal base shell so that the chassis deck comes above the vent holes of the base shell of the tube. This arrangement impedes the proper flow of air and does not improve circuit isolation (A). Do not mount above the chassis a tube with a metal base shell, as the same conditions apply as before (B). In addition, no holes should be provided for the air to pass from the lower to upper compartment other than the passages through the socket and the tube base.

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A HIGH POWER LINEAR AMPLIFIER USING THE NEW EIMAC 3-1000Z

The Eimac 3-1000Z is a compact power triode designed for zero bias, class B r.f. and audio application. Grounded-grid operation is attractive as a power gain as high as twenty can be obtained in a cathode driven circuit. At a plate potential of 2500 volts, two kilowatts PEP input may be run, with intermodulation distortion products -35 decibels or more below maximum PEP level.

Shown in the drawing is the schematic of an all-band (3.5 - 29.7 Mc) amplifier designed around the 3-1000Z. A tuned cathode circuit is employed to achieve minimum distortion and ease of drive (1), and the popular bandswitching pi-network output circuit is used to match coaxial antenna feed systems.

The amplifier may be driven by any sideband exciter having a power output of approximately 65 watts. Drive is monitored by a grid current meter placed across a low impedance r.f. shunt located between grid and ground at the tube socket. The plate current meter is placed in the B-minus lead to the power supply. A simple diode voltmeter is used to indicate relative r.f. output, and is used for tuning purposes. During standby periods, the 3-1000Z is biased close to cutoff by the 50K resistor in the filament return circuit. The resistor is shorted out by the external VOX relay, grounding the center-tap of the filament transformer.

The new Eimac SK-510 Air-System Socket and SK-516 Chimney are recommended for use with the 3-1000Z. The older SK-500 socket may be used, provided care is taken to see that the contact pins move freely about, and do not place lateral strain on the tube pins. Flexible leads should be used with either socket to allow free pin movement.

At a plate potential of 2500 volts, peak plate meter current will be about 400 ma. under voice conditions, and grid current will approximate 125 ma. With carrier injection, the amplifier should be loaded to the single-tone operating conditions shown in the data sheet.

The amplifier may be constructed on a 14" x 17" x 4" aluminum chassis. The tuned input circuit and filament components are mounted below deck, and the plate circuit components atop the chassis. The chassis is pressurized by the addition of a bottom plate, and air forced into the chassis by a "squirrel cage" blower is exhausted through the air socket. The plate circuit components and tube are enclosed in a TVI-proof screen made of perforated aluminum sheet. Overall height of the amplifier is 14½".

(1) "The Grounded Grid Linear Amplifier", QST, August 1961, page 16.

3-1000Z LINEAR AMPLIFIER--PARTS LIST

C1--Three gang b.c. capacitor with sections in parallel. J. W. Miller #2113.

C2--500 μμfd., 10 KV. Jennings Radio Co. #UCSL-500 variable vacuum capacitor.

C3--1500 μμfd., 0.03" spacing. Barker & Williamson #51241.

L1--(See drawing) Mounted beneath chassis in close proximity to tube socket.

L2--Filament choke. Barker & Williamson FC-30. Windings connected in parallel.

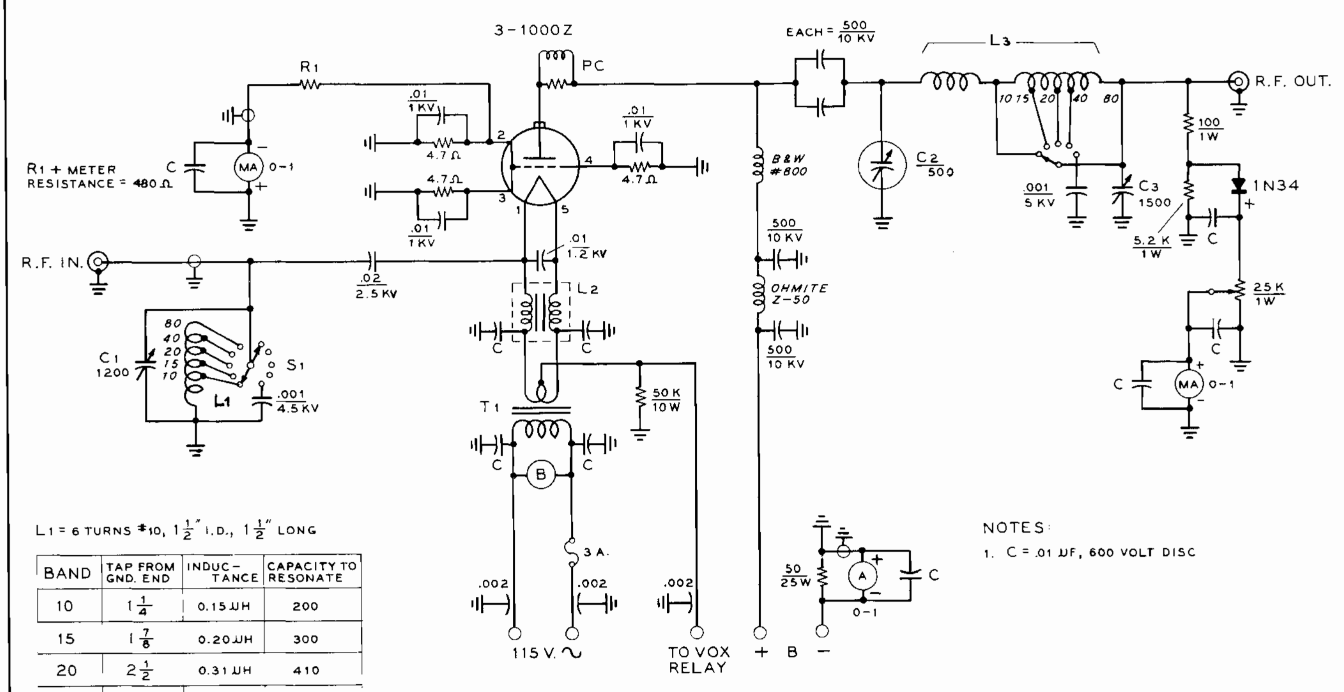
L3--Barker & Williamson #852 all-band coil assembly.

PC--Three 150 ohm, 2 watt composition resistors in parallel, shunted by three turns #12, 3/4" long.

T1--7.5 volts at 22 amperes. Stancor P-6457.

Blower: 20 cubic feet per minute, or greater. Dayton #1C-180, or Ripley #81.

Grid Meter: 0-1 d.c. milliammeter (55 ohms internal resistance), with multiplier. Full scale reading is 300 ma.



L1 = 6 TURNS #10, 1 1/2" I.D., 1 1/2" LONG

BAND	TAP FROM GND. END	INDUC-TANCE	CAPACITY TO RESONATE
10	1 1/4	0.15 μH	200
15	1 7/8	0.20 μH	300
20	2 1/2	0.31 μH	410
40	4 1/8	0.62 μH	825
80	6	1.25 μH	1650

The 4-1000A in Grounded Grid

Fig. 1—K9LKA's kilowatt 4-1000A grounded-grid amplifier. Meters across the top of the panel are, from left to right, for plate voltage, relative r.f. output, plate current and grid current. The band-switch control is in the center, flanked by the plate tuning control and capacitor switch S_2 on the left, and the output loading control on the right. Along the bottom are the filament switch, panel lamp and fuse; r.f.-indicator sensitivity control, and the input tuning control.

Most high-power triodes available at surplus prices do not have a sufficiently high amplification factor to permit zero-bias operation. Tetrodes may be converted to high-μ triodes by connecting the screen to the control grid. However, in the case of most tetrodes, this connection results in excessive control-grid dissipation at the driving-power level required to obtain normal rated output. The 4-1000A is one of the few exceptions to this rule[1] and is also one that is available in usable condition at relatively low cost from a number of sources. The triode connection results in considerable circuit simplification, especially in grounded-grid operation, since regulated bias and screen supplies are eliminated and neutralization is not required.

Zero-Bias Triode Operation in a 1-Kw. Linear

By LARRY KLEBER,* K9LKA

Many construction articles describe radio gear that is almost impossible to duplicate with facilities available to the ordinary ham because of unusual mechanical requirements. Complicated gearing, chain drives or special metal shapes that require power tools found only in machine shops sometimes cause an otherwise excellent article to be passed by. In addition to the mechanical problems, cost is frequently completely out of reach for the would-be constructor.

Here is a kilowatt linear amplifier covering 10 through 80 meters that has several features to recommend it to the fellow who wants to increase power. First of all is the cost. Using all new parts, except the meters which are readily available from used- or surplus-equipment sources, the total expenditure will be less than \$150 plus the cost of the tube. If you are willing to do some horse trading, scrounging and junk-box raiding, you can do it for considerably less. Type 4-1000As from broadcast or police radio transmitters are readily available at prices from \$20 to \$50. Surplus JAN tubes are listed by several *QST* advertisers, and they are regularly offered in Ham-Ads. Remember, the Eimac 4-1000A is built like a Mack truck and, once you have acquired one of these tubes in good condition, you can expect years of satisfactory service if you don't abuse it by overdriving the grid. That is why a grid-current meter is mandatory.

Secondly, construction is extremely simple. All mechanical work can be performed with ordinary hand tools. An electric drill will cut the con-

* 922 Whitney Blvd., Belvidere, Illinois.

[1] The Eimac data sheet on the 4-1000A as a grounded-grid triode qualifies this by adding, ". . . if a plate voltage of at least 3000 volts is used." — *Editor.*

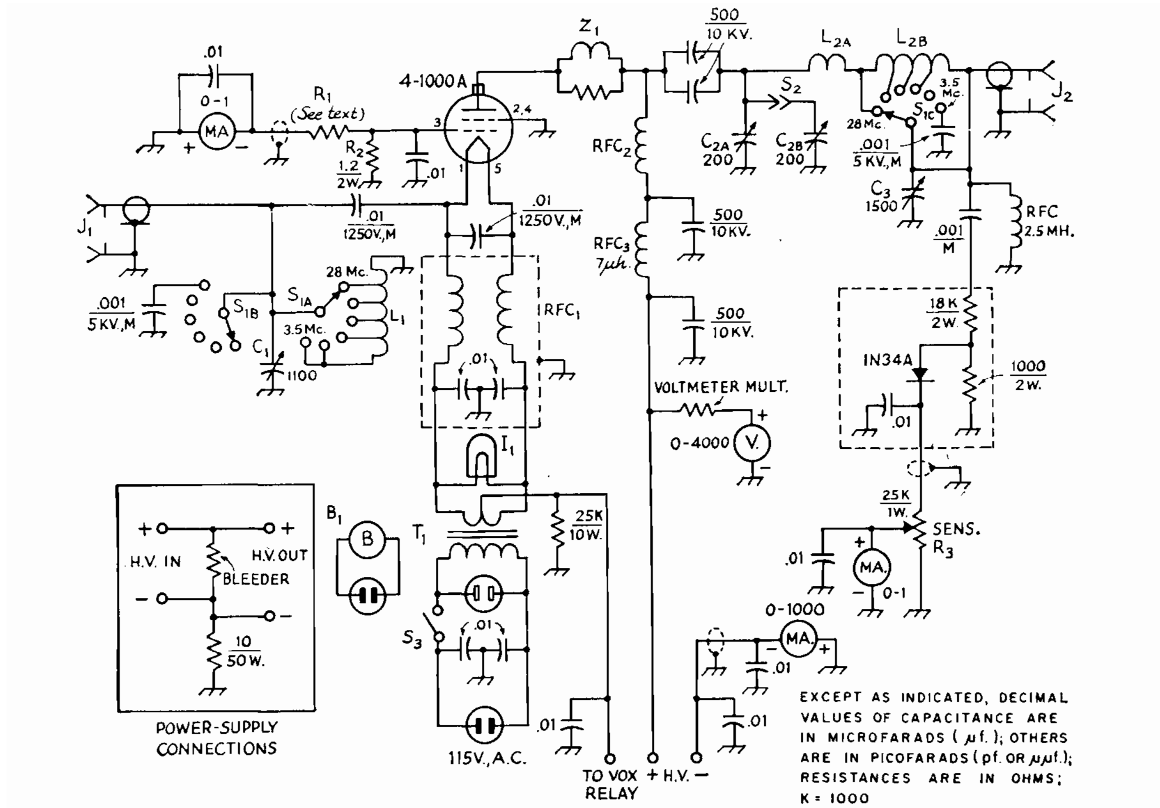


Fig. 2—Circuit of the 4-1000A grounded-grid amplifier. The 500-pf. 10-kv. fixed capacitors are TV doorknob type; others are 1-kv. disk ceramic, except M indicates mica.

B_1—Centrifugal blower, 60 c.f.m. at 0.6-inch static pressure (Ripley 8472).

C_1—Triple-section broadcast-replacement-type variable, 365 pf. or more per section, sections connected in parallel.

C_2—Dual air variable, 200 pf. per section, 7000 volts (Johnson 152-503/200CD70).

C_3—Air variable, 0.03-inch plate spacing (Cardwell PL-8013 or B & W 51241).*

I_1—6-8-volt panel lamp.

J_1, J_2—Chassis-mounting coaxial receptacle (SO-239).

L_1—6 turns No. 10, 1½-inch diam., 1½ inches long, tapped at 1¼, 1⅞, 2½, and 4⅛ turns from ground end.

L_2—Approximately 14 μh., tapped at 7, 3.5, 2.5 and 1.75 μh. (Barker & Williamson 850A bandswitching inductor).

R_1—Approx. 27 ohms; see text.

R_2—Made up of four 4.7-ohm ½-watt carbon resistors in parallel.

R_3—Linear control.

RFC_1—30-amp. bifilar filament choke (B & W FC30A).

RFC_2—Solenoid r.f. choke (B & W 800).

RFC_3—Solenoid r.f. choke (Ohmite Z-50).

S_{1A-B}—Single-section double-pole six-position ceramic rotary switch, 60-degree index (CRL 2551).

S_{1C}—Heavy-duty single-pole six-position rotary switch (part of L_2 coil assembly, modified as described in the text).

S_2—See text.

S_3—S.p.s.t. toggle switch.

T_1—7.5-volt, c.t., 21-amp. filament transformer (Stancor P-6457, Chicago F-725).

Z_1—2 turns No. 8, ½-inch diam., shunted by three 150-ohm 1-watt carbon resistors in parallel.

* The Cardwell capacitor is listed in the 1963 Allied catalog. The B & W capacitor, which is identical, is not stocked by B & W as a retail item, and may or may not be available at any particular time, depending on manufacturing needs. It is advisable to check with B & W before ordering from this source.

struction time considerably, but it is not an absolute necessity. The meter holes can be cut with a bit brace, or with a hand drill and file. Best of all, every single component is standard merchandise and is readily available. Your favorite ham supplier may not have every item in stock, but he should be able to get any of them for you in a hurry.

Triode Operation

The 4-1000A may be connected for high-μ triode operation by placing the grid and screen elements at the same d.c. and signal potentials; in this case, both are grounded. This connection offers several advantages for sideband operation. First, no grid-bias or screen-voltage power supplies are needed. In addition, the drive level of this grounded-grid stage is compatible with the power-output level of modern sideband exciters. Finally, neutralization is not required.

The Circuit

The circuit of the amplifier is shown in Fig. 2. Excitation is fed to the filament through a 0.01-μf. 1250-volt (working) mica capacitor. A ceramic capacitor is not suitable for coupling since it will not stand the current. The cathode coupler, consisting of C_1 and L_1, does an excellent job of input matching. RFC_1 is the new B & W FC-30A bifilar filament choke which is more efficient than the earlier type FC-30. With the center tap of the filament transformer returned to ground through an extra pair of contacts on the VOX or antenna relay, the no-signal resting current will be approximately 60 ma. with 3000 volts on the plate. With the relay contacts open on standby, the 25K bias resistor drops the plate current to a negligible value.

A B & W type 850-A coil-switching unit is used in the pi-network output circuit. The type 852, incidentally, is not suitable for use with the 4-1000A, since it is designed for a much lower plate load impedance. Its use would not only require much higher input and output capacitances, but would also result in an abnormally high-Q circuit in this amplifier. Instead of an expensive vacuum variable for the tank capacitor, C_2 is a split-stator air unit with 0.175-inch plate spacing. To reduce the minimum circuit capacitance on the higher-frequency bands, one section of the dual capacitor is used for 10, 15, and 20 meters; the second section is switched in parallel with the first for the lower frequencies.

The variable output capacitor C_3 is a 1500-pf. unit with 0.03-inch plate spacing. This provides sufficient capacitance for the phone end of the 80-meter band. However, more capacitance will usually be required for the low-frequency end of this band, and this is provided by connecting a fixed 0.001-μf. mica capacitor in parallel with C_3 in the last position of S_{1C}.

Parasitic Suppression

Several different makes of chokes were tried at RFC_2 in conjunction with many different resistance-inductance combinations in the v.h.f. suppressor Z_1. However, it was found practically impossible to completely eliminate parasitic oscillation on all bands until the B & W type 800 choke was tried.

Metering

Grid current is monitored very simply. The control grid is grounded through four 4.7-ohm ½-watt composition resistors in parallel, by-passed by a 0.01-μf. disk ceramic capacitor. The RC combination serves to hold the control grid very close to ground potential. Grid current is monitored by measuring the voltage drop across the resistors with the 1-ma. grid meter, calibrated 0-300 ma. full scale, and a series resistor.

A simple way to determine the value of the series resistor R_1 is to place a regular milliammeter with a scale of 200 ma. or more from the VOX relay terminal to ground. Apply excitation, and substitute resistors at R_1 until both meters have the same deflection at 150 ma. As an example, the Weston Model 301, 1-ma. meter requires a 27-ohm series resistor.

Plate current is measured by a 0-1-amp. d.c. meter shunted across a 10-ohm resistor in the negative high-voltage lead. This resistor is incorporated in the power supply, not in the amplifier itself. The 50-watt rating gives an ample safety factor, since the power dissipation would not exceed a few watts should the ammeter open up. Notice that the negative terminal of the supply must not be grounded except through the 10-ohm resistor.

A plate voltmeter has a definite place in this amplifier, or in any other amplifier where the d.c. input runs 900 watts or more, since it is required by FCC regulations. Even if you run less than 900 watts, it is reassuring to know exactly what your input is at all times.

To continuously monitor the r.f. output level of the amplifier and to aid in efficient tuning, a simple r.f. voltmeter has been incorporated in the circuit. Absolute readings are not necessary, so provision has been made for varying the sensitivity by adjustment of R_3.

Component Modification

Some of the components require minor modification before mounting. The last rotor plate and the last stator plate of the rear section of the tank capacitor C_2 are removed. This is section C_{2A} in the diagram, which is used alone on the higher frequencies. The operation is simple and requires no special tools. The alteration reduces the minimum capacitance to permit a more favorable Q on 10 meters. To further reduce the minimum circuit capacitance, the stators of C_2 are moved farther away from the chassis by mounting the capacitor in an inverted position; that is,

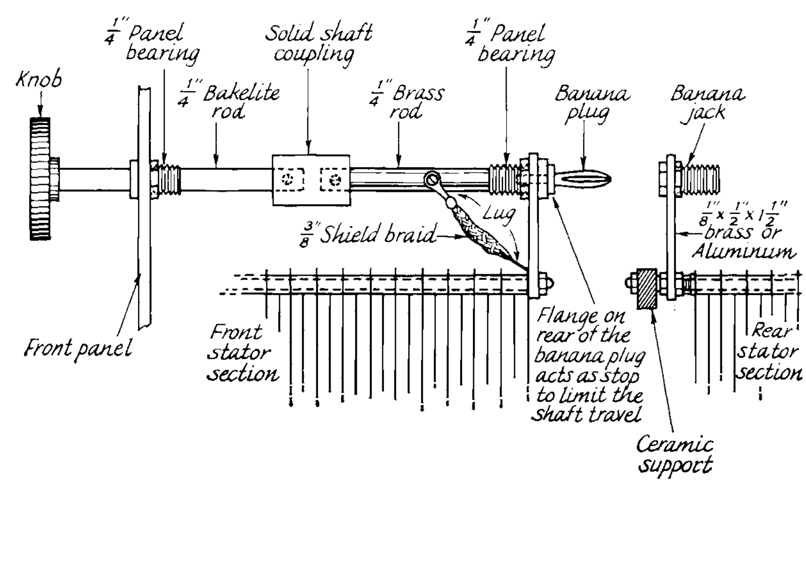


Fig. 3—Sketch showing details of the tuning-capacitor switch, S_2. The stator sections are connected in parallel when the panel control knob is pushed to engage the plug in the jack.

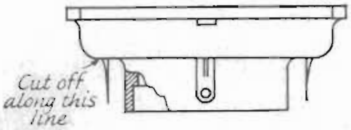


Fig. 4—Sketch showing how the lower portion of the tube socket is cut off.

with the stators on top. The mounting feet of the Johnson capacitor are easily moved to permit mounting in this manner, since the capacitor frame has duplicate mounting holes.

Fig. 3 shows the device used for S_2. Similar metal brackets are attached to adjacent ends of the stator-assembly rods of the dual capacitor. The bracket on the rear end of the front capacitor section (C_{2B}) carries a ¼-inch panel bearing through which a 3-inch length of ¼-inch brass rod slides. One end of this rod is drilled and tapped to accept the threaded shank of a banana plug. The other end of the brass rod is coupled to a 3⅛-inch length of ¼-inch bakelite rod which passes through another bearing in the panel to the control knob. The shaft coupler should be of the rigid type, either metal or ceramic. To assure good contact between the stator of C_{2B} and the banana plug, a piece of ⅜-inch flexible copper braid is used to connect the two directly, rather than to depend on the sliding contact at the bearing.

The banana jack is mounted on the other bracket. Be sure that the two brackets are drilled identically so that the plug and jack may be lined up accurately.

One other slight modification was made in the capacitor before mounting. A small triangular bracket was mounted inside the rear frame plate, that is, between the capacitor sections. This was fastened in place using the same screws which hold the ceramic stator bar against the frame plate. The upper point of the triangle extends sufficiently above the frame plate to allow mounting a 1-inch ceramic pillar. After the components were mounted on the chassis, the open end of the 10-meter section of L_3 was removed from the coil assembly, turned end for end, and fastened between the ceramic end plate and the ceramic pillar. A short length of ¼-inch copper tubing, also fastened to the ceramic pillar, connects the coil to one side of the blocking capacitors. Another short length of tubing connects the rear stator terminal of C_{2A} to the same point.

It will be noted that the 0.001-μf. fixed output capacitor requires an additional switch position. Fortunately, this is not difficult to provide, since there is already a hole for an extra stationary contact in the ceramic end plate of the B & W coil unit. All that is necessary is to obtain a switch contact from B & W [2] for one dollar (or make a reasonable facsimile) and mount it in the spare hole.

The socket for the 4-1000A is Eimac's new plastic type SK-510 (amateur net $6.50). It is designed primarily for duct connection to a blower. For the pressurized-chassis ventilating system used here, you can improve the air flow by cutting off the "nose" of the socket with a hacksaw, as shown in Fig. 4. Remove the socket contacts while this operation is performed, to avoid damaging them. Use extreme care in sawing. Although the socket is made of a tough plastic, unusual stress or strain may cause it to break.

You will note that the socket has slots next to the pins, right in the side of the molded fixture. To ground the two screen leads, pass a ¼-inch copper ground strap through the slot and solder it to the bottom of the screen contact inside the socket; then ground the strap to the chassis at the point where it emerges from the socket. The grid bypass capacitor should be installed in the same manner. One lead passes through the slot and is soldered to the bottom of the grid contact, while the other lead is grounded to the chassis. The leads should be only ¼-inch long.

Construction

The 14 × 17 × 4-inch chassis is made up of a pair of SeeZak [3] R414 rails (4 by 14 inches), a pair of R417 rails (4 by 17 inches), and two P1417 panels (14 by 17 inches). Standard 13 × 17×4-inch chassis are readily available, of course, but the extra inch of depth provided by the SeeZak units is necessary to accommodate C_2 which has a length of 13$\frac{1}{16}$ inches. Machining of the front and rear chassis walls and the top deck is greatly simplified by using these handy rails and panels. No more trying to get big fingers and tools into small corners. You can do all of the drilling and cutting on flat plates, and then assemble your chassis.

[2] Barker & Williamson, Bristol, Penna. Mention 850A type number when ordering.

[3] SeeZak products are available from Radio Shack Corp., 730 Commonwealth Ave., Boston 17, Mass., Terminal Hudson Electronics, 236 West 17th St., New York, N. Y., and California Electronics Supply, Los Angeles, among others.

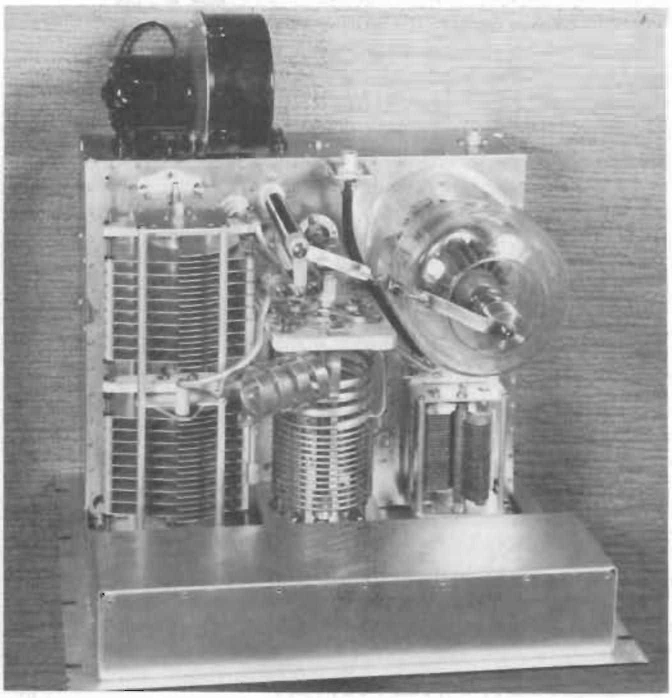


Fig. 5—Plan view of the 4-1000A grounded-grid amplifier. This view shows how the position of the 10-meter section of L_2 is changed.

Cathode Coupler

Place S_1, L_1, and C_1 close to the tube socket, as shown in Fig. 6. In this amplifier, Millen type 39005 universal-joint couplings were used between the shaft of C_1 and the front panel to allow the control to be placed symmetrically in respect to others on the panel. Even though the shaft and rotor of C_1 are at ground potential, use an insulated shaft coupling to couple the indicator dial to avoid the possibility of setting up a spurious tuned circuit. If you don't gang the input and output band switches, as described presently, use an extension shaft on the input switch so that the switch can be placed close to the tube socket.

Ganging the Switches

It is not difficult to gang S_{1A-B} and S_{1C} to provide single control. This can be accomplished by means of a National type RAD geared right-angle shaft coupler. A Johnson rigid ceramic shaft coupler (type 104–252) is attached to the tail shaft of the B & W coil unit. A short length of ¼-inch brass rod couples the gear end of the right-angle drive to the ceramic coupler. S_{1A-B} is mounted below deck with its shaft extending through a clearance hole in the chassis so that the shaft can be lined up with the shaft of the right-angle drive. The two shafts are coupled together by means of a ceramic semiflexible coupler (Johnson 104–262). Since the switch on the B & W coil unit has 60-degree indexing, S_{1A-B} must have the same indexing, rather than the more common 30-degree indexing. The 60-degree switch is, however, a standard item in the manufacturer's catalog. A 30-degree switch may be used, of course, if ganging is dispensed with.

Wiring

As the photographs indicate, very little actual wiring is required. The positive high-voltage lead enters the rear of the chassis through a Millen high-voltage connector where it immediately connects to the first 500-pf. bypass capacitor. RFC_3 is mounted between this capacitor and a feedthrough insulator which is connected to one side of the voltmeter multiplying resistor. The feedthrough carries the high voltage through to the top of the chassis where it connects to the second 500-pf. capacitor mounted on the chassis, and to the bottom end of RFC_2. A tapped ceramic pillar insulator threads onto the top terminal of this capacitor. The two blocking capacitors are suspended from a short copper strap fastened across the top end of the insulator, and a second strap connects them to the top of the r.f. choke. The parasitic suppressor Z_1 is inserted at the center of a copper strap connecting the top of RFC_2 to the plate cap of the tube.

Since the high-C input circuit carries considerable current, the r.f. wiring should be done with reasonably heavy wire (I used No. 10). This includes the short between the 80-meter contacts of S_{1A}.

A lead attached to the stator of C_3 passes down

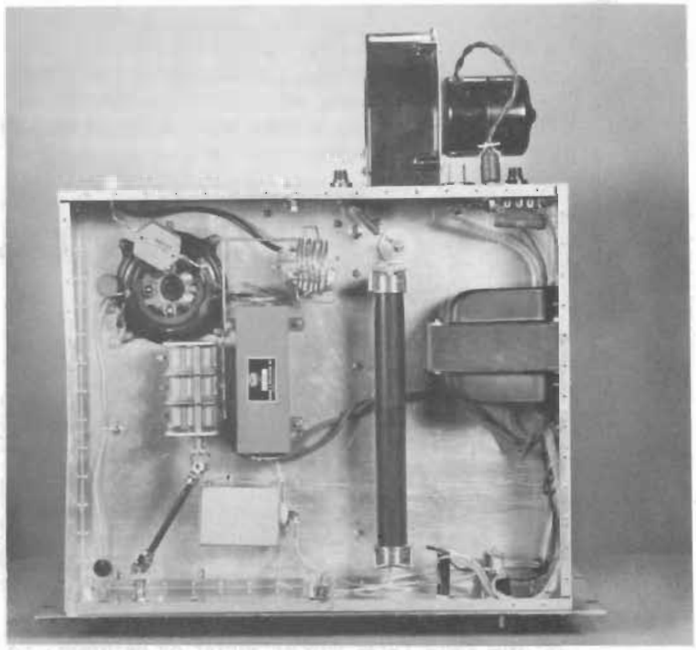


Fig. 6—Bottom view of the 4-1000A amplifier. The filament transformer and voltmeter multiplier resistor are to the right. The input coil, L_1, is at top center, supported on S_{1AB} by its leads. Input capacitor C_1 is operated through a pair of universal-joint shaft couplers so that the capacitor may be placed close to the tube socket without upsetting panel-layout symmetry. The small shielding box (2¾ × 2⅛ × 1⅝-inch Minibox), below the bifilar filament choke, houses the r.f. output-indicator diode and associated components.

through the chassis via a second feed-through insulator to the box below containing the r.f. output-indicator components. A short section of RB-8/U connects the stator of C_3 to J_2. Be sure to ground both ends of the outer conductor.

Blower Mounting

Don't compromise on the blower. The 4-1000A requires 60 c.f.m. at 0.6 inch of static pressure. Some so-called 60-c.f.m. blowers aren't worth their salt when you try to pressurize the chassis. The blower suggested does an excellent job in this respect, and is priced quite reasonably.

Be sure to place the blower well away from the tube socket. If it is placed too close, it will create a pressure wall across the bottom of the socket which will tend to restrict the flow of air through the base and chimney.

An a.c. receptacle is set in the rear apron of the chassis and a short cord from the blower motor plugs into it.

The Panel

The panel is a standard 15¾ × 19 × ⅛-inch unit of aluminum. The four meters are in line across the top. A 4 × 17 × 3-inch aluminum chassis fits over the back of the line of meters to shield them from r.f. fields. It is held in place by eight No. 6 sheet-metal screws inserted from the front. Shielded meter leads (Belden 8882 wire) are brought up from below chassis through rubber grommets in the chassis and in notches filed in the bottom front corners of the meter enclosure.

The panel is fitted with chrome handles (Bud type H9113) for lifting the amplifier in and out of the rack mounting. They also serve to protect the controls if it becomes necessary to place the unit face down on your workbench for service.

The lettering was done with Tekni-Cals, and the engraved plates are obtainable from Central Engravers [4] at 5 cents per letter.

The Shielding Enclosure

The two ends and the back of the shielding enclosure are made of 0.51-inch solid sheet aluminum, while the top is made of perforated sheet of the same weight. One of the SeeZak P1417 panels is used for the bottom cover. Aluminum angle stock, ½ inch by ½ inch, is used to join the pieces with the help of ¼-inch No. 6 sheet-metal screws spaced every two inches. All of the above pieces, including the angle stock, may be obtained cut to size if desired.[5]

Adjustment

After checking out the filament circuit and grounding the center tap of T_1, reduce the sensitivity control of the r.f. voltmeter to near minimum. Select the proper band with S_1 and apply excitation. Adjust C_1 for a grid current of approximately 150 ma. Apply plate voltage and load, and resonate the output circuit with C_2. With a plate voltage of 3000 and grid current of 160 to 170 ma., alternately adjust C_3 and C_2 to increase the plate current to 300 ma. or slightly over. In observing the r.f. voltmeter, you will note that maximum output does not always occur at the point of resonance as indicated by the dip in the plate current.

The amplifier may be checked for linearity as described in the *Radio Amateur's Handbook*.

I am very grateful for the technical advice and suggestions of Bill Orr, W6SAI, and George Stinson, W9KDK. Their analysis of the problems encountered, as well as their suggestions for changes during construction, made this a much better amplifier, and a pleasure to build. Operating at an input of 1 kw. or less, this amplifier actually "coasts" and will give you years of trouble-free service. QST

[4] 529 South State, Belvidere, Illinois.

[5] From Dick's, 62 Cherry Ave., Tiffin, Ohio.

"PULSE TUBES" IN AMATEUR SERVICE

Certain vacuum tubes designed for high voltage pulse service are now available to the radio amateur through "surplus" channels. Some of these tubes (such as the 6C21, which is a variation of an earlier transmitting tube) are suitable for amateur service. The 6C21 is roughly equivalent to the Eimac 1000T triode and the data sheet for the latter may be used to arrive at the proper operating parameters for the 6C21. Other pulse tubes, such as the 4PR60A, 715A-C, 4PR250C, X643A-F, X556D and Y-158 have no earlier r.f. prototype, as the tubes were developed solely with pulse service in mind. This family is designed to be used in high voltage service (up to 50 kilovolts) and the internal geometry is such that the tubes do not perform efficiently at plate potentials normally encountered in amateur operation. In particular, the 4PR60A type can be damaged by excessive screen dissipation when lower than normal plate voltages are used. These tubes, therefore, must be considered impractical for amateur service.

PLANAR TRIODE TUBES FOR UHF AMATEUR SERVICE

Planar triodes (or lighthouse tubes) are well suited for amateur use in the UHF spectrum. Various surplus versions, such as the 2C40, 446B and 2C39 have been used at frequencies up to 1300 Mc. A new planar triode, the 3CX100A5 is now available for improved service in this frequency region.

The 3CX100A5 is relatively unknown to the amateur fraternity, but its older relative, the 2C39A, has been long a favorite UHF tube of the "surplus hounds". The 3CX100A5 is an improved, modern version of this old World War II tube, dressed up in a brand-new ceramic and metal envelope. This tube, and its twin, the 7289 fit into the amateur picture very nicely as a straight-through amplifier, a doubler, or as a tripler in the frequency range of 1000 to 3000 Mc. In addition, it is not expensive, and various glass versions of the older 2C39/2C39A/2C39WA can often be obtained at give-away prices on the surplus market.

The following data covers grounded grid operation of the 3CX100A5 as a UHF power amplifier and multiplier. Because of the high power gain of the tube, grounded grid circuitry is desirable, since intercoupling between the input and output circuits is reduced to a minimum, and neutralization is not required. This data can be used to estimate the performance of the 2C39 glass family of tubes by noting that the useful power output of this style tube will be somewhat less (depending upon the frequency) by an amount up to 25-percent at 2.5 kMc as shown on the graphs.

Various 3CX100A5 tubes were run in a coaxial cavity capable of tuning from 1000 mc (1 kMc) to 3000 mc (3 kMc). A series of measurements was made using a representative sample of standard production tubes, with the tube under test operating as an amplifier, doubler and tripler. Drive and output power were carefully measured for each test, and appropriate filters were used to eliminate feedthrough and harmonic power. In the case of the amplifier configuration, the feedthrough power was useful output and therefore was measured as such. When the tube is operated as a doubler or tripler, the feedthrough power is at the driving frequency and is undesired. In actual use, it is necessary to eliminate this power from the output circuit of any grounded grid frequency multiplier. High-Q tuned circuits or wave filters will do the job.

The 3CX100A5 as a Grounded Grid Amplifier

A graph of average tube performance is shown in figure 2 of the UHF grounded grid amplifier circuit. Grid drive, grid bias, and plate loading were adjusted to provide maximum power output while maintaining the plate current at 100 ma. Average potentials for tubes tested are indicated on the graph. At 1300 mc, for example, the 3CX100A5 is capable of a power output of about 47 watts at an efficiency of 47%. The power gain of the tube is 8 decibels, indicating a required drive level of about 7.5 watts as measured at the input of the coaxial fitting to the cavity. Power output gradually decreases as the frequency of operation is raised. At 2400 Mc, power output (at 100 watts input) drops to 25 watts, and grid driving power increases to 10 watts. Power gain at this frequency is 2.5. The comparative curve for the 2C39A tube over the same frequency range is shown by a dotted line on the graph.

The 3CX100A5 as a Grounded Grid Doubler

The same sample of tubes used in previous tests was used in performing the gain and power output measurements for the doubler configuration. Excitation was applied at half-frequency and the same precautions were followed as in the case of the amplifier. Test conditions were adjusted for maximum power output at a plate current of 100 ma. At 1300 Mc, power output is about 27 watts, with a circuit efficiency of 27%. Power gain is 5.4 decibels. Accordingly, a drive power of about 8 watts is required. At 2400 Mc, power output is 13 watts, with a drive power of about 9 watts. Power gain drops to unity at a frequency of about 2700 Mc. The tube is still useful as a frequency doubler to this frequency, however, as a power output of 10 watts can be obtained. 2C39A curve is shown as dotted line on the graph (figure 3).

The 3CX100A5 as a Grounded Grid Freauency Tripler

The same tubes and general test techniques were used to determine the operating parameters of this tube as a frequency tripler. Drive power was applied at one-third output frequency, and the circuit was adjusted for maximum power output. At 1300 Mc, 17 watts of power were obtained with about 10 watts driving power. At 2400 Mc, power output was about 8 watts, with about 12 watts drive power (figure 4).

Socket and Collets for the Planar Tubes

A complete socket for the 3CX100A5/2C39 type tube is a rare bird indeed: most equipments have the tube sockets built directly into r.f. cavities. Collet rings for the tube, however, may be bought from Instrument Specialties Co., Little Falls, N.J. The catalog numbers are: Plate collet, #97-70; grid collet, #97-72; cathode collet, #97-76; filament collet, #97-280. The Braun Tool & Instrument Co., 140 5th Avenue, Hawthorne, N.J. can supply the following: Plate collet, #134-53; grid collet, #134-51; cathode collet, #165; filament collet (none). A complete socket assembly may be purchased from Jettron Products, Inc., 56 Route 10, Hanover, N.J.

Conclusions

The planar triode tubes of the 3CX100A5 family perform well in the frequency range encompassing the 1215 Mc and 2400 Mc amateur bands. Efficiency is good, considering the frequency of operation. "Radio Engineering" by F.E. Terman indicates that a doubler will provide about 65% of the power output of a straight-through amplifier, and a frequency tripler will provide about 40% of the power output of the amplifier. These figures agree closely with the data shown herein. A pair of 3CX100A5's should make quite a dent in the 1215 Mc band! See you on the high end!

Robert I. Sutherland, W6UOV

Type	Manufacturer	Comment
2C39A	Eimac, Machlett	Glass or ceramic construction
2C39WA	"	"
6897	General Electric	Not interchangeable electrically in most high frequency sockets with 3CX100A5 or 7289
2C41	Machlett	Not interchangeable physically with 3CX100A5 or 7289
381	Eimac, Machlett	3CX100A5 type, rated for pulse service (Eimac-glass or ceramic)
3CX100A5	Eimac, Machlett General Electric	All ceramic construction
7289	Eimac, Machlett, General Electric	Identical to 3CX100A5
3CX100F5	Eimac	Identical to 3CX100A5 except heater voltage is 26.5 volts

Figure 1. Planar Triode Tubes

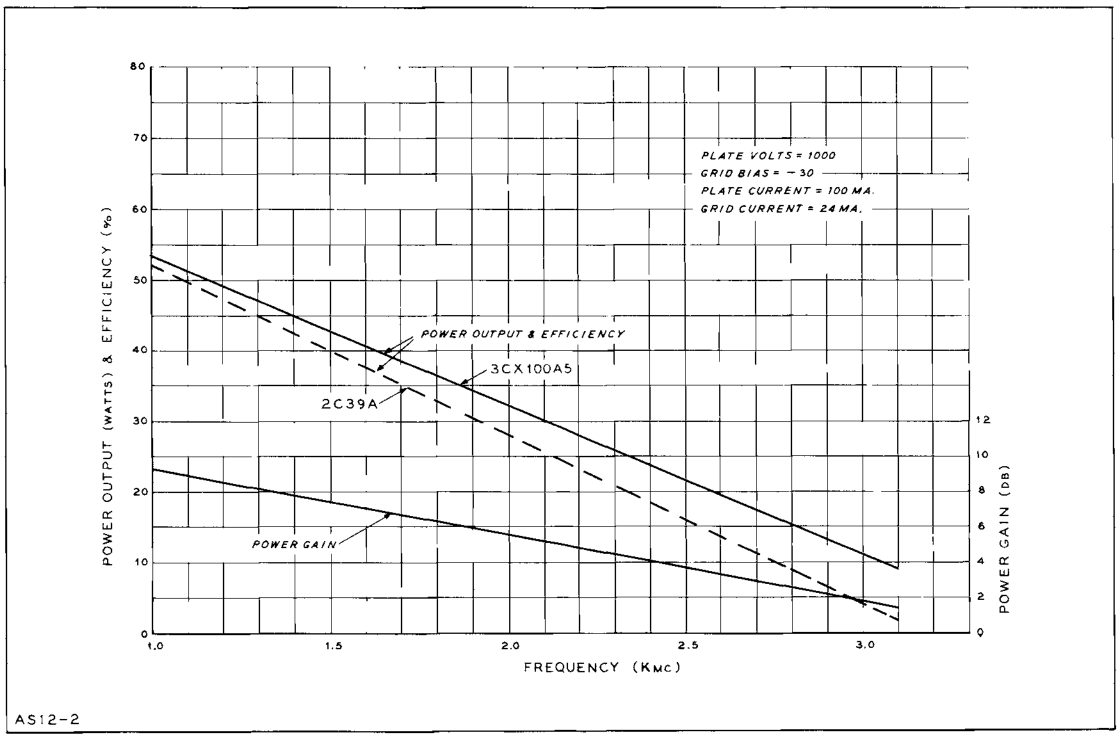


Figure 2. 3CX100A5 grounded grid amplifier; typical power output, gain and efficiency. Glass 2C39A power output shown by dashed line.

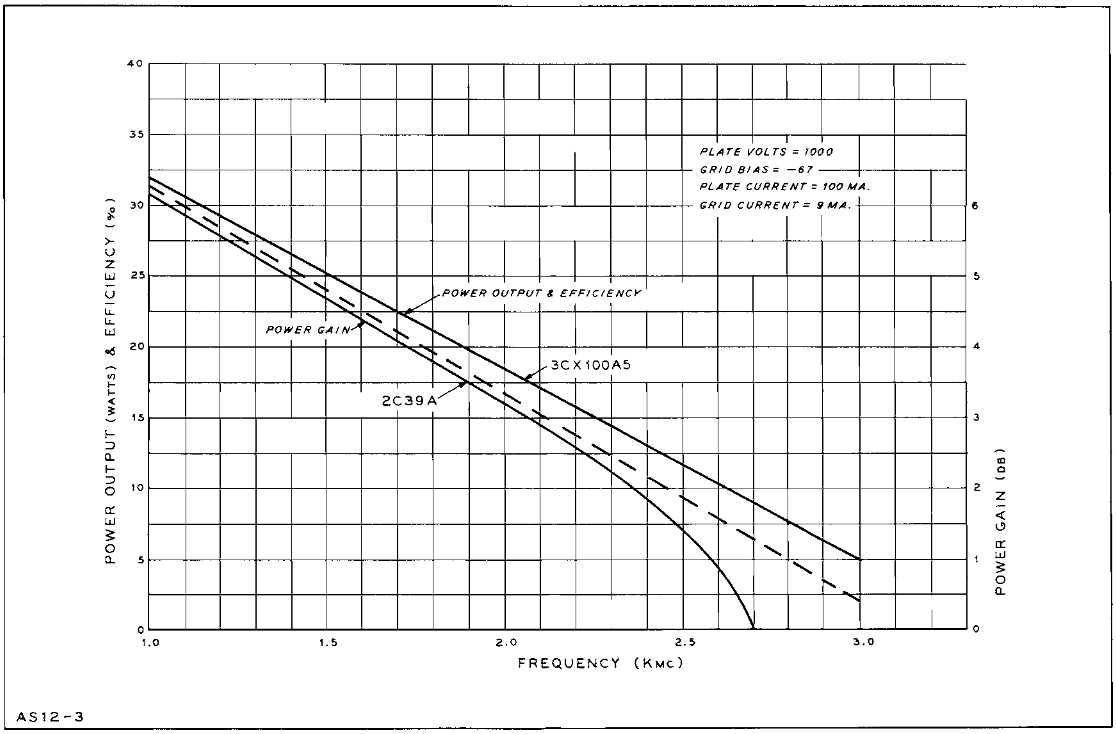


Figure 3. 3CX100A5 grounded grid doubler; typical power output, gain and efficiency. Glass 2C39A power output shown by dashed line.

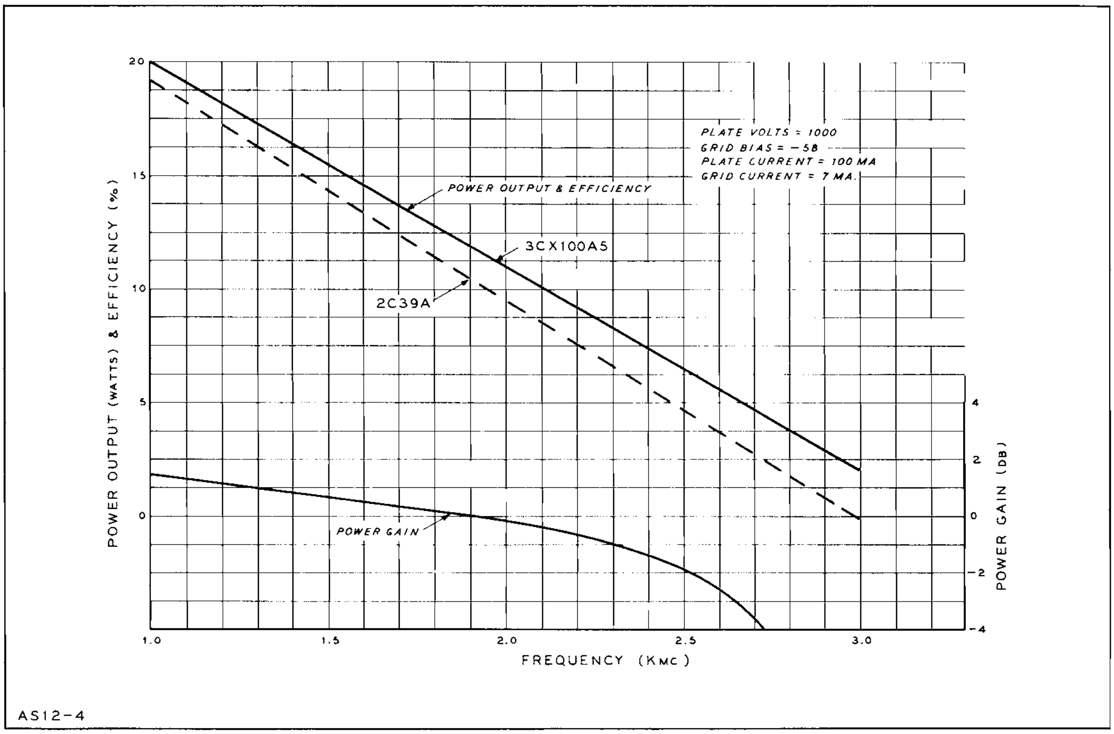


Figure 4. 3CX100A5 grounded grid tripler; typical power output, gain and efficiency. Glass 2C39A power output shown by dashed line.

amateur service newsletter
W6SAI

UNDERSTANDING TETRODE SCREEN CURRENT
Significance in R.F. Amplifier Adjustment and Operation

by David D. Meachum, W6EMD

This article discusses the behavior of screen current in a tetrode r.f. power amplifier using fixed screen voltage, and explains why a screen-current meter is a better indicator of operating conditions than a plate-current meter. Particular reference is made to the adjustment of AB_1 linear amplifiers.

Perplexing screen-current behavior has probably disturbed many amateurs, particularly single-sideband operators. The need for a thorough discussion of the subject has prompted this article. Class AB_1 operation has been chosen for discussion because of its current popularity as a means of achieving good linearity and TVI-free operation. The information given herein assumes grid-driven conditions, but it applies equally well to cathode-driven tetrodes operated Class AB_1 with normal d.c. voltages on the grid and screen, provided that grounded-grid characteristic curves are used for computations.

Screen Characteristics

Fig. 1 shows a set of constant-current characteristics for a typical 4CX300A. The term "constant current" is used because the lines plotted are lines of constant plate, screen, or grid current. The grid-voltage scale appears on the left axis and plate voltage is shown horizontally. These curves depict instantaneous values of plate and screen current for any given grid- and plate-voltage conditions. In this reproduction, the grid-current lines are omitted because grid current is not drawn in Class AB_1 operation. The curves are valid only for fixed screen voltage (350 volts in this case).

Inspection of Fig. 1 will reveal that the lines of constant plate current are nearly horizontal, whereas the constant-screen-current lines are tilted upward from left to right and are concentrated in the left-hand region of the plot. This is generally true for all tetrodes and accounts for the fact that the screen-current meter is the most sensitive indicator of resonance. This important fact will be explained subsequently.

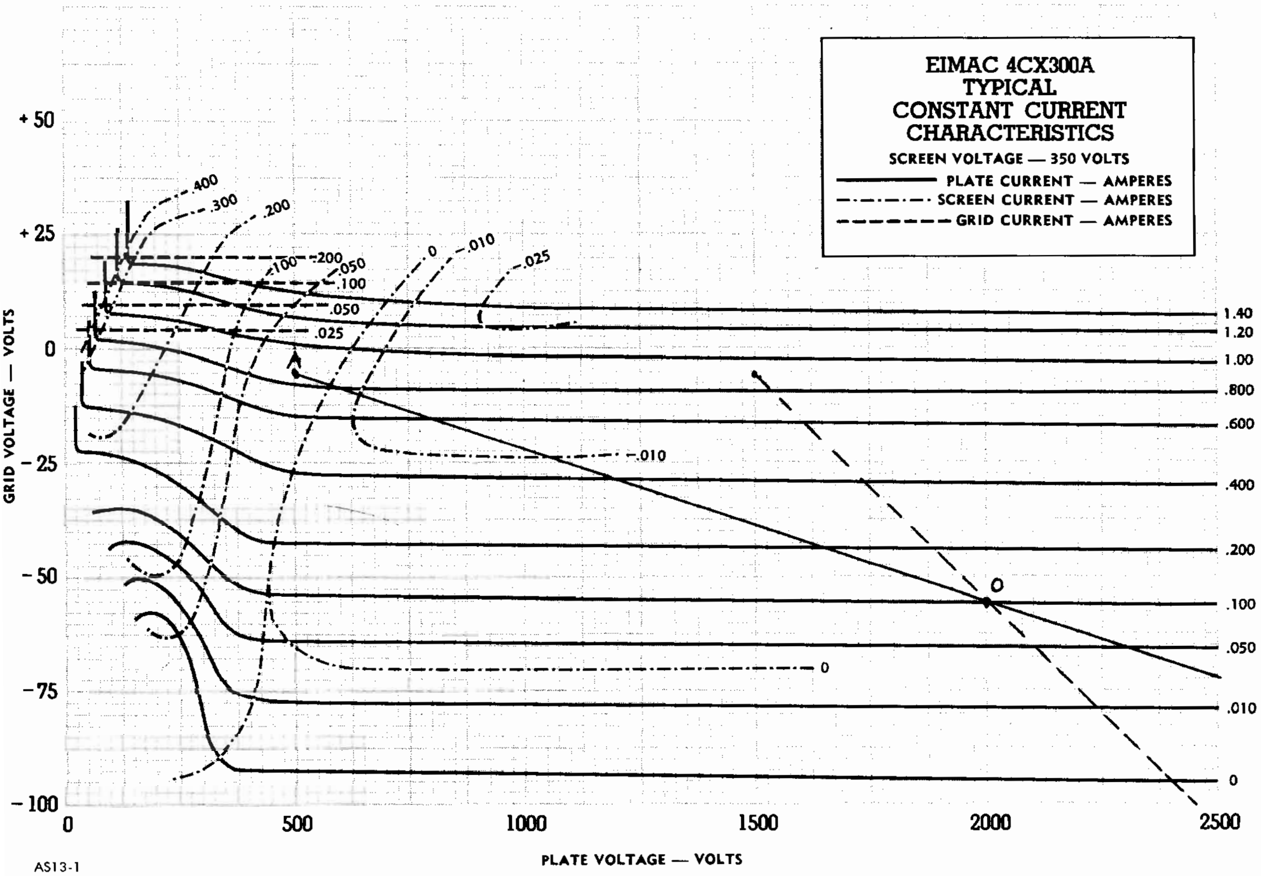


Fig. 1 - Typical constant-current characteristics for the EIMAC 4CX300A tetrode

1 This is different from the usual load line associated with audio calculations using plate characteristic curves.

2 OA is actually only half the operating line length. The other half continues from O out beyond the right-hand edge of the chart for an equal distance and represents the effect of the negative half-cycle of grid driving voltage as it swings down to -105 volts and back to point O again. This half of the operating line is not important since the tube does not "work" during the negative half cycle.

Let us plot a typical operating line[1] on our set of curves, as in Fig. 1. Point O (at -55 volts on the grid in this case) is the operating point at which the tube rests with zero r.f. grid drive. Straight line OA represents a tuned r.f. circuit load (a pure resistance at the operating frequency)[2]. As 100 volts peak-to-peak grid drive is applied, the first positive half cycle can be represented by a point moving along the operating line from O to A and back to O again. During this half cycle, the grid-voltage swing from -55 volts has caused the plate current to swing from the value at point O (100 ma.) up to the value at point A (850 ma.) and back to 100 ma. again. At the same time, the plate voltage swings from 2000 volts down to 500 volts. The a.c. plate current is made up of all the instantaneous values intercepted by the point traveling along the operating line. The same is true of screen current. During the other 180 degrees of the driving cycle, our point merely travels from O down the slope through cutoff to a point opposite -105 volts on the grid-voltage scale and back to point O again along the operating line. Thus, the negative-going grid voltage swings the plate current down to cutoff (for a small portion of the cycle). Plate voltage continues on up to 3500 volts and back down again due to the fly-wheel action of the plate tank circuit.

Drive and Tuning

Now that we can predict exactly what the screen and plate current will be for any instantaneous point during the grid-voltage cycle, let us ask some more probing questions. What happens when we cut our grid-driving voltage in half? The answer is simple. The length of our operating line is merely cut in half! The grid voltage swings to only one half the original peak-to-peak amplitude and the operating point O is still the center of the new operating line length. Now what happens if we detune the plate tank circuit? Detuning the plate circuit actually changes the plate load impedance. How does this appear on our set of curves? It tilts or rotates the operating line about the operating point O. As the load impedance is lowered (detuning from resonance), the operating line[3] assumes a steeper angle (a zero-impedance load would be represented by a vertical operating line).

As "seen" by the tube, the act of tuning to resonance amounts to increasing the load impedance to a maximum value consistent with the degree of antenna loading selected. Thus, the operating line will have minimum slope at resonance. Notice the angle at which our typical operating line in Fig. 1 cuts the constant-plate-current lines. It's a small angle. As the plate tank circuit is tuned to a point out of resonance, the operating line might assume the position indicated by the dashed line[3] (lower impedance). Note that the angle between the dashed line and the plate-current lines has not changed radically, and that our moving point will still intercept essentially the same plate-current values. This is precisely the reason that plate current in a tetrode is not a good indicator of resonance (very little dip). Look at the screen current. It consists of zero or even negative values in the out-of-resonance position. At resonance, though, it is positive. Thus, a peak in screen current indicates resonance.

[3] The tank-circuit impedance would no longer appear resistive at the operating frequency, but would contain a reactive component. Under these conditions, the operating line becomes an ellipse whose center is point O and whose major axis is represented by a dashed line.

During the rotation of the operating line while tuning, its length actually changes, since it is confined vertically only by the constant peak-to-peak amplitude of the grid-driving voltage (two imaginary horizontal lines, one at -5 volts and one at -105 volts). The length increases as resonance is approached and reaches a maximum at resonance. As the length increases, point A penetrates the heavy-screen-current region and the d.c. screen current reaches a sharp peak at resonance.

Loading

What happens if we change the antenna loading? This merely changes the plate-load impedance (still resistive). Again, the effect is to tilt the operating line about the operating point. As the load impedance is lowered (more coupling), the operating line assumes a steeper angle (such as the dashed line). It is easy to see that as loading increases, screen current decreases. Thus, screen current is also an indicator of loading. Screen current varies somewhat from tube-to-tube of a given type, but if each tube is loaded to the same value of screen current at resonance (with the same drive) power output differences will be small, and loading and linearity will be essentially the same.

D.C. Meter Readings

During the r.f. cycle, our point traverses the operating line and intercepts many different instantaneous values of screen current and plate current. The average of all these values is what the d.c. meter in the circuit reads. The fundamental frequency component of plate current is utilized in the plate circuit to produce output (except in a multiplier where use is made of a harmonic component of plate current). For a given operating line, both of these values can be calculated[4]. Suffice it to say that for Class AB_1 operation, the d.c. meter reading is approximately one third the peak value of current at the top of the operating line, and the fundamental component of plate current is approximately one half the peak value.

Tune-Up Procedure

Contrary to somewhat popular opinion, a linear amplifier should never be loaded for maximum power output. Loading should be set to obtain a pre-determined value of screen current under single-tone or inserted-carrier driving conditions. Ideally, loading should be set for minimum distortion - a rather difficult feat to practice. It is recommended that the amateur try to duplicate as nearly as possible a given set of data-sheet conditions as presented by the tube manufacturer. These typical operating conditions are usually given for peak-envelope operation (single-tone or inserted-carrier) and represent the maximum input on c.w. or the peak-envelope-power input (not meter peaks) on single sideband. After adjusting drive, tuning, and loading to duplicate a given set of conditions, the single tone (or carrier) is removed and the single-sideband audio gain is adjusted so that grid current is never drawn and the condition adjusted for above is never exceeded on peaks. The peak-to-average ratio of d.c. plate current (as read on a fluctuating meter) varies, with the individual voice, from about 2:1 to over 3:1. Thus it is normal on voice peaks for the plate-current meter to read no more than half the value of current obtained in the maximum static single.tone condition.

[4] By the use of the EIMAC Tube Performance Computor, Applications Bulletin No. 5, which is based on the method presented by Chaffee in the "Review of Scientific Instruments," October, 1936.

A straightforward tune-up procedure consists of the following steps:

1. Ensure that the tetrode amplifier is neutralized and free of parasitics.

2. With recommended heater, plate, and screen voltages applied, adjust the d.c. grid bias to obtain the recommended zero-signal value of plate current. This value affects linearity and plate dissipation.

3. Connect a suitable dummy load and set the loading control for rather heavy loading.

4. With a single-tone source, gradually increase the drive from zero to a value that produces a significant though small change in screen current.

5. Resonate the plate tank circuit by tuning for a peak (in the positive direction) in screen current.

6. Resonate the grid tank circuit (if any) by watching for a peak in plate current.

7. Now increase the drive until either the desired value of single-tone screen or plate current is reached (whichever is reached first).

8. Without drawing grid current, adjust loading, plate-tank tuning, and drive level to duplicate as nearly as possible a given set of data-sheet peak-Envelope conditions. Remember that plate current increases with drive, whereas screen current peaks at resonance and decreases with heavier loading.

After matching a set of data-sheet conditions, the amplifier is ready to connect to an antenna. With a suitable antenna connected, it should be easy to repeat the operation obtained in Step 8 above by merely adjusting plate-tank tuning and loading with the same drive level as before. Now set up for voice single-sideband drive and adjust the audio gain for the highest level possible without drawing grid current on voice peaks or flat-topping (check this with a scope).

Reverse Screen Current

Most transmitting tetrodes employing oxide-coated cathodes exhibit negative screen current under certain conditions of operation. This is nothing to get alarmed about - it merely means that on the average, more electrons are leaving the screen than are being intercepted by the screen. This results because of secondary electron emission at the screen grid. Small values of negative screen current are not detrimental to tube operation and are quite normal for some tetrodes. Such values usually appear under heavily-loaded conditions or during the idling condition.

Large values of negative screen current are abnormal and should be avoided. Excessive secondary emission usually results in higher values of intermodulation distortion. This condition also prevents an accurate determination of screen dissipation.

Protection

Screen protection can take many forms. Before using a given circuit, it should be analyzed to ensure that it satisfies the two basic criteria for screen protection. First, the circuit connected to the screen must be capable of maintaining the proper screen voltage in the presence of moderate negative d.c. screen current, or normal positive values of current. Second, the protective circuitry must not allow a condition of excessive screen current (positive or negative) to persist, since this causes excessive screen dissipation and resultant tube failure.

The first of these two criteria can be easily satisfied by the use of a bleeder resistance connected directly from the screen to ground, in combination with a suitable well-regulated power supply. The bleeder resistance should be made equal to the screen voltage divided by the largest negative d.c. screen current to be expected for the particular tube used. This eliminates any power supply problems (soaring voltage) when "supplying" negative screen current.

Complete screen protection satisfying both criteria can be obtained by adding a screen-current overload relay to a bleeder and regulated-power-supply combination. The overload relay will protect the screen against excessive currents, either positive or negative, and the regulated power supply will maintain the screen voltage at the proper value as the d.c. screen current varies. The bleeder resistance from screen to ground will not allow the screen voltage, in the presence of negative screen current, to rise above the proper value. This bleeder is good insurance, since even some regulated power supplies react in an undesirable manner when subjected to a negative-current load.

When using a screen-current overload relay one can easily provide for manual resetting in the event of an overload. This feature allows time to consider why the overload occurred and prevents repeated successive overloads. Using an s.p.d.t. relay, merely connect the armature to the positive supply through the coil (with the usual pull-in-adjusting potentiometer shunting the coil). Connect the normally-closed contact to the screen through the screen-current meter and the normally-open contact through a resistor to ground[5]. Adjust this resistor so that the current through it will hold the relay closed, once it has been tripped. First, of course, the pull-in shunt should be adjusted for pull-in at the value of screen-bleeder current, plus screen current, that produces maximum rated screen dissipation. Now, with this circuit it will be necessary to shut off the screen supply (or push a circuit-breaking series reset button) to reset the overload relay after an overload has occurred.

In contrast to the protective scheme outlined above, voltage-regulator tubes offer a simple and nearly foolproof method of screen-current protection. Their use will completely satisfy the first criterion and also the second criterion insofar as positive current overloads are concerned. Since excessive negative current is uncommon, one may elect to disregard protection against its occurrence. VR tubes then become an inexpensive and practical solution for the amateur.

The VR tube solution consists of an appropriate combination of VR tubes (to add up to the desired screen voltage) connected in series to ground and fed from a high-voltage source through an adjustable dropping resistance. The screen bypass capacitor from screen to ground and a screen-current meter from screen to the top of the VR-tubes string complete the circuit. Adjust the dropping resistance so that the VR string extinguishes at or slightly lower than the value of screen current that produces maximum rated screen dissipation. R.F. screen-current peaks will be supplied by the screen bypass capacitance and the VR tubes will "see" only the d.c. component. Now, excessive positive screen current will extinguish the VR tubes, lowering the screen voltage, The VR tubes will supply normal positive current values while maintaining screen voltage at the desired value. Negative currents will not change the voltage, but will merely increase the current flowing through the VR tubes.

[5] See Evans, "Screen Protection and More," QST, October, 1961.

In conclusion, it should be obvious to the amateur that a screen-current meter is a vital necessity in modern transmitters employing tetrodes. By proper interpretation of screen-current readings, one can easily tune to resonance and properly load the tetrode amplifier. The plate-current meter is useful only as an indicator of drive level and average plate-input power (knowing the plate voltage). One more meter - for grid current - is useful but not absolutely necessary. A one-milliampere meter in the grid circuit will warn the operator by a slight kick when grid current is being drawn on voice peaks.

Triode Tubes as Linear R.F. Amplifiers

Many of the common triode transmitting tubes will perform well as linear r.f. amplifiers for single sideband operation. Unfortunately, little operating data for this class of service is available for these tubes, and the radio amateur does not always have the operating curves or other data available to derive proper circuit potentials necessary to achieve low distortion linear operation. The amateur owning a high dissipation transmitting triode, therefore, is often at a loss as to the proper operating parameters for linear service, or he might not know if the tube is really suited for this class of operation. Then, too, good transmitting equipment exists, designed before the advent of SSB, that may be easily converted to single sideband linear service, provided the correct operating potentials can be determined. The purpose of this article is to provide rough guidelines whereby any triode tube may be easily and quickly evaluated as to its useability as a linear r.f. amplifier.

Important Tube Characteristics for Linear Service

Triode tubes[1] may be operated either in grid- or cathode driven service, and may be run in Class-AB-1, Class AB-2, or Class B Mode. The problem, then is to determine **which** triode tubes are most suitable for linear service and in **what** circuit configuration they work best, and finally, to establish the operating potentials which will provide satisfactory results for the user.

Plate dissipation and **amplification factor** are two triode tube characteristics which, in general provide the information to establish the work of the tube as a linear amplifier and which may also suggest the proper circuitry. Other tube characteristics, of course, enter the picture, but by merely examining these two main features, a triode tube may be easily evaluated for linear amplifier service.

Plate dissipation is important because it determines the maximum power limit that the tube may achieve under a given degree of operating efficiency. Linear amplifier efficiency commonly runs between 50- and 65- percent with the remainder of the power being lost as plate dissipation. As a rule of thumb, therefore, 50- to 35- percent of the maximum permissible power input to the tube represents the rated value of maximum plate dissipation. Twisting this idea around, it can be seen that a triode tube having, for example, 450 watts of plate dissipation is probably good for a power input of 900 to 1250 watts, depending upon operating efficiency. Class AB-1 operating efficiency runs close to 50-percent. Class AB-2 efficiency is about the same as Class B; in the vicinity of 65-percent. Thus, knowing the plate dissipation of the tube, we can easily determine approximate maximum power input and power output by first establishing in which class of service the tube is to be operated. It is wise to use "steady state" or single tone condition when making this rough determination of power levels. It is tempting to think that

1 - This discussion deals solely with triode tubes. Similar conclusions may be drawn about tetrode and pentode tubes, but these are not within the scope of this article.

the peak envelope power (P.E.P.) rating may be greater than this level (and it may--!) but this reserve power capability varies between tube types, and can only be determined by experiment at a later stage of the game.

Small transmitting tubes having low values of plate dissipation (3C24, 35T) afford little worth in linear amplifier service as the power capability of these small "bottles" is quite low. The picture changes rapidly as the power capability of the transmitting tube increases, however, and large triode transmitting tubes having high values of plate dissipation are often quite reasonable in cost and are likely candidates for linear service.

Amplification Factor of a triode tube expresses the ratio of change of plate voltage to a given change in grid voltage at some fixed value of plate current. It is determined primarily by the density of the grid structure and the grid-plate spacing of the tube. Amplification factors are expressed in terms of mu (u), and values of u between 8 and 300 are common for triode transmitting tubes. **High-u** tubes are those having an amplification factor of 30 or more, **medium-u** tubes have amplification factors in the range of 10 to 30, and **low-u** tubes fall in the range of 3 to 10.

Amplification factor for many commonly used transmitting triodes is given in most tube manuals and in the transmitting tube section of **The Radio Amateur's Handbook**. This characteristic may be determined experimentally by operating the tube in a quiescent state at normal plate voltage. A small change in the resting bias is made, thus slightly changing the plate current. The plate current is then returned to the original value by making an appropriate change in plate voltage. An amplification factor of 20, for example, means that if the grid potential is changed by one volt, it will take a plate voltage change of about 20 volts to restore the plate current to the original value.

Armed with a knowledge of plate dissipation and amplification factor, an intelligent choice of circuitry may now be made.

Class AB-1, Class AB-2, or Class B Grid Driven Service

These classes of operation are defined by the operating potentials applied to the tube. In general, **any** triode tube in **any** conventional linear amplifier circuit can probably work in **any** of these three modes. Results, however, depend upon circuit design considerations which may be extremely stringent in order to provide a satisfactory degree of linearity for a specific tube in a given circuit. Thus, in order to relax circuit considerations while reaching a desired degree of performance, certain tubes and circuits have been chosen which are widely used in SSB equipments while other tubes and circuits have been cast aside as being economically undesirable.

The Class AB-1 Amplifier is defined as one wherein grid current does not flow over any portion of the operating cycle. That is, grid excitation is held below that value at which the applied **peak** signal level is less than the value of grid bias on the tube. Plate current flows for more than 180 degrees of the r.f. cycle but less than 360 degrees. Once the bias value is exceeded by the driving signal, grid current will flow when the grid is driven positive with respect to the cathode, and the amplifier passes into the **Class AB-2** operating mode. As the peak driving signal level is increased and the bias reduced to substantially cutoff, grid current flows over half the operating cycle, and the amplifier is now operating in the **Class B** mode. As the grid driving signal increases, passing the amplifier stage through the successive operating modes, the instantaneous plate voltage swing increases, and electrode voltages applied to the tube must be varied to allow optimum circuit efficiency and low distortion in each mode of operation. Good and valid reasons exist for each mode of operation, and while it is tempting to jump to the conclusion that all tubes should be operated Class B mode for highest efficiency and output, such is not the case! Other factors affect the choice of operating mode and in doing so, also affect the circuitry and potentials to be used with a particular tube.

In general, a low-u triode tube is a preferable choice for a Class AB-1 grid driven linear amplifier. It is easier to obtain maximum plate current swing with this type of triode tube because it is impossible to drive the grid into the positive region of operation (Figure 1). A high-u tube of equal plate dissipation must be driven into the positive grid mode of operation to obtain output comparable to that of a low-u tube driven just to the point of grid current. Of the many triode tubes, the 304TL and the 211 (VT-4C) are common low-u types that perform well in Class AB-1 service. Other low-u triodes that are satisfactory for grid driven linear service are the 75TL, 100TL, 250TL, 806, 450TL and 750TL. Even though large values of driving voltage are required for these tubes, little driving power is required, as the grid never draws current and only circuit losses require that the driving stage supply power to the linear amplifier in question. High-u triode tubes are not recommended for class AB-1 service because the grid must be driven into the positive (grid current) region before appreciable power output can be obtained.

The **Class AB-2** amplifier is defined as one wherein the grid current is drawn over a portion of the operating cycle, yet the plate current flows for more than 180°. Grid current flows because the exciting signal has a peak value greater than the fixed operating bias of the tube, and the grid assumes a positive potential when the peak signal exceeds the fixed bias level. Electrons intercepted from the cathode to plate current flow by the positive grid create the grid current, and the power required to create the necessary positive grid swing is derived from the exciting signal that drives the linear amplifier. The impedance the grid presents to the input circuit under conditions of grid current is a function of the ratio of instantaneous grid voltage to current which varies in a nonlinear manner. Thus, unless the r.f. power source (the exciter) has extremely good output voltage regulation, and unless the linear amplifier is properly designed to present a fairly constant load to the exciter, waveform distortion will invariably result whenever this class of amplifier loads the exciter during those intervals of the operating cycle when the grid of the tube is driven positive with respect to the cathode. (**The Class B** linear amplifier represents an extension of the Class AB-2 mode, wherein grid driving power requirement is greater and the demand placed on the driving source is even more stringent).

Class AB-2 grid driven operation of triode tubes should be approached with caution because of this problem, as most radio amateurs do not have the test equipment to properly evaluate and adjust their linear amplifier for this mode of operation. It is possible to swamp a portion of the driving signal by means of low value noninductive resistors placed across the grid circuit of the linear amplifier, and by the use of a high-C grid tank circuit achieve the best possible drive-signal regulation. Both of these methods have been used with good results when properly applied.

In most instances, the increase in output power gained from shifting from Class AB-1 to Class AB-2 operation may not be justified, and more power output may be gained in an easier fashion by operating two tubes in parallel in the Class AB-1 mode. Driving requirements and bias supply regulation problems are thus simplified. These remarks apply equally well to Class B grid driven linear amplifiers, and their use should also be tempered with caution.

Class AB-2 and Class B Cathode Driven Service

High-u triode tubes may be used to advantage in cathode driven (so-called grounded grid) linear amplifier circuits (Figure 2). In this mode of operation, the grid is bypassed to ground and normal Class AB-2 or Class B bias is applied to the stage. No bias is necessary, of course, in the case of zero bias tubes, such as the 811A, 3-400Z and 3-1000Z, and the grid of the tube may be grounded directly. A high-u triode tube should be employed for this style of amplifier, and the inherent feedback of the grounded grid stage combined with the use of a tuned cathode input circuit insure a minimum of driver waveform distortion. Use of a high-u tube is suggested for two reasons: First, the inherent cathode-plate shielding of a high-u tube is better than that of a comparable low-u tube; and second, a high-u tube provides better gain per stage and requires less drive because of less feedthrough power. Tubes such as the 811A, 805, 75TH, 250TH, 450TH and 6C21 work well in this circuit. Resistive loading of the driving circuit is not required as long as there is sufficient "Q" in the cathode tank because of the constant "feed through" power load on the exciter. For most triode tubes, a tuned input circuit having a resonating capacitance C-1 of about 15 mmfd per meter of wavelength is sufficient.

Low-u tubes, on the other hand, require extremely large driving signals in the cathode driven mode, and stage gain is extremely small. Thus, the use of these tubes in this particular circuit configuration is not recommended.

In summary, then, triode tubes to be used for linear amplifier service should have a large plate dissipation, and the power output to be expected from the tube will run from approximately once to twice times the rated plate dissipation. Moreover, high-u triode tubes perform better in cathode driven, class B circuits. Medium-u tubes, falling in the shadowy region (u = 10 to 20) usually are easier to get working in cathode driven circuits, as a general rule.

Bias Supplies

Bias supplies for linear amplifier stages that draw grid current must be capable of good regulation so that the fixed bias does not vary as grid current increases. Shown in Figure 3 is a simple bias supply that will provide a regulated bias variable over the range of -20 to -80 volts. Regulation is 0.001 volt per milliampere of grid current. Between -30 and -80 volts the supply will regulate grid current up to 200 milliamperes. Between -20 and -30 volts, maximum grid current is restricted to 100 milliamperes. The adjustable resistor R-1 is set to produce about 20 milliamperes current through the first regulator tube.

A regulated supply for the -100 to -600 volt range is shown in Figure 4. The tap switch of this supply permits rough bias adjustment over the range, while the potentiometer permits a fine adjustment to be made. Maximum permissible grid current runs from about 100 milliamperes in the vicinity of -100 volts to about 25 milliamperes in the -600 volt region.

Operating Potentials for Linear Service

Once the circuit and class of service have been determined for the particular tube at hand, the proper operating potentials for the tube must be determined. Luckily, much of this data is at hand, although in a disguised form. Most data sheets provide Class AB-1 or AB-2 audio data, usually for push-pull operation. As the tube doesn't know if it is being driven by an audio signal or an r.f. signal, this data applies to a significant degree to r.f. linear amplifier service. For a single tube, it is only necessary to divide the indicated currents by two, as the currents are for two tubes. (Actually, only one tube in the push-pull audio service is "working" at one time, but the current meters register currents that are averaged for the two tubes). For example, Figure 5 provides operating data for a 304TL, grid driven, Class AB-1 r.f. linear amplifier, used in the circuit of Figure 1. A plate circuit Q of 15 should be used, and the grid current of the amplifier should be high-C (Q of 15 or better) in order to take care of accidental grid current peaks. Normal grid current is zero. In order to properly load the exciter, it is necessary to swamp the exciter output with a non-inductive load so that the exciter develops nearly full rated output when grid current just starts to flow in the linear stage. The load may be placed across the coaxial line between the exciter and amplifier and consist of a number of 2 watt composition resistors arranged in series-parallel to present a near-52 ohm load. Total wattage rating of the resistor bank should be equal to about one-half the P.E.P. output of the exciter.

Figure 6 provides operating data for a 304TH, cathode driven, Class AB-2 r.f. linear amplifier used in the circuit of Figure 2. At 3000 volts plate potential the 304TH is good for a P.E.P. input close to one kilowatt, and two tubes in parallel can provide the "so-called" two kilowatts P.E.P. Figure 7 provides data for the 450TH in cathode driven, Class AB-2 r.f. linear amplifier service.

It should be noted that steady state conditions are given, and P.E.P. operation should be held to these limits, unless the oscilloscope tells the operator that the tube may be "pushed" a bit before peak flattening or distortion occurs. Use of the oscilloscope should be tempered with caution, however, as it is nearly impossible to read distortion or peak flattening on the "scope" until the degree of intermodulation distortion approaches a level near -20 decibels below one tone of a two tone test signal. By this time, the operator will probably receive a brick through the shack window! To be on the safe side, then, meter readings of grid and plate currents should be one-half or less of those indicated in the tables for voice peaks.

FIGURE 5

304TL, GRID-DRIVEN, CLASS AB1 LINEAR AMPLIFIER

D.C. Plate Voltage	1500	2000	2500	3000	volts
D.C. Grid Voltage*	-118	-170	-230	-290	volts
Zero Signal D.C. Plate Current	135	100	80	65	ma
Single Tone Max. D.C. Plate Current	280	275	245	225	ma
Max. D.C. Input	420	550	615	675	watts
Max. Drive Power	0	0	0	0	watts
Plate Load Impedance	1270	2650	4250	6000	ohms
Max. Output	128	245	305	365	watts

FIGURE 6

304TH, CATHODE-DRIVEN, CLASS AB2 LINEAR AMPLIFIER

D.C. Plate Voltage	1500	2000	3000	volts
D.C. Grid Voltage*	-65	-90	-145	volts
Zero Signal D.C. Plate Current	130	100	75	ma
Single Tone Max. D.C. Plate Current	480	380	320	ma
Max. D.C. Input	720	760	960	watts
Max. Drive Power	70	55	60	watts
Cathode Input Impedance**	195	260	385	ohms
Plate Load Impedance	1850	3000	5500	ohms
Max. Output	510	530	715	watts

FIGURE 7

450TH, CATHODE-DRIVEN, CLASS AB2 LINEAR AMPLIFIER

D.C. Plate Voltage	1500	3000	4000	volts
D.C. Grid Voltage*	0	-50	-85	volts
Zero Signal D.C. Plate Current	50	200	150	ma
Single Tone Max. D.C. Plate Current	400	450	335	ma
Max. D.C. Input	600	1350	1340	watts
Max. Drive Power	70	105	70	watts
Cathode Input Impedance**	262	322	350	ohms
Plate Load Impedance	2200	4100	6400	ohms
Max. Output	416	992	1000	watts

NOTE: 1500 volt operation is zero bias service

*Adjust to give stated zero-signal plate current.

**Fundamental frequency component. High-C, tuned cathode tank should be employed to obtain lowest intermodulation distortion.

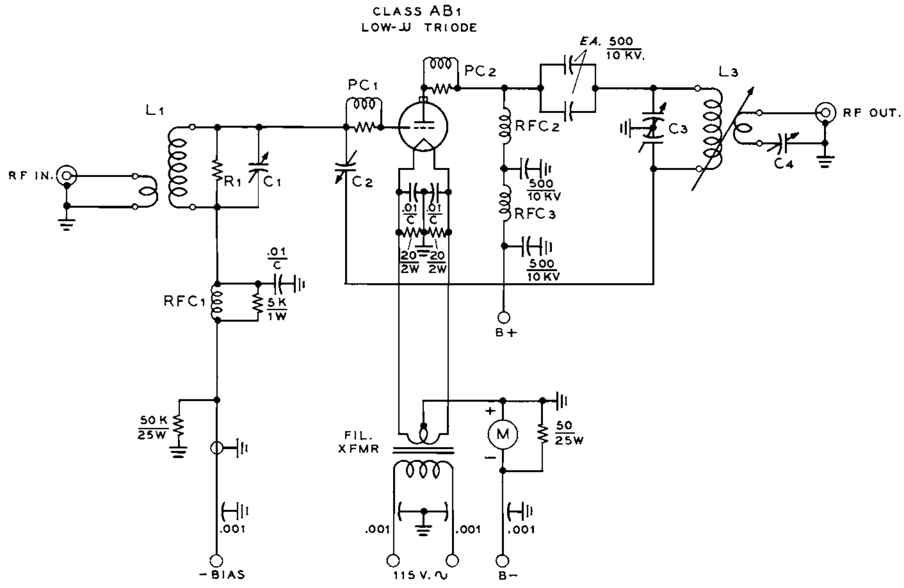


FIGURE 1

TYPICAL GRID DRIVEN, CLASS AB-1 TRIODE LINEAR AMPLIFIER CIRCUIT

C-1, L-1:	Grid input circuit (Q = 15). Ratio of L to C chosen so as to match impedance presented by R-1, the loading resistor.
C-2:	Neutralizing capacitor. Capacitance approximately twice the grid-plate capacitance of the triode tube. Breakdown voltage equal to three times the d.c. plate voltage.
C-3, L-3:	Plate output circuit (Q = 15). Ratio of L to C chosen to match r.f. load resistance of amplifier tube. Handbook charts for Class C Service may be used, provided single tone plate current is used in formulas.
C-4:	Series tuning capacitor for link circuit. Approximately 6 uufd per meter of wavelength (See Chapter 6, A.R.R.L. Handbook, or "Radio Handbook").
PC-1, PC-2:	Parasitic suppressors. Try two 100 ohm, 2 watt composition resistors in parallel, wound with 5 turns, ½-inch diameter #16 wire, spacewound.
RFC-1:	2-½ mH (National R-100). RFC-2: High voltage choke. B & W #800 or equivalent RFC-3: VHF choke. Ohmite Z-50 or equivalent.
Note:	20 ohm, 2 watt composition resistors placed across filament capacitors to "de-Q" bypass resonances.

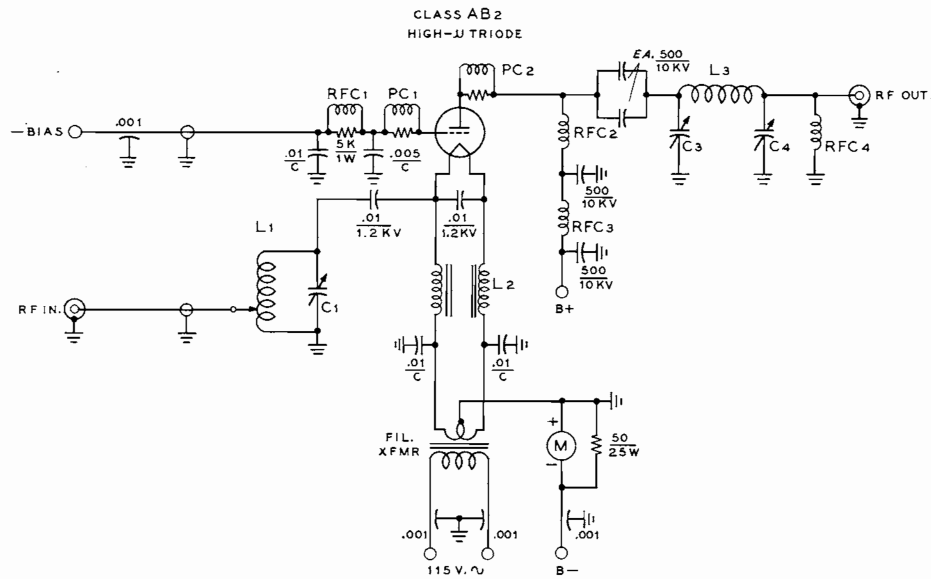


FIGURE 2

TYPICAL CATHODE DRIVEN, CLASS AB-2 TRIODE LINEAR AMPLIFIER CIRCUIT

C-1, L-1: Cathode input circuit(Q greater than 2). Resonates to operating frequency with approximately 20 uuf per meter of wavelength.

C-3, L-4, C-4: Pi-network output circuit (Q = 15). Ratio of L to C chosen to match r.f. load resistance of amplifier tube. For appropriate charts, see A.R.R.L. Handbook and "Radio Handbook". Use single tone plate current in formulas.

L-2: Filament choke. B & W or equivalent.

PC-1, PC-2: Same as Figure 1. RFC-1, RFC-2, RFC-3: Same as Figure 1.

RFC-4: 2½ mH (National R-100).

Notes: Adjust drive tap on L-1 for minimum SWR on coaxial line to exciter. Positive terminal of bias supply is grounded. Grid of tube may be grounded for Class B, zero bias operation with triodes having u of 200 or better.

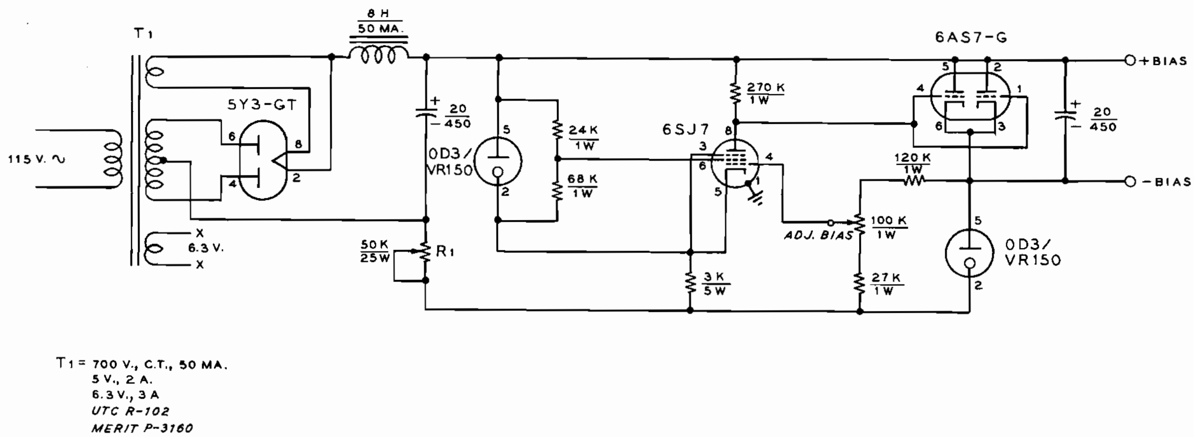


FIGURE 3

LOW VOLTAGE REGULATED BIAS SUPPLY

This simple supply provides a regulated bias voltage variable over the range of -20 to -80 volts. Regulation is 0.001 volt per milliampere. Between -30 and -80 volts, the supply will regulate grid current up to 200 ma. Below -30 volts, maximum grid current is restricted to 100 ma.

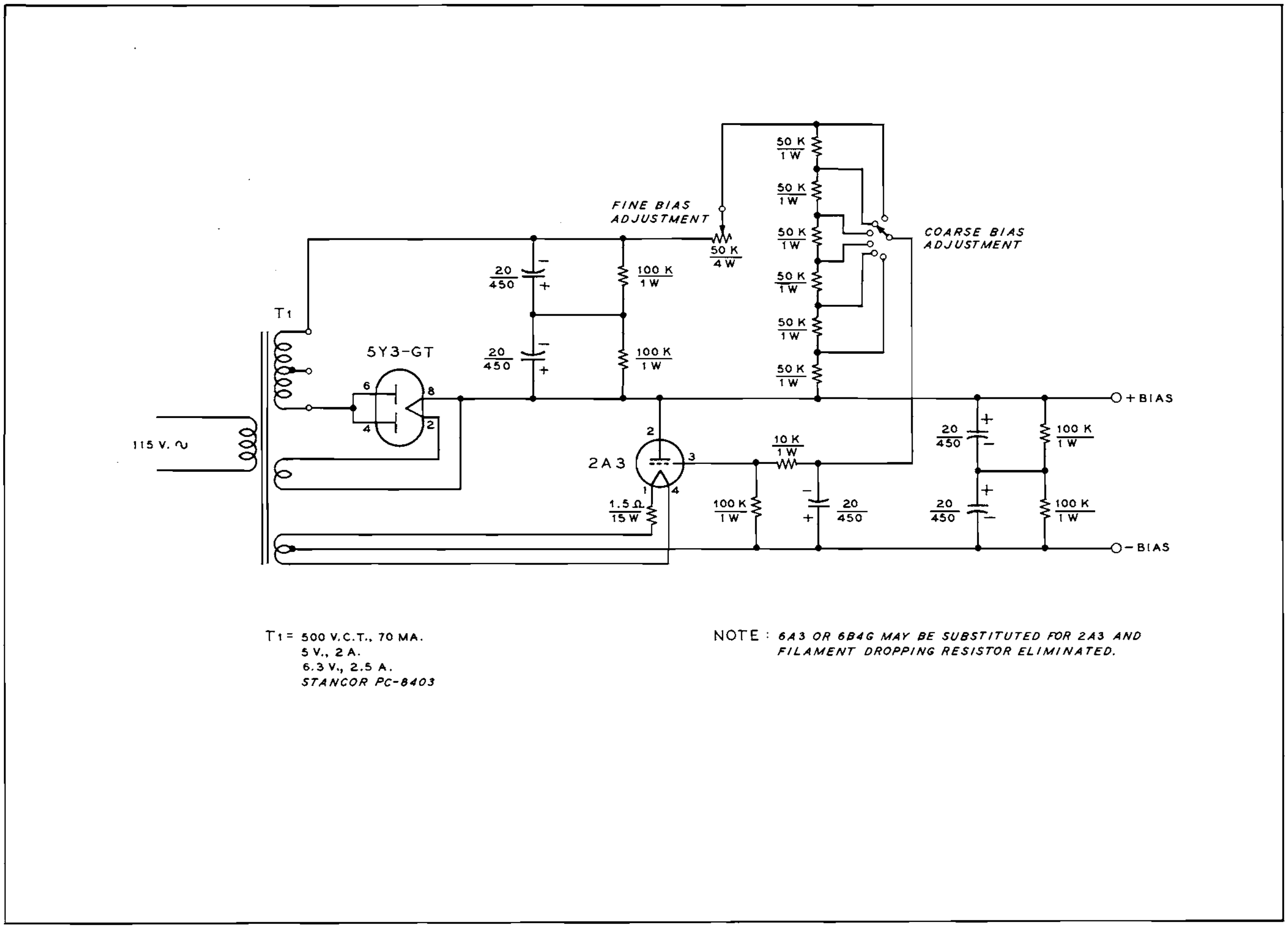


FIGURE 4

MEDIUM VOLTAGE REGULATED BIAS SUPPLY

This series regulated supply acts as a variable bleeder resistor which automatically adjusts its resistance to a value such that grid current flowing through it will develop a constant voltage across the supply terminals. The tap switch permits rough bias adjustment over the range of about -100 to -600 volts, while the potentiometer permits a fine adjustment to be made. Maximum permissible grid current is about 100 ma in the vicinity of -100 volts, dropping to about -25 ma. in the -600 volt region.

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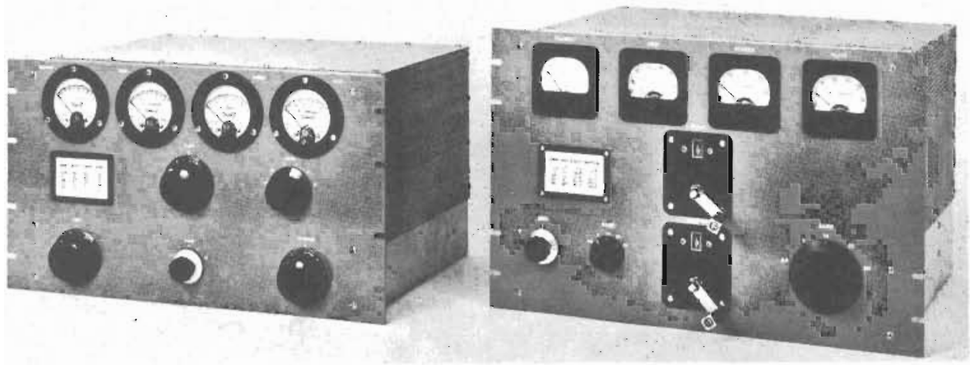
The Pi-L Plate Circuit in Kilowatt Amplifiers

BY RAYMOND F. RINAUDO,* W6KEV

Improved Harmonic Suppression for Multiband Systems



Two amplifiers built to the same general circuit design and layout. The one on the left uses a pair of 4-125As, that at the right a pair of 4-250As. Both are capable of a kilowatt input on c.w. The 4-250As can handle envelope peaks of 2 kilowatts on sideband.

The Pi-L Plate Circuit in Kilowatt Amplifiers

An extra L network tacked on the pi-network tank helps get rid of those harmonics that multiband antenna systems are only too capable of radiating. This article shows how the circuit was applied to two kilowatt amplifiers — both of which are worth your attention because of their construction ideas, too.

BY RAYMOND F. RINAUDO,* W6KEV

During the past ten years the pi network has become almost the standard plate tuning and loading circuit for a radio-frequency power amplifier, whether it ends up with a 6146 or a pair of 4-400s. This came about quite naturally when TVI became a problem because the pi network lends itself very nicely to band switching, with tuning and loading done with capacitors — the capacitors, plate switch and coil being located in a comparatively small shield enclosure. All of this was had along with reasonably good harmonic attenuation: second harmonic down 35 to 40 db. and higher harmonics further attenuated.

However, along with the popularity of the pi network, we have had the development of the three-band beam, the multiband dipole and the multiband vertical. While the multiband antenna is a god send to those with limited acreage, in which category the vast majority of us fall, it serves to bring up another problem because we now find that 35- or 40-db. attenuation of the second harmonic at the amplifier is quite often not enough. The multiband antenna is all too ready to radiate that 20-meter harmonic when the amateur is actually transmitting on 40. Obviously, the antenna under discussion is of the type which requires no tuner between the transmitter and the feed line. A solution to the problem is to put a filter in the transmission line which will pass only the frequencies in one band. But then, when changing to a different band, another filter must be substituted and some of the ease of band change has been lost.

Another way in which the situation can be improved is to use a pi-L network. The pi-L will give 10 to 15 db. more attenuation of the second harmonic than will the pi [1] and even more attenuation of the higher harmonics. This circuit has been used in some commercially-built amateur equipment such as the Collins KWS-1. Further improvement can be had by designing the amplifier plate circuit for a higher loaded Q. For example, raising the loaded Q from 10 to 20 will increase the harmonic attenuation by 6 db. Unfortunately, one runs into the law of diminishing returns here; the losses in the plate coil begin to be large enough to cause serious heating, and a loaded Q of 20 is near the practical upper limit in most cases.

* Eitel-McCullough, Inc., San Carlos, Calif.

[1] *Fundamentals of Single Side Band*, Collins Radio Company.

The 4-125A Amplifier

With the harmonic problem in mind, a design was worked out in late 1958 for an amplifier which was to replace the pi-network final then in use. The requirements were as follows:

1) Operation from 3.5 to 28 Mc., band switched.

2) Power input of 1 kw. with 2500 volts on the plate.

3) R.f. power output to feed into a 50-ohm coaxial load.

4) Standard 19-inch rack mounting with a minimum practical panel height.

5) Amplifier enclosed in a shield and incoming power leads bypassed for TVI.

6) Harmonic radiation via the feed line to be minimized.

7) A minimum amount of cash to be involved.

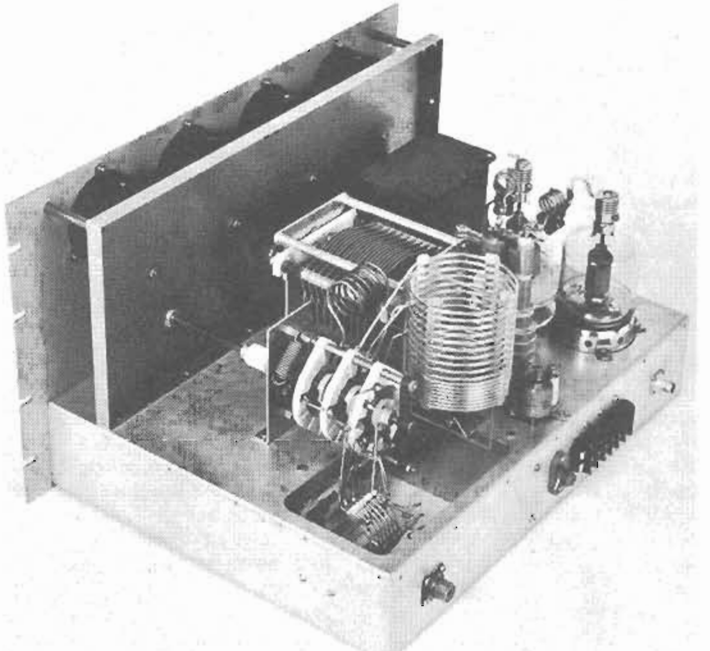
The result of the above design is the 4-125A amplifier shown in the photographs.

The amplifier uses a pair of neutralized 4-125As in parallel. The grid circuit is tuned, fairly high C, and makes provision for bridge neutralization via bypass condenser C_2, Fig. 1. The plate circuit is a pi-L network with an operating Q of 15, and plate current is shunt fed. Individual meters are used to measure grid, screen and plate currents and filament voltage.

The amplifier is built on a 13 × 17 × 3-inch aluminum chassis behind a 10½ × 19-inch panel. The meters are excluded from the r.f. field by a 7 × 17-inch aluminum sub-front panel which is set back two inches from the front panel. The resulting enclosure, which is above the chassis and screened by perforated aluminum, is 11 by 17 by 7 inches. The underside of the chassis is divided into two units by a shield running from front to back. The grid compartment is 10 by 13 by 3 inches and the output compartment containing the loading capacitors and L net coil is 7 by 13 by 3 inches. The bottom of the chassis is covered by perforated aluminum sheet to allow convection air currents to cool the tubes. No blowers or fans are needed to cool the 4-125As, provided that cooling air is allowed to flow freely past the tubes.

In keeping with requirement (7), maximum use was made of the surplus markets and trades with fellow hams, and the author's own junk boxes were given a thorough going over. No real compromise was made by the use of inferior components, but inevitably several of the parts used are either not too commonly-available surplus items or are once-standard parts which are no longer manufactured. But for each of these, a standard commercial part exists which is as good as or better than the one used and will fit into the space available. The standard part is the one given in the parts list. That the use of used and surplus parts paid off is attested to by the fact that the immediate cash outlay was less than $20! On the other hand, if the reader wants to build the amplifier using all new parts, the cost will be approximately $235, including tubes.

As mentioned previously, the grid circuit operates with fairly high C. Approximately 300 $\mu\mu$f. is used on 3.5 Mc., 150 $\mu\mu$f. on 7 Mc., and proportionally smaller amounts for the higher-frequency bands. A large tuning capacitance is used so that there will be a minimum of clipping of the waveform of the driving signal when the grid is driven positive. A distorted waveform at the grid of an amplifier will mean more harmonic



Chassis view of the 4-125A amplifier. The plate tuning capacitor is at the center. The pi coil for 3.5-21 Mc. is vertical. The 28-Mc. pi coil is mounted horizontally between the band switch and the tank capacitor. The plate r.f. choke and neutralizing capacitor are partially hidden by the plate coil. The filament transformer is at the far end of the chassis between the 4-125As and the sub-front panel.

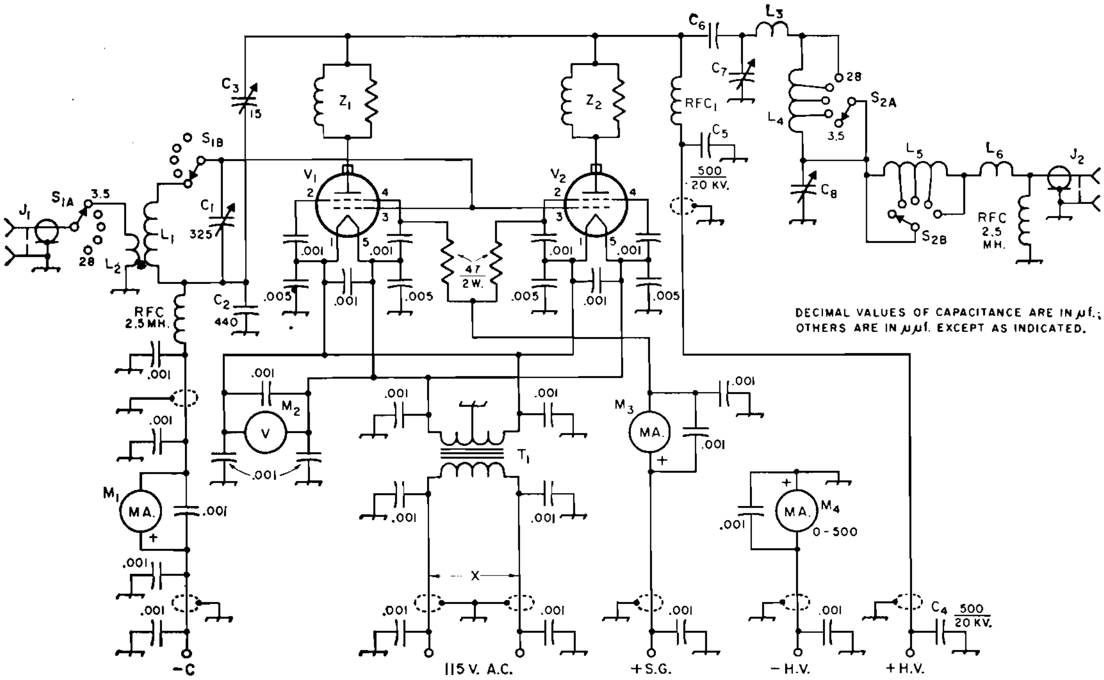
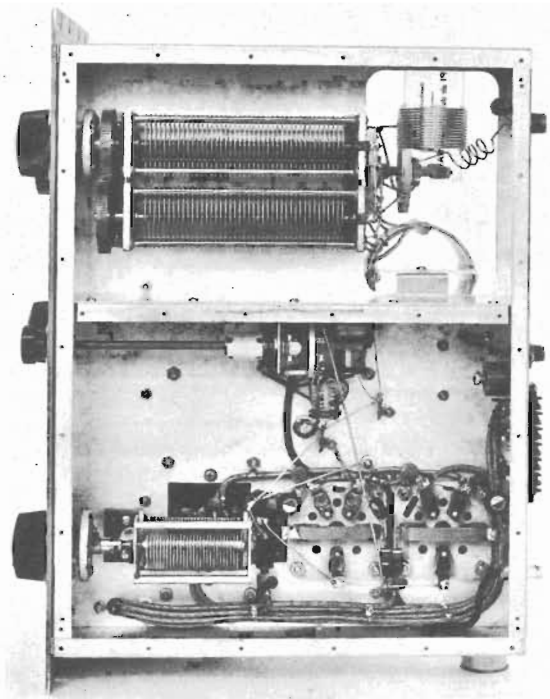


Fig. 1—The amplifier circuit. Either 4-125As or 4-250As may be used at V_1 and V_2. See specification below for circuit values that differ with the two types. "X" indicates point where the two cooling-fan motors are connected in the 4-250A amplifier. Shielded wiring in supply leads is continued up to the bypass capacitors nearest the r.f. circuit. All 0.001- and 0.005-μf. capacitors are disk ceramic, 1000-volt rating.

C_1—320 μμf., 0.0245-inch spacing (Hammarlund MC-325-M).
C_2—440 μμf., silver mica (two 220-μμf. in parallel).
C_3—Disk neutralizing, 2.2–15 μμf. (Millen 15011).
C_4, C_5—500-μμf., 20-kv. ceramic (Centralab TV-20).
J_1, J_2—Coaxial chassis-mounting connectors.
L_1—3.5 Mc.:32 turns no. 20, ¾-inch diam., 16 turns per inch (Air Dux 616T).
7 Mc.: 14 turns No. 20, ¾-inch diam., 16 t.p.i. (Air Dux 616T).
14 Mc.: 11 turns No. 18, ⅝-inch diam., 8 t.p.i. (Air Dux 508t).
21 Mc.: 9 turns No. 18, ½-inch diam., 8 t.p.i. (Air Dux 408T).
28 Mc.: 6 turns No. 18, ½-inch diam., 8 t.p.i. (Air Dux 408T).
L_2—3.5 Mc.: 4 turns insulated hookup wire at cold end of L_1.
7 Mc.: 3 turns same.
14, 21, and 28 Mc.: 2 turns same.
M_2—0–8 or 0–10 volts a.c.
M_4—0–500 ma. d.c.
RFC_1—Transmitting choke (B & W 800, National R-175A, Raypar RL-100).
S_2—Ceramic, 2 poles, 5 positions (Radio Switch Corp., Marlboro, N.J.); see text.
V_1, V_2—4-125A or 4-250A.
Z_1, Z_2—4 turns No. 12, ½-inch diam., ½ inch long, with four 220-ohm, 2-watt composition resistors in parallel.

	For 4-125As:	For 4-250As:
C_6—	0.001-μf., 20-kv. ceramic (two Centralab TV-20s in parallel).	0.002-μf., 20-kv. ceramic (four Centralab TV-20s in parallel).
C_7—	250-μμf., 3000-volt variable (Johnson 154-9).	300-μμf., 10-kv. variable (Jennings UCS-300).
C_8—	0.001-μf., 2000-volt variable (two Johnson 154-3 in parallel, ganged).	0.0012μf., 3000-volt variable (Jennings UCSL-1200).
L_3—	6 turns No. 10, 1-inch diam., 1½ inches long.	See text.
L_4—	Vari-pitch Air Dux 2408D4, modified as described in text.	Illumitronic Pi Dux No. 195-2, modified as described in text.
L_5—	Indented Pi Dux 1411A, modified as described in text.	Vari-pitch Air Dux 1608D6, tapped as described in text.
L_6—	4 turns No. 14, ½-inch diam., 1¼ inches long.	6 turns No. 12, 1-inch diam., 1 inch long.
M_1—	0–50 ma. d.c.	0–100 ma. d.c.
M_3—	0–100 ma. d.c.	0–200 ma. d.c.
S_1—	1 section, 2 poles, 5 positions (Centralab 2505).	2 sections, 1 pole per section, 11 positions, 5 positions used (Centralab YD sections with P-270 index assembly); see text.
T_1—	5 volts, 13 amp. (Triad F9A or F15U).	5 volts, 29 amp. (Stancor P-6492).



Below the 4-125A chassis. The two loading capacitors at the top, ganged together by means of gears, are separated from the grid circuit by an aluminum shield running from the front to the rear of the chassis. The L net coils are directly behind the loading capacitors. The grid band switch and coil are at about the center of the chassis. The grid tuning capacitor is mounted off the chassis by means of bakelite blocks and is directly under the filament transformer.

signal in the plate circuit and, hence, a more difficult job to suppress it. For example, in Class C, B or AB_2 operation, during the portion of the cycle that the grids of two 4-125As in parallel are driven positive, the grids look like a resistor of about 1200 ohms to the tuned grid circuit, and that portion of the cycle will be distorted unless precautions are taken to prevent it. Waveform clipping is minimized by using plenty of tuning capacitance in the grid circuit. Of course, for AB_1 operation, the amount of C is not important because the grid is never driven positive and looks like an infinite resistance to the tuned circuit.

The grid tank circuit uses individual coils for each band. A link coil of insulated hookup wire is wound over the cold end of each coil. The hot ends of the coils and the links are switched by S_1, a 2-section, 5-position switch having one wa'er. This switch is mounted on the underside of the chassis by means of an aluminum bracket. The coils are mounted between the appropriate switch terminal and a tie point and are oriented so that there is a minimum of coupling between them.

The tube sockets are mounted on the underside of the chassis, and spring clips on the top of the chassis held by the socket mounting bolts ground the metal tube base shield. Bypassing of the screen and filament terminals is done in the more-or-less standard way. The screen terminals on each socket are connected together by a ⅜-inch wide strip of thin copper. Each screen terminal is then bypassed to the nearest filament terminal with a disk ceramic capacitor. A disk ceramic capacitor is connected between the filament terminals and another is used to bypass one side of the filament to ground. The remaining filament terminal is grounded with a short, heavy lead. Grounding one side of the filament has been found to be helpful in eliminating v.h.f. parasitics. Those who expect to use the amplifier for linear service should use bypass capacitors to ground on both filament terminals and ground the filament transformer center tap as shown in Fig. 1. A slightly cleaner signal will result. A 47-ohm, 2-watt carbon resistor is used to feed screen voltage to the screen terminal of each socket and is a parasitic preventive measure.

The pi-L plate tank circuit is made up of individually available coils, capacitors and switch. The switch, which is mounted on the chassis with an aluminum bracket, is made by Radio Switch Corporation of Marlboro, New Jersey. The switch used was bought on the surplus market and has three wafers, each wafer with six contacts. Since the wafers were already there it was decided to make use of them by paralleling the contacts on two wafers and using the parallel combination to switch the coil in the pi portion of the network. The circulating current in the pi coil is about twice as high as that in the L coil. However, the current rating of the switch is 20 amperes, so a single section is all that is really needed to handle the pi coil switching. Also, because six contacts per wafer were available, the sixth contact was used to provide a 3.8-Mc. position; that this is not necessary can be seen by the later description of the 4-250A amplifier. If the builder wishes to retain the 6-position band switch, he should order a Model 86 switch, standard bearing, non-shorting, 30-degree detent, with two Type A wafers. If a 5-position band switch will do, then the builder should order a Model 86 switch, standard bearing, non-shorting, 30-degree detent, with one Type B wafer. The second switch, by virture of having only one wafer, will cost about three dollars less. The coils used in both the pi and the L are home-brew for 28 Mc. Illumitronic Engineering Pi Dux coils are switched in for the lower-frequency bands. It is of interest to note that as originally built, the pi coil was a Pi Dux 2007A, which is wound with No. 12 wire. After a bit more than two years' use, two of the turns shorted because coil heating had softened the polystyrene insulating supports. The damaged coil was replaced with a Pi Dux 2408D4, which is made of No. 10 wire. An r.f. choke completes the output circuit to ground for d.c. as a safety precaution.

Drive power is fed into the amplifier through a BNC coaxial receptacle and the output power is taken out by means of a U.H.F. receptacle.

Plate voltage is fed in through a Millen 37001 high-voltage terminal.

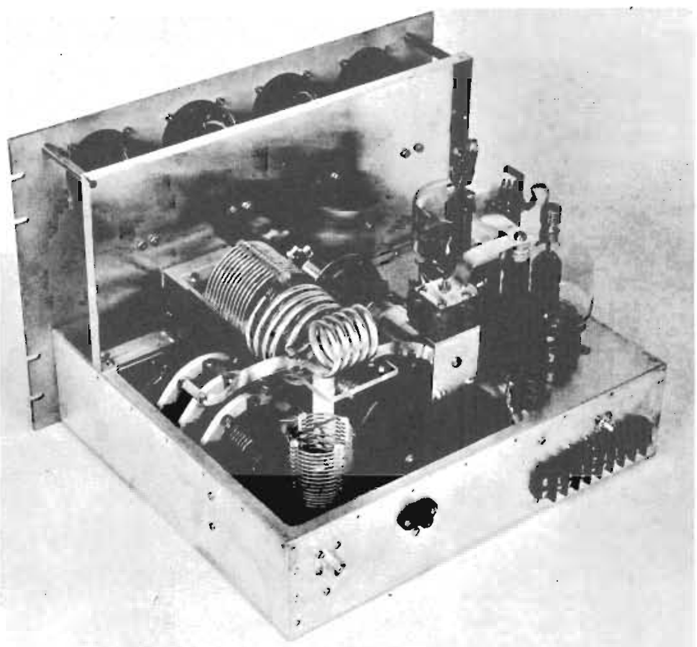
When the amplifier was completed, it was first tested for parasitics without suppressors of any kind. As is almost always the case with a tetrode or pentode amplifier, it oscillated merrily in the v.h.f. range — at about 150 Mc., as a check with the grid-dip meter showed. The parasitic was killed by the installation of suppressors, Z_1 and Z_2, in the plate lead to each tube. The test for parasitics is to operate the amplifier with reduced plate and screen voltage and no fixed bias on the control grid, but using a grid leak of about 5000 ohms to ground to develop bias if the amplifier breaks into oscillation. No drive is used and no load is connected to the output. With this amplifier the plate voltage was set at 1000 volts and the screen voltage increased until the plate current was about 200 ma. and the tubes were dissipating about 100 watts each. At this point the screen voltage was between 150 and 200 volts. If an amplifier can be operated in this manner with no current showing on the grid meter, with no change in plate current, and with no detectable r.f. in the amplifier as the grid and plate tuning, loading and band-switch controls are tuned through their full range, then the amplifier can be considered adequately stable. This is a much more severe test than the one often made where full plate and screen voltages are applied and bias is reduced until the tube or tubes are dissipating full rated power with no excitation.

With the components used, the amplifier will operate with up to 3000 plate volts in Class C c.w. or Class AB_1 linear, or up to 1500 volts for Class C a.m. plate-modulated service. Screen voltage for Class C c.w. or plate-modulated a.m. is 350 volts; for AB_1 linear it should be 600 volts. Grid bias should be -100 to -150 volts for Class C c.w., -210 for Class C, plate modulated, and approximately -95 volts for Class AB_1. The exact value of bias for AB_1 should be adjusted for the required idling plate current for the voltage used. Recommended values are as follows, for two tubes: 2000 volts, 85 ma.; 2500 volts, 70 ma.; 3000 volts, 60 ma.

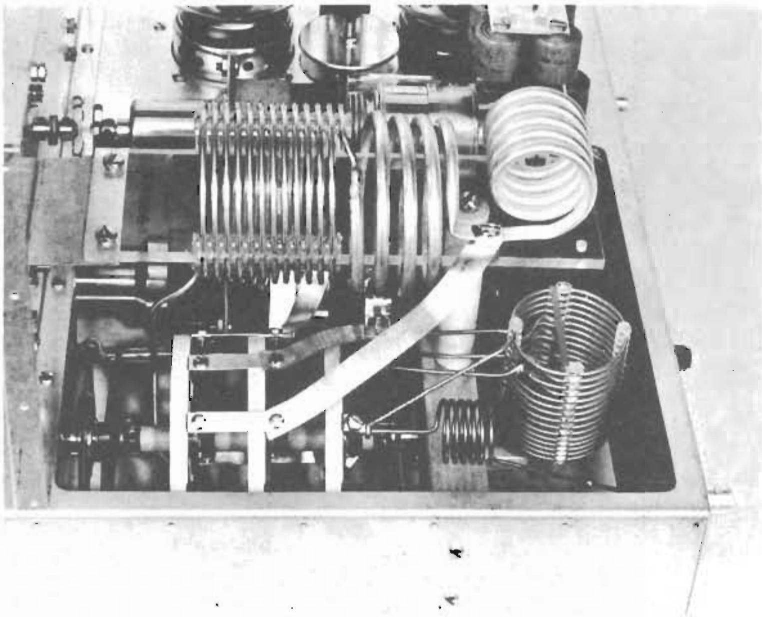
The screen voltage for Class C operation should come from a separate supply of reasonably good regulation. A series dropping resistor from the plate supply is not recommended. For Class AB_1 service, the screen voltage should be well regulated. While an electronically-regulated supply can be used, the simplest method is to use a string of VR tubes in series with a resistor from the plate supply. The reason that Class AB_1 permits this simple method of getting screen voltage is that the screen current excursions are not very great and are well within the capability of VR tubes.

Both the plate tank coil, L_4, and the L-network coil, L_5, are mounted on the chassis by means of aluminum angles bolted to the plastic mounting strip furnished with each coil. L_4 is modified and tapped as follows: Turns are removed from the close-wound end until 18 turns remain. Starting from the end of the coil which has the wide-spaced turns, the 21-Mc. tap is at 2 turns, the 14-Mc. tap at the 4th turn, the 7-Mc. tap at the 8th turn, and the 3.8-Mc. tap (if used) at the 16th turn. Since the 16th turn is in the close-wound portion of the coil, it is much easier to make the tap if the turn on each side is pushed in toward the center of the coil.

The L-network coil, L_5, is modified by removing turns until 13 turns remain. Starting from



In the 4-250A amplifier the band switch is in a chassis cut-out with the pi coil above it. The 28-Mc. L coil and the vertically-mounted 3.5-21 Mc. L coil are between the switch and the rear of the chassis. The plate blocking capacitors are mounted on a bracket held by the vacuum variable plate-tuning capacitor at the center. The plate r.f. choke and its bypass capacitor are beside the 4-250A nearest the rear of the chassis.



Close-up of the 4-250A pi-L plate tank coils and band switch. The horizontally-mounted 6-turn wire coil is L_6, and the vertically-mounted coil is L_5.

the L_6 end, taps are placed as follows: 21-Mc. tap at 2 turns, 14-Mc. tap at 5 turns, 7-Mc. tap at 7 turns, 3.8-Mc. tap (if used) at 12 turns.

Non-standard items used are the grid tuning capacitor from the surplus market, the neutralizing capacitor, which National Radio Company no longer makes, the plate tuning capacitor, no longer made by E. F. Johnson, and the loading capacitor, C_8, which is made up of two capacitors taken from a surplus BC-653 transmitter. The two E. F. Johnson units specified for C_8 will simplify the ganging of the two because they have the shaft out the back as well as the front. The two surplus capacitors did not have this feature and, consequently, gears had to be used for ganging.

The tuning and loading adjustments of the pi-L plate circuit are exactly the same as with a pi network. Plate circuit loading is increased by reducing the capacitance of C_8. Whenever the loading capacitance is changed, the plate circuit must be retuned to resonance with the plate tuning capacitor, C_7.

When the amplifier is first tested, it should be neutralized. The neutralizing capacitor, C_3, is adjusted so that there is about one-half inch spacing between the two plates; then, with plate and screen voltages off and a load connected, excitation is applied and the grid circuit is tuned to resonance. The excitation level is set so that the grid current is only a few ma. Then plate and screen voltages are applied and the plate circuit is tuned to resonance. Plate-circuit resonance is best indicated by the peaking of the screen-grid current as the plate tuning capacitor is tuned through resonance. The loading control is adjusted so that the screen current is about 60 ma. If the plate input is less than desired, increase the grid drive and plate loading until the correct plate current is flowing with screen current at 60 ma. The plate circuit must be retuned to resonance with each change of loading.

The check for neutralization is to tune the plate circuit through resonance, observing both screen and grid currents. When the amplifier is correctly neutralized, the grid-current meter will show a small current peak at the same setting of the plate tuning capacitor that gives a peak in screen current. Neutralization should be done on the 21-Mc. band.

After the amplifier has been neutralized it should be checked for parasitic oscillations, using the procedure given previously. In some cases, parasitics will make it difficult to find the correct neutralization setting. But if construction details are followed, particularly those pertaining to bypassing and the installation of suppressors, parasitic oscillations should not be a problem.

The 4-250A Amplifier

Quite some time after the 4-125A amplifier had been completed and had been operating satisfactorily, a design for a de luxe version was worked out. In this case, the requirements were the same as before except that the rig had to be capable of 2-kw. p.e.p. input for sideband service, and all the parts used were to be currently-available new items. The result is the 4-250A amplifier shown in the photographs.

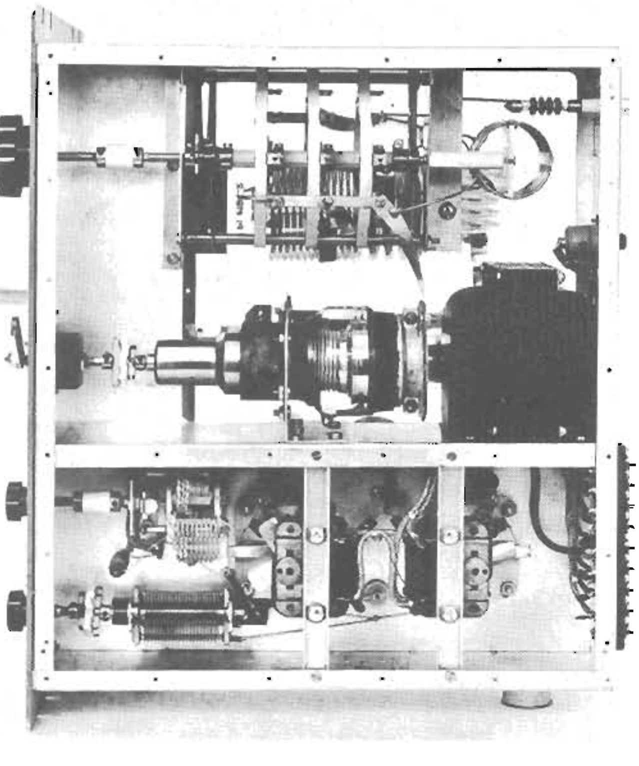
The 4-250A amplifier uses essentially the same circuit as the 4-125A version. However, the plate circuit was designed for an operating Q of 18 instead of 15, in order to take advantage of the heavy-duty plate coil and switch which were to be used. An examination of the photographs shows the similarity of the two rigs in the mechanical layouts and the method of making the shield enclosures. Because all new parts were used, the second amplifier turned out to have a better appearance both inside and out than did the first one.

The 4-250A amplifier is built behind a standard 19-inch rack panel 12¼ inches high. The chassis is 17 by 15 by 4 inches and the shield enclosure above the chassis is 17 by 12 by 7¾ inches. The vertical sub-panel is set back three inches from the front panel. The grid-circuit compartment is 6¼ by 15 by 4 inches and is separated from the rest of the under-chassis space by a shield which runs from front to back. The remainder of the underside of the chassis is opened up to the upper compartment by cutting out that portion of the chassis top. This increases the available space for the plate-circuit components and makes it much easier to connect the various parts together. The filament transformer is mounted in the plate-circuit area on the under-chassis shield which forms the grid compartment. Both the shield enclosure above the chassis and the bottom cover are made of perforated aluminum, which allows convection currents to help keep the tubes and parts cool. The 4-250As require forced-air cooling of the base, and small Barber-Coleman fans are used to blow air directly upward at the tube base pins and through the holes in the tube socket and tube base.

The grid circuit coil turret is made up of commercial coil stock and rotary switches. Two rotary switch wafers are used where only one is really needed to do the necessary switching. The coils are mounted between the wafers on the switch lugs and the axes of all the coils are parallel. The arrangement used with the 4-125As where coils for adjacent bands are at right angles is better, because odd resonances in unused coils are less likely to cause trouble. However, this arrangement has been perfectly satisfactory in performance and is more rugged mechanically.

The cooling-fan motors are mounted on home-brew shock mounts to reduce noise. Rubber grommets with the same spacing as the motor mounting studs are mounted in the support channels which hold the motors, then a sleeve of length equal to the thickness of the grommet is slipped into each grommet. A large washer is placed on each side of the sleeve before the mounting screw is passed through and threaded into the motor mounting stud. Two shock mounts are needed for each motor. The grommet size used is that which fits into a ⅜-inch hole. A ¼-inch diameter sleeve ¼ inch long is just the right size to fit the grommet hole. There is no reason, though, why larger grommets and sleeves cannot be used.

Vacuum variable capacitors are used for plate tuning and loading. These require 24 and 30 turns, respectively, to cover the full capacitance range. Counter dials which read each tenth of a turn are used to drive them. The dials are made by Gates Radio Company, Quincy, Illinois, part No. M3401F. These were chosen because they are r.f. tight and do not require much space behind the panel.



Bottom view of the 4-250A amplifier. The plate band switch, at the top, is mounted on aluminum brackets. The vacuum variable loading capacitor is at the center and the filament transformer is between it and the rear of the chassis. The bracket which supports the loading capacitor also supports the plate tuning capacitor. The grid band-switching turret and tuning capacitor are at the front of the grid compartment. A cooling fan is mounted directly below each tube socket.

The plate-circuit switch is made by Radio Switch Corporation. It is a Model 88 with 36-degree detent and three Type A wafers. Two of the three wafers are paralleled and switch the pi coil. The remaining one handles the L coil.

The Illumitronic coils used in the plate circuit both require modification. The 28-Mc. pi coil should be removed and replaced with one of slightly greater inductance consisting of 5 turns of 3/16-inch copper tubing, 1⅝ inches in diameter and 2 inches long. The remainder, L_4, of the pi coil should be modified by removing turns from the wire end, leaving 12½ turns. Turns are removed from the close-wound end of the L coil, L_5, until 15 turns remain. The 28-Mc. L coil, L_6, is home-brew. The taps on the pi coil are placed as follows: 28 Mc.: junction of L_3 and L_4; 21 Mc.: 2¾ turns from the 28-Mc. tap; 14 Mc.: 5¾ turns from the 28-Mc. taps; 7 Mc.: 9¾ turns from the 28-Mc. tap. The taps on L_5 are as follows: 28 Mc.: at junction of L_5 and L_6; 21 Mc.: 3 turns from the 28-Mc. tap; 14 Mc.: 5 turns from the 28-Mc. tap; 7 Mc.: 9 turns from the 28-Mc. tap. An r.f. choke is used to complete the d.c. circuit to ground at the coax output connector as a safety measure should the plate blocking capacitor, C_6, break down.

A type BNC receptacle is used to feed drive power into the amplifier and a type C receptacle at the output. The d.c. plate voltage is fed into the amplifier via a Millen high-voltage terminal, type 37001.

Many combinations of plate, screen and bias voltages can be used, as a look at a tube-data sheet will show. The following voltages are typical:

	C.W.	*A.M. Phone*	*AB_1 Linear*	
Plate	2500	2500	3500	volts
Screen	500	400	555	volts
Grid	−150	−200	−105*	volts

* Set to give 45-ma. plate current per tube with no drive power.

The tune-up procedures are the same as for the 4-125A. Also, the amplifier should be checked for parasitics as described previously. Best linearity is achieved by increasing the loading on the amplifier until the power output just starts to fall off; during this adjustment, the drive power is held constant.

In operation, there is little to choose between the two rigs for the c.w. man. At 1-kw. input on c.w., the amplifiers handle identically; however, the 4-250As are easier to drive. For a 2-kw. p.e.p. input on s.s.b., the 4-250A amplifier stands alone. Which version the builder chooses depends upon his requirements as balanced against the necessary cash outlay. It should be pointed out that a third version combining the better or less-expensive components of the two designs presented could be built around 4-250As and result in an amplifier not costing much more than the strictly economy 4-125As.

Design of the Pi-L Network

The design of the pi-L tank circuits has been covered before in excellent articles presented in *QST*.[2,3] However, two different approaches are again presented here for those who would like to apply the circuit to transmitters of their own design.

The first method is to use values of components for the pi network with which the builder is already familiar and alter them suitably to make the pi-L work. Figs. 2A and 2B show a pi and a pi-L network, either of which will match a power amplifier tube to a 50-ohm load.

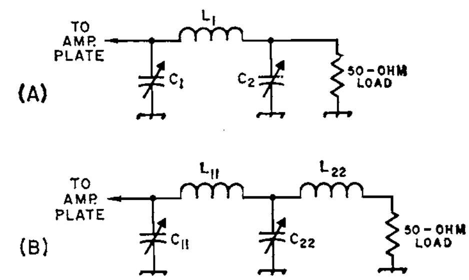


Fig 2

First, the capacitance and voltage rating of C_1 and C_{11} are exactly the same for both circuits. The capacitance of C_{22} will be about one-half to two-thirds that required for the pi capacitor, C_2. The voltage rating of C_{22} must be three or four times that required for C_2. The inductance L_{11} will be greater than L_1 by about 25 per cent. The inductance L_{22}, which has no direct counterpart in the pi, will have an inductance of about one-third to one-half of L_{11}. The circulating currents in L_{11} are the same as in L_1; therefore, a coil made of a wire size suitable for a pi net will also be good for a pi-L. The currents flowing in L_{22} are much smaller than those in L_{11}, so it can be made of smaller wire. For example, if L_{11} must be made of No. 10 wire, L_{22} could be made of No. 14 or 16.

This approach will allow the intrepid experimenter to convert his present pi-network output circuit to a pi-L without much pain. But for those who prefer a more formal method, the following is offered:

Just as the pi is designed as two L networks placed back-to-back, the pi-L is designed as three L nets placed back-to-back. In Fig. 3, a pi-L tank circuit is broken down into its

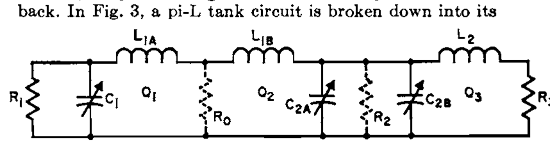


Fig. 3

three equivalent Ls. The first L matches the desired tube load resistance R_1 to a resistance R_0 and is composed of C_1 and L_{1A}. The second L matches R_0 to the resistance R_2 and is made up of L_{1B} and C_{2A}. The third L matches R_2 to the load R_3 (the transmission line) and consists of C_{2B} and L_2. R_1 is determined from the approximate formula:

$$R_1 = \frac{E}{2 \times I}$$

where E = plate voltage applied to the tube
and I = plate current in amperes.

First, the value of Q_1 is selected. Q_1 is the operating Q of the plate circuit and is usually chosen to be between 10 and 20. Knowing R_1 and Q_1 the capacitive reactance X_{C1} of the plate tuning capacitor C_1 is calculated from:

$$X_{C1} = \frac{R_1}{Q_1}$$

Also, calculate R_0 from:

$$R_0 = \frac{R_1}{Q_1^2 + 1}$$

Then, calculate the inductive reactance X_{L1A} of the inductance L_{1A} from:

$$X_{L1A} = R_0 Q_1$$

This completes the calculation of the reactances for the first L network.

[2] Miedke, "Pi and Pi-L Design Curves," *QST*, November, 1955.

[3] Grammer, "Simplified Design of Impedance-Matching Networks," *QST*, March, April and May, 1957.

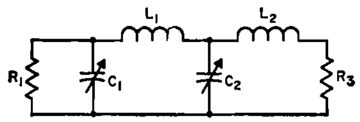


Fig 4

Before proceeding with the second L network, which consists of L_{1B} and C_{2A}, the value of R_2 should be selected (R_2 must always be greater than R_0 and R_3). Although it is possible to arrive at operating values for L_{1B} and C_{2A} by first selecting Q_2 (the Q of the second L network), it is best, from an equipment designer's viewpoint, to calculate R_2 to match the voltage capability of available tuning capacitors. This is done from:

$$R_2 = \frac{E^2}{P}$$

where E = r.m.s. voltage across R_2
P = Amplifier power output in watts.

Because the peak voltage must be considered when determining capacitor voltage breakdown (peak voltage equals 1.41 times r.m.s. voltage) and some safety factor is desired, it is best to let E equal one-half the capacitor breakdown voltage. For a kilowatt transmitter, it is suggested that 1000- to 2000-volt capacitors be considered. Convenient values of power output can be calculated by assuming an efficiency of 75 per cent for a c.w. or plate-modulated amplifier, and 60 per cent for a linear. Don't forget that for an a.m. phone rig, the power output at the crest of a 100-percent-modulated envelope is four times the carrier output.

Having calculated R_2, proceed with determining Q_2 from:

$$Q_2 = \sqrt{\frac{R_2}{R_0} - 1}$$

Calculate X_{C2A} (capacitive reactance of C_{2A}) from:

$$X_{C2A} = \frac{R_2}{Q_2}$$

Calculate X_{L1B} (inductive reactance of L_{1B}) from:

$$X_{L1B} = R_0 Q_2$$

Next, the capacitive and inductive reactances for the third L network, C_{2B} and L_2, are calculated. First, calculate Q_3, the Q of the third L net, from:

$$Q_3 = \sqrt{\frac{R_2}{R_3} - 1}$$

where R_3 is the load that the amplifier will be working into, usually 50 ohms for coax feed lines. It can be almost anything else but must be less than R_2. Then determine X_{C2B}, the capacitive reactance of C_{2B}, from:

$$X_{C2B} = \frac{R_2}{Q_3}$$

Then calculate X_{L2}, the inductive reactance of L_2, from:

$$X_{L2} = R_3 Q_3$$

Since the inductances L_{1A} and L_{1B} are in series, these are combined in one coil, L_1. The inductive reactance is equal to the sum of the separate parts

$$X_{L1} = X_{L1A} + X_{L1B}$$

Similarly, the two capacitors C_{2A} and C_{2B} are in parallel and are combined in one capacitor, C_2. X_{C2}, the capacitive reactance of C_2, is obtained by

$$X_{C2} = \frac{X_{C2A} X_{C2B}}{X_{C2A} + X_{C2B}}$$

The actual values for the capacitors and coils can be determined for any frequency from:

$$C = \frac{10^6}{2\pi f X_C}$$

and $$L = \frac{X_L}{2\pi f}$$

where C = Capacitance in $\mu\mu$f.
L = Inductance in μh.
f = Frequency in Mc.

The complete pi-L network with the combined inductances and capacitors is shown in Fig. 4. QST

Reprinted from July 1962, *QST*

amateur service newsletter
W6SAI

A 2 KILOWATT P.E.P. LINEAR AMPLIFIER FOR SINGLE SIDEBAND SERVICE

The new Eimac 3-1000Z is an air cooled power triode designed for zero bias, class B, r.f. and audio service. It is particularly well suited for single sideband linear amplifier service, delivering 1360 watts of single tone power at a plate potential of 3000 volts. At this power level intermodulation distortion products are 30 db or more below peak power level. At a plate potential of 2500 volts, two kilowatts p.e.p. input may run, with intermodulation distortion products 35 decibels or more below peak power level.

Shown in the drawing is a schematic of an all-band (3.5-29.7 Mc) grounded grid amplifier designed around the 3-1000Z. A tuned cathode circuit is employed to achieve minimum intermodulation distortion and ease of drive*, and a bandswitching Pi-L network is used in the plate output circuit to obtain maximum harmonic reduction and efficient power transfer to low impedance coaxial lines. Use of the Pi-L configuration reducds second harmonic transmission by over 15 db compared to a simple pi-network and permits the use of a less expensive network loading capacitor.

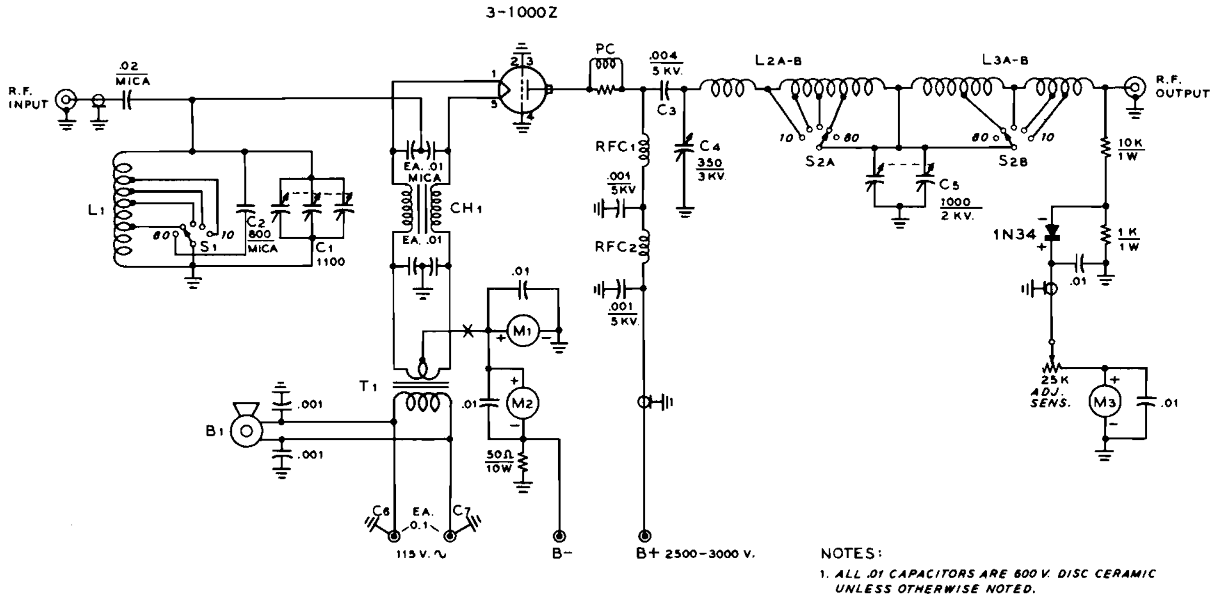
The amplifier may be driven to full output by any single sideband exciter having a p.e.p. output of 65 watts or better. Drive is monitored by a grid current meter (M-1) placed in the grid-cathode return circuit. The grid terminals of the 3-1000Z are grounded directly to the metal chassis for maximum circuit isolation. The plate current meter (M-2) is placed in the B-minus lead to the power supply. A diode voltmeter (M-3) indicates relative r.f. output, and may be used for tuning purposes.

The amplifier can be constructed within a cabinet measuring 10½" high, 15" wide and 14" deep. Alternatively, a standard 14" x 17" x 4" aluminum chassis may be used. The tuned input circuit (L1, C1, C2) and filament choke (CH1) are mounted below deck, and the plate circuit components are placed atop the chassis. The chassis is pressurized by means of a bottom plate. Air is forced into the chassis by a "squirrel cage" blower and then exhausted through the air system socket. Plate circuit components are enclosed in a TVI-proof screen made of perforated aluminum sheet. The meters should be removed from the enclosure or, if placed within the enclosure, are shielded from the r.f. field by means of a metal cap placed over the rear of the meter.

* "The Grounded Grid Linear Amplifier," QST, August, 1961, page 16.

At a plate potential of 2500 volts, the amplifier is loaded under single tone conditions to a plate current of approximately 800 ma. Grid current should be about 250 ma. A ratio of 3.2 milliamperes plate current to one milliampere of grid current indicates proper loading. Finally, the antenna should be overcoupled slightly at this point so as to drop power output about two-percent to achieve a condition of maximum linearity. Under voice conditions, peak plate meter current will be about 400 ma., and peak grid meter current will be about 125 ma. for a p.e.p. input level of 2 kilowatts (plate potential of 2500 volts). To withstand the instantaneous plate current peaks, a minimum of 20 microfarads capacitance in the plate power supply is recommended.

Figure 1. Schematic of 3-1000Z Amplifier



Parts List:

B1--20 cu. feet/min. at socket. Dayton #1C-180.

C1--Three gang b.c. capacitor with sections in parallel. J. W. Miller #2113.

C2--800 $\mu\mu$fd mica. Use .0005 $\mu\mu$fd and .0003 $\mu\mu$fd in parallel. Cornell

C2--800 $\mu\mu$fd mica. Use .0005 $\mu\mu$fd and .0003 $\mu\mu$fd in parallel. Cornell-Dubilier type 4, or type 9. 1200 volt d.c. test.

C3--Four .001 μfd, 5KV Centralab type 858-S capacitors in parallel.

C4--350 $\mu\mu$fd. 3 KV. Johnson #154-10 (type 350E30).

C5--Two 500 $\mu\mu$fd, 2KV in parallel. Johnson #154-3 (type 500E20). Capacitors are driven together by gears, or may be ganged with metal shaft coupling.

C6-7--0.1 μfd, 600 volt Sprague "Hypass" capacitor, type 80P3.

CH1--Four windings. 50 turns each #10 e. Two layers, bifilar wound on ½-inch diam. ferrite rod. Winding length is 4 inches. Rod is 4-3/8" long. Lafayette #M-333. Lafayette Radio Co., New York. A Barker & Williamson type FC-30 choke may be substituted. Windings paralleled to make two coils.

L1--9 turns #10 e., 1" diam., 1½" long. 40 meter tap 4½ turns, 20 meter tap 2½ turns, 15 meter tap 1¾ turns, 10 meter tap ½ turn plus 2" lead. (All taps measured from "hot" end). Taps adjusted so circuit resonates with tube in socket with following total capacitance: 3.5 mc, 1660 $\mu\mu$fd; 7 mc, 830 $\mu\mu$fd; 14 mc, 415 $\mu\mu$fd; 21 mc, 240 $\mu\mu$fd; 28 mc, 210 $\mu\mu$fd.

M1--0-300 ma. M-2--0-1000 ma. M-3--0-1 ma.

PC--Three 100 ohm, 2 watt composition resistors in parallel. Shunt coil is portion of plate lead strap, two turns, 5/8" diam., 1 inch long.

RFC--1-Solenoid choke, 1 ampere, 8 μh. Barker & Williamson #800.

RFC-2--VHF choke, Ohmite #Z-50.

S1--Single pole, 5 position ceramic switch. Centralab #CRL-2501.

S2A-B--Two pole, 5 position ceramic switch. Radio Switch Corp. Model #86 shorting switch, two section. Type H wafers (one circuit, one pole, 12-position, including "off"). Sixty-degree detent, standard bearing.

T1--7.5 volts at 21 amperes. Stancor #P-6457. Chicago-Standard Transformer Co.

Socket and Chimney--Eimac SK-510 and SK-516 respectively. See tube data sheet for cooling requirements of 3-1000Z.

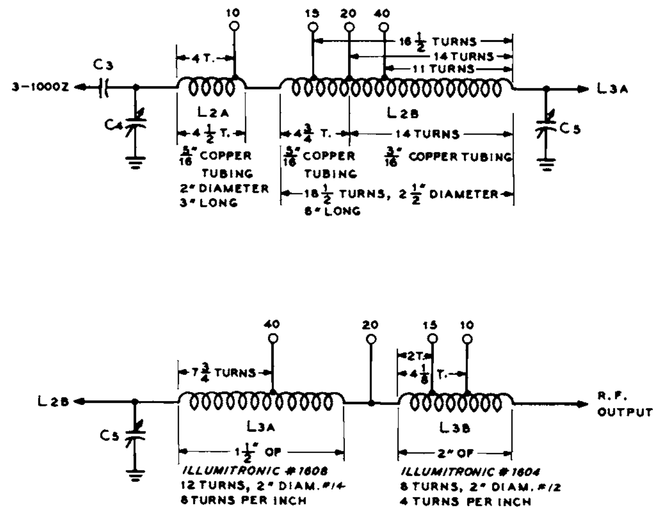
Anode connector--Eimac type HR-8.

Figure 2. Detail of Pi-L Network Coils

PLATE PI-NETWORK CONFIGURATION

Q=15

ITEM	AMATEUR BAND				
	80	40	20	15	10
C4 (UUF)	343	172	86	58	43
L2A-B (UH)	9.2	4.06	2.03	1.35	1.02
C5 (UUF)	891	446	223	149	112
L3A-B (UH)	8.34	4.17	2.08	1.39	1.04



Makeup of the Pi-L plate network coils is shown in figure 2. Coils are wound and positioned in the amplifier then taps adjusted with the aid of a grid-dip meter to provide circuit resonance with capacitance values of C4 and C5 as listed in the table. The tube should be in the socket and connected to the plate circuit for these measurements. Coil taps are made by flexible, ¼-inch copper strap, looped around coil tubing and fixed in place with 4-40 machi

round coil tubing and fixed in place with 4-40 machine screws and bolts. When correctly positioned, joints are soldered with a heavy iron or blow torch and coil is removed and silver plated. The 10 meter tap is positioned near end of coil L2A so that resonance is obtained at 29.7 mc with plate tuning capacitor about 5° meshed. Coils are arranged at right angles to each other to reduce mutual coupling to a minimum.

Plate parasitic suppressor (PC) is required to eliminate tendency to VHF oscillation caused by high gain of 3-1000Z, together with residual inductances in plate and filament circuits. Too many coil turns will cause suppressor to overheat when amplifier is operated on 10 meters, and too few turns will fail to completely "de-Q" the parasitic circuit. In general, two to four turns seem to be correct in most cases.

During standby periods, the 3-1000Z may be biased close to cutoff by placing a 50,000 ohm, 10 watt resistor in series with the lead connecting meter M-1 to the filament center-tap (point X, figure 1). The resistor may be shorted out by external VOX relay contacts.

The new Eimac SK-510 Air-System Socket and SK-516 Chimney are recommended for use with the 3-1000Z. The older SK-500 socket may be used provided additional air is applied to the socket and care is taken to see that the contacts are free to move about, and do not place lateral strain on the tube pins. In either case, the SK-516 chimney should be used.

amateur service newsletter
W6SAI

RECOMMENDED RATINGS: 6C21
FOR CLASS C AND CLASS AB2 CATHODE DRIVEN SERVICE

Electrical Characteristics

Filament Voltage	7.5 volts
Current	15 amperes
Amplification Factor	30 average
Direct Inter-electrode Capacitances	
Grid-plate	4.3 μμf
Grid-filament	9.5 μμf
Plate-filament	2.0 μμf

Radio Frequency Amplifier and Oscillator - Class C

Maximum Ratings	Plate-modulated	Telegraphy
DC Plate Voltage	4500	6000 volts
DC Plate Current	500	600 ma
Plate Dissipation	300	450 watts
Grid Dissipation	50	50 watts
Maximum Frequency	40	40 Mc full ratings

Typical Operation - One Tube

DC Plate Voltage	3000	4000	4500	3000	4000	5000	volts
DC Grid Voltage	-300	-350	-400	-200	-250	-350	volts
DC Fixed Bias	-150	-200	-230	-150	-200	-250	volts
DC Plate Current	380	340	345	500	450	450	ma
DC Grid Current	60	43	50	95	85	90	ma
Peak RF Grid Voltage	490	525	585	400	410	570	volts
Grid Resistor	2500	3500	3500	2500	3000	3500	ohms
Driving Power (Approx.)	30	23	29	35	35	46	watts
Grid Dissipation	14	10	12	21	18	24	watts
Plate Power Input	1150	1360	1550	1500	1800	2250	watts
Plate Dissipation	300	300	300	450	450	450	watts
Plate Power Output	850	1060	1250	1050	1350	1800	watts

Radio Frequency Amplifier, Cathode Driven, Class AB2
Typical Operation - One Tube

DC Plate Voltage	3000 volts
DC Fixed Bias	-50 volts
DC Resting Plate Current	80 ma (approx.)
DC Single Tone Plate Current	450 ma
DC Single Tone Grid Current	62 ma
Driving Power, Single Tone	105 watts
Plate Power Input, Single Tone	1350 watts
Plate Dissipation	450 watts
Power Gain	8.5
Input Impedance	320 ohms
Plate Load Impedance	4100 ohms

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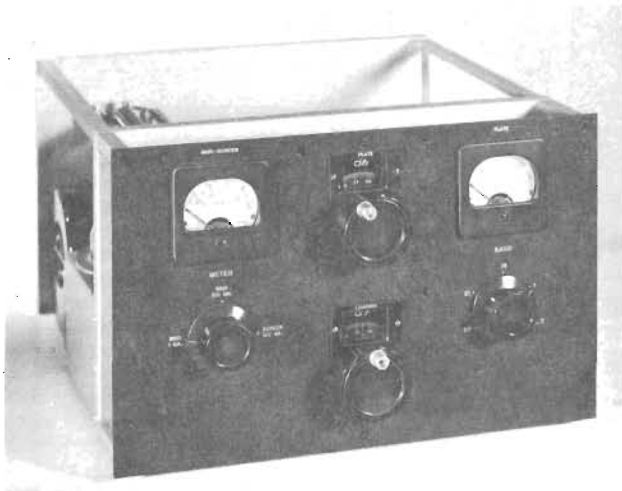


There is a pleasing symmetry to the control layout on the 10 × 15½-inch panel. The grid circuit is untuned, so the only r.f. controls are the band switch, plate tuning, and loading. Separate meters are provided for plate and screen currents, with the screen meter also used as a grid-current monitor. The amplifier, 15 inches deep, contains filament transformer and cooling fan in addition to the r.f. circuits.

Compact AB_1 Kilowatt

Single-tube amplifier runs 1000 watts input on s.s.b., c.w. or a.m. as a linear amplifier with no grid current. A new high-power tube designed specifically for AB_1 operation makes it possible.

BY RAYMOND F. RINAUDO,* W6KEV

BECAUSE IT IS the almost universal practice to generate an s.s.b. signal at a low level and then amplify it to the required output with one or more linear amplifiers operating Class A, AB_1, AB_2 or B, the linearity of the amplifying stages is all important. The stages following the best s.s.b. generator can turn a clean signal into one which is distorted and unnecessarily broad. Thus the need for truly linear amplifiers.

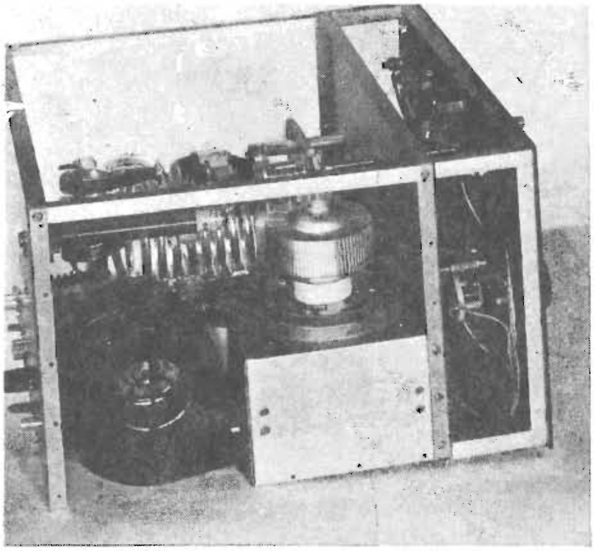
While the individual designer has his choice as to the class of operation in which the amplifier will run, Class AB_1 has several desirable characteristics. Because the control grid is never driven positive the very serious problem of adequate driver regulation never has to be faced, as it does if the mode of operation is AB_2 or Class B. In addition, no driving power is required for the tube: only the grid circuit losses must be supplied.

It should be pointed out that most tetrodes and many triodes appear as a resistance of 200 to 500 ohms from grid to cathode when the grid is positive. During the part of the r.f. cycle when the grid is negative the resistance is infinite. A driving source that can supply either an infinite resistance or a load of a few hundred ohms, without distortion of the voltage wave form in either case, would have to have very low internal resistance. A working approximation is usually achieved by making the tuned grid circuits of r.f. amplifiers extremely high C. In audio amplifiers it is obtained by using low-plate-resistance driver tubes plus a step-down transformer.

Class AB_1 amplifiers compare very favorably in efficiency with AB_2 and Class B. In fact, overall amplifier efficiencies, which take into account the losses in the tube and the circuit, are usually of the order of 55 to 65 per cent. It is only when compared with Class C operation that AB_1 represents a significant lowering of efficiency.

It is for this reason that some of the older tube types do not look particularly attractive in s.s.b. service. In the past almost all transmitting types were designed for optimum service in Class C amplifiers. This optimum provided a balance between plate current and plate dissipation; the higher efficiency realized required less plate dissipation capability for a given input power. In contrast, a tube designed especially for AB_1 application would be expected, for a given output, to have a higher plate-dissipation rating than we have become accustomed to.

* c/o Eitel-McCullough, Inc., San Bruno, Calif.



Vertical chassis construction is used, as this view from the tube side shows. The air-system socket is mounted on the 6 by 6-inch top of an aluminum enclosure 4 inches high, with the chassis pan forming one wall. When the bottom plate is in place this forms a pressurized area for forcing air from the blower through the socket. The socket chimney has been removed in this photograph to show the 4CX1000A tube.

The 4CX1000A Tetrode

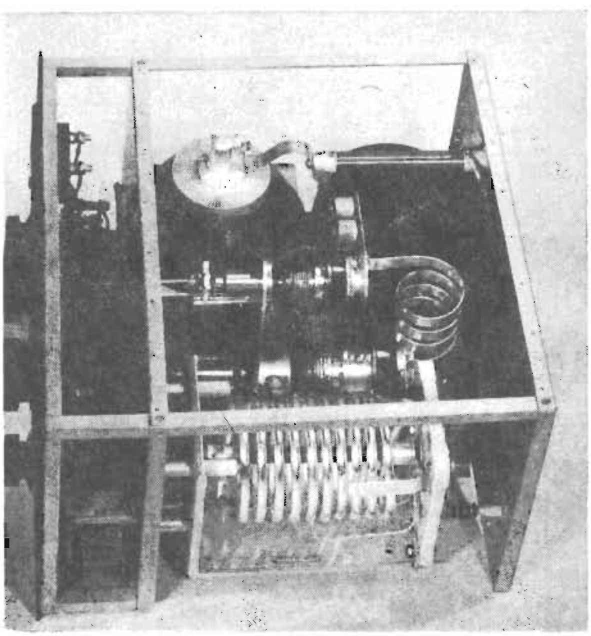
A tube designed to have exceptionally good linearity in Class AB_1 r.f. amplifiers is the newly announced Eimac 4CX1000A. It is a power tetrode of all ceramic and metal construction having an external anode capable of dissipating 1000 watts with 35 cubic feet of air per minute blown through the cooling fins. The filament requires 6.0 volts at 12.5 amperes to heat the oxide coated cathode. With the usual tetrode connection having the cathode and screen at r.f. ground, the grid-to-cathode capacitance is 85 $\mu\mu$f., plate-to-ground is 12 $\mu\mu$f., and grid-to-plate is 0.02 $\mu\mu$f. In spite of the low feed-back capacitance, the very high transconductance of 37,000 micromhos makes neutralization necessary if a tuned grid circuit is used. The maximum ratings are: plate voltage, 3000; plate current, 1 ampere; screen dissipation, 12 watts; control grid dissipation, zero watts.

The power output will vary with the type of service for which the tube is used. For single side band suppressed carrier single tone, the output is 1680 watts for 2700 watts input at the maximum plate voltage of 3000. If the driving signal is an amplitude-modulated carrier, either single or double side band, a carrier output of about 300 watts can be expected from a kilowatt input. If a c.w. signal is being amplified then the output power would be approximately 600 watts. Since for a.m. phone or for c.w. the carrier or key-down conditions apply in measuring power input, it is the legal power-input limit that largely determines the output power. In commercial service the capability is considerably greater.

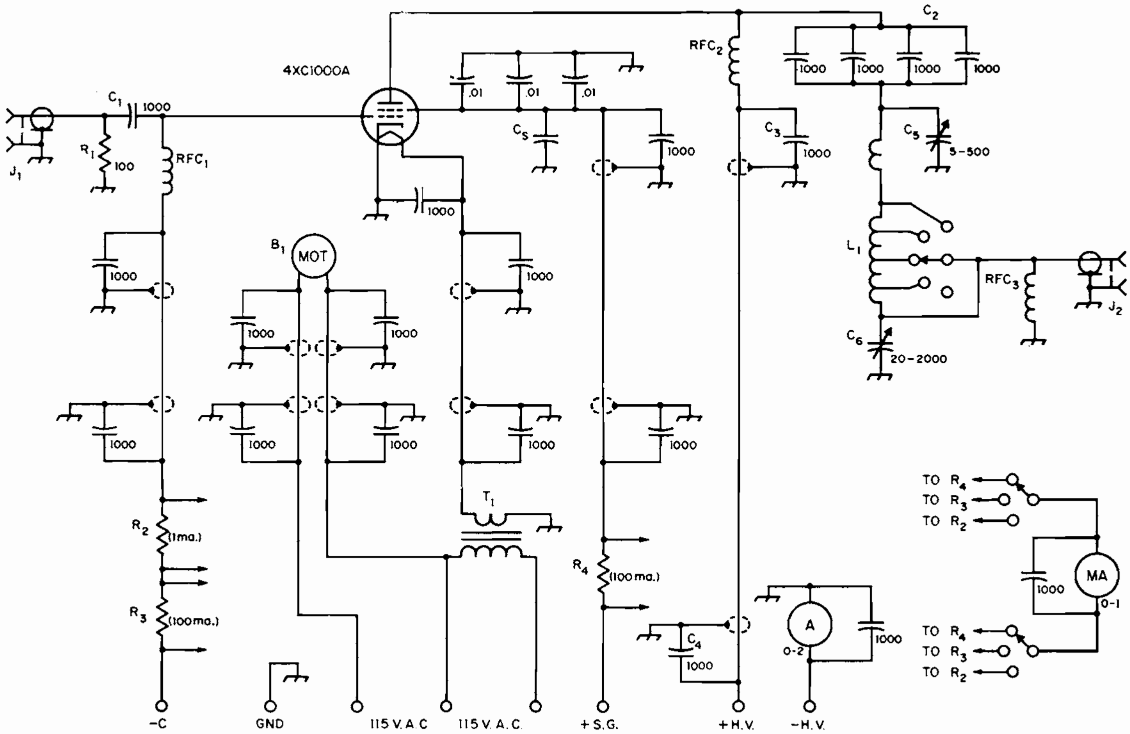
The connection to each element is made by means of three metal tabs or ears which protrude through the side of the envelope at 120-degree intervals around the circumference. The screen tabs are nearest the anode; the control grid, cathode plus one side of the heater, and heater follow in order to the bottom. Ham ingenuity will make it possible for some to build their own sockets but most will use the Eimac SK-800 which has a built-in screen by-pass capacitor. The height of the tube is just under 4¾ inches, and the diameter approximately 3⅜ inches.

The use of ceramics instead of the usual glass for the envelope makes the 4CX1000A much more rugged mechanically and makes possible a higher operating temperature. The first feature is very handy for the time the prized bottle rolls off the table onto the floor!

It will be noted above that the control grid is rated at zero dissipation. In designing the tube for AB_1 operation the location and number of grid wires was not hampered by compromises such as would be necessary if the grid were called upon to handle power. Consequently, a large number of fine wires were closely spaced to the cathode to give an unusually sharp-cutoff grid



This view from the tank-circuit side shows the tapped pi-network coil and the vacuum input and output capacitors. The capacitors are mounted on an aluminum bracket fastened to the tube compartment. The plate blocking capacitor—four units in parallel—mounts on a plate fastened to the hot terminal of the input tuning capacitor. The plate choke is mounted on the rear wall. The chimney is around the tube in this photograph.



Parasitic Suppressor may be required in plate lead.

Fig. 1—Circuit diagram of the amplifier. Unless otherwise indicated, capacitances are in $\mu\mu$f., resistances are in ohms. Capacitors not listed below are 600-volt disk ceramic.

B_1—Blower motor.

C_1—1000-$\mu\mu$f. mica.

C_2—Four 1000-$\mu\mu$f. ceramic in parallel, 5000-volt rating (Centralab 858).

C_3, C_4—1000-$\mu\mu$f. ceramic, 5000 volts (Centralab 858).

C_5—5-500-$\mu\mu$f. vacuum variable (Jennings UCSL 500 3KV).

C_6—20-2000-$\mu\mu$f. vacuum variable (Jennings UCSL 2000 2KV).

C_s—Built-in socket bypass, 1450 $\mu\mu$f.

J_1, J_2—Coax receptacles, chassis mounting.

L_1—Pi-network tank assembly (B & W 852).

R_1—100 ohms, noninductive, to dissipate at least 15 watts (see text). Can be assembled from 2-watt composition resistors in parallel or series-parallel.

R_2—Approx. 1000 ohms (should be 20 or more times meter resistance).

R_3, R_4—Adjusted to shunt 1-ma. meter for 100 ma. full scale; approx. 0.5 ohm in average case.

RFC_1, RFC_3—2.5-mh. r.f. choke.

RFC_2—Solenoid choke, 500 ma. (B & W 800).

T_1—Two 6.3-volt, 6-amp. transformers parallelled.

Note: Power lead for blower motor is brought out separately for resistance control of speed during stand-by.

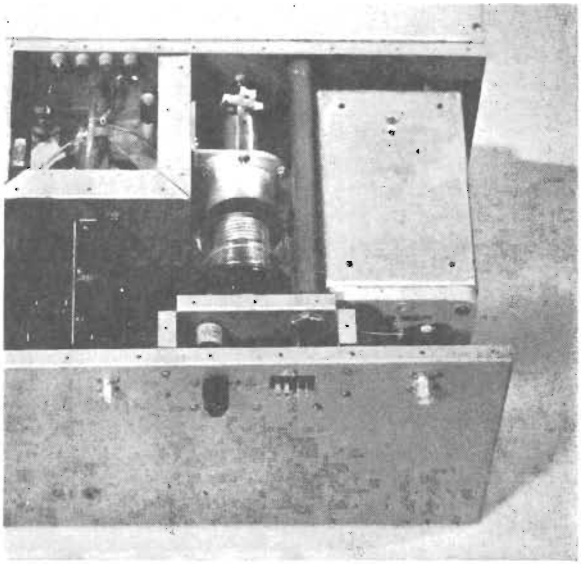
voltage–plate current characteristic. Thus linearity is maintained near cutoff.

While the tube is capable of powers in excess of the legal amateur limit it is quite legal to have peak inputs in amateur service well in excess of a kilowatt if the average power does not exceed that figure. (If there are doubters, please read the excellent article by Byron Goodman, "Linear Amplifiers and Power Ratings," in August 1957 *QST*.) In such cases the tube cathode is asked to supply quite high currents and must be capable of such operation if linearity is to be maintained.

A Compact Amplifier

The tube is a relative midget in size and the challenge to design a small amplifier of high power capability could not be resisted. So the amplifier shown in the photos, contained in a package measuring 10 inches high by 15½ inches wide by 15 inches deep, came into being. The r.f.-tight enclosure is 12 inches front to back, with a 3-inch space between the front panel and shielded box. Not shown in the photographs are the perforated aluminum U-shaped cover, which forms the top and two sides, or the solid sheet of aluminum that completes the shielding on the bottom. The space between the front panel and the main shielded enclosure is out of the r.f. field and so was not made r.f. tight. In spite of the compactness there is no crowding of parts.

The plate circuit is a conventional pi network. However, some of the components do represent a departure from those usually seen in high-power amplifiers. The blocking capacitor is made up of four 1000-$\mu\mu$f. ceramic units in parallel, resulting in a capacitance about double that normally used.



This bottom view gives a glimpse inside the grid compartment, upper left. R.f. input is through the coax connector on the rear wall and a short length of coax into the shield box. Power leads come in through the socket and high-voltage connector at the center, where they are enclosed by a small aluminum shield mounted on the rear wall. All except the high-voltage lead and leads to the blower motor go through the conduit (running alongside the bottom of the tank coil assembly) to the front of the unit. The high-voltage lead goes through shielded wire to the plate choke. Those to the blower are also shielded.

This was done because of the low impedances involved in the low-voltage high-current application. The plate tank inductor has much less inductance than the standard B & W 850A although physically the same size. The unit used was designed specifically for this low impedance application by Barker & Williamson, and it is understood that it is now available, carrying the number 852. The plate choke is the recently announced B & W 800. The Jennings variable vacuum capacitors contributed immeasurably to the compact construction, and here again the 500-$\mu\mu$f. input capacitor is higher in capacitance than usually expected. The high C is necessary at 3.5 Mc. to maintain the operating Q of the circuit. The low inductance of these capacitors helps considerably in the elimination of parasitic oscillations.

The grid circuit represents a departure from the usual practice by having no tuned circuits. As was mentioned previously, AB_1 operation precludes driving the grid positive and so the voltage stabilizing influence of a high-C circuit is not needed. Instead, a 100-ohm resistor is used in the r.f. circuit between grid and cathode. This also represents very heavy loading of the grid and makes neutralization unnecessary. When using the grid bias indicated for typical operating conditions, -55 volts, the power lost in the resistor is 15.1 watts and is the total required driving power. For those who would like to terminate the transmission line from the driver in a 50-ohm resistor, the driving power would be 30.3 watts. The photograph of the under side of the unit shows two noninductive wire-wound resistors which make up the 100-ohm load; these have since been replaced by a bank of carbon resistors.

If the driving power requirement of this untuned arrangement can not be tolerated, a tuned circuit can be added. In such case the only power needed is that required to supply the tuned-circuit loss. Neutralization, of course, would become necessary, and the usual bridge circuit is the logical choice.

The front panel shows that two meters are used, though one is dual purpose. The plate current meter has a full-scale reading of two amperes; however, the maximum plate current that can be drawn is 1 ampere using the single tone test (into a dummy load). The dual-purpose meter is one milliampere full scale and is used in combination with a switch and shunts to read grid current at 1 ma. full scale, grid current at 100 ma. full scale, or screen current at 100 ma. full scale. The one-milliampere scale is used to monitor s.s.b. AB_1 operation so as never to drive into grid current. The 100-ma. grid current scale seems to be (and is) in direct contradiction to the statement that the control grid can dissipate no power. The truth is that from ½ to 1 watt can be handled, but this leaves no margin of safety. The rating of zero dissipation still stands.

Although AB_1 operation minimizes the generation of harmonics, standard TVI-proofing techniques are used throughout. All leads leaving the shield enclosure not normally carrying r.f. are shielded and bypassed at both ends. Leads to the front panel from the compartment that shields the power-input socket are carried through the r.f. enclosure in a length of ½-inch conduit.

Two filament transformers in parallel are used to supply heater power. This was done because no single transformer of suitable capacity was available to fit into the space allotted. The transformers have a total capacity of 12 amperes to supply a heater requiring 12.5 amperes. However, the overloading is considered negligible.

In operation the amplifier has proved to be quite stable. The 100-ohm resistor between grid and ground undoubtedly contributes a great deal to this stability. However, a change in layout, even though minor, could alter the picture. As always, each new design must be checked for parasitics and be debugged if necessary. Slight changes in an old design in effect make it a new one.

The author wishes to thank Vern Olsen, W6INJ, for the use of the photographs which show the construction of the very neat amplifier built by him.

amateur service newsletter
W6SAI

A Compact 500-Watt Transmitter for 50 Mc.

BY WILLIAM I. ORR,* W6SAI AND RAYMOND F. RINAUDO,* W6KEV

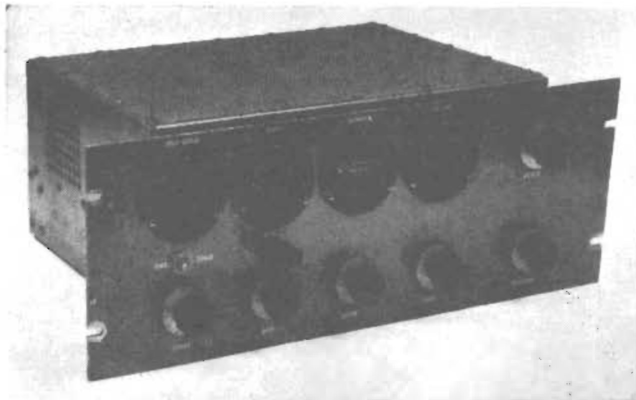


Fig. 1—This 500-watt transmitter packs a punch on 50 Mc.! It is completely enclosed in an aluminum box to reduce TVI problems. The top and bottom are made of perforated aluminum and the holes on the left side of the enclosure are for air intake to the blower. Lower controls are (l. to r.): drive (R_1), oscillator (C_1), doubler (C_2), amplifier grid (C_3), and loading (C_5). Plate tuning (C_4) is at the upper right.

A Compact 500-Watt Transmitter for 50 Mc.

BY WILLIAM I. ORR,* W6SAI AND RAYMOND F. RINAUDO,* W6KEV

Six meters is here to stay. Regardless of the ups and downs of the sunspot cycle, this band remains popular and has heavy usage in most parts of the United States. Many six-meter operators have expressed a desire to build a simple, foolproof, high-power transmitter exclusively for 50-Mc. operation. The design shown herewith has been in service for over two years, has proved to be a steady and dependable performer and is recommended to those six-meter operators who wish to "step up to the very best!"

It has been found from the writers' experience that power levels up to 500 watts or so are readily achieved at 50 Mc. with physically small equipment and reasonable power supplies. Above this power level, however, the cost and difficulty of assembling high-power gear seems to rise as the square of the power input! In particular, the amount of expensive "iron" required in power-supply and audio equipment seems to be entirely disproportionate to the 3-decibel power gain achieved when going from 500 watts to a kilowatt. The extra 3 decibels can more easily be achieved in the antenna.

Circuit Description

The transmitter is a three-stage crystal-controlled unit capable of running 625 watts input c.w. or 300 watts a.m. phone on the 50-Mc. band. Provision is made for v.f.o. control. Only three transmitter stages are required and precautions have been taken to prevent radiation of undesired harmonics or spurious emissions that might cause TVI or BCI.

* Eitel-McCullough, Inc., San Carlos, Calif.

The tube line-up, Fig. 2, is: 6AG7 oscillator multiplier, 6V6GT frequency doubler, and a 4CX300A power amplifier. The crystal oscillator is a modified Pierce circuit, with a crystal connected between the grid and screen of the pentode tube, and the plate circuit tuned to the third harmonic of the 8-Mc. crystal. If desired, 6-Mc. crystals may be substituted without circuit changes, the plate circuit then being tuned to the fourth harmonic of the fundamental frequency. Maximum frequency stability is achieved by a combination of low crystal current and oscillator screen voltage regulation. A 0A2 regulator tube holds screen voltage at 150 on the oscillator stage. The 25-Mc. harmonic output is capacity-coupled to the following doubler stage.

The doubler uses a 6V6GT in a conventional circuit. Its screen voltage is adjustable at the front panel by R_1. This control determines the drive level to the final amplifier. The 50-Mc. output of the doubler stage is applied to the grid circuit of the 4CX300A amplifier stage by link coupling through a short length of coaxial line. Inductive coupling through two tuned circuits helps to minimize the coupling of 25- and 75-Mc. energy and higher harmonics to the final-amplifier grid.

The 4CX300A final-amplifier stage uses a simple pi-network plate circuit. The tube is neutralized by a capacitive bridge circuit composed of the various internal capacitances of the 4CX300A, the neutralizing capacitor, C_7, and the 300-pf. silver mica bypass capacitor in the ground leg of the tuned grid circuit. The 4CX300A is mounted in an Eimac SK-710 socket containing a low-inductance screen bypass capacitor, C_8, especially

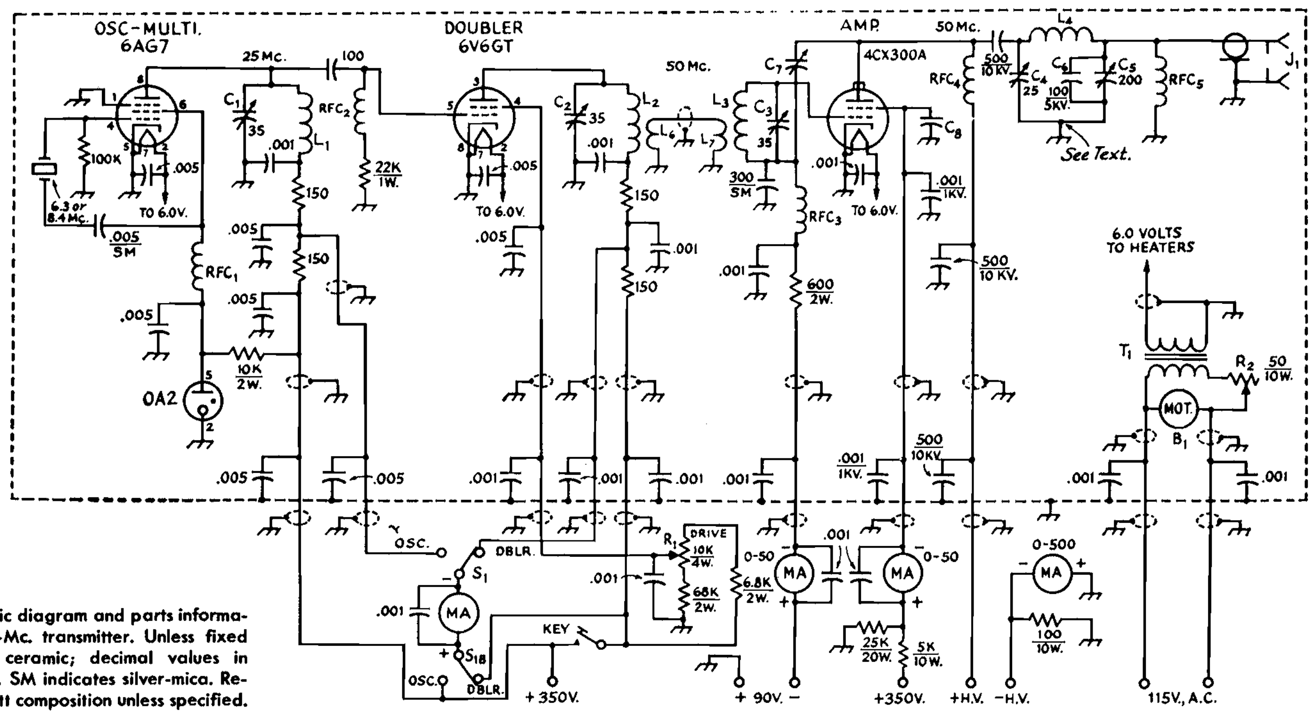


Fig. 2—Schematic diagram and parts information for the 50-Mc. transmitter. Unless fixed capacitors are ceramic; decimal values in μf., others in pf. SM indicates silver-mica. Resistors are ½-watt composition unless specified.

B_1—Phonograph motor to drive blower; see text. Blower should deliver 10 c.f.m. or more.

C_1, C_2, C_3—35-pf. variable (Johnson 149-2 or 35R12).

C_4—25-pf. variable, 0.075-inch spacing (part used was surplus; suggested replacement Johnson 154-11).

C_5—200-pf. variable, .02-inch spacing (surplus, from BC-375 tuner. Johnson 155-6 suitable).

C_6—100-pf. 5000-v. (Centralab 850S-100N).

C_7—Neutralizing capacitor; see text and Fig. 4.

C_8—Screen bypass, built into socket.

J_1—Coaxial receptacle.

L_1—12 turns No. 16 tinned, ½-inch diam., 16 t.p.i. (Air-Dux 416T).

L_2—7 turns No. 16 tinned, ½-inch diam., 8 t.p.i. (Air-Dux 408T).

L_3—4 turns like L_2.

L_4—4 turns ⅛-inch copper tubing, 1-inch diam., ¾ inch long.

L_5—3 turns No. 16 tinned, wound on 50-ohm 1-watt resistor.

L_6, L_7—Links of insulated wire around cold ends of L_2 and L_3, 1 turn each. Can be made from ends of connecting 50-ohm coax, with outer conductor removed.

R_1—10,000-ohm 4-watt control.

R_2—50-ohm 10-watt slide type.

RFC_1—2.5-mh. r.f. choke (National R-100).

RFC_2—44-μh. r.f. choke (Ohmite Z-14).

RFC_3—7-μh. r.f. choke (Ohmite Z-50).

RFC_4—3-μh. r.f. choke. 48 turns like L_1.

S_1—D.p.d.t. toggle.

T_1—6.3 volts, 4 amp. Adjust R_1 for 6.0 volts at heater terminals.

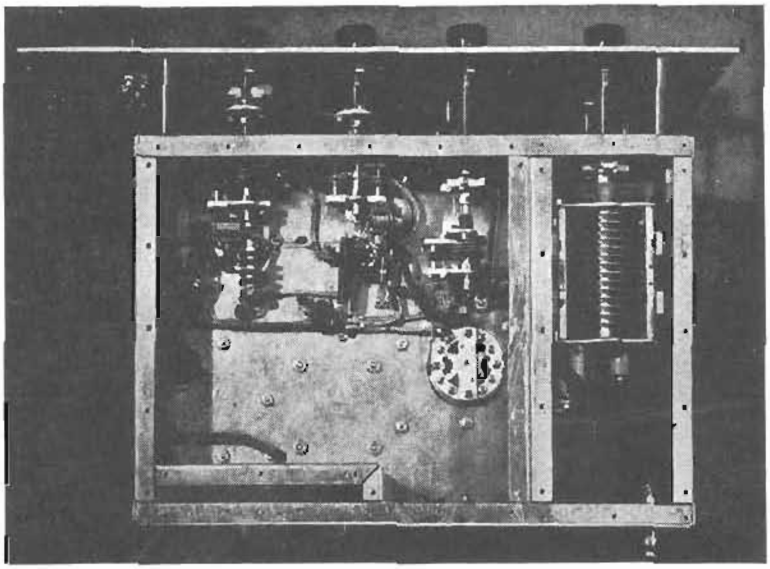


Fig. 3—Under-chassis view of the transmitter. The blower exhaust port is at the lower left, just above the terminal strip shield. All power leads are run in shield braid, with the center conductor bypassed to the braid at each end. The right-hand compartment houses the final-amplifier tank circuit. The 4CX300A socket is at the center of the chassis, behind the grid-tuning capacitor, C5. Under-chassis area is pressurized by a solid plate, removed for photography.

designed for v.h.f. operation. In addition, the socket screen terminal is bypassed with a 0.001-μf. disk ceramic capacitor.

Four panel-mounted meters are used to indicate the currents drawn by the various stages. The left meter measures the plate current of either the oscillator or doubler. A double-pole double-throw switch, S_1, connects the meter across the appropriate 150-ohm shunt. Individual meters are used in the grid, screen and plate circuits of the final amplifier. The plate-current meter, right, is placed in the negative lead from the high-voltage supply, requiring that the negative lead not be connected to ground at any point in the supply.

The oscillator and doubler stages are operated from a 350-volt 150-ma. power supply which also provides the screen voltage for the amplifier stage. C.w. keying is best accomplished by breaking the plate and screen voltage of the doubler stage at the keying point indicated in the schematic, allowing the oscillator to run continuously. A suitable *LC* filter must be used at this point to suppress key clicks, and the use of a keying relay is recommended to remove high voltage from the key. This is preferred to keying the cathode of the 6V6GT.

Transmitter Construction

Reliability is the watchword in the design of this transmitter, and the construction technique used is the tried-and-true relay-rack style, designed to remove as much thermal heat from the equipment as possible. Nearly 200 watts is dissipated by the transmitter when running "wide open," so special pains are taken to insure that all components run cool, and well within their ratings. As the complete transmitter has to be shielded to reduce unwanted radiation, the problem of heat radiation is an important one.

The transmitter is enclosed in a home-made aluminum box 14 inches wide, 10 inches deep, and 6½ inches high. The enclosure is supported behind a standard 7 by 19-inch rack panel by means of four 2-inch metal pillars, thus providing space for the various switches and meters to be mounted behind the panel and out of the strong r.f. field. The top of the enclosure and the small portion of the bottom directly below the final tuning capacitors are covered by perforated aluminum sheet. The chassis portion of the enclosure has a solid aluminum sheet bottom plate, making possible chassis pressurization by the blower, the air escaping through the SK-710 socket and the SK-606 chimney and cooling the 4CX300A anode. A small aluminum shield (lower left corner, Fig. 3) was placed in one corner of the underchassis area to prevent the radiation of harmonics via the terminal-strip connections. Shielded wire is used for all power leads, and each terminal of the barrier strip is bypassed.

The blower used to cool the 4CX300A is a "home-brew" device. A squirrel-cage blower is driven by a phonograph motor which turns at about 3300 r.p.m. The unit moves enough air to cool the tube adequately up to about 150 watts plate dissipation, but would be insufficient for the full 300-watt dissipation rating. However, in neither the keyed c.w. mode (625 watts input) or the a.m. mode (300 watts input) does the average plate dissipation exceed 125 watts, so full dissipation is not required. Although the blower is rated at 10 cubic feet per minute, chassis back-pressure reduces the air flow to less than 4 c.f.m.

The neutralizing capacitor, C_7, is formed by placing a ½-inch square aluminum plate adjacent to the anode of the tube, the separation being about ⅜ inch. This capacitor may be seen in the top-view photograph, Fig. 4, appearing as a small bracket between the 4CX300A and the front panel. The plate is mounted on the top of a ceramic feed-through insulator, which brings the neutralization lead from beneath the chassis.

Component Layout and Wiring

Parts layout is not especially critical. The stages are placed so that the tuning controls are evenly spaced across the front panel, even though the chassis is offset slightly from the center line of the panel (see under-chassis photograph). The blower and motor are placed in a corner of the chassis, with the blower exhaust duct passing through the deck of the chassis. All under-chassis power and filament wiring is done with shielded wire, and r.f. leads are made short and direct. The interstage link circuit is made of a short length of RG-58/U, 52-ohm coaxial line, with the outer shield of the line grounded at both ends. Flexible couplings and panel bushings are used on the tuning-capacitor shafts to insure that the controls turn evenly and smoothly. Capacitors C_1 and C_2 are grounded to the chassis, but capacitor C_3 is mounted on an insulating block, as its rotor is above ground.

The plate circuit of the 4CX300A employed a home-made choke, RFC_4, air-wound over a ½-inch-diameter form. After a period of time, it was found that the choke had a tendency to sag a bit, so a section of Air-Dux coil was substituted for the home-brew choke (see parts list).

The pi-network capacitors, C_4 and C_5, are mounted by their front frames to a ¼-inch Plexiglas or phenolic sheet measuring 3½ by 4½ inches which, in turn, is supported from the front wall of the enclosure by one-inch metal bolts and pillars. The capacitors are driven by insulated couplings and thus the rotor shafts are floating above ground. The rear frame of each capacitor is then grounded to the chassis by means of a ½ inch wide copper strap running from the frame to the side of the chassis box via the shortest possible route. Ground loops are thus avoided, and the 6-meter pi-circuit tunes in a normal manner, "just like on the d.c. bands!"

The amplifier plate coil, L_4, is supported on two one-inch ceramic insulators bolted to the rear frame of the tuning capacitors, and connections to the coil, capacitors, and plate blocking capacitor are made with ½-inch copper strap. The auxiliary loading capacitor, C_6, and grounding choke, RFC_6, are mounted on the rear of the variable loading capacitor, C_5. A short length of RG-58/U 52-ohm coaxial line connects the loading capacitor to the coaxial receptacle mounted on the rear wall of the enclosure. The outer braid of the line is grounded at the receptacle and also at the common point on the chassis wall to which the tuning capacitors are grounded.

Transmitter Operation

Once the transmitter has been completed, the wiring should be checked for possible errors before the equipment is given the "smoke test." A filament supply of 6.0 volts at about 4 amperes is required. It is important to note that the filament voltage of the 4CX300A is *6.0 volts, not 6.3 volts,* and the voltage should be held within plus or minus 5 per cent of the design voltage if best tube life is to be obtained. Borrow a good, one-per-cent laboratory-type voltmeter to check your filament voltage and adjust it to 6.0 volts, or a little under. The 6AG7 and 6V6GT will work well at this slightly-lower-than-normal filament voltage. The blower should be turned on with the filament voltage to make sure that the stem of the 4CX300A remains cool, and the bottom plate should be in position to pressurize the chassis.

For initial tune up, the screen and plate voltages are removed from the 4CX300A. Remember that screen voltage should not be applied to the amplifier stage unless plate and bias voltages are also on the tube. It is permissible to apply or remove all voltages simultaneously, but screen voltage should never be applied *before* grid bias voltage and plate voltage are on, or screen current and dissipation will be excessive. This sequence of operation applies to all tetrode tubes, including small receiving tubes.

The 90-volt bias supply for the 4CX300A must be capable of withstanding about 30 ma. of grid current without serious loss of voltage regulation. A simple 50-ma. supply with a VR90 regulator tube will do the job. Adjust the series resistor to pass maximum current (40 ma.) when the supply is disconnected from the transmitter.

After the tuned circuits have been set to their approximate resonant frequencies with the aid of a grid-dip oscillator, an appropriate crystal is placed in the panel socket and high voltage applied to the oscillator and doubler stages, and bias to the 4CX300A. The drive potentiometer, R_1, is set for minimum screen voltage on the 6V6GT and the tuned circuits are resonated for an indication of grid current on the 4CX300A grid meter. The drive control is adjusted for about 5 ma. amplifier grid current.

The 4CX300A stage is next neutralized in the normal manner, the neutralizing capacitor being adjusted so

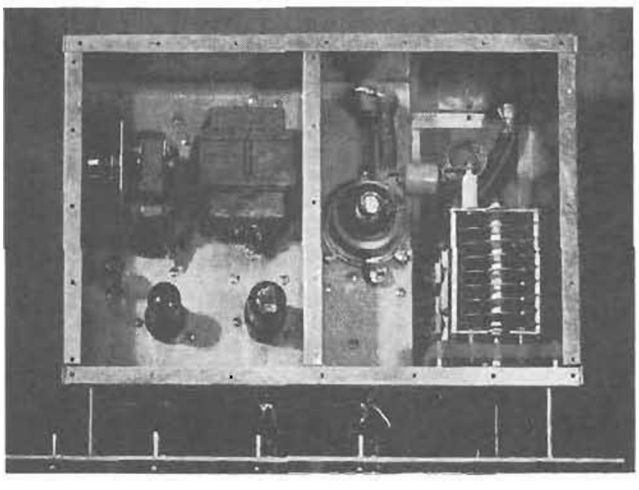


Fig. 4—In this view the blower and filament transformer are seen in the left-hand compartment near the rear of the chassis. Chassis plate, side pieces and shields are made from flat aluminum with flanges folded in a metal brake. In the left compartment are 0A2 regulator, 6AG7 oscillator, and 6V6GT doubler. The 4CX300A amplifier is on the edge of the chassis in the right-hand compartment, along with the plate-circuit components.

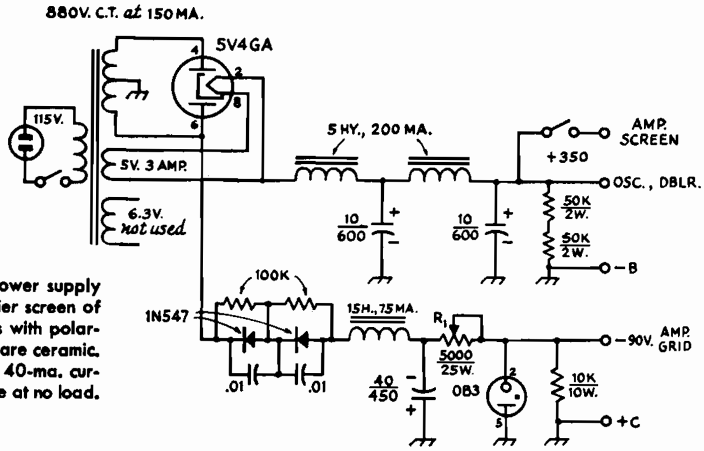


Fig. 5—Circuit diagram of the power supply for the exciter stages and amplifier screen of the 50-Mc. transmitter. Capacitors with polarity marked are electrolytic, others are ceramic. The slider on R_1 should be set for 40-ma. current flow through the regulator tube at no load.

that a minimum of grid current variation is noted when the plate tuning capacitor, C_4, is tuned through resonance (loading capacitor C_5 is set at maximum for this adjustment). Alternatively, a small pilot lamp or other r.f. indicator may be coupled to the coaxial antenna jack and the neutralizing capacitor adjusted for minimum feed-through power. Once the stage is properly neutralized, little or no change on the grid meter will be noted as the amplifier plate circuit is tuned through resonance.

When all is working properly, the transmitter may be connected to the antenna feed line or to a dummy load, and grid drive, screen and plate voltage applied to the amplifier stage. Start with the drive control set near minimum. Increase the drive slightly and establish resonance in the plate tank circuit. Drive and loading are now increased until the typical operating conditions indicated in the table are achieved. The amplifier stage should never be operated in a lightly-loaded condition unless the drive level is reduced. With a little practice, you'll soon observe that grid, screen and plate currents are a function of both grid drive and antenna loading. If loading is too light, grid and screen current will be high. If the loading is too heavy, grid and screen current will tend to be low. Screen current, it should be noted, is an extremely sensitive indication of proper balance between drive and loading. In fact, under conditions of heavy loading, the screen current of a tetrode may reverse. This phenomenon, common to tetrode tubes, makes it dangerous to rely upon a screen-dropping resistor or a series regulator to supply the screen voltage unless a bleeder is connected from screen to cathode. The bleeder should draw at least 15 milliamperes. Resistor R_2 in the screen circuit of the 4CX300A fulfills this requirement. Since it is placed before the screen meter, the bleeder current does not register.

TABLE I

Typical Operating Conditions for the 4CX300A

Circuit	*Class C* Telegraphy	*Class C* Telephony	*Class AB_1* Linear, S.S.B.
Plate Voltage	2500	1500	2500
Screen Voltage	250	250	350
Grid Voltage	−90	−100*	−55
Plate Current, ma.	250	200	100–250
Screen Current, ma.	16	20	4
Grid Current, ma.	25	14	—
Driving Power, watts	2.8	1.7	—
Plate Input, watts	625	300	625
Power Output, watts	500	235	400

* −90 volts fixed, plus 10 volts rectified bias.

For complete operating data, see information accompanying the tube, or write the authors.

Modulating the Transmitter

For a.m. phone, the final amplifier may be plate modulated. Plate potential is reduced to 1500 or 1000 volts. A pair of 811As, operating at the same voltage, will make a satisfactory modulator. As with other tetrodes, it is necessary to provide some modulating voltage to the screen element of the 4CX300A. A choke of 10 henrys (50-ma. rating) in series with the screen supply circuit, between R_2 and the meter, will provide adequate screen modulation so that 100-per-cent plate modulation may be achieved.

Using an External V.F.O.

An external v.f.o. may be used with this transmitter by changing the 6AG7 oscillator stage to a frequency multiplier and injecting a 6- or 8-Mc. frequency into the panel crystal socket. To do this, it is only necessary to ground the crystal socket pin connected to the 0.005-μf. screen-blocking capacitor. The output of the v.f.o. may now be injected into the crystal socket via a short length of coaxial cable. The inner conductor of the cable is terminated to the socket pin attached to the 6AG7 grid terminal, and the shield of the cable is attached to the grounded socket pin. An old FT-243 crystal holder may be used for the plug on the end of the cable. If one of the 4-40 bolts holding the crystal socket to the panel is reversed and allowed to project beyond the panel, a small clip on the shield of the cable may be used to ground both the shield and the proper socket pin by clipping it to the projecting bolt. Other methods of connecting a v.f.o. are shown in July 1963 *QST*, page 16.

amateur service newsletter
W6SAI

The 4CX350A Radial Beam Tetrode

The Eimac 4CX350A is a radial beam tetrode having over twice the transconductance of the 4CX250 type tube. Maximum plate dissipation of the new tube is rated at 350 watts and use of the 4CX350A in new equipments can often eliminate an amplifier stage in practical circuit design while allowing greater safety factor in operation.

The 4CX350A tube is designed for linear amplifier, broadband operation, and distributed amplifier service. It is normally operated with zero grid current and grid dissipation is limited to zero watts. The design features which make the 4CX350A capable of maximum operation without driving the grid into the positive region also make it necessary to avoid positive grid operation. This tube family, therefore, is not rated for class B or class C operation.

The accompanying data sheets compare the 4CX350A with the 4CX250B. Note that the 4CX350A requires less than half the grid drive of the 4CX250B for maximum output signal. On the other hand, higher screen voltage is required for the 4CX350A and more cooling air is required to attain the higher plate dissipation level. Filament current of the 4CX350A is higher than that of the 4CX250B. In addition, the input capacitance of the 4CX350A is appreciably greater than that of the 4CX250B because of the modified grid structure.

Under proper circumstances, substitution of the 4CX350A for the 4CX250B in existing equipment may be done to achieve a substantial saving in grid driving voltage. Such substitution, however, cannot be considered unilaterally, because of the different screen and grid electrode voltages, higher filament current, and difference in interelectrode capacitance between the two styles of tubes. The 4CX350A, therefore, should not be considered as a "high power" substitute for the 4CX250B but as a unique, high transconductance tetrode useful in broadband, low drive devices where high power gain, and maximum plate dissipation are desired. Substitution of one tube type for the other in existing equipment must be decided by a complete study of the system to see if such a change is feasible or justified.

amateur service newsletter
W6SAI

Intermodulation Distortion in Linear Amplifiers

BY WILLIAM I. ORR,* W6SAI

Although there has been much talk about intermodulation distortion in linear amplifiers, a search of available literature brings to light very little in the way of factual data. Here's down-to-earth dope on what linear-amplifier tubes can and can't do.

It is common communication practice to generate a single-sideband signal at a low power level for reasons of economy and then to ampilfy it to the desired strength by the use of one or more linear amplifier stages. The intelligence is contained in amplitude variations in the signal, and it is imperative that the linear stages amplify this intelligence with as little distortion as possible. Strictly speaking, an ideal linear amplifier is one in which the output envelope amplitude is at all times directly proportional to the input envelope amplitude. Amplitude distortion results when the magnitude of the output signal is not strictly proportional to that of the driving signal. This class of distortion (which is the principal type encountered in linear amplifiers) includes *intermodulation distortion*, a particularly interesting type of amplitude distortion encountered in single-sideband service. In passing, it should be noted that intermodulation distortion (abbreviated IMD) occurs only in a nonlinear device driven by a complex signal having more than one frequency. As speech is made up of multiple tones (or frequencies) and as the perfect linear amplifier has yet to be built, the situation leading to IM distortion exists in most s.s.b. amplifiers. Once the intelligence-bearing signal has been generated, the amplitude relationships existing in the intelligence must be faithfully retained or the s.s.b. signal will blossom into a broad, fuzzy caricature of itself, and the unlucky user of the nonlinear equipment will find his on-the-air popularity waning. Intermodulation distortion, therefore, is important to the s.s.b. operator, and the cause and effect of this unloved and unwanted mutilation of s.s.b. signals will be discussed in this article.

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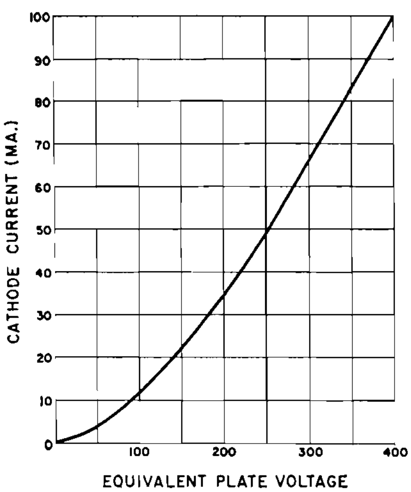


Fig. 1—The electron flow (cathode current) in a vacuum tube is a nonlinear function of the equivalent plate (or plate and screen) voltage and is described by the 3/2-power law. This curve illustrates typical electron flow, which plays an important part in establishment of tube linearity.

The Vacuum Tube and Linearity

The vacuum tube is the heart of the linear amplifier, and the amplifier is designed about it.[1] In addition to the tube, the amplifier is composed of auxiliary equipment — resistors, capacitors, inductors, etc. — chosen to permit the tube to operate in the most linear manner possible consistent with various restrictions imposed by economic, physical and electrical limitations. The auxiliary equipment may be considered to be made up of passive circuit elements while the vacuum tube is thought of as an active element by means of which the desired power gain is accomplished. The passive circuit elements are entirely linear and they affect circuit operation only insofar as they determine the operating parameters of the tube. The linearity of the tube is open to question. The more linear the tube, the less stringent the demand placed upon the circuitry to achieve a desired degree of over-all linearity. The results obtained are a balance between excellence and economy.

The vacuum tube utilizes electrons emitted from a hot cathode by impressing upon them an electric field which varies with time. During the passage of the electrons from cathode to plate, the field is manipulated in such a way as to alter the number of electrons arriving at the plate of the tube. The electric field reacts in a predictable way that may be accurately described by Maxwell's equations. The electron flow (or cathode current) is a 3/2 power function of the applied electrode voltages. This so-called "3/2-power

[1] This discussion applies to vacuum tubes. Similar conclusions may be drawn about transistors, but such conclusions are not within the scope of this article.

law" of Child and Langmuir is theoretically valid for uniform tube geometry and holds true for any space-charge-limited electron flow under the influence of an external field (Fig. 1). The 3/2-power law is not a linear function, and in practical tubes the cathode current is not a straight-line function of grid voltage. Further, practical tubes depart from the 3/2-power law to some extent, depending upon tube geometry, space charge, electron interception by grids, and emission limitations.

The relationship between the electric field and cathode-current flow within the tube described by this natural law plays an important role in the establishment of tube linearity. In practical amplifiers, for example, the magnitude relationship between input and output signals is not perfectly constant at all signal levels within a given range. The relationship defining *amplifier* linearity is termed the *envelope transfer function*, and ideal and typical transfer functions are shown in Fig. 2. The fundamental cause of a non-ideal, nonlinear amplifier transfer function may be traced directly to the nonlinear relationship between the plate current and grid voltage of the tube employed in the amplifier. This relationship *approximates* the 3/2-power law throughout the operating region above cutoff.[2] An examination of intermodulation distortion reveals the importance of significant cathode-current departure from this fundamental law as regards amplifier linearity.

Intermodulation Distortion Measurement Techniques

Leaving the vacuum tube for a moment, it is useful to examine means of testing tuned linear amplifiers for distortion. One such means is to apply two equal-amplitude r.f. signals of different frequency to the input circuit and then to measure the relative strengths of the output signals and the accompanying intermodulation products.[3] This combination of input signals is often called a *two-tone test signal*. The action of the test signals beating with each other in the typical "nonlinear" linear amplifier having amplitude distortion produces intermodulation distortion, and the purpose of the two-tone test is to create this action under controlled conditions and to measure it. Maximum limits of intermodulation distortion have become an important specification determining the excellence (or lack thereof) of linear amplifiers and tubes.

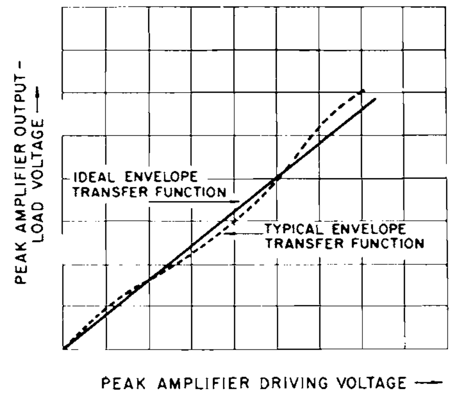


Fig. 2—Amplifier linearity is defined by the envelope transfer function. Departure from linearity is illustrated by curvature of the function (dotted curve) and may be traced directly to the nonlinear relationship between cathode current and electrode voltage shown in Fig. 1.

A practical test technique is to employ a two-tone, low-distortion test signal to drive a linear amplifier, and to use a spectrum analyzer to display a sample of the output signal of the amplifier (Fig. 3). A spectrum analyzer is a precision panoramic receiver having high resolution and capable of resolving signals separated in frequency by only a few kilocycles. The presentation of a portion of the spectrum in which the tests are taking place is given on a long-persistence cathode-ray tube. If the IMD products of the two-tone test signal are known and the amplifier under test is run with no feedback, at a frequency low enough to remove side effects due to circuit uncertainties, the IMD products of the tube under test may be readily determined by visual inspection of the picture on the screen of the spectrum analyzer. Equally important is the fact that the test is reproducible, and that the tube may be operated under any combination of electrode voltages and loads.

A block diagram of a typical IMD test experiment is shown in Fig. 4. The low-distortion signals are generated by separate stable r.f. oscillators operating on 2000 and 2002 kc., respectively, their outputs being carefully combined in a special isolator which prevents the oscillators

[2] Cutoff may be thought of as that amount of grid bias required to reduce the idling plate current of a vacuum tube to virtually zero.
[3] "The Grounded Grid Linear Amplifier," Orr, Rinaudo, Sutherland; *QST*, August, 1961, pages 16–21.

Fig. 3—*QST* authors and prominent DXers W6KEV (standing) and W6UOV examine data plotted by Eimac Intermodulation Distortion Analyzer. General-purpose equipment permits IMD measurements to be made on a wide variety of transmitting tubes in either grid- or cathode-driven configuration. IMD products are seen on screen of panoramic analyzer.

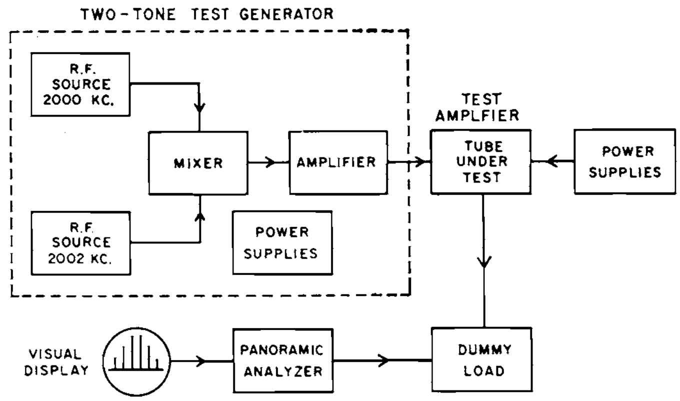


Fig. 4—Block diagram of Intermodulation Distortion Analyzer of Fig. 3. Low distortion two-tone r.f. signal is generated at 2 Mc. and applied to test amplifier. The output of the amplifier is dissipated in a dummy load and a portion of the output signal is examined on the screen of a high resolution panoramic analyzer. Distortion products as low as −60 decibels below peak power may be seen and studied.

from "seeing" each other. The resultant two-tone signal is amplified by successive class A stages until the desired driving level is reached. The two-tone generator shown in the photograph is capable of delivering a test signal having IMD products more than 60 decibels below the two-tone signals, at a power level up to 700 watts.

The tube under test is placed in a test amplifier operating at 2000 kc., and capable of permitting various electrode voltages and r.f. loads to be

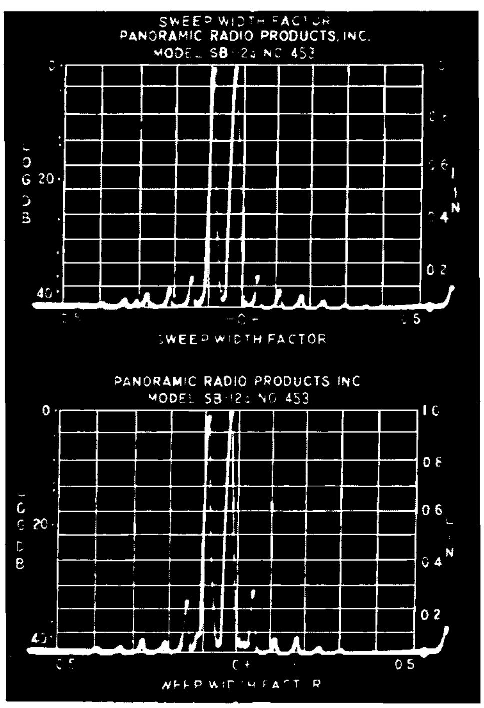


Fig. 5—Typical display on screen of IMD Analyzer. Top: Two test tones are seen at the center of the screen, with IMD products evenly displaced on either side of test signals. Third-order products are 35 db. down in amplitude from two-tone signal, and 5th-order products are 40 db. below test signals. Higher-order distortion products may also be seen. Bottom: Equipment parameters adjusted to raise third-order products and to drop fifth-order products. The linear amplifier may be adjusted to enhance or reduce various distortion products, if desired.

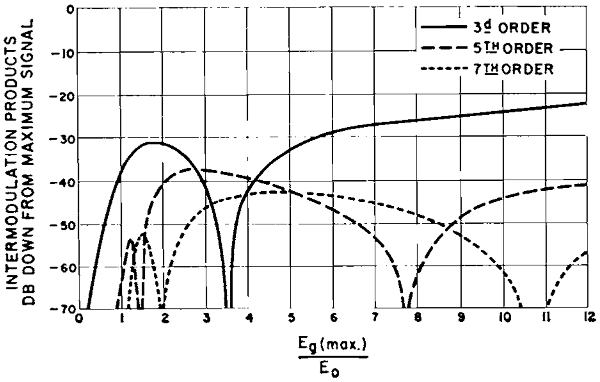
impressed upon the tube at the convenience of the operator. The output of the test amplifier is dissipated in a dummy load and a small portion of the output signal is applied to a panoramic analyzer having a dynamic range of 60 decibels. The two-tone test signal, along with spurious IM products, may be seen on the screen of the instrument, separated on the horizontal frequency axis by the difference in frequency between the two test signals (Fig. 5). A reading is made by comparing the amplitude of a specific intermodulation product with the amplitude of the two equal test tones in the output signal. For convenience, the ratio between *one* of the test signals and *one* of the IM products (there are always two of the same order) is read as a power ratio expressed in decibels below the test-signal level. It is equally correct, and the absolute answer is the same, if the ratio of the sum of the powers of the *two* test tones to the sum of the powers of the *two* IM products of the same order is used. It is equally valid to express IM relative to peak-envelope power, (p.e.p.) provided it is done by taking the ratio of p.e.p. to the square of the sum of the two IM products of the same order.[4] Referring IM to p.e.p. carries the additional information that the IM is specified for conditions of maximum signal level. Peak envelope power occurs when the two test tones are instantaneously in phase.

Measurements made on a wide variety of power tubes, from small to large, filamentary types and oxide cathode, triodes and tetrodes, in grid- and cathode-driven service, have shown conclusively that the magnitudes of the intermodulation distortion products are significantly affected by almost everything: changing heater or filament voltage by only a few per cent; slight shifts in bias voltage, idling current, screen voltage, plate or grid tuning; neutralization, loading — all these factors and others even more obscure enter into the determination of intermodulation distortion.

This might be a melancholy and discouraging picture, but it is a fact of life and is one of the

[4] Expressions of IM without reference to conditions of measurement and techniques are — as expressed by Poo-Bah in "The Mikado" — "merely corroborative detail, intended to give artistic verisimilitude to an otherwise bald and unconvincing narrative." Unfortunately, a trend seems to be developing in this direction. The reader is hereby warned.

Fig. 6—Intermodulation distortion products may be predicted mathematically. This universal family of IMD curves applies to all perfect tubes obeying the 3/2-power law. The curves are plots of IMD level (Y axis) referred to the driving signal expressed as a ratio of drive to operating bias. As the drive is increased, the various IMD products pass through maxima and minima. Misleading conclusions of amplifier performance may be drawn if the equipment happens to be tested near a cusp on the IMD curve, where a particular product drops to an extremely low level. The whole operating range of the equipment must be examined to draw a true picture of IMD performance.



major roadblocks in joint industry efforts (working through the auspices of the Electronic Industries Association with the active cooperation of the U.S. Navy) to set up standards and testing procedures in order to establish a common yardstick for all to follow in vacuum tube IMD testing, rating and equipment design.

Mathematical Analysis

IMD products may be calculated by several methods.[5] The results of different valid mathematical techniques are in good agreement with each other, and also agree in general with data obtained from two-tone tests conducted with the IMD analyzer. A theoretical family of IMD curves of a perfect tube obeying the 3/2-power law is shown in Fig. 6. This universal family of curves applies to all tubes, regardless of operating parameters or tube type. Changes in electrode potentials and circuit values (and even changes in tube type) will produce characteristic curves of this general configuration, but of course the signal level at which particular value of distortion occurs will be different in each case.

In Fig. 6 intermodulation distortion products, expressed in decibels below the output level of the tube, are plotted along the Y axis. The ratio of the two-tone driving signal $E_{g(max)}$ to operating bias, E_o (relative to cutoff voltage) is plotted along the X axis. When E_o is zero, the tube is biased at cutoff (class B). Ratios of $E_{g(max)}/E_o$ greater than one, but less than infinity, represent the possible range of class AB operation. Starting on the curve at the no-signal point ($E_{g(max)} = 0$), the IMD products are nonexistent. As $E_{g(max)}$ is increased, the IM products increase throughout the range of class-A operation and into the class AB region, until a maximum IM distortion figure for the 3rd-order products of about -30.7 decibels is reached at an $E_{g(max)}/E_o$ ratio of about 1.7. The 3rd-order product then drops to zero (minus infinity) again for a ratio of $E_{g(max)}/E_o$, of about 3.5, after which the IM product again increases, gradually rising to a level near -20 decibels for class-B operation. Fifth-order and 7th-order (and higher-order) products follow this same general behavior, compressed along the X-axis, and are shown in dotted lines on the graph.

The results of this theoretical study imply that the amount of intermodulation distortion in any vacuum tube that follows the basic 3/2-power law is predictable; further, that such distortion is inescapable and is independent of tube type. Moreover, the study indicates that the perfect 3/2-power tube will provide 3rd-order IM products no better than -20 to -30 decibels below maximum power output, and that the IM product varies markedly with drive level, dropping to zero at various points in the dynamic operating range. Thus, the perfect tube, obeying a fundamental law of physics, is a mediocre performer from a linearity point of view. As far as IM distortion goes, it is a poor device to use in equipment

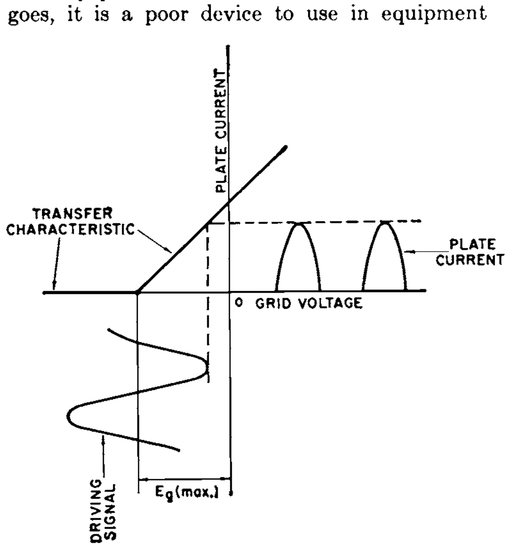


Fig. 7—An ideal tube transfer characteristic departs from the 3/2-power law. The ideal characteristic shown here consists of two linear portions, with the operating point set at the intersection. Half-wave plate current pulses are converted to sine waves by the flywheel effect of the plate tank circuit. Poor tank circuit Q, therefore, will have adverse effect on over-all linearity.

[5] "Approximate Intermodulation Distortion Analyses." Report CTR-173 by R. E. Cleary, Collins Radio Co., Cedar Rapids, Iowa; "Linear Power Amplifier Design," W. B. Bruene, *Electronics*, August, 1955; "Linearity Testing Techniques for SSB Equipment," Icenbice and Tellhaver *Proc. I.R.E.*, December, 1956, pages 1775-1782. "Intermodulation Distortion in High Powered Tuned Amplifiers," R. C. Cummings, Consultant, Eitel-McCullough, Inc., San Carlos, California.

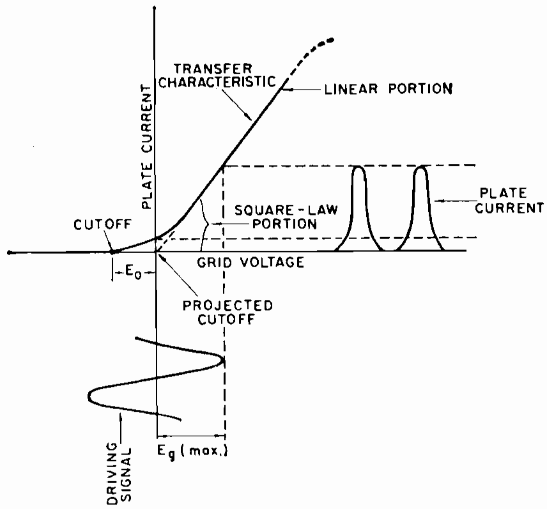


Fig. 8—Another ideal transfer characteristic for a linear tube consists of this form of curve, where the central portion is straight and the lower portion resembles a parabola. Practical tubes exhibit transfer characteristics of this general class, the upper portion of the curve showing additional curvature resulting from saturation of the electron stream in the grid-plate area of the tube. Plate current pulses are converted to sine waves by flywheel action of plate tank circuit.

designed for linear amplification of intelligence-bearing signals.

A Study of Practical Linear Amplifier Tubes

Does this theoretical study actually mean that all tubes are poor linear amplifiers or that it is impossible to achieve IM distortion products of a better order than −20 decibels? Not at all. The study concerns itself with a *perfect* tube that implicitly follows the 3/2-power law. Of course, there is no such device, and *practical* tubes (i.e.: tubes that can be manufactured) depart from this law to a greater or lesser extent. The practical tube, in general, shows an improvement in over-all linearity as a result of departure from the 3/2-power law. The practical tube, in addition, does not have a definite value of cutoff grid voltage, it does not have constant amplification at all points within the structure, and current deviations and amplification variations occur with changes in plate voltage. Current intercepted by the screen and control grids modifies the plate characteristic, and the "constants" that express the 3/2-power law vary with actual operating conditions. Theoretically, IM distortion as a result of this law should be independent of tube type. We know from experimental data that such is really not the case, as practical tubes exhibit transfer characteristics departing markedly from the 3/2-power law. In many instances, an improvement in linearity occurs when the tube departs from this law. For example, an ideal transfer characteristic for a tuned amplifier is shown in Fig. 7, consisting of two linear portions with the operating point set at the intersection. The resulting plate current consists of rectified and amplified half sine waves, the plate tank circuit converting this misshapen wave into an equivalent sine wave by virtue of the fly-wheel effect. The equivalent sine wave is directly proportional to the input signal at all amplitude levels from zero to the maximum value shown.

Alternatively, distortionless linear amplification may be achieved from another transfer characteristic having, instead of the discontinuity exhibited in the first example, a smooth curve of the form shown in Fig. 8. The operating point of the tube is chosen at projected cutoff. Ideally, the curved portion of the transfer characteristic should be a portion of a so-called "second-order" curve (a half-parabola, to be exact). A characteristic such as this is termed *square law*. Distortion products added to the exciting signal by such a

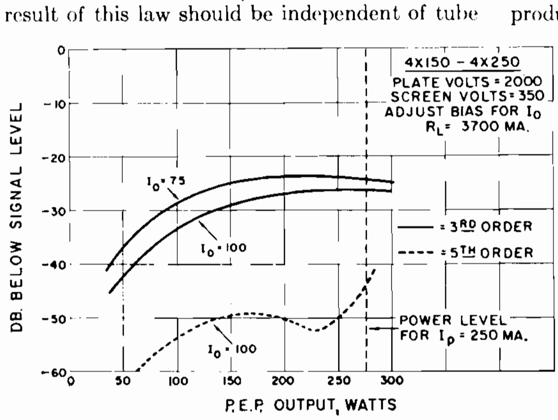


Fig. 9—A family of IMD curves for the 4X150–4X250 external-anode tubes. These curves are representative of this type of tube, and are typical for tubes made by different manufacturers. Intermodulation distortion products average about −25 decibels below peak signal, for 3rd-order products, while 5th-order products average −43 decibels below peak signal. These curves are representative of most small transmitting tubes of this type. Changes in loading or circuit parameters will alter shape and position of the curves.

curvature can be filtered out of the output signal by the tuned plate tank circuit *because all of these products fall in the harmonic regions of the exciting signal.* A distortionless replica of the input signal is thus available at the output circuit of the amplifier. Other transfer characteristics exist which also will provide lower-distortion output. Practical tubes departing from the 3/2-power law (wherein the exponent in the expression is 3/2, or — expressed as a decimal — 1.5) have exponents ranging from 1.3 to 3.4. This range covers quite a spectrum of possible tube performance! A practical tube may have a transfer-characteristic exponent falling somewhere between 1.5 (3/2-power law) and 2 (square law); its transfer characteristic would approximate the curve of Fig. 8, wherein the central portion is fairly linear and the lower portion resembles a parabola. The upper portion of the characteristic may show additional curvature resulting from saturation of the electron stream in the grid-plate area of the tube. That is to say, the grid or screen "robs" the plate of the greater portion of the available electrons and causes a corresponding drop in plate current.

Intermodulation tests run on tubes having this general transfer characteristic show distortion products generally in agreement with the 3/2-power law. Shown in Fig. 9 are IM curves based upon typical measurements made on the 4X150 — 4CX250 family of external-anode tubes. With fixed values of plate and screen potential and plate load impedance, measurements were made at two levels of resting plate current over the operating range of the tube. At the recommended value of resting plate current, the 3rd-order IM products rise gradually and smoothly as power is increased to the maximum value of 500 watts (referred to a single-tone plate current of 250 ma.) until at this value the products reach a level of −26 db. below the p.e.p. signal. Decreasing the resting plate current to 75 ma. will degrade the IM curve by several decibels, as shown. Fifth-order products at the recommended value of plate current are below −43 db. at maximum plate current level. The addition of 10 decibels of negative feedback to a circuit employing this style of tube will reduce the IM products below the values shown by approximately 10 db., so

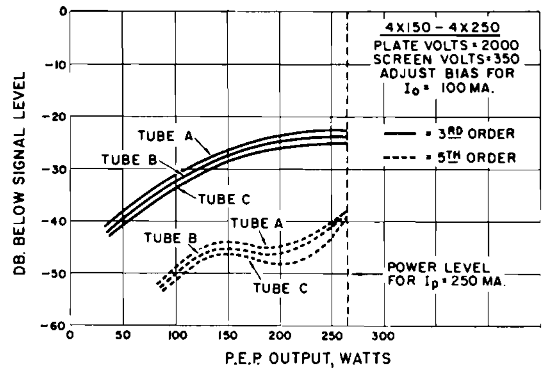


Fig. 10—Intermodulation distortion products vary from tube to tube of the same type, and also vary tube to tube as operating conditions are changed. Small "receiving-type" transmitting tubes are usually poorer than these curves by five to ten decibels.

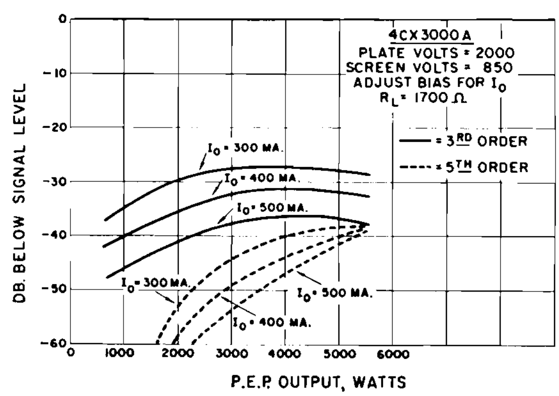
equipment with feedback designed around this tube (other factors being equal) should be able to reach the region of −35 db. IM distortion at full power. Individual tubes (and similar tubes made by different manufacturers) will vary from these curves by two to three decibels. Fig. 10 shows the variation in IM products between three tubes under fixed operating conditions. Changes in loading or other parameters will alter the shape and position of these curves.

Referring back to Fig. 6, tubes of this type are operated under conditions corresponding to a ratio of $E_{g(max)}/E_o$ in the range of 2 to 3 at maximum signal, and therefore distortion must pass through the third-order product maximum of about −31 db. within the operating range. Actually, maximum distortion appears near the 70% to 100% power level and is of the order of −25 db. or so. These curves are quite representative of most power tubes employed in amateur equipment, common varieties of transmitting tubes falling in the "minus twenty" to "minus thirty" decibel intermodulation range. Judicious use of feedback with these tubes will allow IM distortion products to fall in the "minus thirty" to "minus forty" decibel range.

Recent tubes specifically designed for linear

(Continued on page 154)

Fig. 11—Eimac 4CX3000A, specifically designed for linear-amplifier service shows substantially better IMD products by virtue of departure from 3/2-power law. With resting plate current of 500 ma., 3rd- order products are down better than 35 db. from peak signal level.



amplifier service show a decided improvement in IM distortion figures. Fig. 11 shows typical IM values for the 4CX3000A. Depending upon the value of resting plate current, the 4CX3000A is capable of delivering a p.e.p. output of 5.5 kilowatts with 3rd intermodulation products as low as −36 db. or more below the signal level. With the addition of ten or fifteen decibels of feedback, the construction of a high-power, high-quality linear amplifier having IM products in the region of −50 db. below the signal level becomes a reality.

Conclusion

The idealized vacuum tube is a 3/2-power-law device, and theoretically is incapable of low intermodulation distortion, except for low-efficiency class-A operation. By careful design of tube parameters aimed to remove the tube from this style of operation, a considerable improvement in linearity is possible. Even so, the circuit designer must follow the tube manufacturer's recommended operating conditions closely (voltages, loading, etc.) to achieve optimum performance, as these recommendations are usually based on exhaustive tests. Most of us, unfortunately, simply do not have comparable test setups to allow us to know what improvement or (more likely!) degradation will take place when operating conditions begin to wander, or when a desire to "push the tube to the limit" overcomes common sense.

As the state of the art advances, more and better tubes — designed for linear amplifier service — will appear on the market, making the equipment manufacturer's job an easier one, and helping the equipment operator to have a high-quality, low-distortion signal on the air. QST

Acknowledgements: *Grateful thanks to the following amateurs who offered helpful criticism and suggestions during the preparation of this article: Bob Morwood, K6GJF; Bob Sutherland, W6UOV; Ray Rinaudo, W6KEV; Bill Foote, K6BCM; Bill McAulay, W6KM; all of* Eitel-McCullough, Inc., *and Don Norgaard, W6VMH, of* Hewlett-Packard Co.

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amateur service newsletter
W6SAI

External Anode Tetrode Tubes

Developed by Eimac in 1947, the 4X150A external anode tetrode is the forerunner of a whole family of novel and highly efficient transmitting tubes. Now manufactured by a host of companies, the original 4X150A design is still going strong and is a favorite tube-type with many VHF-minded radio amateurs.

In the past dew years a bewildering number of "4X-variety" tubes have appeared on the market, each type having slightly different characteristics than "grandpappy 4X150A." The purpose of this article is to tabulate these tubes and to note basic characteristics of each type.

The Original 4X150A

The nomenclature of the 4X150A briefly describes the tube. The "4" signifies four active elements (a tetrode), the "X" indicates the external anode (forced-air cooled), "150" indicates 150 watts plate dissipation, and the final "A" indicates this is the first production version of the tube. Recent productions of the 4X150A have included a radically new brazed anode structure having complex cooling fins that allow increased plate dissipation ratings to be achieved. The new 4X150A is operationally equivalent to the 4CX250B for frequencies below 150 Mc. The different versions of the 4X150A are not easy to tell apart, as some transitional "150 watt" tubes have the improved cooling fins but do not have the brazed anode structure. In general, the "old-style" and "new-style" 4X150A tubes may be separated by weight: the 150 watt anode tube weighs approximately 5.2 ounces and the 250 watt anode tube weighs approximately 4 ounces. In addition, the center anode (plate) diameter of the older tube is 15/16" and that of the newer tube is 13/16".

High-voltage heater versions of the 4X150A were subsequently introduced as the 4X150D and 4X150S (ruggedized). These new tubes, plus the redesigned descendants of the 4X150A bear alternate Electronic Industries Association (E.I.A.) nomenclature consisting of four-digit numbers in the seven- and eight-thousand series. Thus, the E.I.A. 7034 is the 4X150A, etc.

The 4X150G (2.5 volt heater) and its offspring (4CX250K) have coaxial terminations and are designed for internal cavity operation at frequencies into the gigacle (kilo-megacycle) region.

The 4X250B Family

The 4X250B external anode tetrode is an evolutionary design featuring ceramic insulation and evolving from the glass-insulated 4X150A. Early 4X250B's were made with a ceramic outer cylinder and a glass base. Later versions are all ceramic. These "4X-series" tubes are rated at 250 watts plate dissipation and a maximum potential of 2000 volts.

The 4CX250B Family

The "4CX family" unilaterally replaces the "4X" ceramic tubes having a glass base. The all-ceramic 4CX250B tube has a plate dissipation rating of 250 watts and is manufactured in several versions as shown on the chart at the end of this article. Lesser known tubes in this family are the 4W300B (a water-cooled tetrode of interest to mobileers) and the Y-210 (a limited production tube having no anode cooler, intended for heat-sink or liquid immersion operation). The 4CX350A is a heavy-duty version of the 4CX250B having 350 watts plate dissipation and higher transconductance. The tube is particularly well suited for Class AB1 rf service.

The 4CX300A Family

The latest offspring of the 4X150A is the popular, rugged 4CX300A. This ceramic and metal, ruggedized tetrode is capable of operation at plate potentials up to 2500 volts. Special purpose versions of the 4CX-300A are currently in production and among these is the 4CX300Y which should be of considerable interest to "side-banders" and others contemplating new equipment. The 4CX300Y resembles the 4CX300A in appearance but has a heavy-duty heater (6.0 volts at 3.4 amperes) which permits unusually high values of plate current to be drawn.

Grounded Grid Operation of External Anode Tetrodes?

Modem, high-gain external anode tetrodes do not perform well when connected in the common "grounded grid" configuration. This family of tubes is charactdrized by high perveance, together with extremely small spacing between the grid bars, and between the grid structure and the cathode. Thus, while performing in an excellent fashion as grid-driven, high-gain tetrodes, these tetrodes are unsuited for pure grounded grid operation.

For proper operation of the external anode tetrode, the screen requires much larger voltages than the control grid. Older tetrodes having lower gain figures tend to have more equal balance between absolute grid and screen currents. When these electrodes of the newer, high perveance, external anode tetrode tubes are tied together, the control grid tends to draw tremendous currents and there is grave risk of destroying it. Peak grid current, for example, in a 4X150A operated in grounded grid configuration, can easily be twice the value of the peak plate current.

It is permissible, however, to operate the external anode tube as a cathode-driven tetrode, with the grid and screen elements at rf ground potentials, but operating at the normal dc potentials. Grid dissipation is minimal and stage gain is greatly increased. Screen dissipation is nearly the same as in the tetrode connection. Greater stage gain can be obtained with this circuit because the driver does not have to supply large screen and grid losses. If it is desired to dissipate some excess of driving power, it should be absorbed in a resistive load.

The Tube Socket

The tube socket for the external anode tube serves a triple purpose. It permits connection s to be made to the elements of the tube, it serves to conduct heat away from the stem of the tube and (in some cases) the socket serves as a capacitive bypass for the screen of the tube. Complete Air-System socket assemblies for all non-coaxial based external anode tubes are available, consisting of socket and air chimney, and these are tabulated at the end of this article. These sockets permit air to be blown axially on the base of the tube, past the base to the envelope, and then over the plate cooler. Use of other than an air socket with external anode tubes is not recommended, as tube temperatures cannot be adequately controlled. Use of a receiving-type loctal socket with 4X150A-style external anode tubes is emphatically not recommended. Dangerously high stem temperatures will be generated from the heat of the filament unless the base structure is cooled by an air blast, as the solid construction of the simple loctal socket blocks the normal flow of air about the tube stem.

It should be emphasized that the heater voltage on the "six volt" family of external anode tetrodes is 6.0 volts, plus or minus five percent, and not 6.3 volts. The operational range of heater voltage is 5.7 to 6.3 volts and for longest heater life, it is recommended that the heater voltage not exceed 6.0 volts.

The user of these tubes (or any other transmitting tubes) should check his heater voltage with a meter calibrated against a 1 percent laboratory meter of known accuracy. Monitoring voltage to five percent with a "garden variety" five percent ac meter proves nothing other than the fact the filament transformer is operating.

The Figure of Merit

A graphical presentation of the mutual conductance for common external anode tubes operating at various plate current levels is given at the end of this article. It can be seen that the 4CX350A and 4CX350F have about twice the transconductance of other various tubes, the 4CX300Y has approximately 30% higher transconductance, while the 7580 and the 7580/4CX259R are about 20% higher than the "common" 4X150A. Transconductance is a useful yardstick in determining the figure of merit of a particular tube (sometimes called the gain-bandwidth factor). It is calculated from:

$$\text{Figure of merit} = \frac{\text{Transconductance}}{2\pi C_t}$$

where C_t is the sum of the imput and output capacitance of the tube.

The figure of merit is a comparative figure and should not be interpreted as an absolute number. The input and output capacitances used in the calculation are average values taken from a number of typical tubes. Highest figure of merit vakues are reached by a combination of high transocnductance and low interelectrode capacitances. The 4CX350A and 4CX350F, having the highest transconductance and reasonably low interelectrode capacity, have the highest figure of merit, while the coaxial and breechblock-based tubes (with their higher capacitances) appear to have lower merit values. However, the coaxial-based tubes perform more efficiently at higher frequencies and are especially designed for cavity operation.

4CX300A FAMILY

EIMAC	EIA	P_o Watts	E_p Volts	F_{max} Mc	Heater V/A	Base	Figure of Merit	Notes
4CX300A	8167	300	2500	500	6.0/2.7	Breech-block	52	Ceramic-metal ruggedized
Y-180	-	300	2500	500	6.0/2.7	Breech-block	52	Nickel-Rhodium plated 4CX300A
4CX300Y	-	-	2500	500	6.0/3.4	Breech-block	53	High Plate current version of 4CX300A
4CX350A	8321	350	2000	500	6.0/3.0	9-pin	143	High transconductance, high current 4CX250B
4CX350F	8322	350	2000	500	26.5/0.57	9-pin	143	Aircraft version of 4CX350A
4CX125F	-	125	2000	500	26.5/0.57	9-pin	52	Aircraft version of 4CX125C

SPECIAL VERSIONS

4CN15A	-	15	2500	500	6.0/3.0	Breech-block	52	Low-duty pulse work, or heat sink cooling
4CN15L	-	15	2000	-	2.1/7.5	9-pin	-	Quick heat cathode
4CS100L	-	100	2000	-	2.1/7.5	9-pin	85	Quick heat cathode Heat sink cooling
4CX125C	-	125	2000	500	6.0/2.7	Breech-block	52	Horizontally finned 4CX300A
4CX125F	-	125	2000	500	26.5/0.56	Breech-block	52	Identical to 4CX125C except for Filament Voltage
4CX250L	-	250	2000	-	2.1/7.5	9-pin	-	Quick Heat Cathode
4CPX250K	-	250	5500	500	6.0/2.7	Coaxial	85	Pulse rated 4CX250K

4X150A FAMILY

EIMAC	EIA	P_o Watts	E_p Volts	F_{max} Mc	Heater V/A	Base	Figure of Merit	Notes
4X150A (old)	-	150	1250	500	6.0/2.6	9-pin	86	Old style anode Weight: 5.2 oz.
4X150A	7034	250	2000	500	6.0/2.6	9-pin	86	New style anode Weight: 4 oz.
4X150D	7035	250	2000	500	26.5/0.55	9-pin	86	Aircraft version of 4X150A
4X150G	8172	150	1250	1200	2.5/6.25	Coaxial	56	UHF and video service
4X150R	8296	250	2000	500	6.0/2.7	9-pin	73	Ruggedized 4X150A/7034
4X150S	8297	250	2000	500	26.5/0.56	9-pin	73	Ruggedized 4X150D/7035

4X250B FAMILY

EIMAC	EIA	P_o Watts	E_p Volts	F_{max} Mc	Heater V/A	Base	Figure of Merit	Notes
4X250B	-	250	2000	500	6.0/2.6	9-pin	85	Ceramic shell, glass based 4X150A
4X250F	-	250	2000	500	26.5/0.56	9-pin	85	Aircraft version of 4X250B
4CX250F	7204	250	2000	500	26.5/0.56	9-pin	85	All-ceramic 4X250F
4CX250B	7203	250	2000	500	6.0/2.6	9-pin	85	All-ceramic 4X250B
4CX250K	8245	250	2000	1200	6.0/2.6	Coaxial	54	UHF and video service
4CX250M	8246	250	2000	1200	26.5/0.56	"	54	Aircraft version of 4CX250K
4CX250R	7580W	250	2000	500	6.0/2.6	9-pin	81	Ruggedized 7580
7580W	7580	250	2000	500	6.0/2.6	9-pin	82	High perveance 4CX250B
4W300B	8249	300	2000	500	6.0/2.6	9-pin	85	Water-cooled 4X250B

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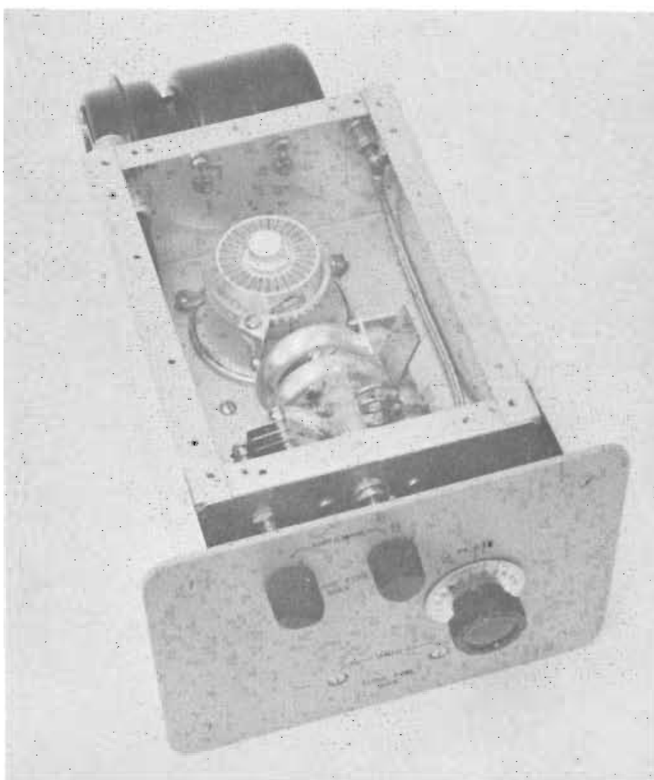
500 WATT AMPLIFIER FOR 144 MC USES EIMAC 4CX250B

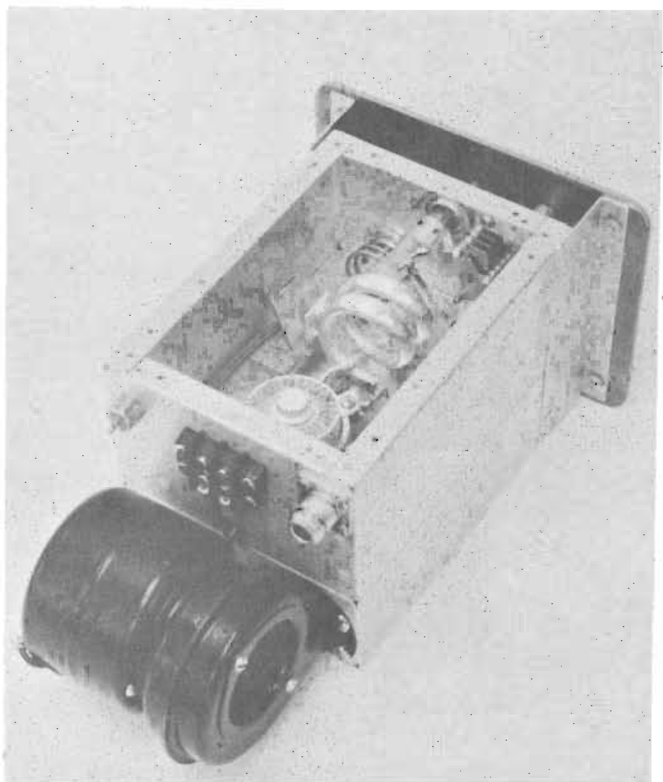
This compact amplifier runs 500 watts input in Class C or AB-1 linear service at 2 meters. Using a single 4CX250B, the amplifier measures 5'' wide, 7'' deep and 5'' high, exclusive of blower. Conventional shunt-fed, parallel tuned tank circuits are used, with input and output links series tuned to reduce unwanted reactance in the coupling circuits.

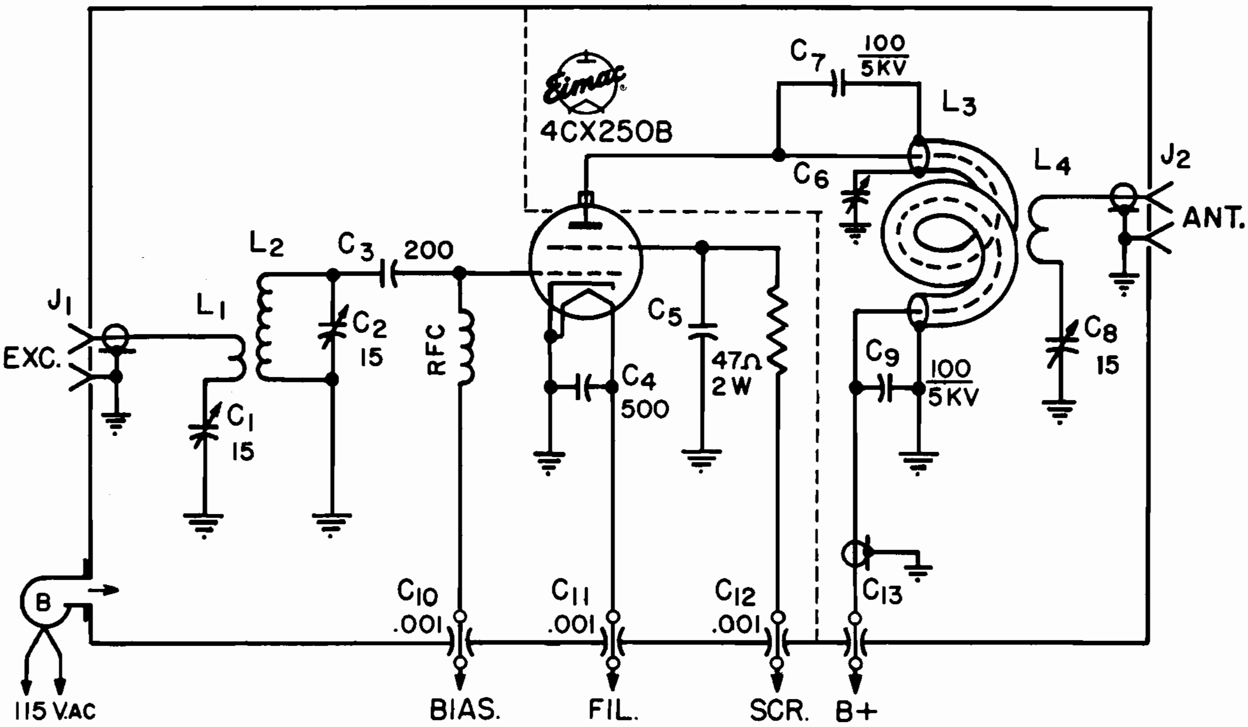
The plate coil is a dual inductor, the outer conductor made of copper tubing and the inner conductor of RG-225/U teflon coaxial line, with the outer braid removed. The copper tubing coil is grounded and dc is applied to the 4CX250B via the inner conductor, eliminating the plate rf choke which adds undesirable capacitance to the circuit and is an erratic performer at 144 Mc.

The RG-225/U is passed through the tubing before it is wound into a coil. The ground end of the tubing is soldered to a small copper plate which is tapper for an 8-32 grounding bolt. The plate blocking capacitor C-7 is bolted to a similar plate soldered to the opposite end of the tubing coil.

Tuning capacitor C-6 is made of two copper plates, one of which is soldered to the tubing coil about 1/2-turn from coupling capacitor C-7. The plate is positioned vertically. The rotor (ground) plate is soldered to a copper ring which encircles a rexolite drive shaft. The ring is grounded by a copper wiper bolted to the chassis. The ends of the antenna coil L-4 pass through holes drilled in a second rexolite rod and serve as terminal points for flexible leads running to capacitor C-8 and the coaxial antenna lead to receptacle J-2. The drive shafts are attached to reduction drive units mounted on the panel.







Schematic, 4CX250B Amplifier for 144 MC.

C-1, C-2, C-8: 15 μμf.
C-3: 200 μμf silver mica.
C-4: 500 μμf silver mica.
C-5: Part of EIMAC SK-600A socket.
C-6: Plate tuning capacitor. Two copper plates about ¾" x 1". One soldered to copper tubing coil in vertical position. The other soldered to a copper ring ring machined for close fit on ½-inch diam. rexolite shaft. Ring is grounded to chasis by wiper arm bolted to chassis.
C-7, C-9: 100 μμf, 5KV. Centralab 850S.
C-10, 11, 12: .001 μf, 500 volt feed-thru.
C-13: .001 μf feedthru, 2.5 KV. Erie 1270-010.

L-1: 5 turns #16, ¼" diam., ½" long.
L-2: 2 turns #16, 3/16" diam., adjust length to resonate with C-2 half-meshed.
L-3: 2 turns, 3/8" tubing, 1-¼" diam., 1-¼" long. See text.
L-4: 2 turns #14, 7/8" diam., ¼" long.
B: Dayton 2C-782.
J-1: UG-291/BU.
J-2: UG-58A/U.
RFC: Ohmite Z-144
Plate collet: EIMAC 008294
Cabinet: two chassis. Bottom 5" x 7" x 2". Top, 5" x 7" x 3" (aluminum).

Amplifier operation is conventional. Initial tuneup should be done at a plate potential of 600 volts or so (screen voltage reduced to 200). When resonance and loading points have been established, plate voltage may be raised to 2000 (1500 volts for modulated service). Typical data for Class AB-1 linear service are:

Plate Voltage	2000	Plate Input Power	500	watts
Screen Voltage	350	Plate Output Power	300	watts
Grid Voltage	-55	Max Screen Current	5	mA
Zero Signal Plate Current	100 mA	(Adjust bias to obtain stated zero signal plate current).		
Max Signal Plate Current	250 mA			

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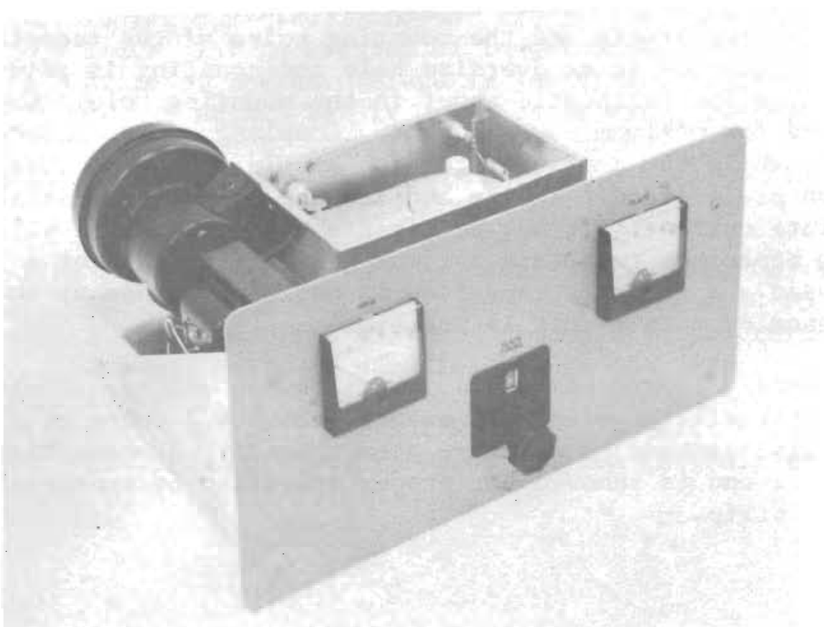
TWO KILOWATT 432 MC AMPLIFIER USES EIMAC 3CX1000A7

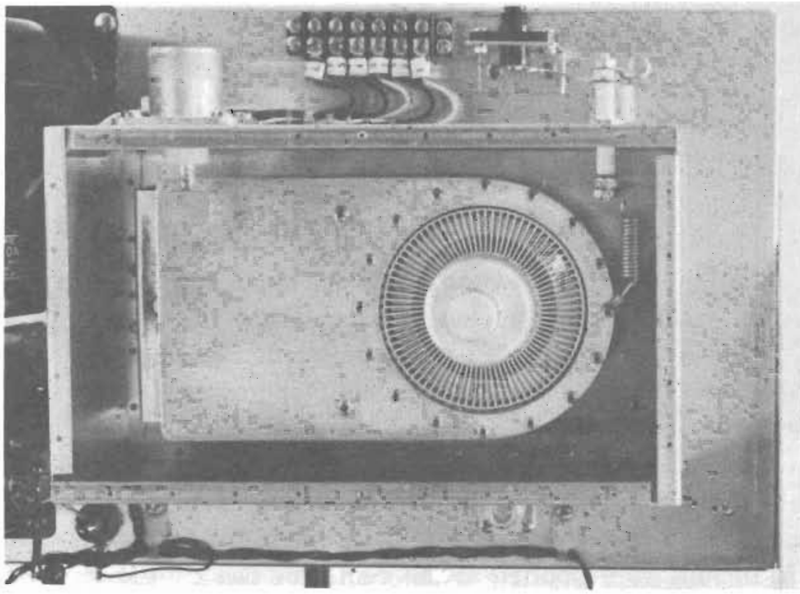
The 3CX1000A7 ceramic zero bias triode runs two kilowatts (PEP or continuous service) in this compact 432 Mc linear amplifier. The 3CX1000A7 is an external anode, ceramic version of the popular 3-1000Z. This experimental circuit uses a half wavelength strip line plate circuit and a tuned input configuration for efficient cathode driven (grounded grid) operation. Power gain of this amplifier is about 8 decibels and overall plate efficiency is approximately 50 percent.

The strip line plate inductor is an 1/8-inch thick copper rectangle 4-3/8'' wide and 7-3/4'' long. The rounded end of the plate encircles the anode of the 3CX1000A7 and is fastened to it by a matching copper collar having flexible finger stock silver-soldered to its inner circumference. The collar is bolted to the strip line which, in turn, is supported at the center by two 2'' high ceramic insulators. The end of the plate opposite the tube is tuned by means of a copper flipper hinged to the chassis and moved to and fro by an arm and worm gear arrangement shown in the illustration. The antenna circuit is capacitively coupled at this end of the strip line.

The EIMAC SK-870 Air Socket and Chimney grounds the multiple grid terminals of the 3CX1000-A7. Four of the filament terminals are bypassed to the ground ring of the socket with 220 $\mu\mu$f silver mica capacitors. The filament terminal to which drive is applied is bypassed to the socket ground bolts by a 3 $\mu\mu$f, 5 KV ceramic capacitor to neutralize the inductance of the internal filament leads of the tube, thereby reducing the s.w.r. on the coaxial line to the driver.



Zero bias triode 3CX1000A7 runs two kilowatts in 432 Mc strip line amplifier. Blower and filament transformer are mounted at end of chassis, with tube and plate circuit enclosed in box atop the chassis.
Plate tuning capacitor is worm driven by counter dial centered on panel.



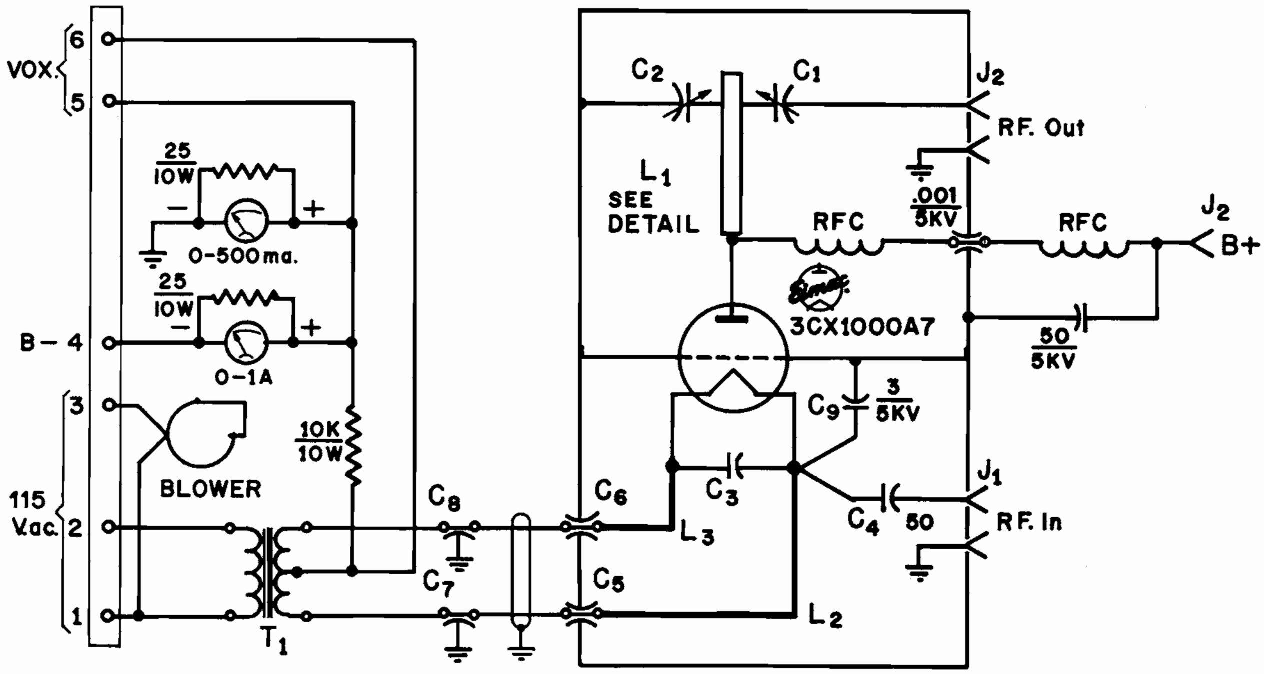
View of strip line circuit. The half wavelength line is supported by two ceramic insulators (see bolt heads near center of plate). Anode of 3CX1000A7 is encircled by copper collar bolted to strip line. Inner circumference of collar is lined with flexible finger stock. Plate r.f. choke is at left, and antenna capacitor plate is mounted to the coaxial plug at right. Edge of plate tuning capacitor is visible below strip line. Grid coaxial receptacle is behind enclosure, on chassis.

The resonant filament lines run from the socket terminals to feedthrough capacitors mounted on a nearby aluminum bracket. From this point, the leads run in shielded braid to the feedthrough capacitors mounted on the chassis deck.

The top plate of the above-chassis enclosure is made of perforated aluminum for proper ventilation. Air is blown into the pressurized under-chassis area and escapes via the tube socket, passing over the anode of the 3CX1000A7 and out the perforated top of the enclosure.

Amplifier Adjustment. The filament lines and plate line may be grid-dipped to 432 Mc with the 3CX1000A7 in the socket and no voltages applied. For initial adjustment a dummy load and power output meter are required, as plate current dip is not a true indicator of performance. Reduced plate voltage and grid drive are applied to the amplifier. The plate circuit is resonated and the coupling capacitor C-1 adjusted for maximum power output. The plate of capacitor C-1 is silver-soldered to the center pin of the coaxial receptacle and the mounting holes of the receptacle are slotted. The receptacle is mounted in an oversize hole and coupling is adjusted by loosening the bolts and moving the receptacle about in the mounting hole. Capacitors C-1 and C-2 are adjusted for maximum power output. Coupling exists between input and output circuits and, while the amplifier remains stable, the grid current varies abruptly depending upon plate circuit tuning and loading. Grid current should be about 40 percent of the plate current. It may be necessary to experiment with the value of the neutralizing capacitor to obtain optimum grid drive. Either a 3 $\mu\mu$f or a 5 $\mu\mu$f capacitor may be used. A variable capacitor is not recommended at this point as the internal inductance of such a unit is too high.

Once the amplifier is properly tuned and loaded, the filament voltage should be reduced to the minimum value that will provide full output (about 4.7 volts or so), as VHF backheating tends to raise cathode temperature above normal. Standby bias is incorporated in the amplifier and is removed for proper operation by shortening terminals 5 and 6 on the power strip.



SCHEMATIC, 3CX1000A7 AMPLIFIER FOR 432 MC

C-1: Coupling capacitor. Copper tab 1" x 5/8" spaced approximately 1/2" from plate line. Tab is supported by copper rod 0.188" diameter, soldered in center pin of coaxial receptacle J-2 (rod may be center conductor taken from RG-17/U coaxial line).

C-2: Tuning capacitor. Copper tab 1" x 4" spaced about 1/4" to 3/8" from plate line. Tab is portion of longer strip bent in an inverted "L" with brass hinge at bottom. Hinge is jumpered with copper shim to provide low impedance ground path. Tab is moved by an excentric arm and 3/8" diameter teflon drive rod driven by worm gear (see drawing).

C-3: 220 μμf dipped silver mica (4 required) mounted from heater terminals to socket ground ring.

C-4: 50 μμf. Centralab 850S-50Z.

C-5, C-6: 200 μμf, 30 amp. capacity. Erie 482-463-10.

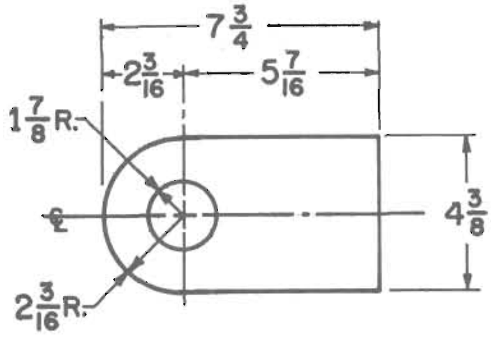
C-7, C-8: 0.01 μf, 30 amp. capacity. Sprague 80-P3.

C-9: 3 μμf. Centalab 855-3Z. Grounded to two thru-bolts of the socket assembly. Connect bolts in parallel and to one side of capacitor.

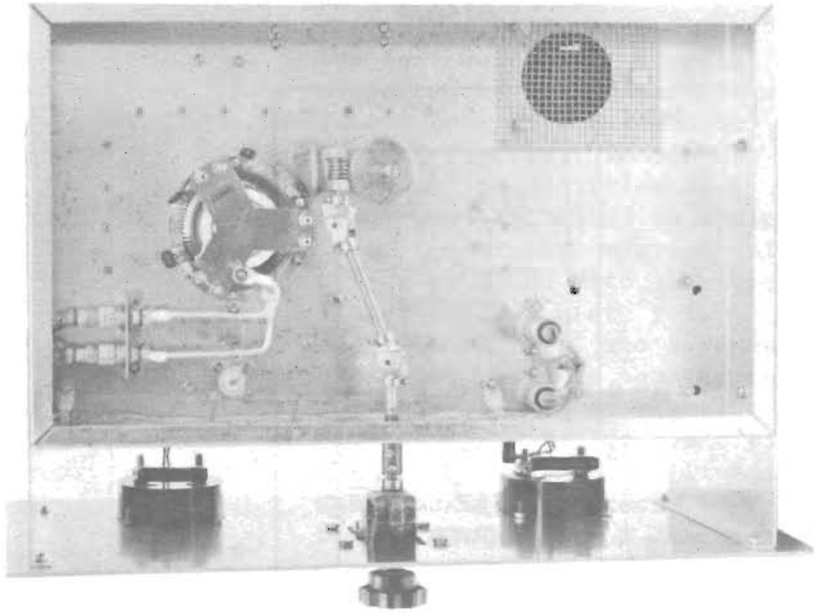
RFC: 15 turns #16, 1/4" diam., 1-1/8" long. Socket: Eimac SK-870.

J-1: UG-58/A. J-2: UG-352/U. T-1: 5 volts at 30 amp. Stancor P-6468.

B: Blower. 80 cubic ft./min. Dayton 1C-180. Box: 6" x 4" x 10".



L1 DETAIL



Filament lines and feedthrough capacitors are at lower left, with grid coupling capacitor below the lines. Filament bypass capacitors are soldered between ground ring of socket and filament terminals. Neutralizing capacitor is soldered between filament terminal (lower left of socket) and the two ground bolts atop socket flange. Plate circuit worm drive is to the side of the socket, and filament feedthrough capacitors are at right.

When antenna coupling is too heavy, resonance indication will be very broad and output will be low. When coupling is too light, a sharp resonance will be noted, combined with severe fluctuations in grid current as the plate circuit is tuned. When properly loaded, maximum power output will be achieved with the plate circuit slightly detuned from resonance point.

The output circuit of the amplifier is designed for RG-17/U coaxial line. Teflon dielectric RG-225/U is recommended for the drive line. At 2 KW input (3000 volts at 667 ma) power output is nearly 1000 watts measured at the load. Drive power is approximately 170 watts. Maximum test power was 3.2 KW (4000 volts at 800 ma) and power output was 1.5 KW. Plate dissipation was 1.7 KW, considerably over rated continuous value.

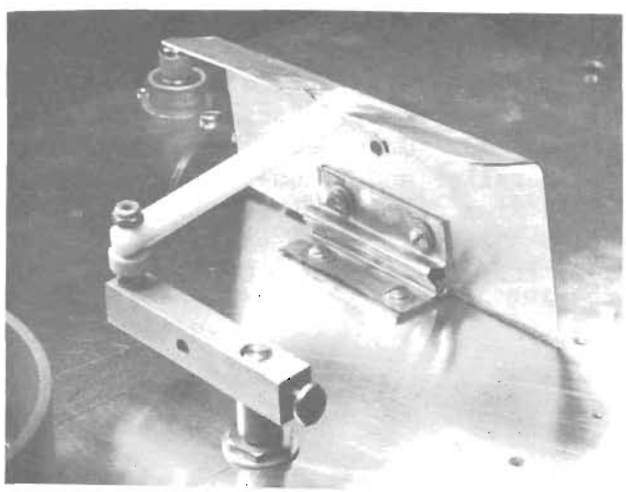


Plate tuning capacitor tab and drive mechanism.

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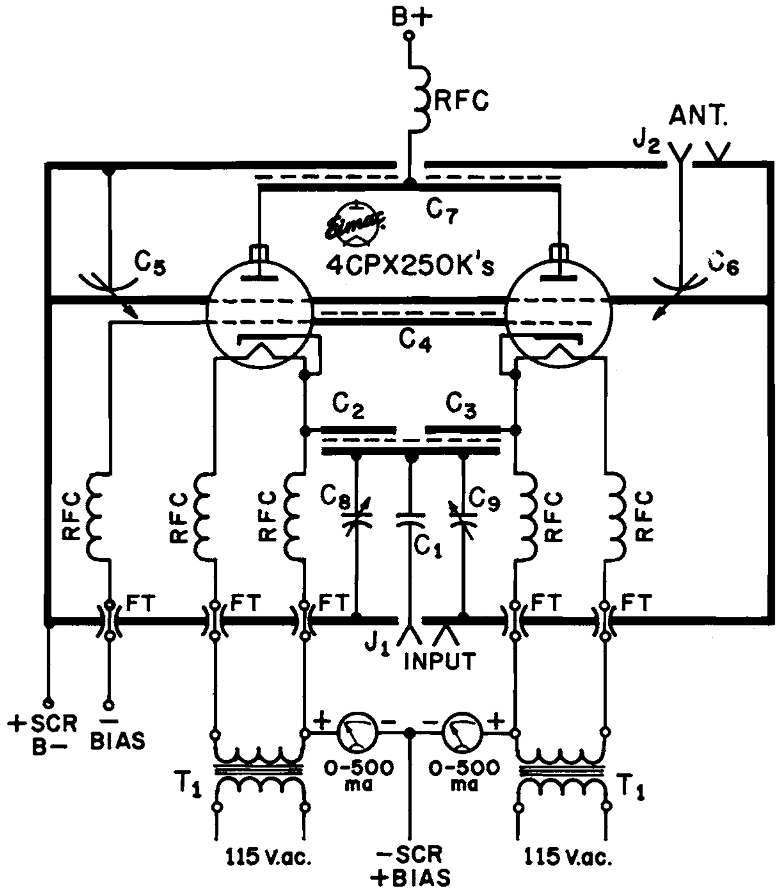
A cavity tank circuit is an attractive alternative to a coaxial or strip line circuit for 432 Mc operation, especially when it is desired to operate two tubes in parallel. The amplifier discussed here employs a plate circuit cavity and is designed for Class AB-1 linear service at the 1 KW PEP level at 432 Mc. Two EIMAC 4CPX250K UHF tetrodes are used in parallel in a cathode driven circuit. The 4CPX250K is a pulse service version of the 4CX250B having a coaxial base and rigidly controlled output capacitance. In addition, the ceramic insulator between screen and plate structures is longer in the pulse tube than in the 4CX250B. For amateur service the electrical ratings of the 4CX-250B apply to the 4CPX250K, and the use of the latter is recommended for UHF service.

The tubes are arranged in cathode driven service as the residual feedback path is degenerative rather than regenerative and the impedance of the path (cathode-plate capacitance) is high. The screens are placed at dc ground potential with screen and bias supplies "floating" below ground. Plate current flows through the screen supply which must be capable of passing 550 mA. As the supplies are in series, a lower voltage transformer for the plate supply may be used.

The amplifier plate cavity is constructed of .064" brass with inside dimensions of 10" x 12-½" x 1-5/16". The 4CPX250K's are spaced 2-½" center to center and the tube centers are 5" from each side of the cavity. The cavity tunes from 420 Mc to 454 Mc with a 2" disc capacitor without changing the unloaded Q (Q = 1600) of the cavity appreciably. If only one tube is used, the cavity should be square (13-½" x 13-½"). Capacitive coupling is used to the antenna circuit and teflon dielectric RG-87A/U cable is recommended for the antenna feed system.

The screen collets are soldered to a brass plate grounded to the bottom of the plate cavity. The grid collets are soldered to a second plate insulated from the cavity and which serves as a bypass capacitor. The cathode circuit is a half wavelength line using capacitive coupling from the driver.

Cathodes are in parallel at the drive frequency but are insulated from the cathode line to permit individual cathode current metering. Filament voltage is reduced slightly below normal to compensate for rf backheating of the cathodes.



SCHEMATIC, 432 MC CAVITY AMPLIFIER USING EIMAC 4CPX250K TETRODES

C-1: 1-1/8" copper disc soldered to J-1.
C-2, C-3: Dual capacitor. Two cathode plates (1-¼" x 5-¼") mounted on base plate (3-7/8" x 5-¼"). Dielectric is .008" thick pressed mica sheet. Cathode collet mounted to each cathode plate and filament pins mounted to base plate.
C-4: Screen capacitor (2-½" x 4-7/8"). Dielectric same as above.
C-5: 2" copper disc capacitor threaded to top of cavity.
C-6: 1-1/8" copper disc soldered to J-2.
C-7: Plate capacitor (3-½" x 10"). Dielectric same as above.
C-8, C-9: Same as C-1, threaded to cavity wall.
RFC: 7" of #16 wire wound into coil 1/8" diam., 1" long.
FT: 500 $\mu\mu$f, 500 working volts feedthru capacitor.
J-1: UG-87/U. J-2: UG-560/U.
T-1; T-2: 6.3 volt, 4 amp. Adjust primary voltage as per text.
Collets:
Plate Finger Stock
Instrument Specialties
97-360KS
Screen Connectors
EIMAC Part No. 149-004
Grid Connectors
Instrument Specialties
97-74S (One finger removed)
Cathode, EIMAC Dwg. #EB-041-003-0200/Filament, EIMAC Dwg. #EB-041-003-0100.

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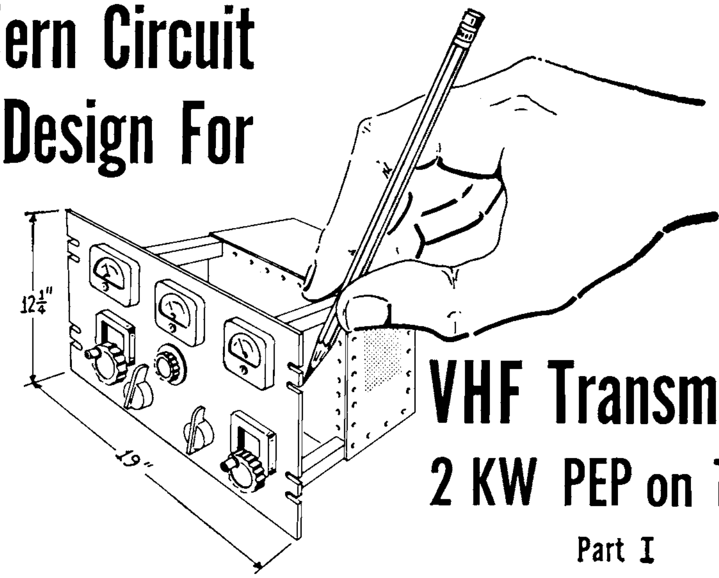
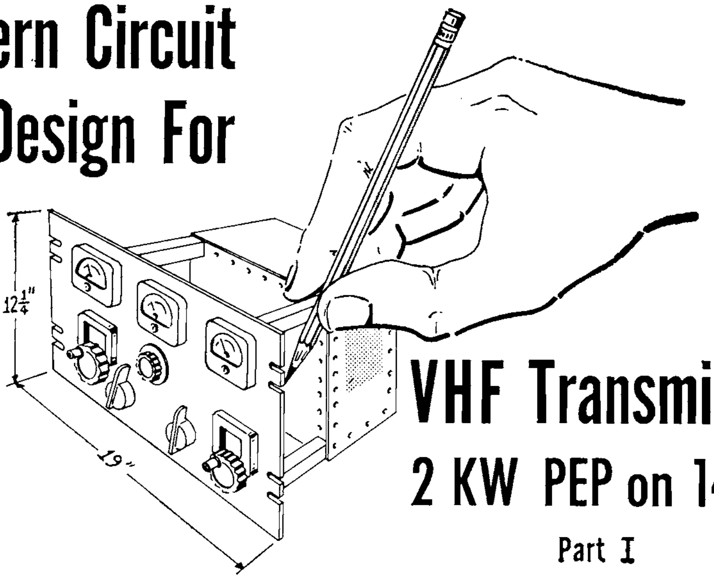
Modern Circuit Design For



VHF Transmitters
2 KW PEP on 144 mc
Part I

Modern Circuit Design For VHF Transmitters



2 KW PEP on 144 mc

Part I

BY H. C. BARBER,* W6GQK; W. I. ORR,† W6SAI; R. RINAUDO,† W6KEV AND R. SUTHERLAND,† W6UOV

Part I describes the design of a two kilowatt p.e.p. amplifier for the serious 2 meter operator. Requiring less than 10 watts of driving power, this efficient amplifier loafs along at the legal amateur power level. Designed for continuous service, a lot of power is packed behind a 12¼-inch relay rack panel. Construction details of this "powerhouse" will be featured in Part II of this two part series.

THE would-be designer of transmitting equipment for the v.h.f. spectrum soon finds that he is operating in a twilight area that falls between microwave techniques and practices associated with the high frequency (h.f.) bands. He realizes, sooner or later, that at some broad, undefined wavelength peculiar things start to happen to h.f. circuitry that otherwise looks deceptively simple on paper. As he progresses from the h.f. int othe v.h.f. region, the attentive amateur soon is aware that bypass capacitors no longer exhibit the normal characteristics they possess at lower frequencies. Short bits of wire assume importance beyond their size. R.f. tends to "leak" through small chassis holes. Components that seemingly are a passive part of normal communications hardware become complex devices that bedevil, and bear little resemblance to the comfortable components that make up h.f. gear.

Viewing the microwave (u.h.f.) region, the amateur finds a new world of waveguides and plumbing. Circuit techniques, equipments, and vocabulary fall into a strange category, and engineering philosophy and hardware of this frequency region are alien to h.f. concepts and techniques.

Between the comfortable world of kilocycles and the alien world of gigacycles are the amateur v.h.f. bands of 144, 220 and 432 mc. What techniques and practices should be used at these frequencies? Are microwave techniques applicable, or is this portion of the v.h.f. spectrum

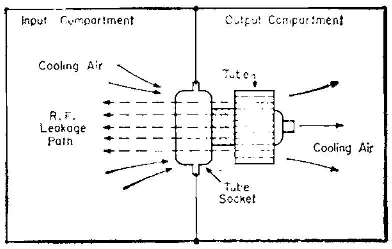


Fig. 1—Socket orfices to permit cooling air to pass across base seals of tube and through anode cooler may also create r.t. leakage path from output to input compartment. Simple "receiving type" sockets have little r.f. attenuation, while most "air-system" sockets afford 20 decibels or so of intra-stage isolation. The new Eimac SK-820 socket and 4CX1000K tube achieve better than 50 decibels of intra-stage isolation below 450 megacycles.

*45 Sherwood Court, Millbrae, California.
†Eitel-McCullough Inc., San Carlos, California.

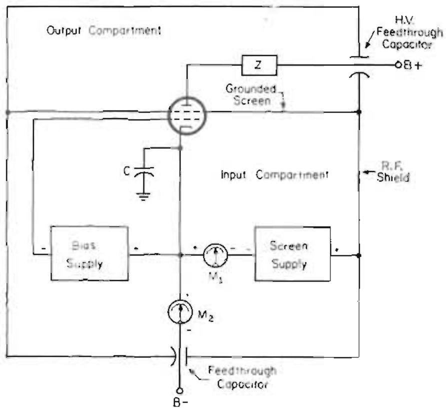


Fig. 2—To achieve maximum isolation between input and output compartments, the screen of the tube may be grounded, with bias and screen supplies placed below d.c. ground potential as shown in this circuit. The screen bypass capacitor is thereby omitted and a cathode bypass capacitor (C) substituted in its place. The B-minus lead is "negative" to the chassis by the amount of the screen voltage. Meter M_1 measures screen current and meter M_2 measures plate current.

merely an extension of the h.f. spectrum, with suitable modifications applied to equipment design and construction?

The answer to both these questions is obscured by realities. The v.h.f. amateur bands can and do use components designed for h.f. service, but these bands fall on the doorstep of the u.h.f. region wherein the components start to assume the size of the radio wave that is being generated. This physical congruence of wave and component calls for techniques and circuitry not normally associated with h.f. equipments.

One obvious solution to this problem is to reduce the size of the v.h.f. hardware so that the dimension of the radio wave is great compared with that of the components. This is commonly done; but there is a limit to this reduction technique, however, since no one has ever invented a way to shrink the *watt* in a corresponding manner. A limiting factor in the design of v.h.f. gear, therefore, is the ability of small components to radiate or otherwise dissipate the heat generated by the power dissipation of active components.

This natural law of thermodynamics becomes a limiting factor at v.h.f. in the design of a one kilowatt (2 kilowatt p.e.p.) amplifier for linear and c.w. service. To begin with, the number of tubes that will accept this power level at this frequency are but a handful. Tank circuits tend to disappear within the tubes, and the problem of dissipating five hundred watts or so within a shielded enclosure containing small tubes and tiny components poses a difficult mechanical design problem. Even at the relative "low" frequency of 144 mc, the use of "garden-variety" tubes and tank circuits provides a marginal solution.

Special tubes and hardware have been designed to work well at v.h.f., and by the proper combination of tube, hardware and circuitry, a reliable amplifier having the aforementioned power capability may be built that "tunes up just like 20 meters" and is capable of continuous, 24 hour-a-day operation. The design of such an amplifier is covered in this article, with some interesting asides directed to problem areas encountered in the construction of a practical amplifier.

Socket-Tube Intra-Stage Coupling

One source of potential trouble in the v.h.f. amplifier is *intra-stage r.f. coupling* (fig. 1). Passage of r.f. energy between two sealed metal compartments placed side by side is zero, but the introduction of power leads and cables in the compartments permits r.f. leakage; and the placement of a tube between the compartments allows the coupling of energy through the socket

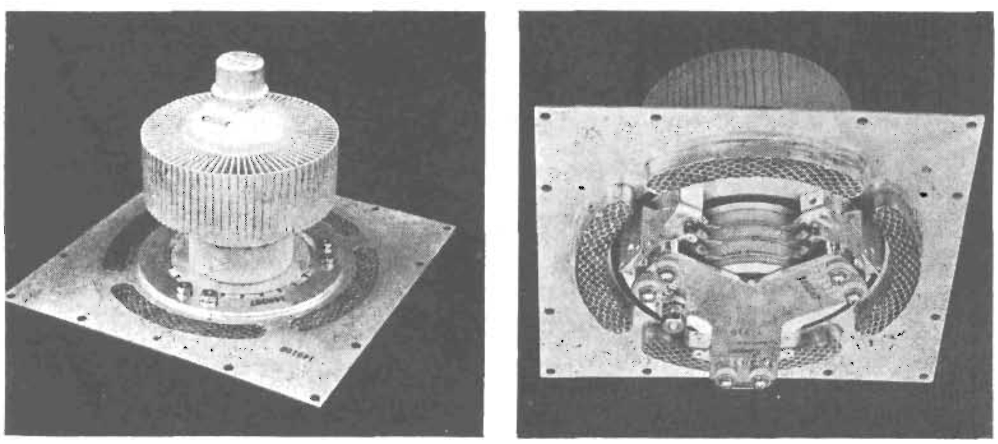


Fig. 3—(Left) Screen terminal of 4CX1000K tetrode is a ring that completely encircles the upper portion of the base structure of the tube. Flexible tabs on matching socket ring insure low impedance ground path from screen element as screen is run at d.c. ground. Cooling air is passed through the socket structure by means of "honeycomb" section of expanded metal (Right) built into socket plate which acts as a waveguide beyond cutoff frequency and provides high degree of attenuation below 450 mc. The socket provides 50 decibels or better intra-stage isolation in the v.h.f. region.

orfices, impairing the erstwhile perfect isolation between the compartments. It is possible to reduce intra-stage coupling via the power leads by proper bypassing and termination of the individual wires. Combining these techniques with the use of a new, improved tube socket and tube base, an intra-stage isolation of —50 decibels or higher may be achieved.

The degree of r.f. coupling between compartments may be determined by injecting a signal into the input compartment at the operating frequency of the gear, and then measuring the residual signal developed in the output compartment. For example, assume each compartment in the illustration has zero r.f. leakage other than the path through the tube and socket. This coupling path has an attenuation of —20 decibels, that is, the signal measured in the output compartment is 20 decibels lower in power than the signal injected in the input compartment. The particular tube in the socket has a power gain of 25 decibels, so it can be seen, without even turning the equipment on, that +5 decibels of positive coupling occurs through the socket-tube path and that the circuit will sustain self-oscillation at the frequency in question.

If the attenuation of the path is increased to greater than —25 decibels, oscillation may still occur under certain conditions of equipment tuning and loading wherein the tube gain may momentarily rise over the nominal figure. Many combinations of tubes and sockets normally used in the v.h.f. region exhibit a high degree of internal coupling since the screening of the tube is not perfect. Also, leads within the tube are long, bypass capacitors are imperfect at the design frequency, and the necessary air path through the socket permits the plate of the tube to "see" the input circuitry.

To provide the proper degree of isolation, the socket-tube intra-stage coupling should provide a degree of attenuation that is at least 20 decibels greater than the circuit gain of the stage, and the internal screening of the tube should be as excellent as the state of the art permits.

The "Grounded" Screen

In order to increase screen isolation at v.h.f. and to insure that the screen is as close to ground potential as possible, it is expedient to ground the screen to the chassis and to supply bias and screen potentials "below ground" as shown in fig. 2. The screen bypass capacitor is thereby eliminated and a cathode bypass capacitor substituted in its place. The screen capacitor usually employed in a tetrode stage is in a critical location in the v.h.f. circuit as it carries almost all of the plate circuit r.f. circulating current and, in addition, introduces regeneration in the circuit if the capacitor exhibits inductive reactance. When this cranky component is removed, the tetrode screen can be physically grounded, and the burden is transferred to the cathode bypass capacitor which carries little circulating current and tends to be a *degenerative* element if it exhibits inductive reactance.

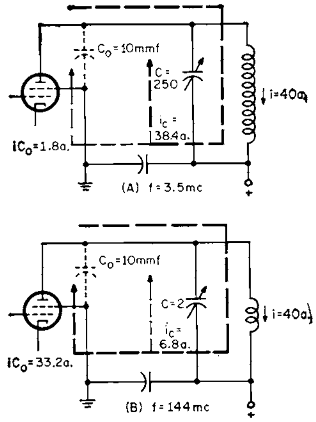


Fig. 4—At v.h.f., the output capacitance of the vacuum tube constitutes a larger proportion of the total tank circuit capacitance than in the h.f. region. Circulating r.f. tank current (i) divides through tube capacitance (C_0) and tuning capacitor (C) in proportion to capacitance of each. At 144 mc, current flowing through tube output capacitance (iC_0) is almost twenty times as great as at 3.5 mc.

Intra-stage r.f. leakage through the socket may be reduced by modiying the air vents between input and output terminations while allowing cooling air to pass across the base seals and through the tube anode.

A "Grounded Screen" Tube and Socket

Shown in fig. 3 are the new Eimac 4CX1000K tetrode and the companion SK-820 socket. The "K" tube is an improved version of the 4CX1000A having a low impedance screen grid terminal ring that completely encircles the upper portion of the base structure of the tube. The 4CX1000K is designed for improved input-output isolation and the screen terminal ring presses snugly against a circular spring-like grounding plate built into the top portion of the SK-820 socket. The screen is thus completely and securely grounded around its circumference by an extremely low impedance path, reducing r.f. intra-stage leakage through the socket to —50 decibels or better at 450 mc.

Cooling air is passed through the socket assembly by means of "honeycomb" sections of expanded metal built into the socket which act as simple waveguides beyond cutoff frequency, and provide a high degree of attenuation to r.f. currents passing between the grid and plate compartments.

Use of the 4CX1000K in the improved socket permits construction of a high power v.h.f. amplifier having excellent intra-stage isolation and that does not require neutralization. Thus, one of the big "trouble makers" in v.h.f. equipment design has been conquered!

The V.H.F. Plate Tank Circuit

The purpose of the v.h.f. plate tank circuit is twofold: First, it provides an impedance match between the tube and the load; and second, it

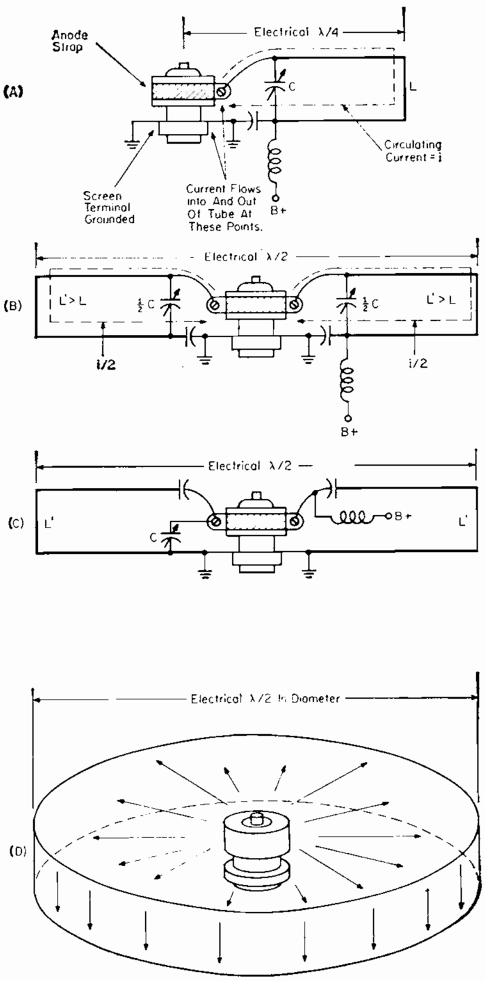


Fig. 5—Simple "quarter-wave" loaded tank (A) may be used as resonant circuit at 144 mc. However the circulating r.f. tank circuit flows into tube at one point and overheating of screen lead and vacuum seal may result from the high concentration of current. Divided tank circuit (B and C) places tube at the high potential point in half-wave line and divides circulating r.f. tank circuit between two sides of the tube. Limiting case is circular cavity (D) with tube placed at center and with circulating r.f. current evenly spread over inner surface of cavity. At 144 mc, the circuit of (C) is an efficient and practical compromise.

attenuates undesired spurious emissions. As the v.h.f. region is approached, the popular h.f. resonant tank composed of a parallel coil and capacitor shrinks in size until it is more practical to employ a different, larger resonant configuration. A capacitance-loaded section of transmission line may be used as a resonant circuit, since v.h.f. short-circuited transmission lines can efficiently replace inductances in tank circuits.[1] Such transmission line circuits exhibit considerably higher impedance at resonance and better Q than can be obtained with "lumped" circuitry. In general, the transmission line tank circuit may be designed for maximum efficiency, for maximum bandwidth, or for mechanical convenience. That is to say, given a reasonable efficiency and line impedance, the physical constants may be "juggled" about to permit a satisfactory circuit to be constructed of the material at hand so that odd sizes of tubing or unusual construction techniques need not be used.

[1] *"Fields and Waves in Modern Radio"*, by Ramo and Whinnery, Chap. 10. John Wiley and Sons.

Output Capacitance

In passing, it should be noted that the output capacitance (C_o) of a vacuum tube assumes great importance in the v.h.f. region as compared to the h.f. spectrum since it constitutes a larger percentage of the total tank circuit capacitance (fig. 4). At 3.5 mc, for example, a reasonable value of plate tuning capacitance (C) might be 250 mmf, of which 10 mmf (or 4%) represents the output capacitance (C_o) of the tube. At 144 mc, the total plate tank capacitance may be 12 mmf, and the tube contributes 83% of this capacitance. In a high-Q tank circuit, the circulating r.f. plate current may reach the order of 40 amperes at the kilowatt level and this current divides through the tube output capacitance and tuning capacitor in proportion to the capacitance of each. Thus, in the case of the 3.5 mc amplifier, 4% of the r.f. plate circuit current (1.6 ampere) flows through the output capacitance of the tube, whereas at 144 mc 83% of the current (33.2 amperes) flows through the tube output capacitance! In the case of a tetrode, the circulating current flows through the screen circuit as shown in fig. 4. The screen leads and screen bypass capacitor and a portion of the tank circuit must carry this extraordinarily heavy r.f. current. Elimination of the screen bypass capacitor removes this cranky component from the high current density path; however, the circulating r.f. current flowing through the tube terminals tends to heat the vacuum seals, often to such a temperature that the tube may be damaged. The current cannot be eliminated, but it may be distributed in such fashion as to minimize seal heating consistent with the shortest possible return path from plate to screen. Placing the screen at absolute ground potential, and using a low inductance screen terminal on the tube aids this task.

The Modified Coaxial Tank Circuit

The condition for resonance of a capacitance loaded transmission line is:

$$X_c = Z_o \tan L$$

Where X_c is the reactance of the loading capacitor, Z_o is the impedance of the transmission line, and L is the electrical length of the line in degrees: ($\lambda/4 = 90°$).

In the general v.h.f. situation, the loading capacitance is the output capacitance of the tube, and line parameters (Z_o and L) are adjusted so as to establish circuit resonance with this capacitance.

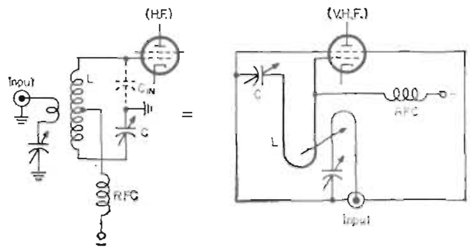


Fig. 6—"Split-stator" h.f. grid tank circuit may be duplicated at v.h.f. by the use of a loaded half-wave line. Input capacitance of tube makes use of quarter-wave line impractical, as the line tends to "disappear inside the tube".

A simple "quarter-wave" loaded line may be used as a resonant tank circuit (fig. 5A) but the circulating r.f. tank current flows through the tube via one small area and overheating of the screen terminal and vacuum seal in this particular area may result. It is possible, however, to achieve the same resonant frequency with one-half the tuning capacitance and a somewhat longer transmission line section (L') as shown in fig. 5B. Combining two of these circuits (fig. 5C) permits the use of the longer line (an electrical half-wavelength long) with the tube placed at the center. Circulating r.f. plate currents now flow through both sides of the tube, providing somewhat better current distribution than before. The longer line is easier to extract energy from because of its greater length, and it is less critical of construction. In addition, the physical mass of metal in the long line can dissipate heat more readily than can the smaller mass of the shorter quarter-wave line. The limiting design, of course, is when the tank circuit completely surrounds the tube in the form of a cavity (fig. 5D), wherein r.f. plate current is equally divided around the circumference of the tube. Such an approach is necessary in the upper regions of the v.h.f. spectrum, but at 2 meters, the simple half-wavelength line discussed here works well, and is simpler and cheaper to construct than the cavity configuration.

An optimum value of line impedance exists, which is a function of the dimensions of the resonant line. In the design of 5C the line is made in coaxial fashion, composed of a circular center conductor (made of a section of readily obtainable copper water pipe) and the outer walls of the shielded plate circuit enclosure. The impedance of a line of these rough dimensions is approximately 60 ohms, and the enclosure dimensions and spacing of the line are chosen to provide minimum surface area of the elements consistent with maximum volume, thus reducing the surface "skin resistance" of the elements.

The V.H.F. Grid Input Circuit

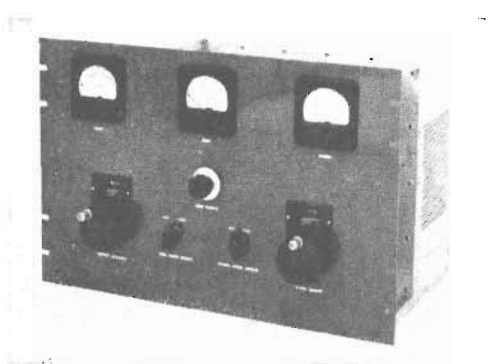
It is possible to use a quarter-wave line for the input circuit in the lower portion of the v.h.f. spectrum. However, the law of diminishing returns is at work: since the tube input capacitance does not shrink as the operating frequency is raised, the line must necessarily be drastically reduced in length to maintain resonance. Broad, low resistance circuits and wide contact areas are very fine for heat radiation but, by their very area, they add residual capacitance to any circuit built around them. Input capacitance of tubes designed for the v.h.f. region tends to run into a respectable figure, and some of the better tubes have input capacitances in the range of 50 to 100 mmf. The input capacitance is the sum of the grid-cathode, grid-screen, and grid-plate capacitances. Considerable ingenuity is demanded of the v.h.f. engineer to construct a high impedance tank resonated by such a value of capacitive loading, but it can be done, even though the quarter-wave tank tends to disappear within the tube!

An effective solution to this problem is to add a capacitance-shortened quarter wave line to lengthen the tank assembly to an electrical half wavelength (fig. 6). A form of "split stator" resonant circuit is achieved, and the driving signal may be easily coupled to the added tank section, as shown in the illustration. Low inductance leads are required, and the tank circuit may advantageously be made of wide copper strap, with multiple strap connections to the grid terminals of the tube to insure proper division of circulating current. A simple series tuned inductive loop can be used to efficiently couple the driving signal into the grid circuit of the v.h.f. amplifier. At best, grid drive is hard and expensive to obtain at v.h.f. and it is poor engineering practice to waste it!

A Practical V.H.F. Amplifier

A high power 144 mc amplifier making use of these design techniques and featuring the 4CX1000K and SK-820 socket will be described in the second part of this article.

[*To be continued*]

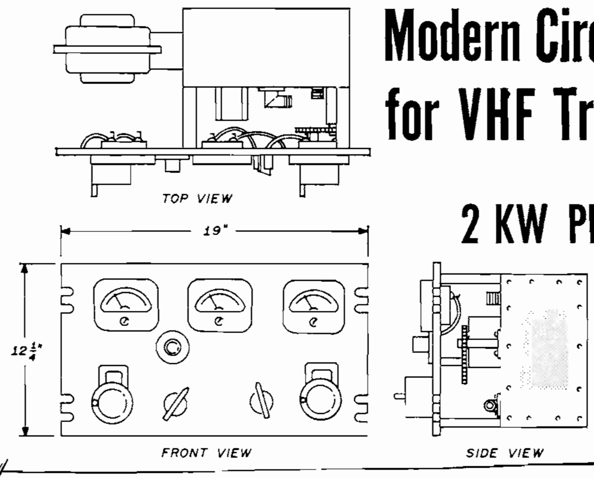


The next part will deal with the construction of the above amplifier. This is a front view of a compact two kw p.e.p. linear amplifier suitable for s.s.b., a.m., f.m., or c.w. It requires less than ten watts of drive to reach the maximum legal power.

Modern Circuit Design for VHF Transmitters

2 KW PEP on 144 mc

Part II



Modern Circuit Design for VHF Transmitters

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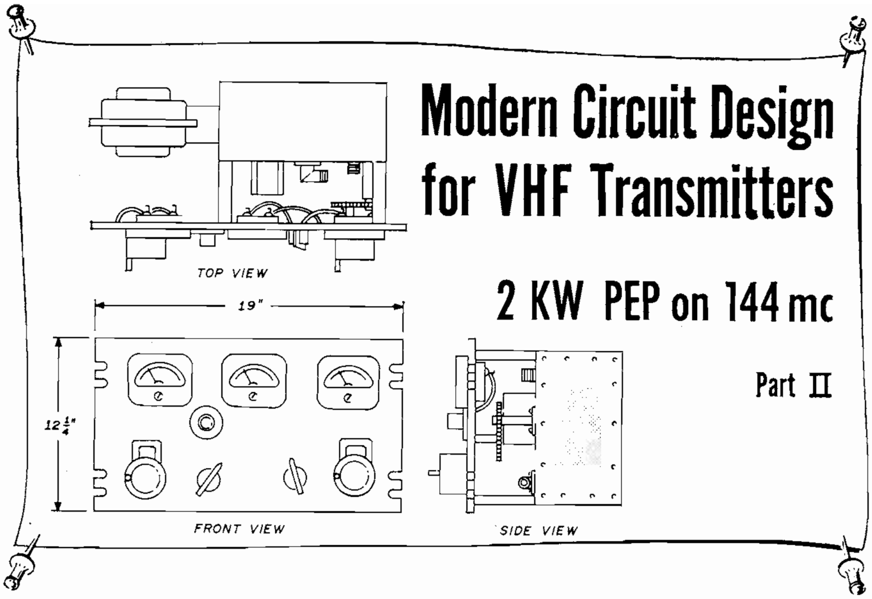

BY H. C. BARBER,* W6GQK; W. I. ORR,† W6SAI; R. RINAUDO,† W6KEV AND R. SUTHERLAND,† W6UOV.

Part II of this article describes the construction of a 2 kw p.e.p. linear amplifier that loafs along at the legal amateur power level and requires less than ten watts of drive.

The serious v.h.f. operator has a need for high power transmitting equipment if he is interested in Oscar IV, meteor scatter, moon bounce or other exotic forms of long-distance v.h.f. communication. In addition, if he contemplates s.s.b. operation, he may be desirous of running the "two kilowatt p.e.p." limit.[2] Equipment capable of achieving this power level on a reliable basis is scarce in the world of hamdom, although such gear is fairly commonplace in the commercial fields. Many instances exist where commercial gear has been patterned after equipment designed for radio amateur use, so perhaps there is justification for purely amateur gear designed around a proven item of commercial equipment.

This article describes a unique 144 mc high-power linear amplifier, patterned after a 5 kilowatt amplifier that has proven itself in space communication work in the v.h.f. spectrum. Scaled down to amateur size and power level, and making use of the new Eimac 4CX1000K v.h.f. tetrode, this compact continuous service kilowatt unit combines high power gain, stability and good linearity in a small package. This amplifier design, moreover, may be further scaled down for use at a lower power level, employing a single tube of the 4CX250B family. Requiring no neutralization, and straightforward in adjustment and tune-up, this linear amplifier is an interesting example of up-to-date v.h.f. design techniques. Even though many amateurs may not care to duplicate this unit in its entirety, many of the unique circuits and layout may be well adapted to other v.h.f. equipments.

The previous article[3] discussed some of the problems overcome in the design of a high power v.h.f. linear amplifier and the techniques that were employed to solve these problems. This article describes the construction of a 144 mc linear amplifier employing these previously discussed techniques and which is capable of continuous operation at a maximum power level of 2700 watts (p.e.p.). This margin allows a reasonable safety factor at the maximum amateur power input limit and insures that the equipment will not blow up at a crucial moment when an existing DX record is about to be shattered.

The 4CX1000K amplifier employs a half-wave linear plate tank circuit and the grounded screen

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[2]The F.C.C. definition of the so-called "two kilowatt p.e.p." limit is: "The maximum d.c. plate power input to the radio frequency tube or tubes supplying power to the antenna system of a single-sideband suppressed-carrier transmitter, as indicated by the usual plate voltmeter and plate milliammeter, shall be considered as the "input power" . . . provided the plate meters utilized have a time constant not in excess of approximately 0.25 second, and the linearity of the transmitter has been adjusted to prevent the generation of excessive sidebands. The "input power" shall not exceed one kilowatt on peaks as indicated by the plate meter readings."

[3]Barber et al., "2 KW PEP on 144 mc," *CQ*, Nov. 1965 page 30.

configuration discussed in the previous article. Overall amplifier efficiency in the Class AB-1 linear mode is unusually high for these frequencies, being of the order of 60 percent or better at a plate potential of 2500 volts. As the practical limit of plate efficiency of a Class AB-1 linear amplifier is no more than 68 percent, it can be seen that tank circuit losses of this unusual design consume only about 8 percent of the output power. Most garden variety v.h.f. tank circuits exhibit losses ranging from 20 to 50 percent.

At a plate potential of 3000 volts the power gain of this linear amplifier is over 22 decibels at the maximum power level, and the peak driving signal power is only 9.5 watts. At a two kilowatt p.e.p. level (3000 volts at 666 ma) the peak driving power is about 6.3 watts. For c.w. or f.m. service at one kilowatt input, the amplifier is capable of 625 watts power output with a driving power of 5.2 watts, and as an a.m. linear amplifier at the one kilowatt input level, a fully modulated 250 watts output may be obtained with only 1.25 watts of drive power.

Thus, in these classes of service, this compact "black box" is capable of better than 22 decibels power gain up to the maximum power level of 2700 p.e.p. on a continuous, 24 hour basis. Cost of all components is less than five hundred dollars, including the tube and socket, so on a watts-per-dollar basis, it is difficult to surpass the high power economy of this modern v.h.f. amplifier package.

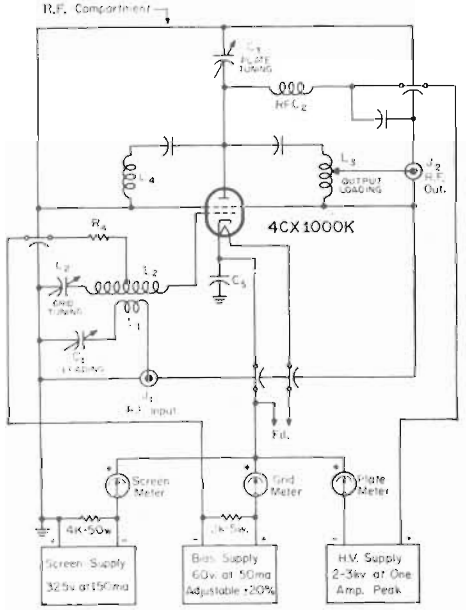


Fig. 7—Simplified schematic of 4CX1000K amplifier. The screen circuit of this v.h.f. amplifier is operated at d.c. and r.f. ground potential with bias and screen supplies placed "below chassis ground". The cathode capacitor (C_5) is an integral part of the Eimac SK-820 socket. Grid circuit components are placed in "r.f. tight" compartment and the special v.h.f. socket prevents intra-stage r.f. currents to flow between plate and grid circuits. The plate tank circuit (L_3, L_4) is a half-wave line with the 4CX1000K placed at the center. All power leads entering the compartments are suitably bypassed to keep the r.f. where it belongs.

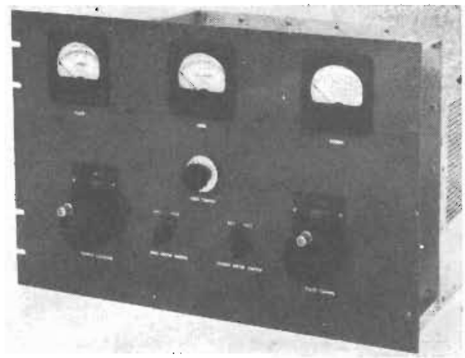


Fig. 8—Front view of a compact two kilowatt p.e.p. linear amplifier for the v.h.f. man. This high power 144 mc linear is suitable for s.s.b., a.m., f.m. or c.w. operation, requiring less than 10 watts drive power to reach the maximum legal amateur power level. The amplifier is mounted behind a standard 12¼" relay rack panel. Across the top of the panel (l. to r.) are the plate, grid and screen milliameters, with the grid tuning capacitor (C_2) centered below. At lower left and right are the counter dials for the OUTPUT LOADING control and PLATE TUNING (C_3). Between the counter dials are the grid and screen meter reversing switches. The amplifier enclosure is spaced behind the panel by sections of rectangular aluminum rod and the 4CX1000K mounts in a horizontal position within the enclosure. At the right end of the enclosure are the ventilation holes above the anode of the tube.

Circuit Features

A simplified circuit diagram is shown in fig. 7. The 4CX1000K power tetrode is operated with the screen at d.c. and r.f. ground potential, with the positive terminal of the screen supply grounded and the screen and bias supplies placed "below chassis ground." This is done to eliminate the screen r.f. bypass capacitor in order to achieve maximum intra-stage isolation. Because of the excellent isolation between input and output circuits, neutralization is neither desired nor required.

The plate circuit is composed of a half-wave, shunt-fed, inductor (L_3, L_4) with the tube positioned at the center of the line. The r.f. output loading tap is placed on one segment of the line. Resonance is established at the operating frequency by plate tuning capacitor C_3, which is a simple two plate assembly with one plate mounted to the anode cooler of the tube.

Two ceramic transmitting capacitors in parallel are used as blocking capacitors for each segment of the plate line inductors. So-called "TV-type" ceramic capacitors should *not* be used, as the circulating tank current is of the order of 40 amperes at maximum power level, and a large portion of this r.f. current flows through the blocking capacitors. The TV capacitors are not rated for such severe operating conditions.

The loaded Q of the plate tank circuit is determined almost entirely by the ratio of plate load impedance to reactance of the output capacitance of the tube. This capacitance is a fixed value and the load impedance is determined by the operating conditions imposed by the tube. Loaded Q therefore is relatively inflexible, although higher than normally encountered at lower frequencies. Even so, good tank circuit efficiency can be readily obtained if the unloaded Q is sufficiently high. It is in this case, being approximately 1000. Unloaded circuit Q of this magnitude is difficult to achieve at lower frequencies as the unloaded Q of good coils usually falls below 400. The loaded Q in this instance must be of the order of 10 to 20 to achieve a reasonable degree of tank circuit efficiency.[4] One disadvantage of high tank circuit Q (usually demanded by high-C v.h.f. tubes) is that a corresponding high value of r.f. circulating current flows. (Restricted bandwidth could also be a disadvantage in some applications). Wide leads having low r.f. resistance are required to carry this current. Ordinary lead solder is *out* as its r.f. resistance is too high, and high temperature silver solder is used for connections in v.h.f. tank circuits operating at power levels in excess of a few hundred watts. Alternatively, the joints may be physically bolted together with brass bolts, provided the r.f. current does not pass through the body of the bolt.

The coaxial output receptacle, J_2, is a *Type N* fitting, with the flexible strap jumper from the moveable tap on inductor L_3 bolted to the center pin of the fitting. Appreciable current flows through the jumper and a soldered joint at the receptacle has sufficient r.f. resistance at 144 mc to run hot enough to melt the solder at full amplifier input. The older SO-239 style coaxial receptacle is *not* recommended at this point, as the matching PL-259 coaxial plug does not make adequate contact to the shield of the coaxial line, which is only soldered at four points within the plug.

The resonant grid circuit is a half-wave line (L_2), capacitance loaded at one end by grid tuning capacitor C_2 and at the opposite end by the input capacitance of the 4CX1000K. Bias is applied to the tube through the grid line, via isolating resistor R_4 placed at the minimum r.f. voltage point of the circuit. Use of a v.h.f. choke at this point instead of the isolating resistor resulted in a violent 10 mc parasitic oscillation which is entirely absent when the resistor is employed.

The point of minimum r.f. voltage is at the electrical center of the half-wave grid line which happens to fall very close to the socket terminals. The input capacitance of the 4CX1000K is approximately 80 mmf and a simple parallel resonant tank tends to "disappear within the tube" at 144 mc. A half-wave line made up of the tube capacitance and lead inductance permits sufficient tank circuit external to the tube to efficiently couple excitation to the tube. To

[4]Tank circuit efficiency is defined as:

$$\text{Eff} = 100\frac{Q_u - Q_l}{Q_u}$$

Where Q_u is the unloaded Q, and Q_l is the loaded Q.

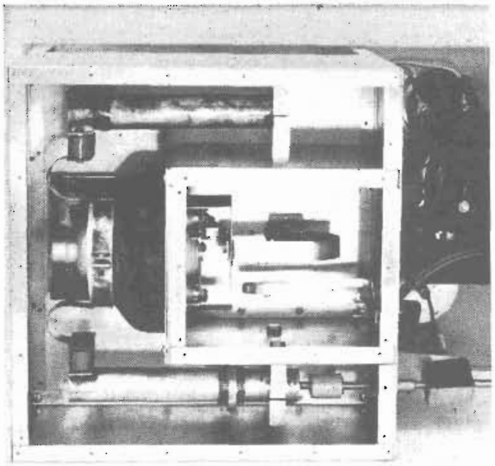


Fig. 9—Interior view of the amplifier enclosure with the rear plate removed. The enclosure normally mounted on its side, is shown in a vertical position. The 4CX1000K, air chimney, and special socket are placed atop the center of the grid compartment, with cooling air passing into the "bottom" of the compartment. The three grid terminals of the 4CX1000K socket are strapped together and the junction is bolted to the U-shaped grid inductor (L_2). The opposite end of L_2 terminates at the GRID TUNING capacitor, C_2, mounted to the front wall of the compartment. Below the grid assembly at one side is the input link (L_1) with LOADING CAPACITOR, C_1, mounted to the wall of the compartment. Mounting brackets for the two plate inductors are affixed to the outer side walls of the compartment. The squirrel-cage blower is mounted with a piece of screening across its mouth to restrict r.f. leakage through the vent opening. At the side of the blower is the right-angle drive and flexible shaft for the OUTPUT LOADING adjustment, visible on one anode inductor. The moveable tap is driven via a threaded rotary shaft integral to the plate line.[5]

satisfy this requirement, the grid line is made of copper strap and has an unloaded Q of about 300. Loaded Q is close to this figure as the grid circuit is swamped only by the circuit losses and input conductance of the 4CX1000K. Grid driving power is theoretically zero. In practice a few watts are required to develop the proper voltage at the grid of the tube because of the circuit losses.

Because of the high value of circuit Q and the appreciable capacitance of tuning capacitance C_2, grid circuit tuning is quite sharp and resonance must be reestablished for frequency excursions of more than a few hundred kilocycles.

Metering and Power Circuits

Separate d.c. milliammeters are used to mon-

[5]A set of blueprints is available at a cost of one dollar which covers, in detail, the various plate circuit components. Write: Amateur Service Dept., Eitel-McCullough, Inc., San Carlos, Calif.

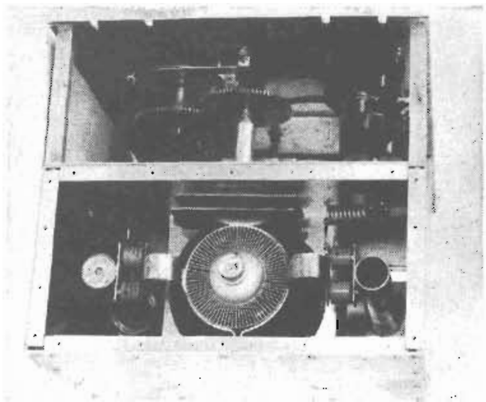


Fig. 10—View into top of plate circuit compartment with the perforated top removed. The PLATE TUNING capacitor (C_3) is immediately in front of the 4CX1000K, driven from the panel-mounted counter dial by two surplus gears. The moveable plate of the capacitor is grounded to the wall of the enclosure by a two-inch wide strap of flashing copper which flexes as the capacitor plate moves closer to the fixed anode plate. The dual plate blocking capacitors are mounted to the sides of the anode strap of the 4CX1000K, and visible at the right is the simple air-wound plate r.f. choke. The bearing of the rotary drive shaft for the OUTPUT LOADING tap is visible at the top of the left plate circuit inductor.

itor grid, screen and plate currents. Normal grid current of the 4CX1000K tetrode is zero in Class AB-1 service; however, small values of current flow during the tune-up process, and grid current excursions up to plus or minus 5 milliamperes may occur. In like fashion, screen current can be either positive or negative in value up to approximately 40 milliamperes or so. Negative current excursions are common in the grid and screen circuits of high gain tetrodes operating in the Class AB-1 mode. To accommodate positive and negative excursions of screen and grid current, zero center meters may be used or a polarity reversing switch can be employed with less expensive positive reading meters. The latter technique is used in this amplifier. Screen and grid metering are accomplished across shunt safety resistors, and rotary meter switches are used in lieu of toggle switches to insure low contact resistance. The safety resistors are placed across the switch arms rather than across the meters so that the circuits are not interrupted during the switching action.

Amplifier Cooling

The 4CX1000K tetrode installed in the SK-820 Air System socket requires an air blast of 25 cubic feet per minute for operation at maximum plate dissipation with inlet air temperatures up to 40 degrees Centigrade. This corresponds to a pressure difference of 0.2 inch of water column. A Dayton #1C-180 blower or Ripley #8472 blower will provide adequate cooling at this level of back pressure. The mouth of the blower is mounted to the under-chassis area as shown in the photograph and r.f. leakage through the blower air vent is prevented by covering the blower opening with a piece of screening. Cooling air passes from the blower into the grid compartment, through the honeycomb vents in the SK-820 air socket, and is conducted to the tube anode by means of a phenolic hood (Eimac SK-806) that slips over the top of the socket assembly. The air is exhausted from the plate circuit compartment through a perforated area in the enclosure located above the tube. The blower is energized as soon as filament voltage is applied to the tube.

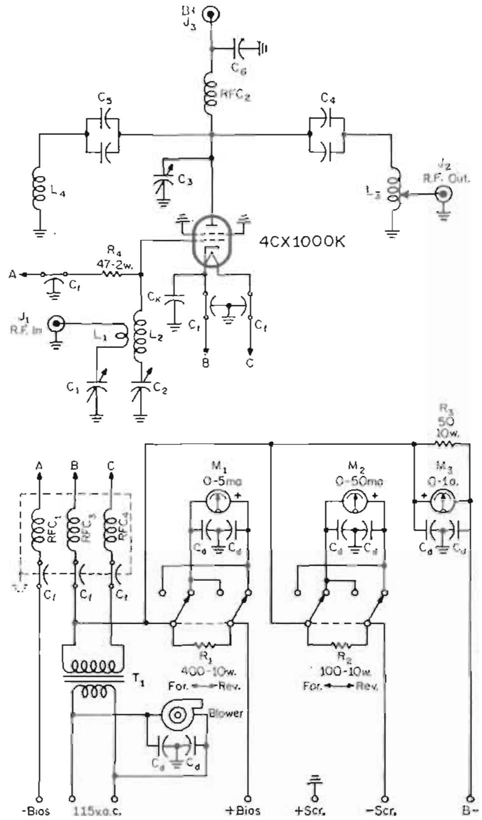


Fig. 11—Circuit of the 2 kw p.e.p. linear amplifier.
C_1—20 mmf, Hammarlund HF-20 or equiv.
C_2—15 mmf, Hammarlund HF-15X or equiv.
C_3—Approx. 3.5 mmf Plate size 2¼" × 4", spaced about ⅜". See text.
C_4—C_5: Each two 100 mmf, 5kv in parallel. Centralab 850S-100 or equiv.
C_6—1000 mmf, 5kv. Centralab 858S-1000 or equiv.
C_d—1000 mmf disc ceramic, 1.6 kv.
C_f—1500 mmf ceramic feedthru. Erie 327-X5U-152M or equiv.
BL_1—25 cubic feet/min. at 0.2" pressure. Dayton #1C-180 or equiv.
RFC_1—1.8 microhenries, Ohmite Z-144 or equiv.
RFC_2, RFC_3, RFC_4—9 t. #12, ½" dia., 1¼" long.
T_1—6.0 volts at 13 amperes. Stancor P-6463 or equiv.
J_1—Receptacle, BNC (UG-1094/U).
J_2—Receptacle, right angle, type N (UG-997A/U).
J_3—High voltage receptacle, Millen 37001 or equiv.
S_1, S_2—D.p.d.t. rotary switch, Centralab 1464 or equiv.

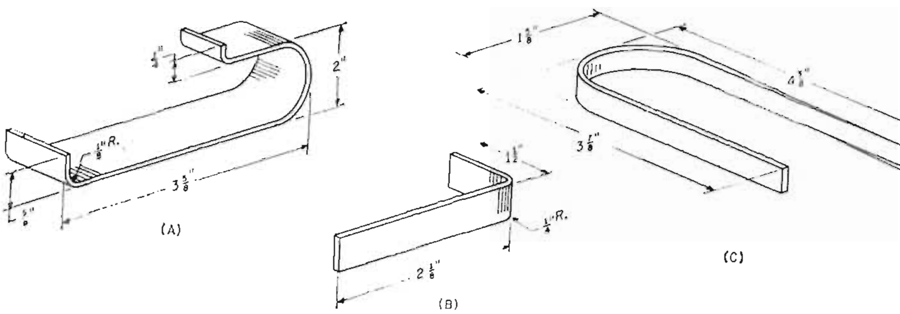


Fig. 12—Structural details for the grid circuit components. (A)—L_2, ⅝" strap, 0.032 material formed to dimensions shown. (B)—Grid connecting strap ⅝" wide, 0.064 material, 3 required. (C)—Input link, L_1, ¼" strap, 0.032 material.

Amplifier Enclosure

The amplifier mounts behind a standard 12¼" relay rack panel, with the 4CX1000K placed in a horizontal position behind the panel (fig. 9). The r.f. enclosure measures 12 inches square and is 6 inches deep. This aluminum enclosure is spaced 4½ inches behind the panel, and in this panel space are located the various meter switches, the meters and counter dials, the filament transformer and a small aluminum box containing the filament and grid isolation networks. The SK-820 air socket is mounted atop the grid compartment box which measures 6 inches wide, 7 inches high and 6 inches deep. The front and back panels of the amplifier enclosure also form the panels of the grid compartment box. As the enclosures are not standard size items, they were made up of 0.062 inches aluminum sheet, formed into simple panels.

Plate Circuit Assembly

The half-wave plate inductor is made up of two 7½ inch lengths of 1⅛ inch outside diameter copper water pipe (fig. 9). The pipes are clamped to the outer walls of the grid compartment box by means of heavy dural blocks, as shown in the photographs. A special sliding fitting is required on one plate line to allow adjustment of antenna coupling. This fitting consists of a copper ring which slides over the pipe and makes contact to the pipe by virtue of a circular section of flexible "finger stock" which is soldered to the inner circumference of the ring. The ring is driven along the length of the pipe by means of a threaded brass rod internal to the pipe, which is controlled from the front panel through an insulated coupling, a right-angle gear drive and a length of flexible shaft.[5]

The anode collar assembly is shown in fig. 13. It is made up of a circle of ¾-inch wide copper strap to which "ears" have been silver soldered. The ears form mounting brackets for the twin plate blocking capacitors. The stator plate of the variable tuning capacitor (C_3) is also silver soldered to this assembly.

The moveable plate of the tuning capacitor is panel driven by means of a simple bearing that translates rotary motion into thrust (fig. 10). The capacitor plate is prevented from turning by a wide ground strap as shown in fig. 10. To preserve panel symmetry, the capacitor is driven off-center from the counter dial by a set of surplus gears.

Grid Circuit Assembly

Three grid terminals are provided on the SK-820 socket and these are connected in parallel by ½ inch wide lengths of ⅝ inch wide copper strap silver soldered to the projecting socket terminals. The junction of the three straps is bolted to the grid inductor (L_2), which consists of a single length of copper strap bent into a hairpin which terminates at the stator lugs of the grid tuning capacitor, C_{12}. The composition isolating resistor R_4 is attached to the junction of the straps which is close to the point of minimum r.f. voltage in the grid circuit.

The input coupling loop (L_1) is about an inch away from the grid inductor and consists of a section of ¼ inch copper strap formed in a hairpin and connected between the input coaxial receptable (J_1) and the loading capacitor, C_1. Aside from the isolating resistor and the two tuned circuits, no other components are mounted within the grid circuit enclosure. Filament and bias leads exit via feedthrough capacitors into a small aluminum box which houses simple r.f. decoupling networks. With a stage gain of over 20 decibels, it is important that no

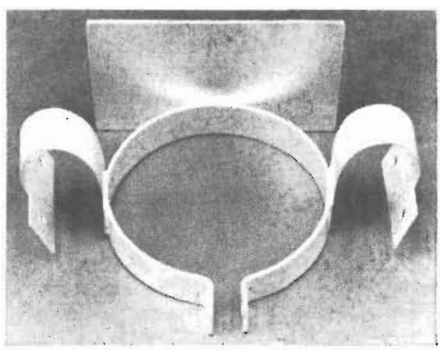


Fig. 13—Anode collar assembly. The anode assembly of the 4CX1000K is made of copper, silver soldered and silver plated. The circular strap encircles the anode cooler of the 4CX1000K and supports one plate of tuning capacitor C_3 and two "ears" which attach to the dual plate blocking capacitors. The capacitor plate measures 2¼" × 4".

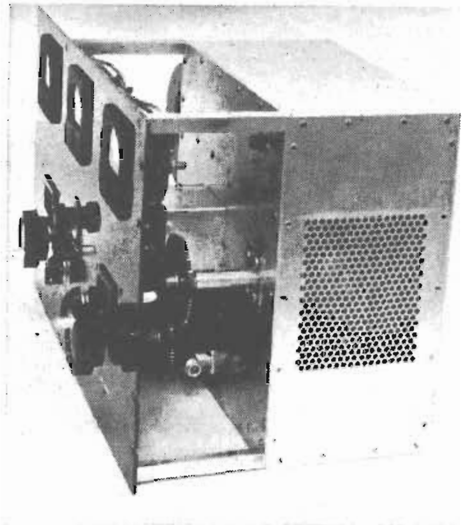


Fig. 14—Oblique view of the v.h.f. amplifier. The r.f. input fitting (J_1), plate circuit gears, and compartment for the power lead filters may be seen in this view. The B plus high voltage connector is located at the top of the enclosure. All meter leads are run in shielded wire, with the shields grounded at each end of the lead and the inner conductor bypassed to the shield with small disc ceramic capacitors at the meter terminals. The end of the plate compartment is perforated to allow exit of cooling air.

r.f. energy be permitted to leak into the grid enclosure. The air inlet at the base of the grid enclosure is covered with screening to reduce r.f. coupling through this opening. In addition, all power leads external to the enclosure are run in shielded braid which is grounded and bypassed at both ends of the lead.

Metering Circuits

Screen, grid and plate circuits are monitored separately by three panel meters. It is useful to observe the ratio of these currents during the tuning process and proper loading can readily be achieved when the operator has sufficient information to adjust the ratio of grid drive to plate impedance. Polarity reversing switches are incorporated in the grid and screen circuits in order to note negative currents often encountered when tetrodes are operated in the v.h.f. region. Meter circuits are returned to a common cathode point, as shown in fig. 7. The complete metering circuitry is shown in the general schematic, fig. 11.

Plate current is measured in the cathode return circuit to the high voltage power supply. It is necessary, therefore, that the negative terminal of the supply be isolated from ground. Both terminals of the bias supply are isolated from ground as is the negative terminal of the screen supply. The bias supply, in addition, is "hot" to ground.

Power Supplies

As is common with all power tetrodes the screen power supply of the 4CX1000K should be well regulated and the supply "bled" to 70 milliamperes or so. Dangerously high plate currents may flow if the screen power supply exhibits a rising voltage characteristic with negative values of screen current.

Voltage stabilization may be accomplished in several ways: a suitable bleeder resistor may be connected across the screen supply or an electronically regulated supply may be used. It is essential to use a bleeder if a *series* regulated supply is used, as such a supply exhibits a high input impedance to negative current. A well regulated, choke input supply capable of 150 milliamperes current capacity and bled to 70 ma is probably the best compromise solution.

The rated heater voltage for the 4CX1000K is 6.0 volts (*not* 6.3 volts). Tolerance is plus or minus 5%, but for longest tube life, the voltage should be held between 5.8 and 6.0 volts. The cathode and one side of the heater are internally connected. Heater voltage and blower should be turned on for three minutes before other operating voltages are applied.

Amplifier Tuning and Adjustment

For initial adjustment, a kilowatt dummy load is connected to the amplifier; bias voltage is applied first and set to approximately —60 volts.

Mode (Class)	Plate Volts	Plate Current (ma)	Plate Input (watts)	Peak Drive Pwr-watts	Power Output (watts)	Max. Grid Current	Max. Screen Current	Grid Bias* (Approx)
AB^1 (s.s.b.)	2000	883	1766	4.5	915	0	20	—53
AB^1 (s.s.b.)	2500	800	2000	4.4	1165	0	21	—55
AB^1 (s.s.b.)	3000	666	2000	6.3	1200	0	18	—56
AB^1 (a.m.)	2500	400	1000	1.2	250	—0.3	—29	—57
B (c.w.)	2500	400	1000	5.2	625	—2.5	21	—80

*Adjust bias for: Resting plate current of 250 ma for Class AB^1.
—Resting plate current of 1 ma for Class B.
—Resting plate current of 200 ma for a.m. linear service.

Fig. 15—Amplifier operating characteristics for the 4CX1000K.

Screen and plate potentials are applied *simultaneously* and the grid bias is adjusted for a resting plate current of 250 milliamperes. A small amount of r.f. drive is applied to the amplifier and the grid and plate circuits brought to resonance. The link loading capacitor C_1 is adjusted for lowest s.w.r. on the coaxial line from the driver. Driving power is slowly increased, while dipping and loading the plate tank circuit. Plate loading and grid drive are juggled until the desired parameters outlined in the "Amplifier Operating Characteristics" (fig. 15) are achieved. Grid and screen currents are sensitive indicators of amplifier operation,[6] and these, together with some form of r.f. output meter, will permit the operator to quickly adjust the amplifier for optimum operating parameters. For best linearity, the amplifier should be slightly overcoupled to the antenna so that the r.f. power output drops about two percent from maximum value at the point of proper loading. Under the two kilowatt p.e.p. rating, the amplifier will deliver 1200 watts to the antenna circuit, with a power gain of over 22 decibels. Intermodulation distortion products are low and the signal, when driven by a clean s.s.b. exciter is a pleasure to to listen to. Neutralization, of course, is not required and the amplifier tunes up "just like on the d.c. bands." ■

[6]Meacham, D., "Understanding Tetrode Screen Current," *QST*, July 1961, p. 26.

THE "STANLEY STEAMER"

A 2-KW. P.E.P. LINEAR AMPLIFIER WITH VAPOR-PHASE COOLING

Get acquainted with a new (in the amateur world, at least) method of cooling high-power tubes. Called vapor-phase cooling, it lets tubes operate at a lower temperature than the more-familiar cooling methods, requires no fans or pumps, and is completely silent in operation. The tube used in the amplifier described here is now commercially available, costs about the same as its air-cooled counterpart.

BY JACK QUINN,* W6MJG

FORCED-AIR cooling is "old hat" to modern radio amateurs who run the maximum power level; during the past decade tubes and components have diminished in size and convection-cooled tubes have given way to forced-air-cooled tubes in amateur gear. In some instances, water-cooled tubes have been featured in specialized equipment. In all cases, however, the nuisance of providing mechanical means of moving the coolant past the tube has been a major headache. Most blowers designed to move appreciable quantities of air have proven to be noisy at best, and a blower with a bad bearing or erratic rotor blades can be intolerable. Thus, by default, high-power operation has come to indicate noisy movement of air past the amplifier tubes.

A recent commercial development in the field of vacuum-tube cooling has been the Eimac vapor-phase cooling system. This article describes this system and illustrates its application in the design of a high-power linear amplifier for amateur service. The few people who have heard of this new cooling technique for transmitting tubes have ignored the principle of vapor-phase cooling by saying, "Oh, that's the system invented by some Frenchman — how can you cool power-grid-tube anodes with boiling water?" Not only can it be done, but in many cases it is far superior to either air or water cooling because it utilizes the highly efficient "latent heat of vaporization" principle.

In 1949 a French engineer, Charles Beurtheret of Compagnie Francaise Thomson Huston (CFTH) in Paris, France, applied the vaporization cooling principle to large external-anode transmitting tubes. He construeted a tube using a thick copper anode with pineapple-like fins, then immersed it in a water-filled boiler. As a result he was able to double the plate dissipation capability over that of a water-cooled tube, and more than triple that of an air-cooled transmitting tube. He also verified his prediction that by using this principle he could build a cooling system that had no pumps, blowers, fans, or, in fact, any moving parts or rotating machinery. Beurtheret's idea resulted in a less expensive, more efficient, and completely-silent cooling system. In 1951 a high-power broadcast station was built and sold to the French Government, the first station to be cooled by steam.

If you have a kilowatt linear, you know how annoying it can be to try to copy a weak c.w. signal over the noise of the air blower in the final amplifier. Most operators tolerate this nuisance as one of the penalties which result from running the legal power limit. When a vapor-cooled amplifier is put into service the major source of noise is eliminated. No motor is required to actuate the cooling system. However, this writer suddenly discovered the noise from the 75-cent fan motor in his exciter now sounded

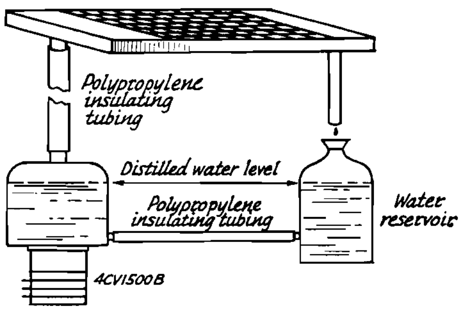


Fig. 1—The vapor-cooling "circuit." Steam generated by boiling water in the tank around the tube anode rises to the condenser, where it is cooled and converted back to water which drips into the reservoir.

*c/o Eimac, Division of Varian Power Grid Tube Division, San Carlos, California.

like a double-decker Greyhound bus in the Holland Tunnel! A pair of wire cutters quickly solved this last remaining objection. Then silence reigned supreme.

How Vapor-Phase Cooling Works

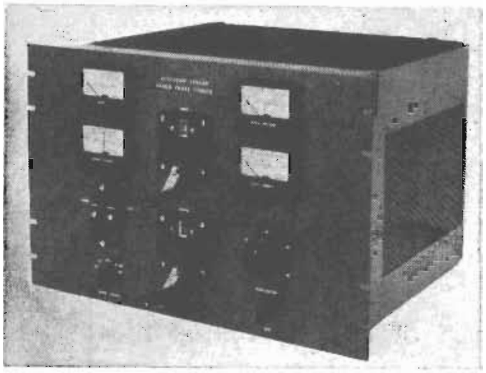
Conventional cooling systems have used forced air or circulating water as a heat-transfer medium. However, these methods have their limitations. Vapor-cooling of power tubes owes much of its appeal to its high heat-transfer efficiency, as shown in the following summary.

Comparison of Cooling Methods

Air — In a forced-air cooling system, air is forced past the external anode fins of the tube to absorb and dissipate the heat. Air is a relatively poor heat conductor, however, and in terms of power densities, forced-air systems are capable of removing only about 50 watts of power per square centimeter of effective internal anode area.

Water — Higher power densities are practical in water-cooled systems. Typically, circulating water removes approximately 100 watts per square centimeter. Thus, a power tube using circulating water as a heat-transfer medium is capable of approximately twice the plate dissipation rating of its air-cooled counterpart. Water temperature must be limited, however, so that steam is not generated inside the tube water jacket, causing localized hot spots which may destroy the tube. In practice, the temperature of water leaving the tube is limited to 70° C. to preclude the possibility of spot boiling. This heated water is then passed through a water-to-air or water-to-water heat exchanger where it is cooled to approximately 40° C. before being pumped over the tube anode again.

Vapor — Vapor-phase cooling systems eliminate some of the disadvantages of both systems by exploiting the latent heat of vaporization of water. Raising the temperature of one gram of water from 40° C. to 70° C. (as in a water system) requires 30 calories of energy. *Transforming one gram of water at 100° C. to steam vapor requires 540 calories.* In a vapor-cooling system, a given quantity of water will remove nearly *twenty times* as much energy as in a water-cooling system. Power densities as high as 500 watts per square centimeter of effective internal anode surface at atmospheric pressure have been attained through vapor-phase cooling. A typical vapor-phase cooling installation consists of a tube with a specially designed anode immersed in a "boiler" filled with distilled water (Fig. 1). When power is applied to the tube, anode dissipation heats the water to 100° C.: further applied energy causes the water to boil and to be converted into steam vapor. The hot vapor is passed through a condenser [1] where it gives up its energy and is converted back to the liquid state. This condensate is then returned to the boiler, completing the cycle.



Apart from the relay-rack panel and chassis, the vapor-phase-cooled 4CV1500B amplifier departs from the conventional in appearance. On top of the enclosure is a heat radiator constructed along the same lines as a car radiator. The water-level indicator and counter dials lend a different touch to the panel layout. Capable of an easy kilowatt average-d.c. input, the amplifier is silent in operation and the tube actually runs cooler than its air-cooled equivalent would with forced-air cooling.

[1] A *capacitor* stores electrical energy. A *condenser* converts steam to water.

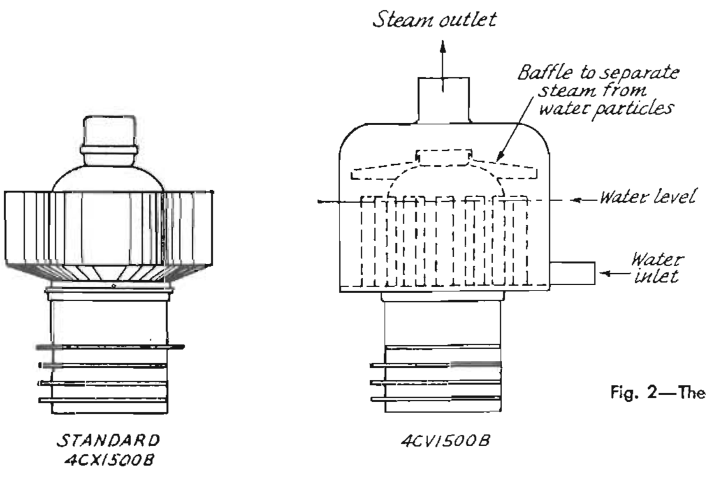


Fig. 2—The "boiler" for the 4CV1500B.

Table 1
RECOMMENDED OPERATING CONDITIONS

Plate Voltage	2900 v.d.c.	2500 v.	2000 v.
Plate Current (static)	300 ma.	300 ma.	300 ma.
Screen Voltage	225 v.	200 v.	225 v.
Control Grid Bias	−34.0 v.	−31.5 v.	−37.0 v.
Plate Load Resistance	2200 ohms	2200 ohms	1600 ohms

R.F. Output Impedance = 52 ohms
Filament 6.0 v. a.c. at 10 amp.

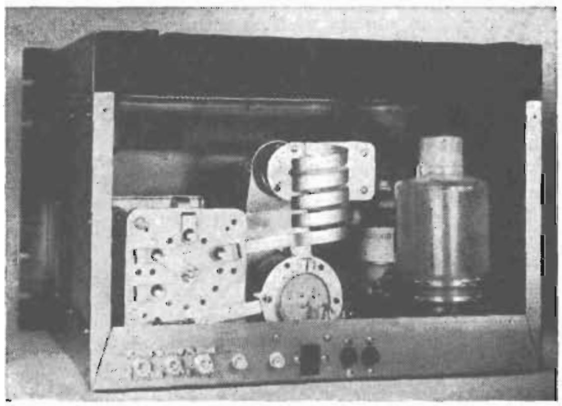
R.F. drive should be adjusted in all cases to a plate-current level of 1 kw. input on c.w. or 2 kw. p.e.p. on single sideband.

150 and 200° C. A properly designed vapor-phase-cooled anode at rated dissipation operates at between 100 and 115° C. maximum. Strange as it may sound, a steam-cooled anode actually runs cooler than most air-cooled tubes.

Amplifier Construction

The front is a standard dural aluminum relay-rack panel measuring 12¼ x 19 x ⅛ inches. The cabinet is 16½ inches wide, 12 inches high and 15 inches deep. A 3-inch section directly behind the front panel furnishes a shielded enclosure for the meters, filament transformer, bias control, meter switch, and water-level indicator as well as the pi-network input and output dial mechanisms. To eliminate TVI, shielded conductors pass from the terminal box at the rear of the cabinet through 1000-pf. feed-through capacitors, then through ½-inch conduit to the section behind the front panel and to the 4CV-1500B tube subchassis.

The front panel contains four 3-inch square meters (Weston Model 1921, black bakelite case). Three are used for monitoring plate and screen currents and plate voltage. The fourth is a 0-1 milliammeter which can be switched to read control-grid current, bias voltage, screen voltage, or to sample the rectified 50-ohm r.f. output voltage. There are two counter dials, for tuning the input and output capacitors of the pi-network. A band switch, bias control, meter switch, and water level gage complete the front-panel layout.



The rear view shows the radiator in place at the top, with the reservoir at the right and the amplifier tube partly concealed by it. The plate tank coil and vacuum variables are easily recognized.

The schematic and parts list in Fig. 3 shows that the r.f. input circuit is untuned and utilizes a 100-ohm 50-watt noninductive wire-wound resistor. Amplifier neutralization is not required when terminating the r.f. drive into this resistor, and amplifier stability is excellent. Only 15–20 watts peak drive is required to produce the full 1-kw. average or 2-kw. p.e.p. input power. A standard r.f. attenuator pad should be used between a 150–200 watt exciter and the amplifier input circuit.

The plate tank is a B&W Model 852 inductor with a "piggy-back" rotary switch coupled to the shaft to automatically select the proper antenna. (This switch was added after the photographs were taken.) Coax connectors are provided on the rear of the cabinet to accommodate 10-15-20-, 40- and 80-meter antennas. A Jennings Model RJ1A vacuum switch serves as the antenna change-over relay to feed the proper antenna back to the receiver. No forced-air filament-seal cooling is required for the tube as it too, is cooled by natural convection and operates at less than 200° C. if proper ventilation is provided.

The tube socket is mounted on a subchassis plenum box. The amplifier is rack mounted, and the area directly under the tube subchassis is open, serving as the air intake. A piece of perforated aluminum covers this opening. Care should be taken in furnishing sufficient air to maintain seal temperatures at or below 200° C. "Templac" colored wax painted on these areas will indicate the temperatures.

This 4CV1500B linear amplifier is the ultimate in amateur equipment. The low-intermodulation-distortion tetrode produces a sharp, clean transmitted signal. The elimination of the air blower noises enables the amateur to receive weak DX signals in the complete silence of the ham shack.

A special debt of gratitude goes to Bob Sutherland, W6UOV, for the use of his shop tools in making the various brackets and cabinetry and for his assistance and words of encouragement.

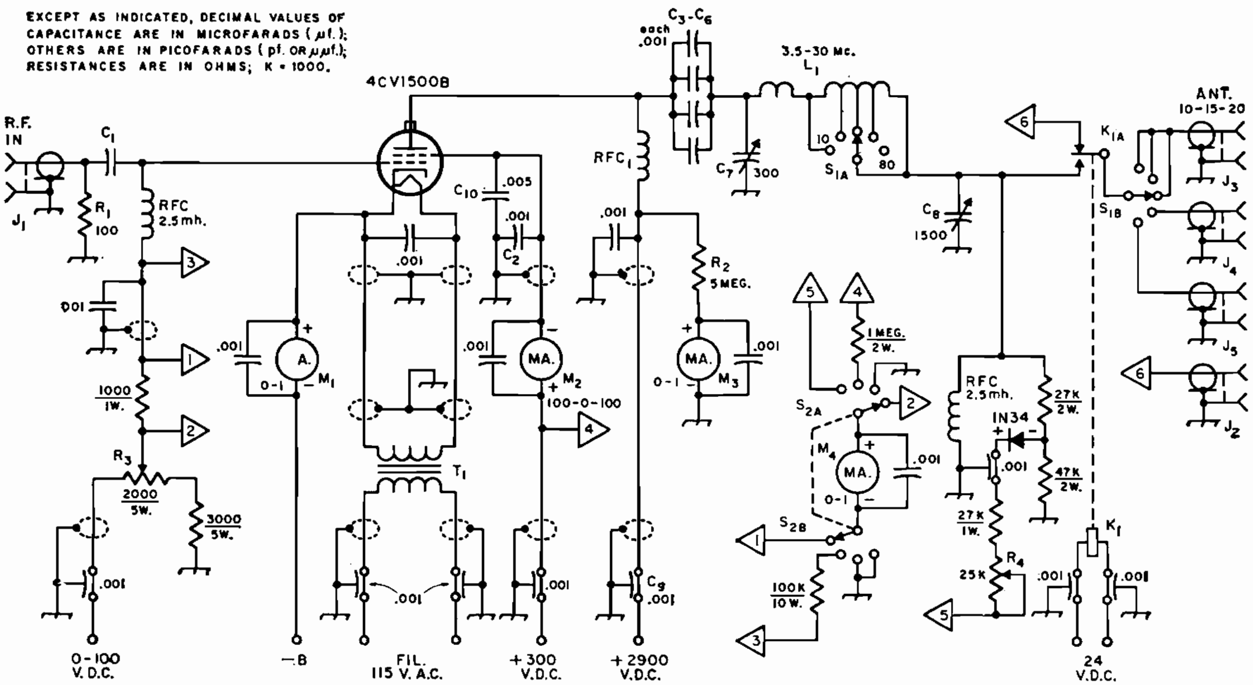


Fig. 3—Circuit diagram of the 4CV1500B amplifier. All 0.001-μf. feed-through type capacitors not listed below are 500-volt rating (Erie CK70AW-102M or equivalent); other 0.001-μf. capacitors are disk ceramic, 500 volts.

C_1-C_6, inc.—0.001 μf., 4000 volts (Centralab 858S).
C_7—300-pf. vacuum variable, 5000 volts (Jennings UCSL).
C_8—1500-pf. vacuum variable, 3000 volts (Jennings UCSL).
C_9—0.001-μf., 4000-volt feed-through (Erie type 2498).
C_{10}—0.0015 μf., 400 volts (part of Eimac SK800B air-system socket).
J_1, J_2—Chassis-mounting coaxial connectors, BNC.
J_2, J_3, J_4—Chassis-mounting coaxial connectors, UG-58A/U or SO-239).
K_1—Vacuum relay, s.p.d.t., 24-volt d.c. coil (Jennings RJ1A).
L_1—Turret tank inductor (B & W 852).
M_1—0-1 d.c. ammeter (Weston 1921).
M_2—Zero-center d.c. milliammeter, 100-0-100 (Weston 1921).
M_3, M_4—0-1 d.c. milliammeter (Weston 1921).
R_1—100 ohms; two 50-ohm, 25-watt noninductive resistors in series.
R_2—5 megohms, 10 watts; five 1-megohm, 2-watt resistors in series.
R_3—2000-ohm, 5-watt linear control.
R_4—25,000-ohm linear control.
RFC$_1$—500-ma. plate choke (B&W 800).
S_1—Part of L_1 (see text).
S_2—Rotary, 2 poles, 1 section, 5 positions (4 positions used).
T_1—Filament transformer, 6 volts, 10 amp.

A Vapor-Cooled Tube

The new highly-linear 4CX1500B was chosen as an experimental vapor-cooled tube. Heavy vertical copper fins were first brazed to the bare anode of the tube, then an integral boiler with the same outside diameter as the regular air-cooled fin radiator was affixed to the tube. A distilled-water inlet of ¼-inch copper tubing was installed at the base of the boiler and a 1-inch diameter steam outlet was placed in the top (Fig. 2).

The 4CX1500B is a recent version of the well known 4CX1000A with improved intermodulation distortion characteristics of at least −40 db. 3rd-order at 1 kw. It has the same external dimensions and appearance; however, the internal geometry has been optimized using new computer design techniques. This new vapor-cooled version has been designated the 4CV1500B. Had a larger boiler been employed, the anode dissipation could have been rated at 3 kw. However, in amateur service this rating could not have been utilized. It was desirable to keep the boiler diameter to a minimum, hence the more than ample 1500-watt dissipation rating. Any standard external-anode tube type could be made with the vapor-phase-cooled anode. The 4CV1500B was chosen only as a vehicle to demonstrate the principle.

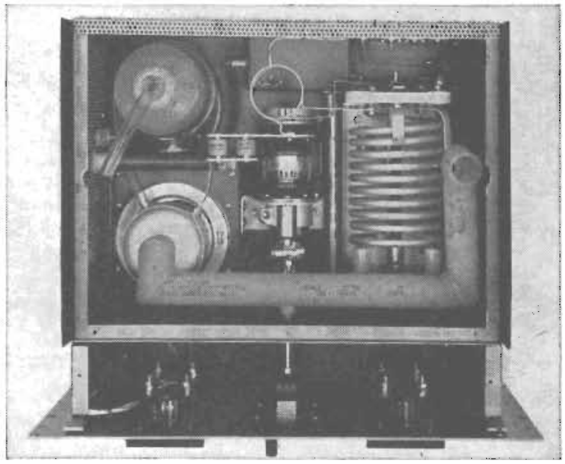
The "Stanley Steamer" Circuit

The circuit and layout of the "Compact AB_1 Kilowatt"[2] amplifier by Ray Rinaudo, W6KEV, was used as the design for the vapor-phase cooling amplifier. It is difficult to improve upon the parts, layout, and circuitry which Ray designed.

Under typical operating conditions this AB_1 linear amplifier has an *average* input of 1000 watts, which results in approximately 400 watts of plate dissipation, assuming 60% plate efficiency. This heat causes the water surrounding the anode to change to steam. Under a slight positive pressure, the steam flows up through the polypropylene plastic insulating tube to the condenser mounted in the lid of the amplifier.

The energy from the steam is dissipated by the convection-cooled radiator and the steam is changed back into the water, or liquid, state. The water then flows by gravity into the plastic reservoir. This reservoir and the 4CV1500B integral boiler are connected together by means of a ¼-inch plastic tube to provide the return input water path and to complete the cycle. The water level in the tube boiler is, of course, dependent upon the level in the reservoir. The level gauge on the front panel is connected to the reservoir and provides a visual check of the system.

Pyrex or polypropylene plastic tubing is utilized for the water inlet and steam outlets, providing d.c. and r.f. isolation between the tube anode and ground. A one-quart plastic container is used to store the distilled cooling water. The water level can vary by as much as ½ inch over



The plastic water pipe helps give the interior of the amplifier the look of a piece of power-house equipment. The water reservoir is in the upper left corner in this view. The radiator, which connects to the plastic tube at the left and the pipe opening at the right, covers this equipment in regular operation.

the length of the anode surface and still supply ample cooling. In actual practice it has been found necessary to add only a few ounces of water every four to five weeks of normal operating time.

A steam condenser, Model #E-56073, measuring 11½ x 16½ x 2 inches, was obtained from the Liberty Radiator Core Company, 250 14th Street, San Francisco, California. It is constructed in the same manner as that of an automobile radiator using several straight-through parallel paths, and is made of copper parts silver-soldered together. Brass or soft-soldered parts should not be used in vapor-cooling systems as the steam will attack such materials. These impurities will contaminate the cooling water and cause high d.c. leakage current. This leakage promotes electrolytic action which in turn attacks the brass or solder joints and results in water leakage. If copper is chosen for all materials which come in contact with the water or steam, none of foregoing difficulties will be encountered. These basic rules have been used over the years in water-cooled systems.

The condenser also forms the r.f. shield and top lid of the amplifier and is cooled by natural air convection. It is capable of fully condensing steam up to anode dissipation levels of at least 600 watts. The condenser is mounted with the steam inlet end slightly elevated over the water outlet end so that the water drains easily back into the reservoir. As mentioned previously, no pumps, fans or blower are required. It is a straightforward, simple, efficient and absolutely silent cooling system.

Many air-cooled ceramic-metal tetrodes are rated at a maximum anode core or seal temperature of 250° C. For longer tube life, most equipment is designed to operate below this maximum, with typical temperatures ranging between

[2] Article published in the November 1957 *QST*.

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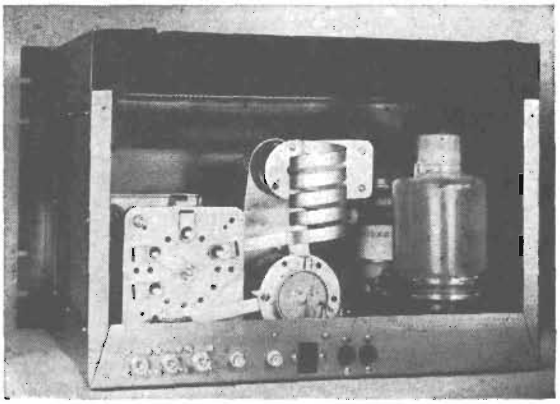

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QST

amateur service newsletter
W6SAI

A 2 KILOWATT 3-400Z LINEAR FOR SIX METERS

BY WILLIAM I. ORR, *W6SAI

The Six Meter DX operator literally twitches in anticipation as he watches the sunspot cycle slowly and inexorably rise from the 1964 low. Will the sunspot count rise high enough to permit the wild DX era of 1958 to be repeated—the era when international 50 mc DX was relatively commonplace? Or will the peak of the forthcoming cycle fall short of the last record-breaking maximum? At this stage of the cycle, no one can be sure of the coming events, and the F_2 propagation history for the future of the six meter band remains a tantalizing speculation.

Even so, other forms of long distance propagation lurk just around the corner of this fascinating band, and the avid six meter DX enthusiast cannot afford to relax, as propagation surprises await the operator who has the equipment, antennas, and "know-how" to fully exploit the interesting possibilities that abound in "Channel One."

The trend to sideband, while slower than on the d.c. bands, seems to be taking place at an accelerated rate on six meters. The advantages of s.s.b. which are apparent on lower frequencies apply equally well to six meters. It is the writer's opinion that the use of a.m. on six meters will gradually be eclipsed by s.s.b. during the forthcoming ascendency of the sunspot cycle. In support of six meter s.s.b. and in anticipation of the forthcoming DX seasons on six meters, this article is presented. Let's go six!

S.S.B. exciters and linear amplifiers are becoming commonplace on six meters. This amateur band, existing in the twilight area between h.f. and v.h.f., exhibiting the characteristics of both, poses some unique problems in the design of suitable transmitting equipment. While v.h.f. construction techniques apply to this band, their unqualified use is somewhat limited as cavities and tuned lines are just too large to be used comfortably at 50 megacycles. On the other hand, employment of conventional h.f. circuit techniques may be "asking for trouble" on this quasi-v.h.f. band. Conventional h.f. components and techniques often turn out to be cranky troublemakers at six meters, bypass capacitors resembling blocks of wood, tank coils overheating and melting to a pasty blob, tubes seemingly difficult to drive and conservative circuitry exhibiting an aggravating degree of instability. Parasitic oscillations race through proven circuits and, if proper design techniques are not used by the would-be six meter enthusiast, he is apt to become discouraged and call it a day, returning to the more serene pastures below 10 meters. In other words, he is forced to "drop the fiddle and the bow: take up the shovel and the hoe."

In no area of the amateur radio spectrum is circuit *finesse* demanded more than in the design of v.h.f. s.s.b. equipment wherein the requirements of amplifier linearity are combined with the problems of circuit stability. It would be fatuous to assume that this combination of problems does not exist at six meters.

This article describes the "case history" of conversion of a 3-30 mc h.f. linear amplifier to six meter operation and, in addition, points out some of the circuit problems that were overcome in the conversion of "conventional circuitry" to quasi-v.h.f. operation. While modification of a specific amplifier package is shown, the discussion applies equally well to similar gear, or to construction of six meter amplifiers in general, and grounded grid amplifiers in particular.

The Grounded Grid Linear Amplifier

The grounded grid amplifier has attained unprecedented popularity for s.s.b. linear service in the high frequency amateur bands. As the name implies, the circuit is one in which the grid of the amplifier tube is at r.f. ground potential and the drive signal is applied between cathode and grid, either element being at the necessary d.c. bias potential. Another (more accurate) name for the circuit is *cathode driven,* as the term "grounded grid" implies the grid is actually grounded to both r.f. and d.c. potentials. This may not be the case when d.c. bias is required, or if r.f. feedback is introduced in the amplifier to enhance the intermodulation distortion figure.

A simplified schematic of a typical h.f. grounded grid amplifier is shown in fig. 1. The

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circuit possesses definite advantages over the grid-driven triode and tetrode circuits. For example, neutralization of the grounded grid circuit is either not required (depending upon the frequency of operation) or, if required, is simply added and easily adjusted. An economy of circuitry is achieved by the use of high-mu triodes that do not require costly screen and bias power supplies. The drive power of the grounded grid stage, moreover, is compatible with the power output of today's s.s.b. transmitters and transceivers and—best of all—a portion of the drive power shows up as "free" power in the output of the grounded grid stage.

In a conventional grounded grid amplifier, such as illustrated, the grid element of the tube is at r.f. ground potential and acts as a shield between the plate circuit and the input circuit. A careful observer will note that this configuration is similar to that of an oscillator and might suspect that the ground grid amplifier will perform admirably as an oscillator if the shielding action of the grid is insufficient to prevent undue feedback of energy from the plate to the cathode circuit. This suspicion is well founded, for this is exactly what happens in the upper reaches of the h.f. spectrum with many grounded grid amplifiers where feedback paths are difficult to control.

While the grounded grid stage is not neutralized in the true sense of the word, the degree of internal feedback through the grid structure of the tube at the lower frequencies is insufficient to permit oscillation to take place. Even so, while the amplifier seems stable, it may still be highly regenerative and, as the frequency of operation is raised, the shielding action of the grid will tend to deteriorate until stage oscillation takes place. This deterioration of intra-stage isolation is primarily due to decreasing cathode-to-plate reactance as the operating frequency rises, and also because of stray electro-magnetic coupling through the interstices in the grid structure itself. In either event, the intra-stage feedback through the grid circuit becomes of increasing importance as the frequency of operation of the stage is raised until, at some critical frequency, the isolation of the grid structure is insufficient to prevent the amplifier from oscillating at a frequency determined by the plate and cathode tuned circuits. In addition, if additional intra-stage coupling exists around the tube, by virtue of inductive or capacitive coupling between input and output circuits, it is possible for the feedback path to be enhanced and oscillation might possibly occur at some lower frequency than normally estimated.

After paying due attention to external feedback paths, oscillation and instability in a grounded grid amplifier may be suppressed by the use of a suitable neutralizing circuit that feeds energy back from the plate to the input circuit so that it is out of phase with the r.f. energy fed back through the grid structure of the tube. A satisfactory neutralizing circuit for a grounded grid amplifier is shown in fig. 2. The

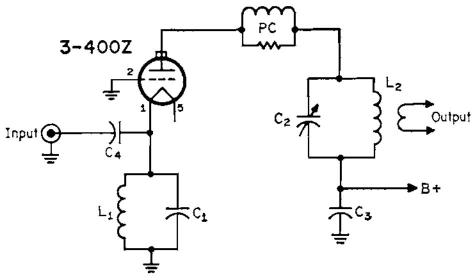


Fig. 1—Circuit of a grounded grid linear amplifier. The input signal is applied to a tuned cathode circuit, L_1, C_1. The circuit is adjusted to resonance with C_1 value of approximately 20 mmf per meter of wavelength (*i.e.*: for 20 meters, C_1 is about 400 mmf). Coupling capacitor C_4 and tuning capacitor C_1 should be able to stand circulating r.f. currents that exist in this circuit. The use of 1250 volt fixed mica capacitors is recommended. A value of 0.001 mf for C_4 is satisfactory for all bands between 80 and 6 meters. Parasitic suppressor PC is made of a noninductive composition resistor placed across a small inductance in the plate lead. For six meters, the inductance consists of a section of the plate strap material (see parts list). If the inductor is too large, the resistor will run hot.

neutralizing coil is connected in such a way as to "buck" the feedthrough voltage and neutralization may be accomplished by adjustment of the neutralizing capacitor or the coupling between the neutralizing coil and the cathode coil.

The Prototype Six Meter Amplifier

When the experimenter "designs from scratch" all the exotic v.h.f. circuit techniques may be employed as desired. However, when an existing design is modified, the builder may be restricted in his attempts at redesign because of various features or peculiarities of the amplifier in question that do not readily fit into the scheme of things to come. Luckily, in this case, the amplifier to be modified proved to be very accommodating as far as the need for radical cir-

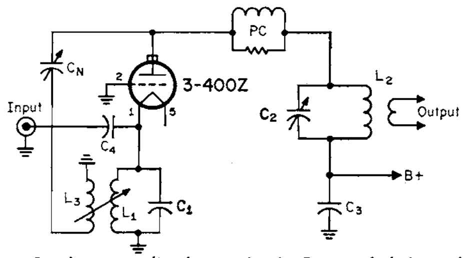
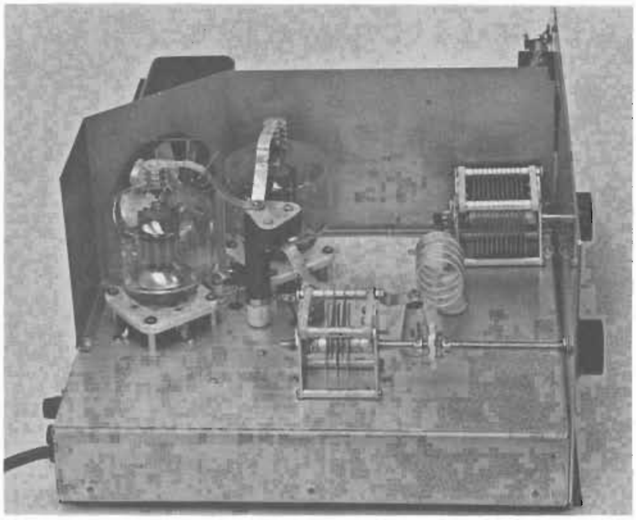


Fig. 2—The neutralized g.g. circuit. Energy fed through from plate to input circuit may cause the grounded grid amplifier to exhibit instability. The addition of a simple neutralizing circuit (C_N, L_3) permits energy to be fed back out of phase with the energy passing through the incompletely shielded grid structure. Neutralizing capacitor C_N is of the order of a few mmf. Neutralizing coil L_3 may be two or three turns placed in the coil of the cathode circuit. Adjustment of coil coupling or capacitance of C_N will neutralize the circuit. Neutralization is achieved when maximum grid current, minimum plate current and maximum power output coincide at one setting of the plate tank circuit.



Side view of the modified SWAN six meter linear amplifier. Conventional low frequency bandswitching circuitry is removed and replaced with six meter pi network tank. The plate tuning capacitor (C_2) is in the foreground. An insulated coupling connects the capacitor to the tuning dial, otherwise the metal shaft extension forms a closed loop closely coupled to the plate tank coil. The loading capacitor (C_3) is at the rear. The connection between the loading capacitor and the coaxial receptacle on the rear of the chassis is made with a short length of RG-8/U coaxial cable. In the modified amplifier, the original antenna changeover relay was used, remounted near the antenna receptacle to reduce lead length.

The three parallel connected 100 uuf, 5 kv plate blocking capacitors are suspended from the top of the plate r.f. choke. One of the parasitic suppressors may be seen above the glass envelope of the left-hand 3-400Z. While Eimac air-system sockets and chimneys are usually recommended for the home constructor, the Swan amplifier used special Johnson sockets (modified for the 3-400Z by reducing spring tension) and a special air scoop and blower system to cool the tubes.

cuit changes: nothing was required that could not be handled in the home workshop with a minimum of tools and test equipment.

The "victim" amplifier chosen for modification to six meters was the *Swan Mark I* linear amplifier covering 3.5-29.7 mc. It makes use of two Eimac 3-400Z tubes connected in parallel with a conventional pi-network circuit. The amplifier is rated for the so-called "two kilowatt p.e.p. level" and had a self-contained power supply. The modification consisted of removing the h.f. input and tank circuitry and substituting six meter circuitry, plus other minor circuit changes that resulted in a stable, maximum power linear amplifier, capable of smooth operation at six meters.

The first step was to simply drop in six meter tuned circuits in the input and plate tank configuration, modify the plate parasitic chokes for 50 mc operation, and apply the "smoke test." As might be expected, the amplifier responded in predictable fashion by "taking off like a bird" with a robust 50 mc parasitic oscillation that brought roars of protest from the small fry, their eyes and ears glued to a TV cartoon program on Channel 2.

The amplifier was immediately disconnected and the various circuits explored with a grid-dip oscillator. Resistance tests indicated that the numerous ceramic disc bypass capacitors in the amplifier, while satisfactory at 10 meters and below, were ineffective at six meters. Replacing or shunting these capacitors with 750 mmf mica units whose self-resonant frequency was above 54 mc was the first step in making the amplifier operable. A transmitting type ceramic capacitor having low internal inductance was substituted for the high voltage disc capacitor at the B-plus end of the choke. The final step was to clean the paint from the chassis, bottom plate and cover shield at the points of contact to insure hat a minimum of r.f. intra-stage leakage existed through the cabinet via the various joints and seams.

During this modification, it was discovered that the primary power cord was "hot" with 50 mc r.f., so the input terminals of the line were bypassed at the fuse connections.

After this series of simple modifications, the amplifier was turned on and loaded to maximum input into a dummy load. With excitation removed and no cutoff bias on the tubes, it was still possible for a weak 50 mc parasitic oscillation to take place, indicating that complete stability had not been achieved. The oscillation was noted by a slight show of grid current at random settings of the plate tuning and loading capacitors and by the fact that maximum grid current, maximum power output and minimum plate current did not coincide at the same tuning condition when the amplifier was in operation. Small indications, it is true, but symptoms that more work remained to be done on the amplifier.

The Neutralizing Circuit

The point had been reached where further modifications to the amplifier would be rather extensive and basic if it was desired to completely stabilize the unit by improving the intra-stage isolation. The 3-400Z is used in grounded grid circuitry in commercial f.m. transmitting equipment up to 100 mc or so without the need of neutralization, but sophisticated shielding and bypass techniques appropriate to those frequencies are a necessity, and probably would be incompatible with the simple and uncomplicated design of the Swan amplifier, which is typical of

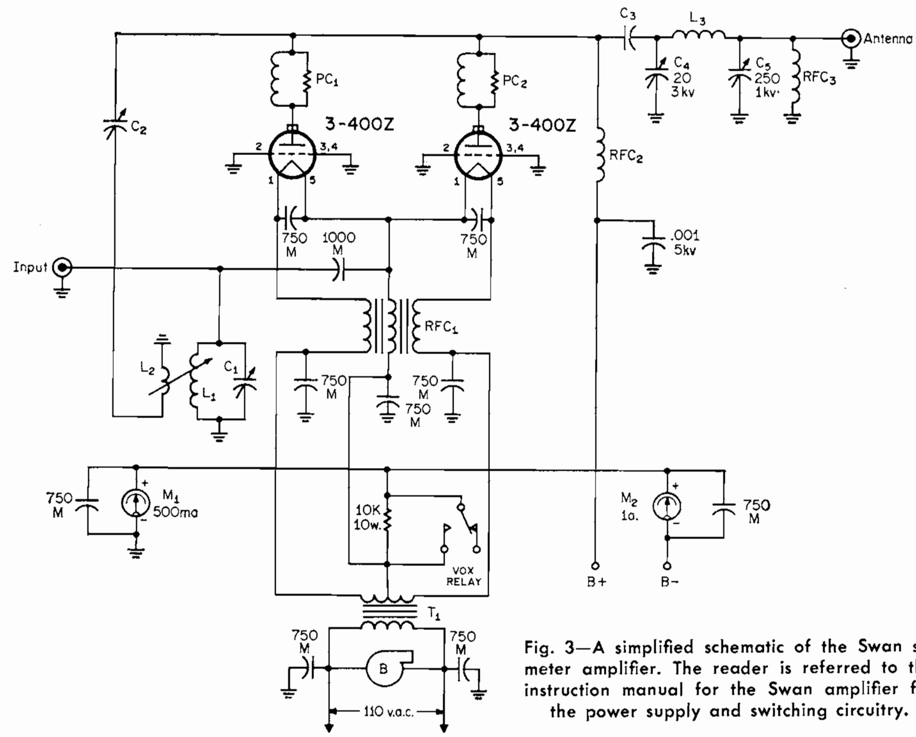


Fig. 3—A simplified schematic of the Swan six meter amplifier. The reader is referred to the instruction manual for the Swan amplifier for the power supply and switching circuitry.

C_1—50 mmf variable mica capacitor. Adjust coil L_1 to obtain resonance with the capacitor 90% compressed.

C_2—Neutralizing capacitor. Copper strap 3½" long and ½" wide mounted on a feedthrough insulator placed between the 3-400Z tubes. The capacitor "sees" the anodes of the tubes. The tubes are mounted about two inches apart, with the capacitor centered between them.

C_3—Three 100 mmf, 5 kv Centralab #850 capacitors in parallel bolted to a triangular aluminum plate atop RFC_2. See photograph.

C_4—20 mmf, 3 kv. Johnson 154-11, reduced to 3 rotor and 2 stator plates. Panel driven with insulated coupling.

C_5—250 mmf. Johnson 154-1.

L_1—3 turns #14 wire, ½" dia., ⅝" long.

L_2—2 turns #18 insulated hookup wire, ½" dia., placed between adjacent turns of coil L_1. Observe the polarity as shown. Adjust the coupling for proper neutralization.

L_3—4¼ turns of copper strap, ⅜" wide, 1⅜" dia., 2½" long.

PC_1, PC_2—50 ohm, 2 watt composition resistor shunted across 1½ of the plate lead (½" wide copper strap). Place a suppressor close to plate cap of each tube.

RFC_1—3½" long by ½" dia. ferrite rod (Lafayette Radio Co., N.Y. MS-333). File a nick in the rod at the correct length and break off remainder. Make triple winding (two #12 e., one #18 e.) consisting of 12½ turns. Wind all three wires at once, under tension. Coat with epoxy cement or nail polish to hold windings in position. (Equivalent of original Swan choke).

RFC_2—140 turns #22 e. wound on a ¾" diameter ceramic form. The winding length is about 3. (Equivalent of original Swan choke).

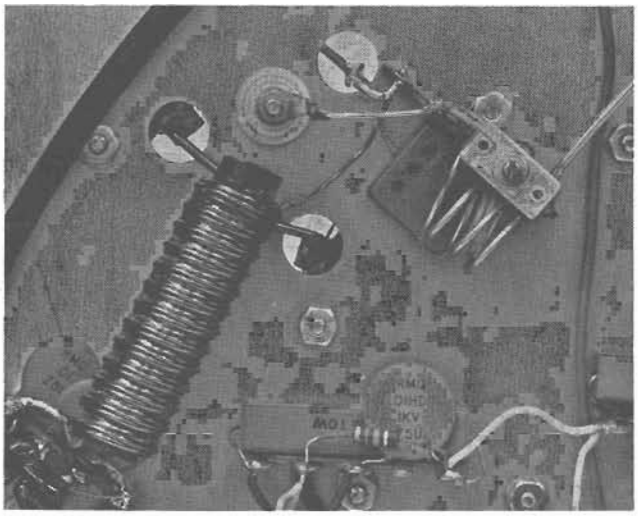
RFC_3—Ohmite Z-50.

T_1—10 volts at 15 amperes. Thordarson 21F146 or equiv. Blower—See text.

good amateur practice below 10 meters. Under the restrictions imposed by the circuit layout and assembly, therefore, it was decided to leave well enough alone as far as intra-stage isolation was concerned, and to complete the job of stabilizing the amplifier by the addition of a neutralizing circuit. The circuit of fig. 2 was chosen and installed. The neutralizing capacitor was fabricated from a length of copper strap mounted mid-way between the 3-400Z's on a ceramic feed-through insulator. The strap provided a small capacitance to the anodes of the tubes by virtue of its position. The under-chassis terminal of the feed through insulator was connected to a small link coil inserted into the cathode coil, as shown in the photograph. The neutralizing coil was mounted so that the polarity of the coil connections could be changed, if need be. Once neutralized, the amplifier proved to be a "winner" and tuned up and operated just as if it was on the "d.c. bands." The whole modification and tuning program took less than six hours and was well worth the effort.

Neutralizing and Tuning Adjustments

Neutralizing and tuning are no problem in the modified six meter amplifier, the schematic of which is shown in fig. 3. The cathode circuit is tuned to 50 mc with a grid-dip oscillator with the tubes in place in the sockets, after which this



Under-chassis view of the cathode circuit of the 6 meter linear. The tuned cathode circuit is at the upper right, mounted to a five lug terminal strip. Directly to the left is the ceramic feedthrough insulator, atop which the neutralizing capacitor is mounted. To the left is the three-winding (trifilar") filament choke. The five volt filaments of the two 3-400Z tubes are connected in series to reduce the filament current passing through the choke. A third wire to the common filament point is required to insure that equal voltage appears across each tube. Additional mica bypass capacitors are placed across original .01 mf disc ceramic capacitors at the cold end of the choke. The coaxial output cable running to the antenna receptacle cuts across the upper left of the photograph.

adjustment may be forgotten. The next step is to place the output loading capacitor at full capacitance and tune the plate tuning capacitor to approximate resonance using the grid-dip oscillator. The amplifier is turned on with appropriate filament and plate voltage but without excitation. The plate tuning and loading controls are varied at random from their previous settings and the grid meter of the amplifier is observed for signs of grid current. Without excitation, grid current is zero, but under conditions of oscillation, ten to twenty milliampere of grid current will be observed[1], and the resting plate current will climb a few milliamperes. With the amplifier oscillating, the neutralizing coil is moved about in the cathode coil until a point is found at which grid current disappears and oscillation stops.

In this operation, the bottom plate must be on the amplifier to complete the r.f. shielding and also to force the moving air through the tube sockets, if air-system sockets are used[2]. These special sockets are not used in this amplifier, but the problem of intra-stage r.f. leakage remains. As the link adjustment is simple, a modification was made that permitted the bottom plate to remain in place. A small hole was drilled in the bottom plate of the amplifier directly below the cathode circuit, and the neutralizing link was moved about by pushing it from below, using a wooden stick as the operative tool. Adjustment took about thirty seconds. The point of neutralization was found to be quite broad and noncritical. Once the correct coil placement was found, the coil was immobilized in position with a drop of "airplane cement," or "coil dope."

As could be guesssed, Kelly's Law[3] was in effect and the coil was first polarized improperly and the amplifier oscillated in a robust manner until the coil connections were reversed. When properly polarized and placed, the little neutralizing coil did the trick: the amplifier was completely stable and maximum grid current, maximum output and minimum plate current all coincided at one tuning adjustment.

Once the amplifier is neutralized, it should be properly loaded to the recommended grid and plate currents. In this case, at about 2500 volts plate potential, the peak (carrier) plate current was 800 milliamperes for two tubes, with about 300 milliamperes grid current. The Swan six meter s.s.b. transceiver had ample output to drive the linear amplifier with power to spare.

Final Thoughts and a Word of Warning

The modified Swan six meter amplifier certainly has proven to be a success, and this modification is recommended to those hardy souls who do not mind "butchering" a piece of commercial gear. For others, the circuitry is satisfactory for home construction from scratch. It must be pointed out, however, that the ventilation system used in the Swan amplifier is of a special design, in which the physical layout of the ventilation fan, the air scoop and the cabinet provide proper air cooling to the base and plate seals of the 3-400Z tubes. Unless the prospective builder has the means at hand to measure the temperature of the glass envelope of the tubes, it is recommended that the standard Eimac air-system sockets and tube chimneys be used, in conjunction with a squirrel cage blower for proper ventilation. Suggested blowers are the Dayton 1C-180, Ripley 81 or Ripley 8472. ■

[1] If vox operated cathode bias is used, such as shown, the bias resistor will have to be shorted out for these tests. The relay may be energized, or the resistor temporarily jumpered with a length of wire.

[2] The Eimac air-system sockets provide improved intrastage isolation and insure maximum cooling. Their use is recommended. In any case, the amplifier should not be turned on its side when the filaments of the tubes are lit as the 3-400Z's are designed only for vertical (base up or down) operation.

[3] Kelly's Law*: If something *can* go wrong, it *will* go wrong.

**Also known as Murphy's Law. Ed.*

Eimac®
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W6SAI

Pi and Pi-L Networks for Linear Amplifier Service

These graphs may be used to determine the values of components in Pi and Pi-L Networks. The graphs cover the most generally used operating Q's, load resistances and antenna impedances. To use the charts it is only necessary to know the plate voltage, peak plate current, the desired operating Q and the transmission line impedance.

Using the Pi Network Charts

A. Choose the amplifier tube(s) to be used. Select the plate voltage and determine the plate current for normal operation from the tube data sheet.

Assume, for example, that a Pi Network is to be designed for a pair of 3-400Z tubes operating at a plate potential of 2500 volts and a PEP input of two kilowatts. Peak envelope plate current is determined by:

$$\text{Peak envelope plate current (in amperes)} = \frac{\text{PEP Watts}}{\text{Plate Voltage}} \qquad (1)$$

$$= \frac{2000}{2500} = 0.8 \text{ ampere}$$

B. Determine the resonant load resistance from:

$$\text{Load resistance } (R_1) = \frac{\text{Plate voltage}}{2 \times \text{plate current in amperes}} \qquad (2)$$

For the case of the 3-400Z's, the load resistance is:

$$\text{Load resistance} = \frac{2500}{2 \times 0.8} = 1560 \text{ ohms}$$

C. Choose the operating Q. Good practice calls for a Q between 10 and 20. A Q of 15 is recommended for linear amplifier service.

D. Choose the antenna transmission line impedance (R_2). These charts are designed for use with either 52- or 72-ohm loads as coaxial cables for these impedances are generally available.

E. Find the reactance of the Pi Network coil from figure 1. For the case of the two 3-400Z's operating with a load resistance of 1560 ohms and a Q of 15, the reactance of the coil is approximately 120 ohms.

F. Find the reactance of the loading capacitor (C_2) from Figure 2. For the case of the 3-400Z's operating with a load resistance of 1560 ohms and a Q of 15, the reactance of the loading capacitor is about 20 ohms.

G. Find the reactance of the tuning capacitor (C_1) from figure 3. For the case of the 3-400Z's operating with a load resistance of 1560 ohms and a Q of 15, the reactance of the tuning capacitor is about 100 ohms.

Summary: For two 3-400Z tubes, operating at a plate potential of 2500 volts with a peak plate current of 0.8 ampere (two kilowatts PEP) and a Q of 15, the values of the Pi Network plate circuit are:

Tuning capacitor (C_1) = 100 ohms
Loading capacitor (C_2) = 20 ohms
Pi Network coil (L_1) = 120 ohms

Note: As a quick check, note that the sum of the reactances of the two capacitors is equal to the reactance of the inductor.

H. Determine the values of capacitance and inductance for the components of the Pi Network. Charts of the figures 2-44 and 2-45 show reactance values of inductors and capacitors in the range commonly used in r-f circuits for the h-f amateur bands. For the reactances determined for the 3-400Z tubes, the circuit components may easily be determined for each amateur band. In the case of the 20 meter band, for example, the values are:

Tuning capacitor (C_1) = 100 ohms = 133 pf
Loading capacitor (C_2) = 20 ohms = 565 pf
(above determined from figure 2-45)
Pi Network coil (L_1) = 120 ohms = 1.36 μH
(above determined from figure 2-44)

Using the Pi-L Network Charts

Figures 3,4,5 and 6 are used to determine the reactance of the components of the Pi-L Network.

A. Choose the amplifier tubes to be used. Select the plate voltage and determine the peak plate current for normal operation as outlined under step 1 for Pi Networks.

Assume, for example, that a Pi-L Network is to be designed for a single 3-1000Z operating at a plate potential of 3000 volts and a PEP input of two kilowatts. Peak envelope plate current (formula 1) is:

Peak envelope plate current (in amperes) = 0.667 ampere

B. Determine the load resistance, as outlined previously in formula 2:

Load resistance (R_1) = 2250 ohms

C. Choose the operating Q. (Let Q = 15).

D. Choose the antenna transmission line impedance. (Let R_2 = 52 ohms).

E. Find the reactance of the tank coil (L_1) from figure 4. For the case of the 3-1000Z operating with a load resistance of 2250 ohms, the reactance of the coil is approximately 215 ohms.

F. Find the reactance of the loading capacitor (C_2) from figure 5. In this case, the reactance is about 47 ohms.

G. Find the reactance of the tuning capacitor (C_1) from figure 3. In this case, the reactance is about 150 ohms.

H. Find the reactance of the loading coil (L_2) from figure 6. In this case, the reactance is about 140 ohms.

Summary: For a single 3-1000Z operating at a plate potential of 3000 volts with a peak plate current of 0.667 ampere (two kilowatts PEP), and a Q of 15, the value of the Pi-L Network plate circuit components is:

Tuning capacitor (C_1)	=	150 ohms
Loading capacitor (C_2)	=	47 ohms
Pi Network coil (L_1)	=	215 ohms
L Network coil (L_2)	=	150 ohms

I. Determine the values of the capacitance and inductance for the components of the Pi-L Network. Charts of figures 2-44 and 2-45 show reactance values of inductors and capacitors in the range commonly used for r-f circuitry for the h-f amateur bands. For the reactances determined for the 3-1000Z tube, the circuit components may be easily determined for each amateur band. In the case of the 80 meter band, for example, the values are:

Tuning capacitor (C_1)	=	150 ohms	=	275 pf
Loading capacitor (C_2)	=	47 ohms	=	900 pf
Pi Network coil (L_1)	=	215 ohms	=	9 μH
L Network coil (L_2)	=	150 ohms	=	6.5 μH

Note: Capacitance values are for resonance with a nonreactive load. It is suggested that the tuning capacitor have about 50% greater capacitance than indicated and the loading capacitor have 100% greater capacitance than indicated.

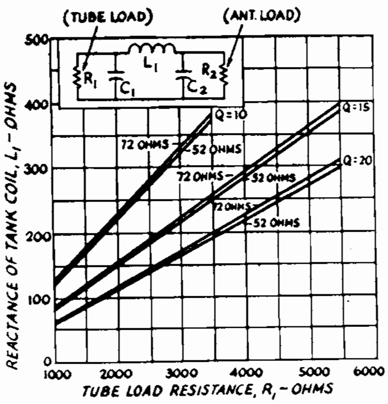

Fig. 1 — **Reactance of tank coil, L_1, as a function of tube load resistance, R_1 (for pi networks).**

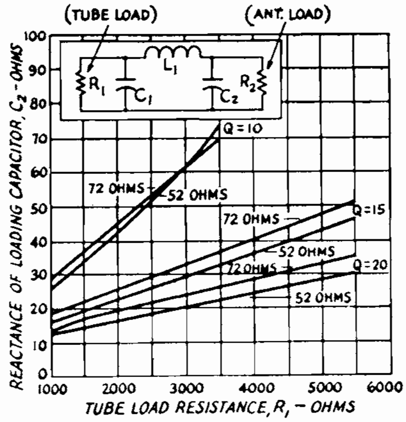

Fig. 2 — **Reactance of loading capacitor, C_2, as a function of tube load resistance, R_1 (for pi networks).**

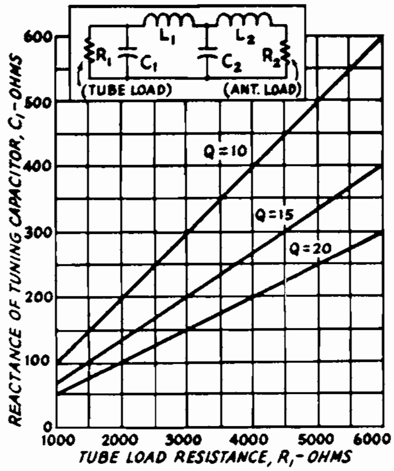

Fig. 3 — **Reactance of tuning capacitor, C_1, as a function of tube load resistance, R_1 (for pi and pi-L networks).**

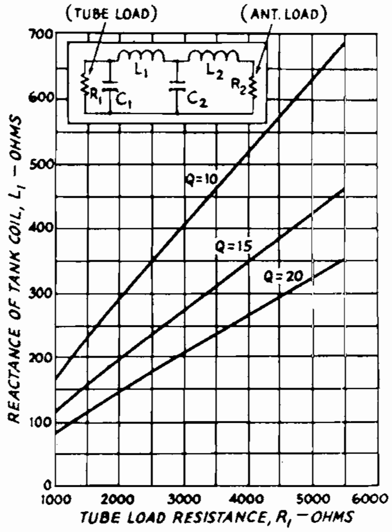

Fig. 4 — **Reactance of tank coil, L_1, as a function of tube load resistance, R_1 (for pi-L networks).**

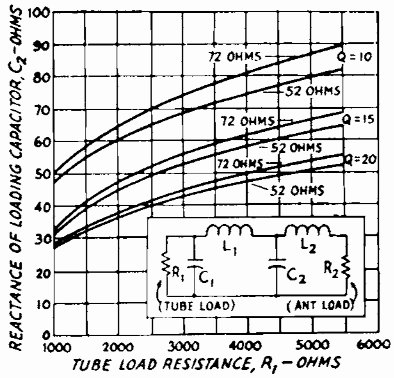

Fig. 5 — **Reactance of loading capacitor, C_2, as a function of tube load resistance, R_1 (for pi-L networks).**

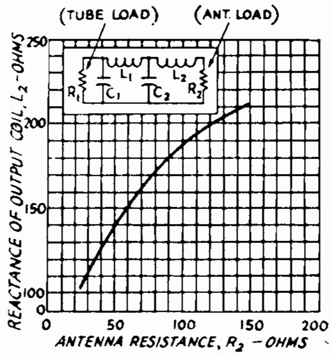

Fig. 6 — **Reactance of loading coil, L_2, as a function of antenna load resistance, R_2 (for pi-L networks).**

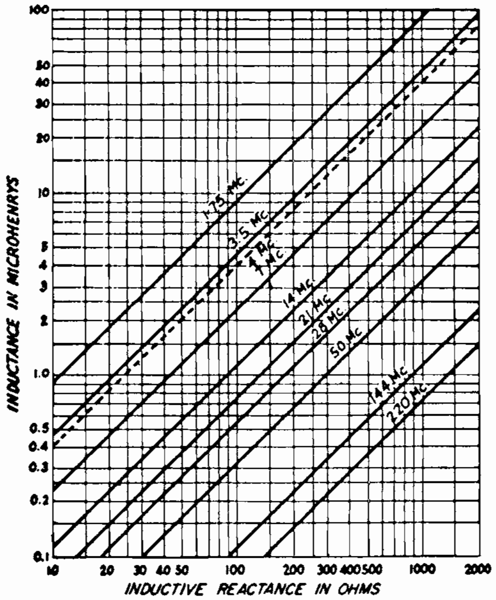

Fig. 2-44—Reactance chart for inductance values commonly used in amateur bands from 1.75 to 220 Mc.

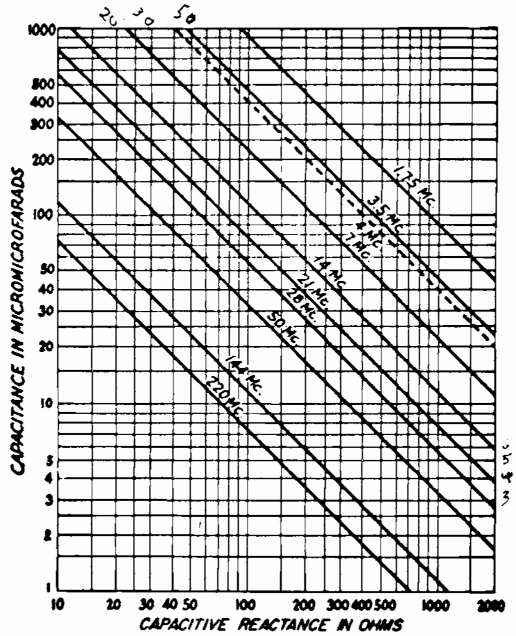

Fig. 2-45—Reactance chart for capacitance values commonly used in amateur bands from 1.75 to 220 Mc.

amateur service newsletter
W6SAI

Forced-Air Cooling of Transmitting Tubes

Some Considerations in the Selection of a Suitable Impeller

BY WILLIAM I. ORR,* W6SAI

We live at the bottom of a vast ocean of air. This invisible, life-supporting elixir provides the equipment designer with an inexpensive and efficient cooling medium for heat-generating devices, such as transmitters and receivers. Over the years, electronic equipment has grown more sophisticated and compact, and the problem of removing heat from the gear has become acute. Until someone miniaturizes the watt, heat-exchange systems will remain one of the major problem areas of equipment design. Aspects of forced-air cooling systems are discussed in this article.

Most electronic equipment generates heat, and this heat must be removed or the equipment will eventually burn up. The heat may be removed by radiation, conduction or convection [1], or by a combination of these methods. This article examines forced-air cooled systems (an efficient form of convection cooling), which are used in commercial transmitting equipments up to the level of tens of kilowatts and, in amateur gear, up to the so-called "two-kilowatt p.e.p." level. Generally speaking, from 20 to 70 per cent of the primary power drain of electronic equipment is dissipated in heat emitted from tubes and components, and the resulting temperature rise must be held within reasonable limits to insure satisfactory life for both the tubes and the other parts in the equipment.

The Air System

Two typical forced-air cooling systems for a power tube are shown in Figs. 1A and 1B. They consist of an air blower, or impeller; a conduit to guide the cooling air to the tube, or a pressurized chassis; the heat radiator of the tube; and an air exhaust exit. By stretching the imagination only a little, this air system can be compared to the electrical series circuit of Fig. 1C, in which each component in the air system is represented by a resistor which has a potential drop across it corresponding to the *back pressure* or resistance [2] that the original component offers to the flow of air. The sum of the back pressures in the air system must add up to the total pressure of air supplied by the blower, just as the sum of these voltage drops in the electrical analogy must add up to the generator voltage. The blower in the air system corresponds to the generator in the electrical system, of course.

The electrical analogy suffers in that the back pressure across a component in the air system does not strictly follow an equivalent of Ohm's Law for electric potential drops. Instead, the

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[1] Quinn, "The Stanley Steamer," *QST*, May, 1966.

[2] The resistance offered to the flow of air may also be expressed in terms of "pressure drop" or "static pressure."

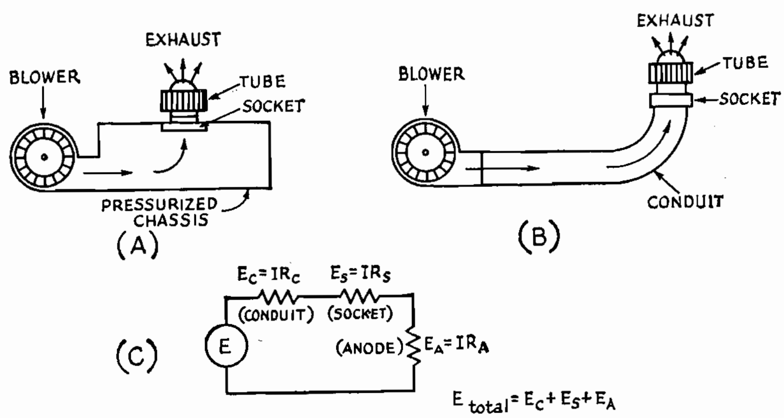

Fig. 1—A forced-air cooling system. In A, the blower is mounted directly on the chassis which is used as a plenum chamber. Air is exhausted past filament and plate seals of the tube. In B, the blower is mounted at some distance from the tube, and cooling air is conducted to the tube via a conduit or hose. C indicates an electrical analogy of the forced-air system. The blower is represented by generator *E*, and various unavoidable back pressures are represented by voltage drops across resistors *Rc* and *Rs*. Useful work (cooling the anode) is represented by voltage drop (E_A) across the tube.

back pressure across an air-system component varies approximately as the *square* of the airflow rate (volume per unit of time). Thus, if the volume demand is doubled, about four times the pressure will be necessary to meet the increased requirement. Even though the analogy is inexact, the transmitter designer who is comfortable in the presence of Ohm's Law for series circuits can gain insight of the action of pressure drops incurred in a forced-air cooling system.

The problem to be solved is that of determining the size and characteristics of an air blower that will satisfy the temperature limitations imposed upon a particular tube type by the manufacturer, and reconciling these limits with available blowers. Maximum operating temperatures and air requirements of forced air cooled transmitting tubes are generally supplied in the data sheet, or provided upon request by the tube manufacturer. This simplifies the problem considerably, as few engineers have the equipment or time to run temperature checks on transmitting tubes. Blower data, too, is supplied by the numerous impeller manufacturers. It remains, then, to translate this available and unfamiliar data into the proper hardware for the system at hand.

Tube Cooling Requirements

Forced-air-cooled transmitting tubes, such as the 4CX250B, 4CX1000A and similar external-anode tubes, require cooling air to be passed from base to anode[3]. Unless otherwise specified in the data sheet, cooling air should flow as long as the tube filaments are lighted. The external anode cooler of tubes of this family is usually composed of a number of copper fins arranged in a circle about the anode core, with the air passing vertically across the surface of the fins. An exchange of heat takes place between the fins and the passing air, the moving air extracting heat from the anode core and holding overall anode temperature at or below the maximum limit. As the air is impeded in its flow through the interstices of the anode structure, a back pressure is created, caused by friction of the air against the fin surfaces, and by turbulence of the air in the anode passages.

The cooling airflow requirement for transmitting tubes may be expressed in terms of the ratio of watts of anode dissipation to tube temperature (in watts per degree Centigrade) as a function of either the mass airflow rate in pounds of air per minute, or the volumetric airflow rate in cubic feet per minute[4]. This information may be expressed in graphic form (Fig. 2), enabling the design engineer to determine the actual cooling-air requirement in terms of specific tube temperature and system back pressure.

[3] Large convection- and radiation-cooled glass tubes (4-400A and 4-1000A, for example) also require forced-air cooling to hold seal temperatures within prescribed limits.

[4] Precise calculation of airflow in cubic feet per minute must take into account air humidity and barometric pressure. Equipment builders often design for a mythical user living in Denver, Colo., who operates the equipment on a hot, humid day.

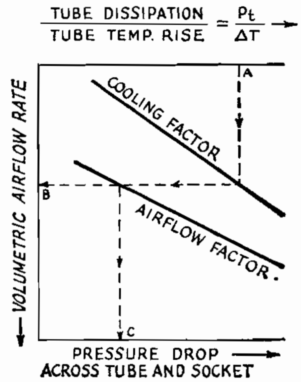

Fig. 2—Allowable temperature rise and dissipation (A) of power tube determine airflow rate (B) from laboratory measurements. Pressure drop (C) across the tube and socket may be measured by a manometer device. The interlocking relationship of cooling requirements may be expressed in graphical form, as shown here.

The total heat to be removed is determined from a study of the operating characteristics of the tube, and includes plate and filament dissipation (plus grid and screen dissipation where applicable). Maximum element dissipation rating is normally given in the data sheet. The operating temperature rise of the tube is found by taking the difference between the maximum measured tube temperature (at the hottest point of the tube) and the maximum inlet air temperature expected. The air requirements expressed by the plots of the cooling factor and the airflow factor are usually given as a pressure drop across the *tube and socket* expressed in *inches of water*, and a corresponding volumetric airflow is defined in *cubic feet per minute* (c.f.m.). This information is necessary to determine the size and speed of the blower required to provide the proper airflow through the system. Volumetric air flow may be calculated or determined by experimental means.

Air-Pressure Measurements

Air pressure in a forced air system may be defined in terms of an equivalent weight of water.

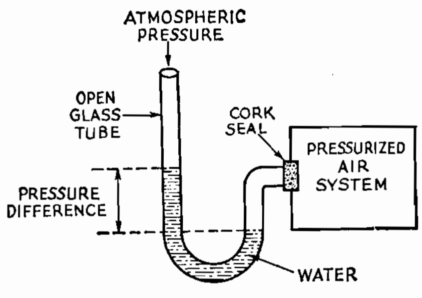

Fig. 3—A simple manometer compares system static pressure with atmospheric pressure. In this drawing, air flows at right angles to manometer input, i.e., into or out of this page. Pressure difference is expressed in "inches of water." Placement of manometer to avoid turbulence in the system should be determined by experiment.

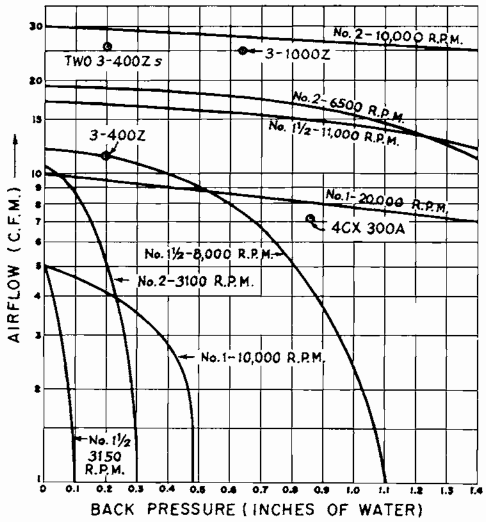

Fig. 4—Typical performance data for No. 1, No. 1½, and No. 2 centrifugal blowers. Performance of blowers of different sizes and speeds can be compared with the cooling requirements specified for various tube types. Notice that requirement points are shown for a pair of 3-400Zs as well as for a single tube. If the requirement point falls on or below the performance curve for a particular blower, that blower will give adequate cooling under the conditions outlined in the text. The curves show that blower efficiency drops rapidly after a critical value of back pressure is reached, and that the blower "windmills" (reaches zero output) at high values of back pressure. High-speed blowers can withstand more back pressure than can low-speed units (notice the curves for 10,000- and 20,000-r.p.m. blowers.

(The weight of a uniform column of water 27.7 inches high is 1 lb. per square inch of column base area.) The measurement is made by means of a *manometer* whose readings are expressed in inches of water (Fig. 3). A simple manometer for shop use may be constructed of a short length of ¼-inch glass tubing bent into a U shape, with one end left open to the atmosphere. The opposite end is inserted in the air system in proximity to the tube socket and *at right angles* to the airflow. Optimum position should be determined by experiment so as to make sure that the manometer is not influenced by eddy currents in the airstream. The bottom portion of the manometer is filled with water and, if the air pressure in the cooling system is equal to atmospheric pressure, the water will rest at equal heights in both vertical sections of tube. Under this quiescent condition, no air moves through the system or, if it moves, it encounters no back pressure. However, if a difference of pressure between the atmosphere and the inclosed air system is created by a blower, the water will be forced up towards the open end of the glass tube by the back pressure of the air moving through the system. The pressure within the duct or plenum, as compared to atmospheric pressure, may be noted by measuring the difference in height (in inches) of the two water columns, as shown in the illustration.

System Pressure Drops

Pressure drop in an air system is caused by physical obstruction to the flow of air, or by turbulence in the air. In the case of a tube anode which contains many fins over which the air must pass, the pressure drop is intentional and useful. Other system drops caused by air friction, pressure drops in the hose or socket, or a change in the air velocity in the system, are undesirable and not useful. All pressure drops caused by these factors must be added to the pressure drop of the tube and socket. Drops caused by an abrupt change in the cross-sectional area of a system include *both* expansion and contraction drops for variations in conduit area, and are additive. While these values may be calculated for a system of known dimensions, it is beyond the scope of this article to cover such calculations. Suffice to say that when the overall pressure-drop and airflow requirements are determined, it is possible to match the requirements to the blower characteristics to achieve satisfactory system cooling.

Blower Characteristics

Air blowers come in many shapes and sizes and some are "good" and some are "poor." The most commonly used impellers in air cooled systems are *squirrel-cage* (*centrifugal*) *blowers*, and *axial fans*. The important characteristics of an air cooling system are the relationship between blower-outlet back pressure (in inches of water) and the airflow (in cubic feet per minute), and these characteristics determine the blower to be used. It is foolhardy to determine the "good" air impellers from the "poor" impellers by intuition.

Graphs of typical squirrel-cage blower performance for various units are given in Figs. 4, 5 and 6. The areas under the curves are regions in which the blower does useful work. It can be seen that as the back pressure rises, the efficiency of the blower decreases until, at some critical value of back pressure, the blower ceases to function as a useful device and merely "wind mills" the air about the impeller blades and cavity. This is termed "blower cutoff." Blowers vary to a great degree in their ability to cope with back pressure: low speed, open axial fans are the least efficient, while high speed squirrel-cage devices have somewhat higher efficiency.

Squirrel Cages and Axial Fans

The typical squirrel-cage blower has a multi-bladed impeller wheel rotating within a tightly fitting housing.[5] Small units normally have the discharge edge of the blade inclined forward, in the direction of rotation. The inexpensive axial fan, on the other hand, has a few, large, wide blades (usually four), slowly rotating in the open air or in a short housing section. More expensive vane-axial impellers have more blades (five or six) and rotate at higher speeds.

[5] The most efficient centrifugal blowers have a housing which closely fits the edges of the rotor. Excessive air gap between the rotor and the rim of the housing destroys the ability of the blower to work into back pressure.

Squirrel-cage blowers are often cataloged according to impeller wheel diameter and rotational speed. Thus a No. 2½ blower has a wheel diameter of 2½ inches, and may be available in a number of speeds, of which 2800, 3100, 6000 and 9000 r.p.m. are common off-the-shelf values. For a given wheel size and design, the c.f.m. delivered is proportional to blower speed, as is the ability to withstand system back pressure. Using the electrical analogy again, it may be said that the "voltage regulation under load" (ability to overcome back pressure) of any blower increases as the impeller speed increases. Unfortunately, as the impeller speed increases, the air noise, motor noise, vibration and unit cost also increase. While a 2800-r.p.m., or even a 6000-r.p.m. squirrel-cage blower may have a tolerable noise level, many 15,000-r.p.m. blowers create sufficient air noise to deafen even the most dedicated DX-contest operator.

Examination of the blower curves shows that a "trade-off" exists between rotational speed and wheel diameter and, generally speaking, a large impeller wheel running at moderate speed will be more satisfactory and less noisy than a smaller wheel running at a somewhat higher speed.

Inexpensive axial fans deliver large volumes of air at rather low rotational speeds and are generally fairly quiet, but suffer more than do squirrel-cage blowers from the effects of back pressure (Fig. 7). *Most small, low-speed axial fans and squirrel-cage blowers cannot move sufficient air into moderate values of back pressure to properly cool modern external-anode transmitting tubes, and their use should be tempered with caution.*

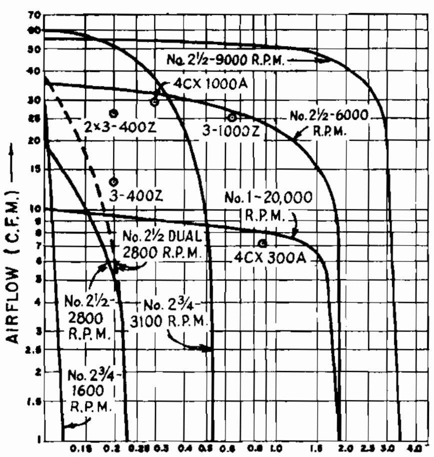

Fig. 5—Typical performance curves, similar to those of Fig. 4, for No. 2½, No. 2¾, and dual No. 2½ blowers. Notice that the No. 2½ 6000-r.p.m. blower will handle the cooling requirements of all of the tube types indicated, since the requirement points fall below the curve for this blower. "Wind-milling" is clearly indicated by the rapid drop of airflow to an unacceptable rate at the higher values of back pressure. Also notice that the use of dual blowers provides more airflow than a single blower of the same type at low back pressures, but does not project the "wind-milling" point.

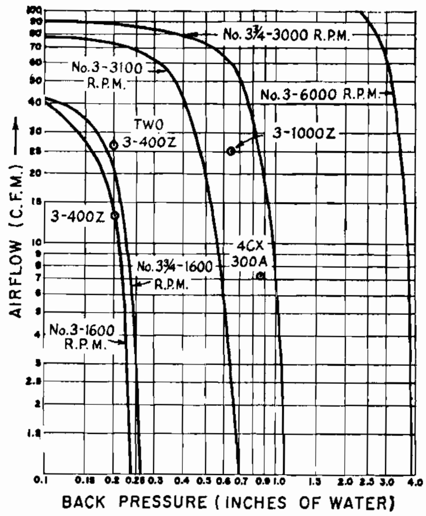

Fig. 6—Typical performance data for 3-, 3½- and 3¾-inch blowers. The low-speed (3100 r.p.m.) No. 3 blower cannot deliver the required flow of air into the back pressure offered by 3-1000Z or 4CX300A cooling systems. The 6000-r.p.m. unit, however, will handle the requirements of any of the tube types indicated, or, in fact, a pair of any of these tubes. Notice that doubling the speed of the blower more than triples the back-pressure capability. Although catalog-rated at "50 c.f.m.," the No. 3¾ 1600-r.p.m. blower is suitable for only low values of back pressure. When the speed is increased to 3000 r.p.m., the same-size blower will handle any of the tubes indicated.

Designing a Forced-Air Cooling System

An application of this design data, as a practical exercise, is the determination of a proper blower to cool a 4CX300A external-anode tetrode in an air-system socket, operating at various values of plate dissipation and 250° C. (maximum) anode temperature. If the ambient (room) air temperature is taken to be 50°C., airflow requirements to hold the tube temperature rise below 200°C at sea level, and at an altitude of 10,000 feet are graphed in Fig. 8. (These curves are based on figures taken from the data sheets for the 4CX300A and 4CX300Y.) If full 300-watt plate dissipation is desired at sea level, the air system must provide 7.2 c.f.m. of air *at the socket* of the tube under a combined tube and socket pressure drop of 0.58 inch of water. At an altitude of 10,000 feet (where the air is thinner), cooling requirements rise to 10.5 c.f.m. at a corresponding back pressure of 0.85 inch.

The additional pressure drops of the cooling system including back pressure developed by the cabinet structure, may be substantial, and must be added to the drop determined for the tube and socket. Unless a manometer is used to check the operation of the complete system, the additional back pressure caused by the duct coupling the blower to the tube and socket is a matter of speculation. If a pressurized chassis is employed having a large, internal open area (plenum chamber) into which the blower works, the additional system back pressure will be obviously less than if the blower has to force air through a flexible hose and around large under-chassis components. Experience has shown that it is generally safe to estimate an additional 50-percent back pressure requirement when the blower works directly into a reasonably clear pressurized chassis area, and this is the most common situation in amateur practice.

Taking the 50-percent extra back pressure requirement as par, an additional back pressure of 0.29 inch of water must be overcome for a total back pressure requirement of $0.58 + 0.29 = 0.87$ inch of water for the system. The use of an inexpensive manometer to verify this educated estimate is recommended in the design of new equipment.

Turning to the squirrel-cage-blower data charts, it can be determined that a No. 1 wheel running at 20,000 r.p.m., or a No. 2 wheel running at a speed of 6500 r.p.m. will do the job, as will a No. 2½ wheel operating at 6000 r.p.m.[6] A No. 3 wheel operating at 6000 r.p.m. is more than satisfactory. The No. 3 wheel running at 3100 r.p.m., however, is unsatisfactory, as the graph of Fig. 6 indicates that the impeller "windmills" above approximately 0.6 inch back pressure, and that its output falls rapidly to cutoff zero slightly above this figure. In the interest of minimum noise it would seem prudent to choose a No. 2½ blower running at 6000 r.p.m. to properly cool the 4CX300A with a suitable safety margin. If blower size is a problem, it may be necessary to use a No. 2, higher speed blower at some increase in noise level.

Glass Tubes

Large glass transmitting tubes (above approximately 200 watts plate dissipation) require moderate amounts of cooling air passed over the filament and plate seals to hold the seal temperature below a safe maximum value. As a large quantity of heat is dispelled by infrared radiation from the hot anode, the air requirements of the glass-style tube are usually less than that amount required for an equivalent value of dissipation from an external-anode tube whose anode temperature is limited by the insulator seal. Proper cooling of the glass tube requires that the air pass over the filament seals and then be guided past the glass envelope by a chimney. The chimney must be transparent to infrared radiation from the tube. Lastly, the air passes over the plate seal and is exhausted from the system.

The 3-400Z zero-bias triode, for example, requires 13 c.f.m. at a back pressure of 0.13 inch at the socket, while the 3-1000Z requires 25 c.f.m. at a back pressure of 0.43 inch at the socket. While the amount of air required is of about the same quantity for comparable values of plate dissipation in external-anode tubes, the back pressure demand is considerably less for the glass envelope design, as the air is not required to flow through interstices of a cooling anode.

[6] The curves shown in Fig. 4 and those following, apply to specific models. All models of the same size number and rotational speed (even those of the same manufacturer) do not necessarily have the same performance ratings.

Referring again to the blower and fan charts, it can be seen that a 3-400Z may be adequately cooled by a No. 2 blower (6500 r.p.m.), or a No. 2¾ blower (3100 r.p.m.), when a 50-percent margin is allowed for extra system back pressure. Two 3-400Zs will require twice the c.f.m. at the same back pressure, or a total of 26 c.f.m. at a system pressure of 0.2 inch. In this case, the No. 2¾ blower (3100 r.p.m.) would suffice.

A single 3-1000Z requires 25 c.f.m. at a system back pressure of 0.64 inch (including the 50 percent safety factor), and a single No. 2½ (6000 r.p.m.) blower will do the job.

Either a single 3-400Z, or a pair, may be cooled by a 4-inch axial fan (2800 r.p.m. or higher), as such a device will work into a back pressure of about 0.2 inch. The 3-1000Z, however, cannot be properly cooled by the axial fans listed in Fig. 7.

In all of these examples, full plate dissipation is assumed, and the proper air-system socket and chimney for the tube in question is employed.

The unknown factor in the determination of the overall air-system requirement is the additional back pressure caused by the conduit system or plenum chamber arrangement. This is the reason that the tube manufacturer is reluctant to specify a particular blower for a given tube, as he does not know the characteristics of the overall air system to be used. If the blower works into a reasonably clear under-chassis area sealed for air leaks, and the air is exhausted through the tube socket, the safety factor of about 50 percent in back pressure mentioned earlier should be satisfactory. If, on the other hand, the blower is placed at some distance, with a connecting hose to the socket, blower requirements may rise by a factor of ten or more. *The only safe way to determine the actual requirements of a given air system*

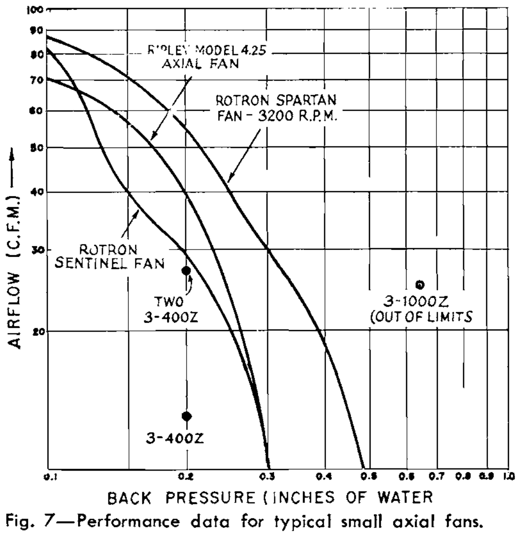

Fig. 7—Performance data for typical small axial fans. Medium-speed axial fans are suitable for a single 3-400Z, or a pair of this type. Axial fans must operate into a plenum chamber that transmits the air to the tube socket without introducing prohibitive additional back pressure. Tube date shown includes 50% extra back pressure, as discussed in the text.

TABLE I

Air requirements and suggested blower data for various air-cooled tubes.
Data is given for single tube, with 50 percent back-pressure allowance.

Tube Type	Socket	Chimney	C.F.M.[5]	Back Pressure (In. Water)[5]	Blower Size	R.P.M.
3-400Z	SK-410	SK-416	13	0.2	3	1600
3-1000Z	SK-510	SK-516	25	0.64	2½	6000
					3¾	3000
4-400A[1]	SK-410	SK-406	13	0.25	3	3100
4-1000A[2]	SK-510	SK-506	25	0.64	2½	6000
					3¾	3000
4CX250B[3]	SK-600 Series	SK-606 Series	6.4	1.12	2½	6000
4CX1000A 4CX1500B[4]	SK-800 Series	SK-806 Series	22	0.3	3	3100
5CX1500A	SK-840 Series	SK-806 Series	47	1.12	3	6000

[1] SK-400 socket requires 14 c.f.m. at 0.37 inch.
[2] SK-500 socket requires 25 c.f.m. at 0.9 inch.
[3] Data applies to 4X150A for 250 w. dissipation.
[4] Air requirement for 1000 w. dissipation.
[5] Sea level requirements.

is to make back pressure measurements with a manometer.

When in doubt as to the air-system requirements, it is wise to provide an oversupply of air at somewhat greater back-pressure values than estimated by a study of system requirements. It is impossible to damage a tube by too much air, unless the tube is blown out of the socket by the air blast! All low-speed blowers and axial fans should be avoided, too, unless a manometer is used to check out the system under full tube-dissipation conditions.

A summary based upon a 50 percent back-pressure safety factor for various tube and blower combinations is given in Table I.

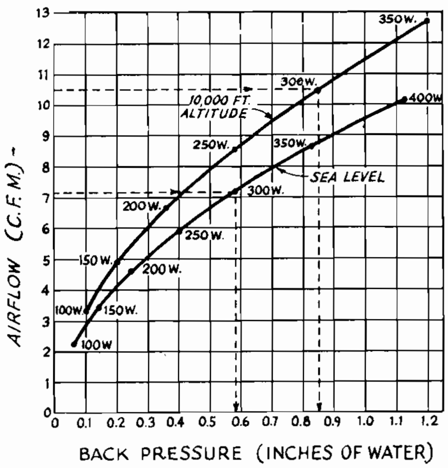

Fig. 8—Typical curves indicating how the cooling requirements increase with an increase in tube plate dissipation. These curves are for a 4CX300A mounted in an air-system socket. The dashed lines point out how the airflow requirements also increase with altitude because of the thinner atmosphere.

Tube Temperature Measurements

Measurement of tube temperature is possible, and the most reliable technique is to use a thermocouple attached to the tube. A somewhat simpler technique for the radio amateur is to determine tube-surface temperature by the use of temperature-sensitive paint.[7] The paint is applied in a *very thin* coat to the tube and dries to a powdery finish after application. At its critical temperature, it melts and virtually disappears. After subsequent cooling, it has a crystalline appearance which indicates that the surface with which it is in contact has exceeded the critical temperature.

Reliable temperature measurements can be made with temperature-sensitive paint only if it is applied in a very thin coat over small areas of the surface to be measured. The substance as supplied by the manufacturer is too thick for use in the presence of forced-air cooling and must be thinned, using only the thinner recommended by the manufacturer. The paint is applied with an air brush or atomizer (or with an aerosol dispenser) in a well-diluted spray, as the amount required to produce a reliable indication is virtually unweighable. A convenient set of equipment for using the temperature-sensitive paints is an atomizer with several vials, each equipped with an airtight cap. One vial may be filled with thinner for cleaning the atomizer, while the re-

[7] Temperature sensitive "decals" are also available

(*Continued on page 142*)

mainder are filled with properly-thinned paint sensitive to several different critical temperatures.

Measurements made with temperature-sensitive paint yield basic information sometimes obtainable in no other way, and are the "ounce of prevention" that is worth a "pound of cure."

Conclusion

Tube-surface temperatures are the ultimate criterion by which cooling adequacy may be judged. As tube life is closely related to surface temperatures, reliable temperature or cooling information is very important to the equipment-design engineer and the radio amateur. The proper choice of air blower is important, especially in cases where a high order of back pressure exists in the air system. Use of a manometer to determine back pressure, as well as the use of temperature-sensitive paint, allow the circuit designer to construct a satisfactory forced-air cooling system at the lowest possible cost.

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The Cathode-Driven Linear Amplifier

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THE cathode-driven, or grounded-grid, amplifier[1] is ideally suited to amateur s.s.b. or c.w. service and seems to be gradually relegating the grid-driven amplifier to the junk box. The attributes of the cathode-driven amplifier are impressive: it has reasonable power gain, it usually requires no auxiliary neutralization below 30 megacycles or so, it offers lower residual circuit capacitance, and parasitic suppression is not difficult. Under certain conditions, moreover, inherent negative feedback exists in this configuration, to the benefit of amplifier linearity. Finally, a portion of the cathode r.f. drive power shows up in the output circuit, thus providing a degree of "free" output power not otherwise available from a conventional grid-driven circuit.

Strictly speaking, the extra output power is not "free," as r.f. power is expensive compared to d.c. plate power and may only be "free" if it is unavoidably available. It is generally referred to as *feed-through* power, but the implication in this term may be misleading, as this portion of the drive power does not appear in the load circuit of the cathode-driven stage until after it is converted to a varying d.c. plate potential effectively in series with the main amplifier power supply. This *converted drive power* performs a useful function in Class AB_2 and Class B linear service by swamping out the undesirable effects of nonlinear grid loading and presenting a reasonably constant load to the exciter[2].

Neutralization and control of grid isolation within the cathode-driven amplifier permit the designer to adapt the basic circuit to the particular operating conditions at hand. Power gain and feed-through power may be varied, and the amplifier can be stabilized for proper operation over a wide frequency range.

The purpose of this article is to examine certain aspects of the cathode-driven amplifier, not widely recognized, that afford additional flexibility and versatility under particular operating conditions, and which permit accurate and complete neutralization to be achieved when needed.

The Basic Cathode-Driven Circuit

First discussed in *QST* in September, 1933,[3] the cathode-driven circuit has generated a considerable body of literature over the past few decades (see bibliography). The circuit is believed to have first been conceived circa 1920 by Ernst Alexanderson of alternator fame. Used about 1938 in European short-wave broadcast and TV service, this unique amplifier configuration became popular in U.S. post-war low-channel TV transmitters about 1944 or so.

The basic cathode-driven circuit is shown in Fig. 1. It may be operated either as an oscillator or as an amplifier by proper choice of components and potentials. The grid of the tube is nominally at r.f. ground potential and the exciting signal is applied to the cathode, or filament. For amplifier service, if it is assumed that the cathode is instantaneously driven positive with respect to ground (the grid), the plate will become more positive with respect to the cathode, and also with respect to ground. The instantaneous plate voltage, in effect, is developed in series and in phase with the exciting voltage, and the driver and amplifier stages may be thought of as op-

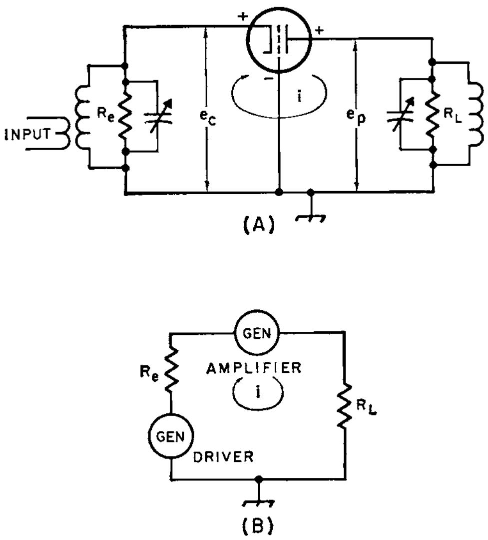

Fig. 1-A—The cathode-driven circuit. Driving voltage (e_c) is applied to the cathode of the amplifier and the output voltage (e_p) appears across the plate load impedance, R_L, in phase with e_c. The grid of the tube is at nominal ground potential. B—The driver and cathode-driven amplifier are in series with respect to the amplifier r.f. voltages. Amplifier cathode current (i) flows through the load resistance of the driver, contributing a degree of r.f. feedback.

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[1] The term "cathode-driven," or "grid-isolation" is preferred over "grounded-grid," as the latter implies that the grid is at r.f. and d.c. ground. This is often not the case.

[2] Pappenfus, Bruene and Schoenike, *Single Sideband Principles and Practice*, McGraw-Hill Book Co., N. Y. (1964).

[3] Romander, "The Inverted Ultra-audion Amplifier," *QST*, September, 1933.

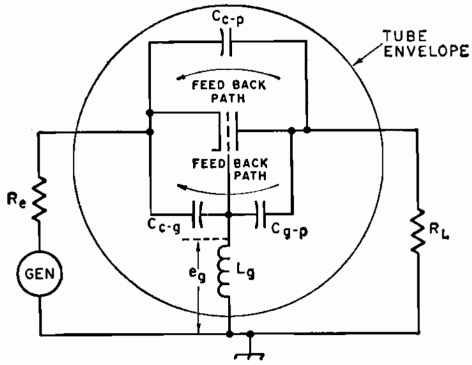


Fig. 2—Distributed constants of cathode-driven tube. Cathode-to-plate (C_{c-p}), cathode-to-grid (C_{c-g}) and grid-to-plate (C_{g-p}) capacitances, together with grid-lead inductance (L_g) make up feedback paths that must be neutralized for proper operation of the cathode-driven amplifier. Two feedback paths enter the picture: the direct path from plate to cathode via C_{c-p}, and a more devious path via series capacitors C_{c-g} and C_{g-p}.

erating in series to deliver power to the load, R_L. The delivered power is the sum of converted drive power and amplifier power, less any power from the driver required by the amplifier grid circuit. A parallel-tuned circuit is used in the cathode of the amplifier to enhance the regulation of the driver stage, to complete the plate circuit r.f. return path to the cathode, and to provide proper driver termination over the operating cycle.[4]

As the cathode-driven amplifier is effectively in series with the driver stage, the output current passes through the load resistance of the driver (R_e), causing a voltage drop across that resistance which opposes the original driving voltage. This indicates that inverse feedback is inherent in the cathode-driven amplifier to some degree if the driver has appreciable load resistance.[5]

Neutralization

The familiar cathode-driven amplifier used in h.f. amateur service is usually not neutralized. That is to say, no external neutralizing circuit is built into the amplifier. This omission has led to the general belief that the "grounded grid acts as a shield" and neutralization is not necessary in any and all cathode-driven amplifiers. The accepted proof of this belief is the fact that most h.f. amplifiers, in most instances, will not oscillate in use. Operation of an unneutralized cathode-driven amplifier in the upper portion of the h.f. spectrum, however, may provide unpleasant surprises. Many amateurs have found to their chagrin that such an amplifier is often a tricky "beast" to tame at 10 and 6 meters.

The reason for the unwanted instability is simple. Wires and leads represent finite inductances, and their position relative to each other and to other circuit components represents capacitance; both these quantitites may have an effect upon amplifier performance. Vacuum tubes have these distributed constants within their envelopes in the form of interelectrode capacitances and lead inductance.

Voltage feedback from output to input through the distributed constants of the tube has a deleterious effect on amplifier performance. The magnitude, phase and rate of change with respect to frequency of this feedback determine the dynamic stability of the amplifier, and control of feedback is termed *neutralization*. The purpose of neutralization of any amplifier, regardless of circuitry, is to make the input and output circuits independent of each other with respect to voltage feedback and the resulting reactive currents.[6] When a cathode-driven amplifier is operated at the higher frequencies, the internal capacitances and the inductance of the grid structure of the tube contribute to the degree of feedback (Fig. 2). To achieve stability, the various feedback paths through the distributed constants inherent in the tube structure must be balanced out, or nulled, in some fashion by neutralization techniques. Proper neutralization may be defined *as the state in which, when plate and cathode tank*

[4] C. E. Strong, "The Inverted Amplifier," *Electrical Communication* (England), Volume 19, No. 3, 1941.

[5] J. J. Muller, "Cathode Excited Linear Amplifiers," *Electrical Communication* (England). Volume 23, September, 1946.

[6] In fact, the cathode voltage is dependent to a degree upon the output voltage, as the input and output circuits are in series.

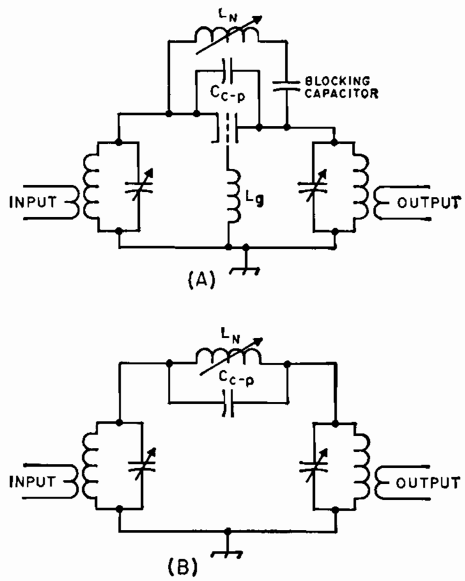


Fig. 3-A—Cathode-plate inductive neutralization. Capacitive feedback path between cathode and plate via C_{c-p} may be neutralized by making the capacitance part of a parallel-resonant circuit tuned to the operating frequency by the addition of L_n. A blocking capacitor is used to remove the d.c. plate voltage from the coil. Neutralization is frequency sensitive. B—Equivalent circuit; high-impedance parallel-resonant circuit nullifies feedback path between input and output circuits via plate-to-cathode capacitance.

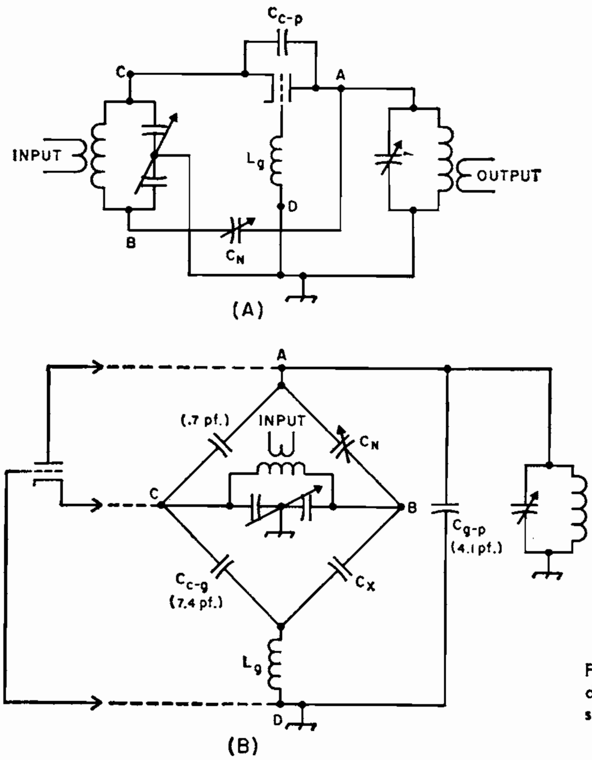


Fig. 4-A—Cathode-plate bridge neutralization. Balanced input provides euqal out-of-phase voltages at points *B* and C. When C_N is equal to $C_{c\text{-}p}$, equal out-of-phase voltages will cancel each other at point *A* and feedback path via $C_{c\text{-}p}$ is neutralized. B—Neutralization circuit redrawn in bridge form, with typical capacitance values for 3-400Z triode shown in parentheses. Bridge is balanced except for capacitance C_x, representing residual capacitance to ground at point *B*. If the balanced input circuit is high-C in comparison to interelectrode capacitances of tube, capacitances $C_{c\text{-}g}$ and C_x are swamped out and bridge may be considered to be balanced.

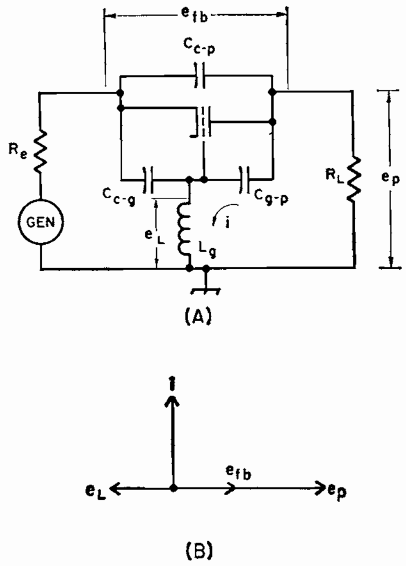


Fig. 5-A—Three-terminal representation of cathode-driven tube. See text for explanation. B—Vector representation of feedback voltages in cathode-driven tube.

circuits are resonant, maximum cathode voltage, minimum plate current, and maximum power output occur. This definition implies that the input and output circuits are independent of each other with respect to common reactive currents, and that tuning of the circuits reveals no interaction.

As the grid of the tube is at nominal ground potential in a cathode-driven amplifier, it appears that this element may act as a screen, or shield, between the output and input circuits and that instability or oscillation due to feedback paths through the interelectrode capacitances of the tube may be avoided, or reduced to negligible values. At the lower frequencies, particularly with respect to well-shielded, low-gain tubes, this belief may be true. However, in the higher-frequency region the practical tube (i.e., the tube that can be built) departs to an important degree from this simplified concept.

Neutralizing the Cathode-Driven Amplifier

Stable operation of the cathode-driven amplifier often requires some form of neutralization when the frequency of operation approaches the upper reaches of the h.f. spectrum. Complete circuit stability requires neutralization of *two* feedback paths, for which separate techniques are required.

The first feedback path involves the cathode-to-plate capacitance, C_{c-p} . Although the capacitance involved is small, the path is critical and requires neutralization. Neutralization may be accomplished either by a shunt inductance (Fig. 3) or by a balanced capacitive bridge circuit (Fig. 4). The first technique consists of connecting a reactance from plate to cathode of such magnitude as to transmit back to the cathode circuit a current equal in value but opposite in phase to the current passing through the cathode-to-plate capacitance. The bridge technique is a version of the well-known capacitance neutralizing circuit used in conventional grid-driven amplifiers to balance out the effects of grid-plate capacitance. The balanced input circuit provides equal out-of-phase voltages to which the cathode of the tube and the neutralizing capacitor are coupled. As the value of the neutralizing capacitor is equal to the cathode-to-plate capacitance of the tube, the voltages are balanced at the junction of the two capacitances, which is the plate termination of the cathode-driven tube. Both capacitances are usually quite small, and the effect of series lead inductance in the bridge circuit is relatively unimportant. Consequently a reasonable bridge balance over a wide frequency range may be obtained with a single setting of the neutralizing capacitance.

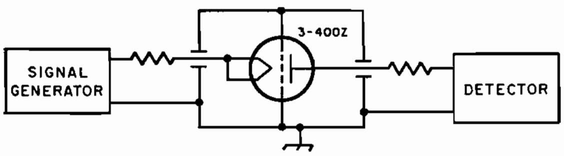
The shunt-inductance neutralizing circuit of Fig. 3, on the other hand, has the disadvantage of requiring adjustment for each working frequency, as the external inductance and cathode-to-plate feed-through capacitance form a frequency-sensitive parallel-resonant circuit at the operating frequency.

Either neutralizing circuit may be properly balanced [7] even though the grid of the tube may not be at actual ground potential because of internal grid inductance, L_g. Intrastage feedback resulting from this inductance requires a separate, unique solution, apart from the neutralizing technique just discussed.

Grid-Inductance Neutralization

The second feedback path in the cathode-driven stage includes the grid-to-plate capacitance, the cathode-to-grid capacitance and the series grid inductance, L_g, as shown in Fig. 2. The grid inductance represents the sum of all possible feedback paths *through* the grid structure, plus the actual series inductance of the grid structure. In practical tubes, there is no possibility of avoiding all inductance in the path between the active grid element of the tube and ground. This path exists because the grid is not a solid, intercepting structure. After all, openings must exist to permit electrons to pass from the cathode to the plate! Capacitance leakage can exist between the cathode and the plate through these openings. In addition, Maxwell's equations state that changing electric and magnetic fields propagate each other through space. In the vicinity of the real grid structure, the electric field about the "input" side of the structure gives rise to currents flowing in the structure which, in turn, cause an electric field to exist about the "output" side of the structure. In addition, electromagnetic coupling through the interleaved grid structure is also observed[8].

[7] With physically large tubes having appreciable series input inductance, in-phase neutralization is often required. This may be achieved by adding external cathode-to-plate capacitance, or by detuning the shunt inductor from the condition of parallel resonance.



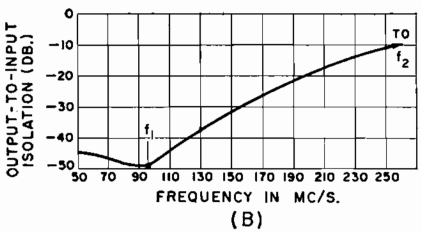


Fig. 6-A—The self-neutralizing frequency of a cathode-driven triode may be measured by observing the transmission properties of the cold tube when treated as a three-terminal network. B—Typical plot of intrastage isolation of 3-400Z triode mounted in test fixture. Self-neutralizing frequency of tube is about 100 megacycles.

These spurious coupling paths result in an apparent r.f. leakage through the cathode-to-grid and grid-to-plate capacitances that is often many times greater than that predicted by actual measurement of the internal capacitances. A simplified picture of this complex path may be seen as an inductance in series with the grid-to-ground path, common to both input and output circuits (Fig. 2). If this path is not neutralized, a voltage e_g appears on the grid of the tube which either increases or decreases the driving voltage, depending upon the value of internal capacitances and grid inductance. With sufficient spurious grid voltage, the cathode-driven stage may oscillate, or be unstable, even though the cathode-to-plate feedback path discussed earlier is completely neutralized.

The voltage e_g on the so-called "grounded grid" is determined by a complex action between the total cathode-to-plate capacitance and a separate low-Q circuit composed of a capacitive voltage divider (C_{c-g} and C_{g-p} in series) together with the grid inductance, L_g. A certain frequency at which these two feedback paths nullify each other is termed the *self-neutralizing* frequency (f_1) of the tube. This frequency usually occurs in the lower portion of the v.h.f. spectrum with small transmitting tubes. All the elements comprising the neutralizing circuit are *within the tube*. However, connecting the tube into the circuit by wiring or socketing will alter this frequency.

The self-neutralizing phenomenon comes about because of a frequency-sensitive voltage balance that takes place within this network, Fig. 5A, and which may be explained by a simple vector diagram, Fig. 5B. The r.f. plate voltage (e_p) causes a current (i) to flow through C_{g-p} and L_g. If the reactance of L_g is small in comparison with the reactance of C_{g-p} (as would be the case below the self-neutralizing frequency), the current i will lead the plate voltage e_p by 90 degrees. In flowing through L_g this current will develop a grid voltage (e_L) which is 180 degrees out of phase with e_p and with the voltage e_{fb} fed back to the cathode via C_{c-p} and series-connected C_{c-g} and C_{g-p}.

At some frequency the voltage e_L developed across L_g will just equal the voltage fed back through the interelectrode capacitances (e_{fb}). The frequency at which e_L is equal to e_{fb} is the self-neutralizing frequency. At this frequency a cancellation of feedback voltages occurs and the complex feedback path is nullified, or "neutralized." (A second, somewhat higher, frequency at

[8] Feedback admittance also is enhanced by the self-inductance of the grid wires, which provides common coupling between input and output circuits. The inductive coupling may partially compensate for the feedback through the cathode-to-plate capacitance. (See Bibliography, item 3.)

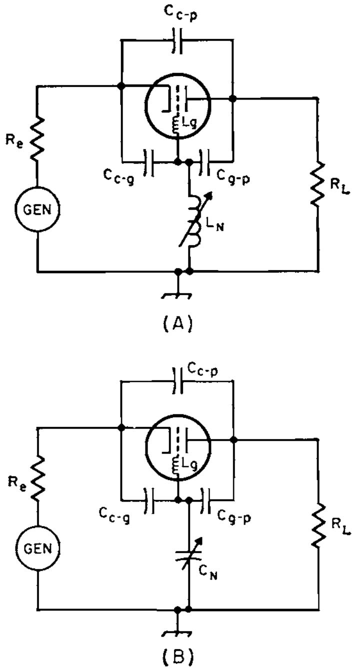


Fig. 7-A—The point of self-neutralization may be shifted lower in frequency by the addition of an inductance (L_n) in series with the grid-to-ground termination of the tube. B—The point of self-neutralization may be shifted higher in frequency by the addition of a capacitor (C_N) instead of an inductor.

which the complex grid configuration is in a series-resonant state with respect to intrastage isolation is called the *grid series-resonant frequency* (f_2) of the tube.)[9]

The Self-Neutralizing Characteristic Curve

The self-neutralizing characteristic of a cathode-driven triode may be determined by treating the tube as a passive three-terminal network and measuring transmission as a function of frequency. The tube is placed in a test fixture which is contrived to insure that the frequency measured is dependent on the tube and socket only (Fig. 6). A signal is applied to the "cold" tube through an appropriate attenuator and a detector is used to measure the transmission voltage through the tube. Investigation over a range of frequencies will produce a typical plot such as shown in Fig. 6B. The point of maximum isolation is the self-neutralizing frequency, f_1. Measurements are not quantitative, as nothing is known about the impedance of the input or output circuits. The relative isolation with respect to frequency, however, is the interesting parameter.

The self-neutralizing frequency (a broad null of several hundred kilocycles) may be moved about by manipulation of the external grid-to-ground circuitry of the tube, or by changing the capacitive feedback path. Or, if desired, a secondary point of neutralization may be created, as described later. If the desired frequency of operation is above the self-neutralizing frequency the voltage developed on the "grounded grid" will be too great and the series grid inductance, L_g, must be reduced, or the feedback path adjusted to establish self-neutralization. If the operating frequency lies below the self-neutralizing frequency, the voltage on the "grounded grid" will be insufficient to cancel the feedback voltage and the series grid inductance must be increased.

The portion of the plot around the point f_1 has been experimentally verified by observing the intrastage leakage (transmission) properties of a 3-400Z zero-bias triode mounted in an SK-510 socket and fixed in a partition in an r.f.-tight enclosure. Observation was over the range of 50 to 250 megacycles, and the self-neutralizing frequency was seen to be in the neighborhood of 100 megacycles (Fig. 6B). Above this frequency, the intrastage isolation gradually deteriorated as the series-resonance frequency, f_2, of the grid element was approached. Near the latter frequency, tube operation is impractical, being further complicated by transit-time effects and other v.h.f. phenomena.

The Self-Neutralizing Frequency

The self-neutralizing frequency of a cathode-driven triode depends to a large degree upon the size of the tube, the interelectrode capacitances, the physical configuration of the grid structure and the inductance of the grid leads and terminals. Below this frequency, the tube can be neutralized by the addition of a small inductor (L_N, Fig. 7) in the grid-to-ground path. Above this frequency, neutralization may be achieved by reducing the reactance of the path by the addition of a suitable series capacitance, C_N. To demonstrate this a variable capacitor was placed in series with one grid terminal of the 3-400Z mounted in the test fixture. At any frequency between f_1 and 250 megacycles the shape of the plot could be altered by adjustment of the capacitor, providing a neutralizing "null," Fig. 8, in the curve of about the same amplitude as observed at the lower frequencies. The Q of the neutralizing circuit (one grid lead plus the capacitor) was considerably higher than the Q of the grid system, and the neutralizing adjustment proved to be rather frequency-sensitive. The original self-neutralizing frequency (f_1) was little altered by the addition of the auxiliary circuit.

A second test conducted on a larger tube (the 3X2500A3, a 2.5-kw. low-μ triode) showed that it could be neutralized on the lower-frequency side of the self-neutralizing frequency f_1 by the addition of a suitable inductor between the grid terminal and ground. Both techniques are shown in Fig. 7.

It should be noted that intrastage self-neutrali-

[9] "Care and Feeding of Power Grid Tubes", application bulletin No. 13, EIMAC, a Division of Varian, San Carlos, Calif.

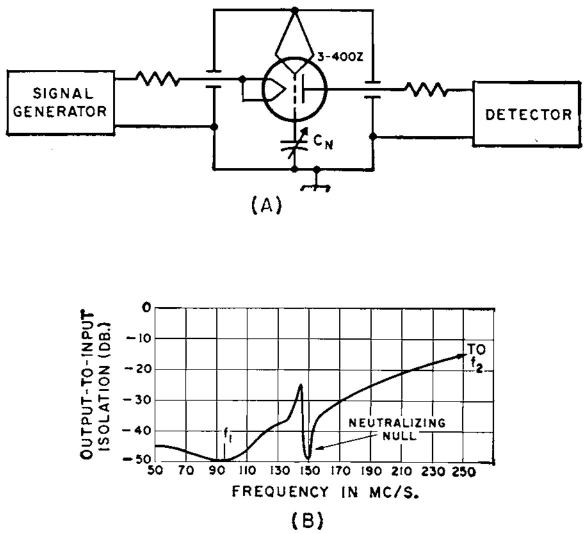


Fig. 8-A—The 3-400Z may achieve neutralization over a wide v.h.f. range by the addition of a series capacitor in one grid lead. Neutralization adjustment is frequency sensitive and must be peaked for maximum intrastage isolation of the operating frequency. B—Plot of intrastage isolation of 3-400Z, showing neutralizing null added by the series grid capacitor. Null may be moved about between f_1 and f_2. A similar neutralizing effect may be obtained at frequencies lower than f_1 by the circuit shown in Fig. 7-A.

zation and cathode-plate neutralization are interlocked. In the lower portion of the v.h.f. spectrum only one technique may be necessary to achieve a satisfactory degree of neutralization, at least as far as amplifier stability goes. At 6 meters, for example, either system will completely stabilize many amplifiers in most situations. At higher frequencies such is not the case, and both feedback circuits may require attention and manipulation to allow the amplifier in question to be properly neutralized.

General Remarks

Conclusions to be drawn as to the degree of intrastage isolation, or as to the requirement for neutralization in a cathode-driven amplifier, tend to be clouded unless backed by measurements made on the equipment, just as is the case with grid-driven amplifiers. In the latter instance, neutralization of the circuit is almost taken for granted. Not so with cathode-driven amplifiers, as adequate isolation and stability have often been achieved at the lower frequencies even with tubes that were not designed for this purpose. It is unwise to jump to the general conclusion that this special situation exists in all cases.

At the lower frequencies, particularly with well-shielded, low-capacitance tubes, neutralization may not be necessary, and this permits the circuit designer to make use of circuit techniques and practices that afford variation of power gain, converted drive power, and degree of inverse feedback to the cathode driven amplifier. Specifically, these parameters may be varied to meet the demands of the system or to adjust the converted drive power requirement of the amplifier to match the available drive power of the exciter. These circuit schemes, however, should not be confused with the separate problems of amplifier neutralization, discussed in this article.

A future article will discuss *super-cathode-driven* and *semi-cathode-driven* circuits. The authors wish to thank W. H. McAulay, W6KM, and R. I. Sutherland, W6UOV, for their help and suggestions in preparation of this article. QST

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Semi- and Super-Cathode-Driven Amplifiers

BY WILLIAM I. ORR,* W6SAI AND WILLIAM H. SAYER,** WA6BAN

Operating conditions for linear amplifiers exist which offer advantages to the circuit designer and equipment user. Power gain and "feed-through" power of the stage may be varied, and reduced intermodulation distortion is achieved by manipulation of the ratio of cathode to grid drive, as discussed in this article.

In a previous article covering problems peculiar to cathode-driven ("grounded-grid") amplifiers[1] it was pointed out that when well-shielded tubes are operated in cathode-driven circuits in the h.f. region, neutralization is not always necessary for achieving circuit stability in properly designed equipment. If required, neutralization could be easily applied in one or more forms. The cathode-driven amplifier, moreover, permits the designer to include a degree of additional negative or positive feedback, in the form of grid driving voltage, to establish desired operating conditions. Specifically, the applied grid voltage may be used to vary the power gain and so-called "feed-through" power of the amplifier and, in a special case for tetrode and pentode tubes, this permits the elimination of the screen supply, screen power being taken from the r.f. drive. Circuits that make use of auxiliary grid-drive voltage are termed *semi-* and *super-cathode* driven. This article discusses the application of these circuits to amateur practice.

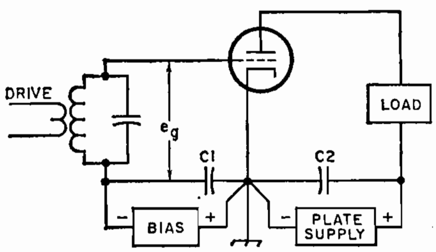

Fig. 1—The grid-driven amplifier. Drive signal (e_g) is applied between grid and cathode. When the grid is positive with respect to ground, plate potential becomes more negative with respect to the cathode (ground). Instantaneous plate voltage is out of phase with grid drive voltage, and the two circuits are common only at the cathode (ground) point. Bias and plate power supplies are considered in the circuit for d.c. and out of the circuit from an r.f. point of view, by virtue of bypass capacitors C_1 and C_2. Class of operation is determined by bias and drive signal voltage levels.

The Grid-Driven Circuit

The grid-driven circuit is a good place to start investigation.

Fig. 1 is a block diagram of a conventional grid-driven triode amplifier. For simplification, neutralization is not shown, and power and r.f. circuits are greatly simplified. The driving signal, e_g, is applied between grid and cathode (ground). In a perfect amplifier, input and output tuning adjustments are independent of each other and the grid and plate voltages are 180 degrees out of phase.

Driving power is the amount of signal power dissipated by the grid, if the grid is driven sufficiently positive to attract electrons from the cathode, plus any power demanded by various circuit losses. The class of operation is defined by bias voltage and driving-signal level. In the case of Class AB_1 operation, grid-drive requirements are very low because the grid is never driven positive and therefore no grid current is drawn. Class AB_2 or class B operation may call for a moderate amount of driving power on positive signal peaks when grid current is drawn. For Class A and B modes of operation, the output waveform is a replica of the input waveform, and the circuit may be used for linear amplification. When the circuit is adjusted for Class C operation (with bias greater than the cutoff value and plate current flowing in pulses less than one-half an operating cycle) the linear relationship between input and output signal no longer exists and the operating parameters are unsuited for linear amplification.

The Cathode-Driven Circuit

Fig. 2A illustrates a triode amplifier, simplified as previously explained, in which the drive signal e_c is applied between grid and cathode, with the grid grounded with respect to the r.f. signal. Operation of this *circuit* is strikingly different than that of the grid-driven configuration of Fig. 1, but *tube* operation is the same. That is to say, when the grid is driven positive in either case, the cathode is driven negative and plate current flows. The mode of operation is, of course, determined as before by choice of bias and drive signal levels.

In the linear mode, if it is assumed that the cathode is driven negative with respect to the grid (r.f. ground), the grid is then positive in relation to the cathode. With a positive grid signal, the plate becomes more negative with

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[1] Orr and Sayer, "The Cathode-Driven Amplifier", *QST*, June, 1967.

respect to both cathode and ground. On the other half of the operating cycle, when the cathode is positive with respect to the grid, the plate becomes more positive in relation to ground. Thus the plate potential responds in like polarity to the cathode-drive signal. During the time that the cathode is driven negative, converted drive voltage is added to the d.c. plate potential, as shown in Fig. 2B. An extra amount of instantaneous plate voltage is developed in series and in phase with the cathode signal. The driver, then, may be pictured as a second plate supply effectively in series with the main plate supply of the amplifier. The portion of *converted drive power* which appears in the plate circuit as additional r.f. output is commonly called "feed-through" power, even though it does not "feed through" anything. The effective d.c. plate-to-cathode voltage on the cathode-driven tube during negative signal excursions of the cathode voltage is the sum of the d.c. plate voltage and the r.m.s. value of the cathode voltage, e_c. During positive signal excursions (when the grid is negative with respect to the cathode) the tube is cut off, so the subtractive portion of the drive voltage during this part of the operating cycle is ineffective.

The plate voltage of the cathode-driven amplifier thus varies over the operating cycle, deviating from the nominal power supply value to a somewhat higher value in accord with the modulation envelope of the drive signal. The value of converted drive power in the plate circuit is approximately the product of the r.m.s. cathode voltage and the d.c. plate current ($e_c \times I_p$). The total drive requirement is the sum of grid-drive power, converted drive-signal power, and grid-circuit losses. Grid-drive power and grid-circuit losses remain relatively constant in either mode of operation, the extra converted grid-drive power appearing only in the cathode-driven mode.

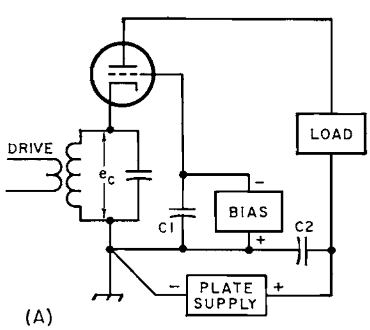
As in the grid-driven case, the cathode-driven amplifier may be operated Class A, B or C by proper choice of bias and drive-signal level. High-μ triodes and *some* tetrodes may be operated in near Class B condition, with zero grid bias and screen grounded. This subtle distinction should again be emphasized: Circuit configuration and operating mode are two separate and distinct things, and the use of the loose, inclusive term "grounded-grid" tends to blur and confuse the distinction. A circuit may be cathode driven, but is not necessarily "grounded-grid" from either an r.f. or d.c. point of view.

Envelope Modulation

Comparison of the operating parameters of grid-driven and cathode-driven circuits utilizing the same tube type in the same class of operation reveals that drive requirements of the tube are identical, with the obvious exception of the converted drive power which is a characteristic of the cathode-driven circuit. When comparing stage gains between the two modes of operation, the additional converted-drive-power requirement of the cathode-driven stage effectively reduces the overall power gain of the circuit and provides a degree of inverse r.f. feedback roughly equal to the reduction of stage gain.

In the case of tetrode and pentode tubes, a portion of the converted drive power is used to supply screen power as well as plate power during negative drive-signal excursions. This is why such tubes operating in cathode-driven service usually have reduced d.c. screen voltage; the remainder of the required screen voltage is supplied by the driving source, reaching the desired maximum value at the peak of the driving signal (an example is the Collins 30S-1 amplifier, which utilizes a Class AB_1 4CX1000A tetrode in this circuit).

R.f. envelope modulation resulting from envelope variations of plate and screen voltage affords a degree of inverse feedback not easily obtainable in a grid-driven stage. A reduction of intermodulation distortion has been observed for various tetrode tubes operated in this fashion, amounting to 3 to 10 decibels improvement in unwanted third-order products.



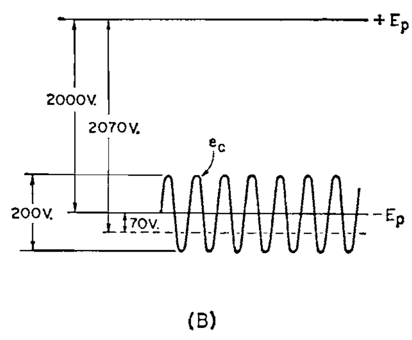


Fig. 2—The cathode-driven amplifier. (A) Drive signal (e_c) is applied between cathode and grid (r.f. ground). When the cathode is driven positive with respect to the grid, the plate potential becomes more positive in relation to ground. Instantaneous plate voltage is in phrase with cathode drive and in series with it, from a d.c. point of view. (B) Effective plate voltage during the negative portion of the cathode drive signal is the sum of the d.c. potential plus the r.m.s. value of the converted drive voltage. In this case, d.c. plate voltage is 2000, peak-to-peak r.f. drive voltage is 200, and r.m.s. drive voltage is 70. The effective plate voltage is 2070.

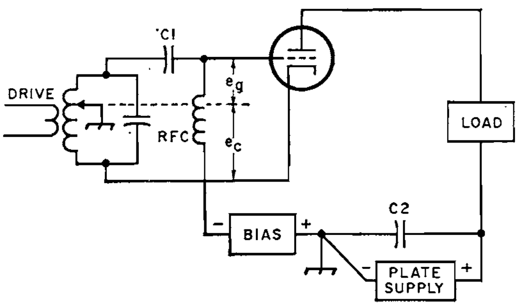


Fig. 3—The semi-cathode-driven amplifier. Auxiliary drive voltage (e_g) is applied to the grid out of phase with the cathode signal (e_c), raising the stage gain and lowering the converted drive power. Total drive requirement is reduced as the proportion of grid to cathode excitation is raised. When e_g is large compared to e_c, the circuit resembles a grid-driven stage, with e_c serving to boost drive level and reduce stage gain over simple grid-driven requirements.

Semi-Cathode-Driven Operation

Operating modes between grid-driven and cathode-driven states are possible by movement of the ground point to positions between the configurations of Figs. 1 and 2. The r.f. ground return is thus electrically placed between the grid and cathode of the tube (Fig. 3). This configuration is termed semi-cathode-driven service. In this mode of operation, a portion e_g of the drive signal is applied to the control grid out of phase with the cathode signal, e_c. While the total grid-to-cathode driving voltage remains the same no matter where the ground point is placed, the ratio of cathode volts to grid volts varies with the position of the ground return. The limiting condition is reached, of course, when the cathode is at r.f. ground and full drive is applied to the grid of the tube. At intermediate points the degree of converted drive power varies directly with respect to the cathode drive voltage. Stage gain is inversely related to cathode drive voltage, and the total drive power is closely related to cathode drive voltage. Thus, stage gain is enhanced and total drive power is reduced as the circuit departs from the cathode-driven mode and approaches the grid-driven mode.

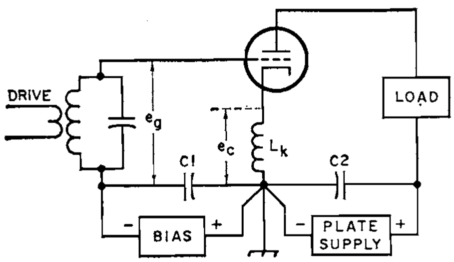


Fig. 4—Grid-driven amplifier having cathode lead inductance. Drive voltage, e_g, flowing through input circuit creates voltage drop (e_c) across cathode lead inductance, L_k, by virtue of cathode r.f. current. Cathode voltage tends to oppose grid drive, lowering power gain of stage and making it more difficult to drive.

In other words, if an auxiliary voltage, out of phase with the cathode signal, is applied to the control grid of a cathode-driven stage it will boost stage gain and reduce converted drive power. This is a very convenient scheme to match the drive level of a linear amplifier stage to the power output of a given exciter, if the output of the latter tends to be marginal.

Looking at the other side of the coin, it can be realized that introduction of out-of-phase cathode-drive voltage into a grid-driven stage will tend to lower the power gain of the stage, making it more difficult to excite, as excitation power must be translated into converted drive power. This is exactly the case in v.h.f. amplifiers having excessive cathode lead inductance across which a portion of the drive signal is developed (Fig. 4). Cathode lead inductance, in other words, robs the v.h.f. amplifier of grid drive because it converts needed excitation into converted drive power appearing in the plate circuit, thus effectively lowering the power gain of the stage and boosting the excitation level required for a given value of power output.

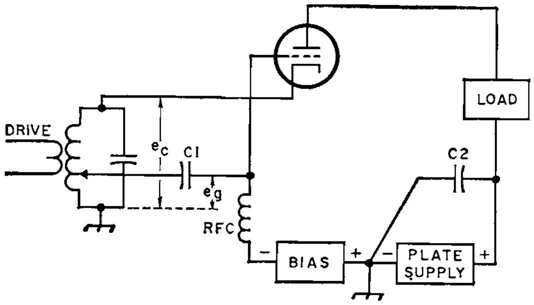


Fig. 5—Super-cathode-driven amplifier. Drive voltage, e_c, is applied to cathode, and a portion, e_g, is applied to the grid in phase with cathode signal. Stage gain is lowered and converted drive power is raised. This circuit may be used to absorb extra driving power of exciter and convert it to plate-circuit power.

By judicious division of the drive signal between grid and cathode of an amplifier stage it is possible to balance the drive requirement with the available power from the exciter. Many modern s.s.b. linear amplifiers make use of cathode-driven circuitry, but the drive requirement is something of a hit-or-miss situation. If the s.s.b. exciter is modest in power output, it is possible to raise the power gain (reduce the converted drive requirement) of a particular "grounded-grid" amplifier by introducing out-of-phase drive voltage into the grid circuit, effectively "matching" the drive requirement of the amplifier to the power capability of the exciter.

Super-Cathode-Driven Operation

Shown in Fig. 5 is a circuit in which a portion, e_g, of the total drive signal is applied in phase to the grid of a cathode-driven amplifier to effectively oppose the cathode voltage. This is

TABLE I

4CX300A, Class B, Typical Super-Cathode-Driven Service	
Plate Voltage	2000
Grid Voltage	0
Screen Voltage	330 (peak)
D.C. Plate Current	
no signal	15 ma.
max. signal	250 ma.
Drive Power	75 watts
Measured Power Output	375 watts
Intermodulation Distortion Products:	
3rd order =	−46 db.
5th order =	−49 db.

4CX300A, Class AB_1, Typical Grid-Driven Service	
Plate Voltage	2000
Screen Voltage	350
Grid Voltage	−55
D.C. Plate Current	
no signal	100 ma.
max. signal	250 ma.
Drive Power	0 watts
Measured Power Output	300 watts
Intermodulation Distortion Products:	
3rd order =	−27 db.
5th order =	−36 db.

termed super-cathode-driven operation. Drive power is increased and stage gain is decreased, as compared to a conventional cathode-driven circuit. It may appear fatuous to design an amplifier which demands more than the minimum driving power; however, this circuit may be used to advantage when it is necessary to absorb excess drive power from the exciter, over and above that value required by normal drive and "feed-through." The circuit, moreover, has other advantages that make it appealing to the circuit designer. An early s.s.b. transmitter design, for example, had series-connected super-cathode-driven low-μ tubes adjusted so that the drive power contributed by the first stage and amplified by the second stage equalled the power supplied by the second stage. Each stage thus contributed 50 per cent of the total output power, permitting the transmitter to make use of four tubes in a two-stage amplifier, neither stage being individually capable of producing the desired power level.

The Super-Cathode-Driven Tetrode

When used with a tetrode or pentode tube, super-cathode service permits the cathode driving signal to serve as a screen power source. Screen-to-cathode voltage (e_{sg}) is supplied on alternate half-cycles of the drive signal as shown in Fig. 6. The control grid may be driven (tied to the cathode) or tapped to a point on the cathode circuit. In the former case, the tube resembles a low-μ triode having an abnormally high converted-drive-power characteristic combined with an unusually low value of static plate current. (Static plate current, of course, is low because static screen voltage is zero.) Operating data for a 4CX300A in this mode are given in Table I. Note the great degree of improvement in intermodulation distortion as compared to grid-driven service. Super-cathode drive requirement is high, but a large proportion of this is converted to output power as indicated.

The super-cathode-driven tetrode circuit of Fig. 6 may be modified by the inclusion of screen and bias voltages to shift the operation to near Class AB_1. Power gain rises and rectified drive power drops as this shift is made. Screen and grid potentials, in fact, may be varied to match the power gain of the stage to a predetermined

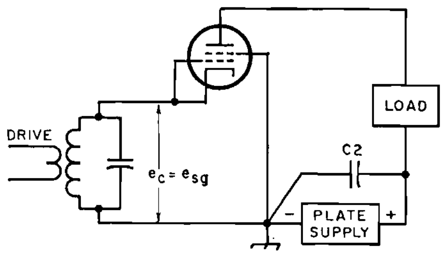
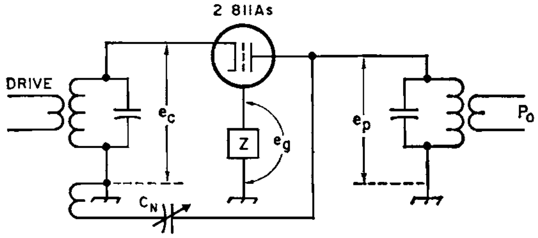


Fig. 6—Super-cathode-driven tetrode amplifier. Tetrode tube may be strapped as a triode with control grid tied to the cathode. Drive voltage, e_c, serves as screen voltage, e_{sg}, since screen is at ground potential. Resting plate current is low as screen voltage is zero with no drive signal. Converted drive power is large, as is total grid drive requirement. Screen and control-grid bias voltages may be added to this circuit to raise power gain of tube and decrease total drive requirement.

TABLE II

Two 811A, Class B, Cathode-Driven, Neutralized (values for two tubes given)		
	$Z = 0$, $e_g = 0$	$Z = 75$ $\mu\mu$f./tube $e_g = 70$ volts, r.m.s.
Plate Voltage	1500	1500
D.C. Plate Current	335 ma.	335 ma.
Power Input	500 watts	500 watts
Grid Current	50 ma.	50 ma.
Plate Load	3000 ohms	3000 ohms
Drive Power (total)	25 watts	50 watts
e_c (r.m.s.)	61	130
e_g (r.m.s.)	0	70
e_p (r.m.s.)	980	1080
Converted Driver Power	20 watts	46.5 watts
Power Output	322 watts	386 watts



drive level, falling between the very low requirement of Class AB_1 service and the rather large Class B requirement specified in Table I. Power gain is set by screen-voltage adjustment, and the static plate current is determined by the bias level.

Plate-Circuit Feedback

The circuits discussed so far are special instances of the general circuit of Fig. 3 where the control grid of a cathode-driven amplifier is lifted above r.f. ground to permit the injection of an auxiliary drive signal. The previously-mentioned circuits are ones in which the feedback voltage is derived from the driving signal. It is also possible to derive the feedback voltage from the output signal of the stage, with the tube included in the feedback loop.

In the circuit of Fig. 7A, the feedback signal is applied to the grid of a cathode-driven stage. Generally speaking, external feedback is not applied to the tube element receiving the drive signal; applying it to separate element minimizes the reaction of the feedback signal upon the driving source. If the feedback is in, or out of, phase with plate and cathode signals, amplifier operation is comparable with that of the super- and semi-cathode-driven circuits discussed earlier.

The degree of feedback is determined by the capacitance ratio C_1/C_2. In normal practice, C_1 is of the order of 1 to 5 pf. and C_2 may fall in the range of 100 to 500 pf. The greater the capacitance of C_2 as compared with C_1 the less will be the feedback signal at the grid of the tube.

This feedback technique is used in the Collins 30L-1 amplifier to match the drive requirement of four cathode-driven 811A tubes to the nominal power output of the S-line exciter (about 100

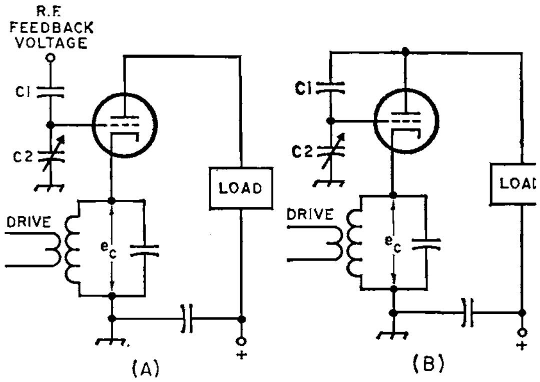


Fig. 7—Plate-circuit feedback. (A) Auxiliary control voltage may be applied to the grid of a cathode-driven stage, either in phase or out of phase with the driving signal. Capacitors C_1 and C_2 form a voltage divider, with grid voltage determined by setting of C_2. (B) Feedback voltage at grid of amplifier may be derived from plate signal, providing negative feedback and increasing drive requirements. Stage gain is decreased and rectified drive-power level is increased. As feedback level is increased, stage must be reneutralized.

watts). The nominal drive requirement of four cathode-driven 811A's is about 50 watts without additional feedback. Sufficient feedback is introduced by the choice of capacitor C_2 to raise the drive requirement of the amplifier to about 100 watts. At the same time, a reduction in intermodulation distortion of about 3 decibels is achieved. The feedback voltage is derived from the plate circuit as shown in Fig. 7B.

It should be noted that use of the grid element of the cathode-driven stage for auxiliary signal injection tends to upset the neutralizing balance of the stage to a degree. This may not be too important with well-shielded tubes used below 30 megacycles, but can become important in the lower reaches of the v.h.f. spectrum. As the power gain of the stage is reduced by decreasing the value of C_2 in Fig. 7, the neutralizing circuit (if any) must be rebalanced for minimum intrastage feedback.

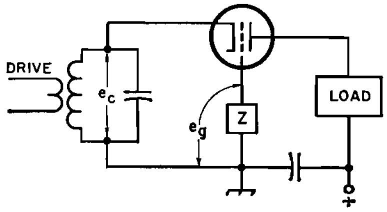


Fig. 8—Grid impedance in cathode-driven amplifier. General case for semi- and super-cathode driven amplifiers is summarized by placement of impedance Z in grid return. Magnitude and sign of Z determine stage gain, converted drive power, and total drive power. For average tubes in h.f. region, Z is usually positive (inductive), making the stage somewhat easier to drive than normal, and also making stage prone to instability and oscillation when external feedback circuits are not controlled. Feedback current (Iz) flows through grid-plate capacitance.

Effect of Grid Impedance

Both the semi-cathode-driven and super-cathode driven circuits may be summarized in the general case shown in Fig. 8, where an impedance Z is placed between grid and ground. Amplifier operation is assumed to be below the self-neutralizing frequency of the tube. It can be shown that when Z is positive (inductive) the amplifier is in a semi-cathode-driven mode and (as compared with a simple cathode-driven amplifier) requires a lower-than-normal value of driving power and exhibits less-than-normal converted drive power. On the other hand, when Z is negative (capacitive) the amplifier is in a super-cathode-driven mode, requiring a higher-than-normal value of driving power and exhibiting more-than-normal converted drive power. An example of an 811A cathode-driven amplifier having both zero and negative grid-impedance characteristics is shown in Table II. The magnitude and sign of Z, therefore, set the stage for operating parameters of the seemingly simple "grounded-grid" amplifier. Practical limits to the manipulation of impedance Z exist, as large values of impedance prevent effective neutralization of the cathode-driven stage.

Envelope-Modulation Circuits

A number of unorthodox linear amplifier circuits have come into vogue in the past decade (the "ZL-linear," the "Class C" linear, the "G2DAF" linear, etc.), all of which utilize some form of envelope modulation. A subsequent article will deal with these interesting circuits.

The authors wish to thank W. H. McAulay, W6KM and Raymond Rinaudo, W6KEV, for their assistance in the preparation of this article. QST

amateur service newsletter
W6SAI

The 3-500Z in Amateur Service

Here's a new zero-bias triode from Eimac that features increased plate dissipation.

The 3-500Z is a heavy-duty power triode of 500 watts plate dissipation. It is exceptionally well suited for use as a class-B amplifier in rf or audio application. It may be used in zero-bias linear-amplifier service at plate potentials up to 3000 volts, eliminating bulky and expensive screen and bias power supplies.

Of particular interest to the radio amateur is the use of the 3-500Z as a grounded-grid (cathode-driven) amplifier for ssb service. One 3-500Z is capable of a PEP input of over 1100 watts, requiring only 30 watts PEP drive power. Intermodulation distortion products at this power level are 30 dB or more below one tone of a two-tone test signal. At 2000 volts, moreover, over 500 watts of power output are obtainable with distortion products better than 38 dB below one tone of a two-tone signal. Typical operating characteristics for the 3-500Z are listed in **table 1.** A data sheet covering operation of the 3-500Z may be obtained at no cost by writing to me.

In cases requiring additional plate dissipation, the 3-500Z may replace the 3-400Z. The forced-air requirements for the two tubes are approximately equal and a blower capable of 13 cubic feet per minute at a back pressure of 0.2 inch is satisfactory for a single 3-500Z. (Use blower size #3 at 1600 rpm. For two 3-500Z's, use blower size #3 at 3100 rpm, or size #2½ at 6000 rpm.)

The zero-signal pate current of the 3-500Z is somewhat higher than that of the 3-400Z. When the 3-500Z is used to replace the 3-400Z, a means of reducing the zero-signal plate current is recommended, particularly if the equipment is power-supply limited. Only a few volts of bias from a low impedance source are required. The simplest way of obtaining well-regulated bias voltage is to place a zener diode in the filament return circuit of the 3-500Z **(fig. 1).**

The 1N4551 zener diode has a nominal voltage drop of 4.7 volts and an impedance of 0.1 ohm, making it ideal for this service. At this value of bias, the zero-signal plate current of the 3-500Z at a plate potential of 3250 volts is reduced from 160 to approximately 90 milliamperes.

The zener diode may be bolted directly to a cool area of the chassis which will act as a heat sink. Additional VOX-selective bias may be placed in series with this zener diode to reduce standby current of the 3-500Z to nearly zero in order to eliminate "diode noise" during reception and conserve standby power **(fig. 2).**

the grid-current meter

It is advisable to monitor the grid current of the 3-500Z as an indicator of correct drive and antenna loading. Too much grid current indicates underloading or overdriving and too little grid current indicates underdriving or overloading, other things being equal. As the grid must be held at rf ground, the grid meter must be introduced in such a manner as not to disrupt this circuit. A simple grid meter scheme is shown in **fig. 1.** Each grid pin is grounded through a .01-pF mica capacitor paralleled with a 3.3-ohm, 2-watt composition resistor. A small dc voltage drop exists across the resistor under normal tube operation. The voltage drop is read by a simple dc voltmeter (M1) calibrated in terms of grid current.

In the example shown, it is desired that the grid meter have a full scale indication of 200 milliamperes. The dc grid-to-ground resistance is about 1.1 ohm and, at a current of 200 mA, a voltage drop of 0.22 volts will be developed. The 0-1 dc milliammeter is converted to read 0.22 volts full scale by the inclusion of a series multiplier resistor. The sum of the resistor plus the meter resistance should total 220 ohms.

3-500Z circuitry

No specific circuits are shown for the 3-500Z, since published 3-400Z circuitry applies equally well to this tube. Two 3-500Z's may be used in place of a single 3-1000Z with appropriate corrections in air flow, filament power requirements and zener bias (if necessary).

ham radio

table 1. Typical operation of the 3-500Z in grounded-grid rf linear-amplifier service.

DC plate voltage	3000	2500	2000	V
Zero-signal dc plate current	160	130	95	mA
Single-tone dc plate current	370	400	400	mA
Single-tone dc grid current	115	120	130	mA
PEP input power	1110	1000	800	W
PEP useful output power	750	600	500	W
Resonant load impedance	5000	3450	2750	ohms
Cathode input impedance	115	100	100	ohms
Intermodulation products (3rd order)	—30	—33	—38	dB

fig. 1. Zener diode bias circuit for the 3-500Z. A 1N4551, 4.7-volt, 50-watt zener diode provides cathode bias for the 3-500Z. Meter M1 (0-1 mA dc) reads grid current of the tube in terms of the voltage drop across the three grid resistors. Meter M2 reads plate current. The multiplier resistor plus internal resistance of meter M1 should total 220 ohms. Grid and filament bypass capacitors are 600-volt mica units (M). Other bypass capacitors are ceramic discs.

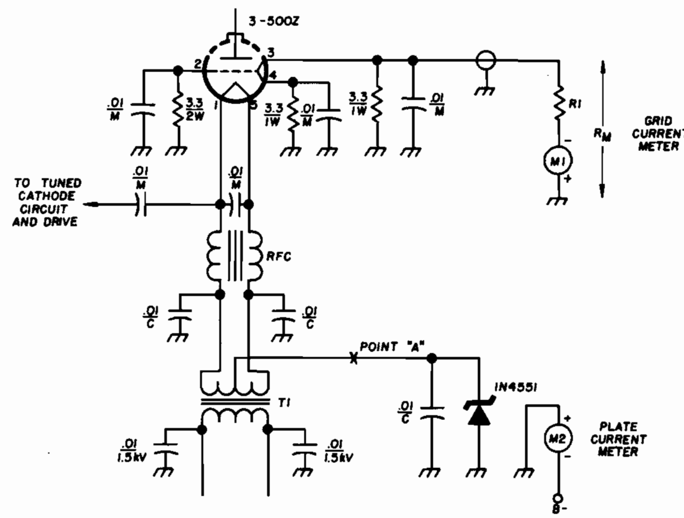
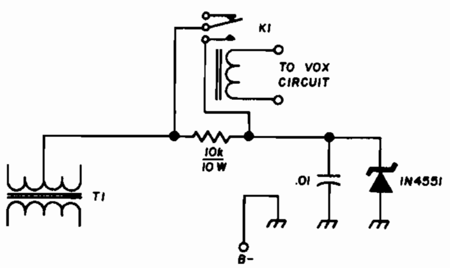


fig. 2. VOX-selective cutoff bias circuit. Additional cutoff cathode bias is added by the VOX relay to reduce standby plate current to near-zero, eliminating "diode noise" in a nearby receiver. The bias is added at point A in fig. 1.



amateur service newsletter
W6SAI

A Comparison of the Eimac 3-400Z and Eimac 3-500Z High-Mu Power Triodes

The new Eimac 3-500Z bears a resemblance to the 3-400Z, being of the same general configuration but somewhat larger, physically and electrically.

The following tabulation lists and compares some of the pertinent characteristics of these tubes. Complete data on each type is summarized in the individual data sheets, obtainable upon request from the Amateur Service Department, Eimac Division of Varian, San Carlos, California 94070.

Electrical	3-400Z	3-500Z	Note
Fil Voltage	5.0	5.0	
Fil Current	14.5	14.5	
Amplification Factor (Average)	200	160	X
Interelectrode Capacitance (Aver)			
Grid-Filament	7.4 pF	7.4 pF	
Grid-Plate	4.1 pF	4.1 pF	
Plate-Filament	0.07 pF	0.07 pF	
Frequency for Maximum Ratings	110 MHz	110 MHz	
Maximum Plate Dissipation	400 watts	500 watts	X
Mechanical			

Base	5 Pin Special	5 Pin Special	
Mounting Position	Vertical	Vertical	
Cooling	Radiation & Forced Air	Radiation & Forced Air	
Heat-Dissipating Plate Connector	Mounted on Tube	HR-6	X
Maximum Operating Temperatures			
Plate Seal	225°C	225°C	
Filament Seals	200°C	200°C	
Maximum Overall Dimensions			
Height	5.25"	5.875"	X
Diameter	3.57"	3.438"	X
Recommended Socket	Eimac SK-410	Eimac SK-410	
Recommended Chimney	Eimac SK-416	Eimac SK-406	X

Typical Operating Data	3-400Z	3-500Z	Note
R-F Linear Amplifier, Grounded Grid, Single Tone Conditions			
Typical Operation:			
D-C Plate Voltage	3000	3000	
Zero-Signal D.C. Plate Current (approx.)	100 ma	160 ma	X
Max. Signal D.C. Plate Current	333 ma	370 ma	X
Max. Signal D.C. Grid Current	120 ma	115 ma	X
Driving Impedance	122 ohms	115 ohms	X
Resonant Load Impedance	4750 ohms	5000 ohms	X
Max. Signal Driving Power	32 watts	30 watts	X
Peak Envelope Plate Output Power	655 watts	750 watts	X
Typical Operation:			
D-C Plate Voltage	2500	2500	
Zero-Signal D.C. Plate Current (approx.)	73 ma	130 ma	X
Max. Signal D.C. Plate Current	400 ma	400 ma	
Max. Signal D.C. Grid Current	142 ma	120 ma	X
Resonant Load Impedance	3450 ohms	3450 ohms	
Peak Envelope Useful Power Output	560 watts	600 watts	X
Intermodulation Distortion Products	-35 db	-33 db	X
Typical Operation:			
D-C Plate Voltage	2000	2000	
Zero-Signal D.C. Plate Current (approx.)	62 ma	95 ma	X
Max. Signal D.C. Plate Current	400 ma	400 ma	
Max. Signal D.C. Grid Current	148 ma	130 ma	X
Resonant Load Impedance	2750 ohms	2750 ohms	
Peak Envelope Useful Power Output	445 watts	500 watts	X
Intermodulation Distortion Products	-40 db	-38 db	X

Cooling Air Requirements	3-400Z		3-500Z	
Anode Dissipation:	Air Flow CFM	Pressure Drop (in-H_2O)	Air Flow CFM	Pressure Drop (in-H_2O)*
300 watts	-	-	6.6	0.023
400 watts	13	0.13	10.3	0.052
500 watts	-	-	13.0	0.082

*inches of H_2O

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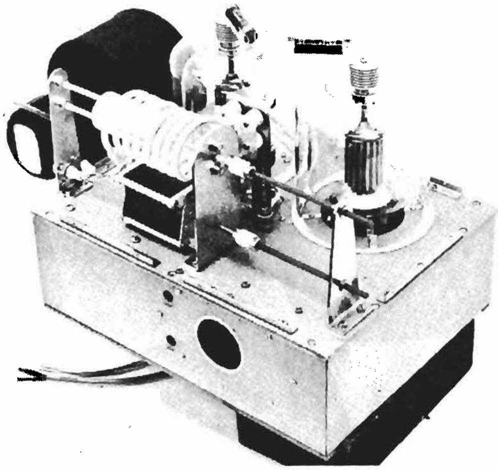


inductively-tuned high-frequency tank circuit

A method for achieving high efficiency in the "shadow region" between 14 and 54 MHz

William I. Orr, W6SAI, Eimac Division of Varian

Radio frequency amplifiers require a certain critical value of plate circuit impedance and Q for optimum performance at any frequency. Design deviations may lead to higher levels of intermodulation distortion or excessive harmonic radiation. While the design requirement may be quite tolerant in some cases, the mechanical assembly of the components and the choice of proper values become increasingly critical as the operating frequency nears the upper region of the hf spectrum. It is as though a "shadow region" exists that's too high for conventional lumped circuit components, yet is too low for conventional vhf linear and stripline techniques. The "shadow region" extends roughly from 27 through 54 MHz.

Above 27 MHz or so, the construction of a conventional high-power plate-tuning circuit having good Q and good efficiency can be vexing, as residual tube and circuit capacitances combine to assume a major portion of the tank circuit capacitance. It's possible, in fact, for this residual capacitance to be much larger than specified for proper design considerations. The unusually high circuit capacitance may lead to unreasonable Q and high circulating tank current, resulting in poor over-all efficiency and excessive heat loss in the tank circuit.

inductive tuning

To achieve good circuit efficiency and proper Q in the upper portion of the hf spectrum, it is convenient to resort to a different mechanical configuration than is commonly used at lower frequencies. One way to overcome problems of efficiency and Q is to reduce residual circuit capacitance to an absolute minimum by removing the cause of the largest portion of this unwanted capacitance: the tank tuning capacitor. Circuit resonance can then be established by including a fixed capacitance combined with a variable inductor.

The inductor can be a fixed, high-Q coil having a low-loss shorted turn introduced into one end. As the turn is moved within the coil, coil inductance is reduced, and resonance is established by correctly positioning the shorted turn **(fig. 1)**. The rf current in the shorted turn is high compared to the coil current; however, if the turn is of homogeneous structure and low-resistance material, turn losses will be small.

Shown in the photographs is a commercial 5-kW input PEP linear amplifier designed for any 500-kHz range between 27 and 54 MHz. The plate tank is a conventional pi network, inductively tuned by a shorted turn within the plate coil. The turn (or "slug") is moved into and out of the coil by a lead screw driven from a counter dial mounted on the amplifier panel. Tank circuit values were derived from pi network charts.[1]

The amplifier uses a pair of parallel-connected 3-1000Z high-mu triodes in a grounded-grid, cathode-driven circuit with zener diode bias.[2] The combined output capacitance of the tubes is approximately 15 pF. Stray circuit capacitance from plate to ground is less than 10 pF, which provides a minimum input capacitance (C1) for the pi network of about 25pF.

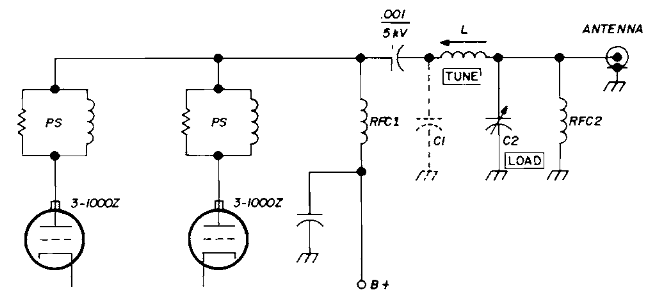


fig. 1. Inductively tuned tank circuit using conventional component values. C1 is the residual circuit capacitance plus tube output capacitance. Resonance is achieved by a variable shorted turn moved inside L.

Additional capacitance can be added for operation at lower frequencies. Ceramic or vacuum padding capacitors can be used for this purpose.

construction

Tank-circuit inductance is calculated for the low-frequency end of the tuning range in the usual manner. As the shorted turn is driven into the plate inductor, its inductance decreases, and resonant frequency increases. A range of about 500 kHz or so can be obtained with this assembly.

Construction details are shown in **fig. 2.** The shorted turn is a section of seamless copper water pipe 1-¾ inches O.D. by 3 inches long. Discs of insulating material are cut to fit into the ends of the pipe and are held in position with small pins or rivets. A

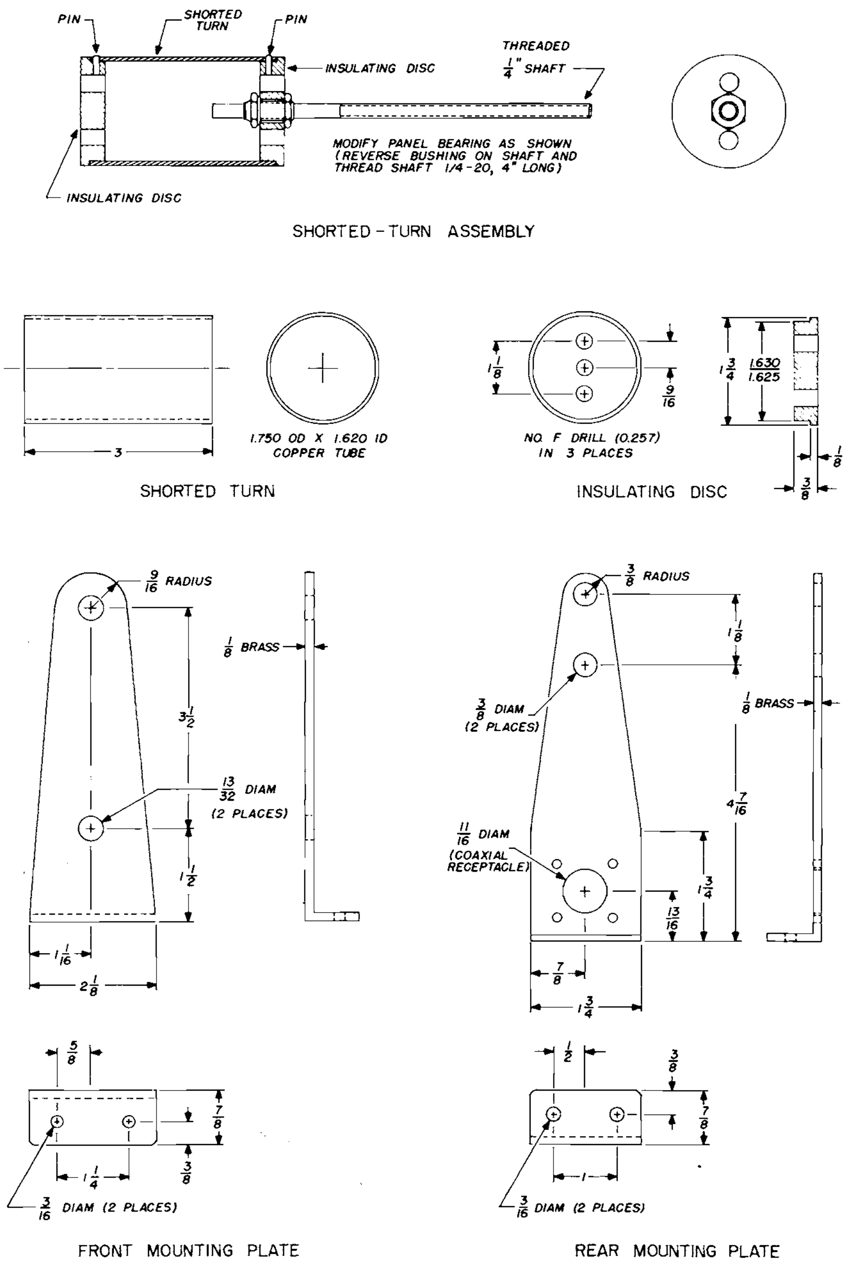


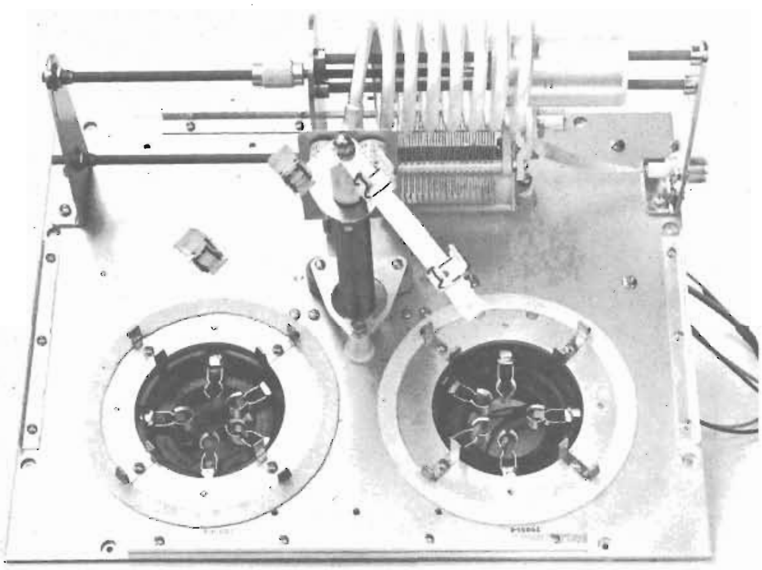
fig. 2. Mechanical details of the shorted-turn assembly.

threaded bushing is attached to one disc, through which the drive shaft extends. The shaft is ¼-inch-diameter copper rod, threaded with a ¼-20 die. Two fiberglass guide rods are mounted between the end support brackets to keep the shorted turn from rotating as it's driven back and forth. The threaded drive shaft is driven from the amplifier panel by an insulated extension shaft and an insulated coupling. The shorted turn is ungrounded at all times, and no moving parts carry rf current.

The plate tank coil is wound with ¼-inch-diameter silver plated copper tubing with an I.D. of 2-¼ inches, providing ample clearance for the shorted turn to move within the coil without danger of arcing. Heavy-duty copper mounting lugs are silver soldered to the coil. Resonance is initially established with a grid-dip oscillator. Indication of resonance is quite broad. Loading is accomplished with the pi network variable output capacitor, C2.

value will cause excessive fundamental rf power to dissipate in the parasitic resistor. The proper value of shunt inductance, while not particularly critical, should be determined for the operational range of the amplifier in each case.

The amplifier shown in the photographs was designed for commercial service and is included in a shielded cabinet as part of a larger package. A similar design using 4-1000A's has been built for commercial ssb service at the 6-kW PEP level—but that's another story. The basic design is adaptable to any high-power amplifier operating in the upper region of the hf spectrum.



top view of the high-power commercial amplifier. Inductively tuned tank is shown at rear. Tuning is the same as with a variable capacitor, except inductance decreases as shorted turn penetrates coil. Parasitic suppressors have been removed for this photo.

parasitic suppression

An important consideration in plate-circuit design is a parasitic suppressor. In this amplifier parasitic suppressors are included in each plate lead. Each suppressor is a 40-ohm, 16-watt *Glo-bar* resistor shunted across an inductor, which consists of a length of plate lead. The value of the shunt inductance is important. A too-small value won't completely suppress the tendency for vhf parasitics, and a too-large

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ham radio

amateur service newsletter
W6SAI

intermittent voice operation of power tubes

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The power capability of a transmitting tube is often the subject of long and heated discussions among amateurs (and even among equipment design and tube engineers). In the past, amazing things have been done to power tubes by daring amateurs who seemingly had an inexhaustible supply of replacement tubes at hand.[1] The tube manufacturer looks upon such goings-on with mixed emotions; he's proud his products can take such a beating, but he shudders at the gross overload he knows is taking place, and he has nightmares when he imagines that such tactics are being done by users who may be ignorant of the basic limitations of vacuum tubes. Sometimes that manufacturer may be his own worst enemy. When he speaks of the ruggedness, long life and reliability of his product, he may unintentionally be inviting some eager-beaver to prove the utter conservatism of his remarks.

Up to now, tube ratings have been based upon an absolute system providing "maximum" ratings and "typical" operating conditions for various classes of service, for use below a certain specified frequency. These ratings are designated as *Continuous Commercial Service* (CCS) and *Intermittent Commercial and Amateur Service* (ICAS). The CCS rating may be defined as, "that type of service in which long tube life and reliability of performance under continuous operating conditions are the prime consideration.[2]" The ICAS rating is defined to include the many applications where the "transmitter design factors of minimum size, light weight, and maximum power output are more important than long tube life." The term "intermittent" is used to identify operating conditions in which no operating or "on" period exceeds five minutes and every "on" period is followed by an "off" or standby period of at least the same or greater duration.

These ratings are of cold comfort to today's radio amateur. The first rating applies to high-reliability service (broadcast, military, etc.) wherein off-the-air time is critical or costly; and the second rating, by its very definition, excludes amateur operation in meaningful terms. Neither classification, moreover, applies to ssb or cw operation. Ssb and cw are more properly expressed in terms of peak-to-average power ratio rather than in terms of "on" and "off" periods.

Before new and meaningful ratings are proposed for today's operational modes, it would be prudent to look for a moment at transmitting tube ratings now in use and examine their validity. Contrary to often expressed belief, maximum ratings and typical operating conditions are not arbitrary figures dreamed up by the manufacturer to avoid answering legitimate questions posed by users. On the contrary, they are the result of careful analysis of tube geometry and of prolonged life tests run on typical production tubes with some guaranteed or expected life in mind. Properly understood, the maximum ratings and typical operating conditions can be employed by the tube user to decided advantage.

transmitting tube ratings

Maximum ratings (or absolute maximum ratings) are those limits within which all tubes of a given type should give satisfactory service and long useful life. Why they are necessary at all and how they are determined are discussed in this article.

The *data sheet* (often suspected of being written by the wise to impress the humble) informs the user of the capabilities and limitations of the tube, both of which are based upon the maximum temperature the elements of the tube can safely withstand for an expected life. Heat, then, is the enemy of unlimited tube life, but heat is the unfortunate consequence of making the tube work. Once the maximum tube capability is determined *a compromise of some kind must be made to establish useful life without exceeding the heat limitations, yet allowing some safety factor for "cockpit troubles."*

maximum plate dissipation

Plate dissipation is limited by the maximum safe temperature of the plate and plate-to-glass (or ceramic) seals of the tube. Generally speaking, the plate will withstand several times its maximum rated dissipation level for a short period of time. Other parts of the tube (glass envelopes, mainly) are greatly affected by the excessive heat radiated by the plate. High level of plate temperature may cause the grid, filament or envelope to become overheated. The grid structure may warp, the filament temperature may rise to an excessively high degree, or the tube envelope may be destroyed. These effects, however, are not instantaneous, and short periods of plate overload do not usually overheat the adjoining tube structure to a damaging extent. However, the user has no way of telling to what degree he can safely exceed the plate dissipation limit, or over what period of time this abuse can take place. The obvious conclusion is that the maximum plate dissipation rating should not be exceeded in continuous operation if long tube life is desired.

maximum plate voltage

The maximum plate voltage point is set at a value above which the internal or external insulators of the tube may arc over, or above which the envelope of a glass tube may be damaged from dielectric losses. Finally, a plate voltage ceiling tends to set a limit to the maximum rf charging current flowing in the plate and screen leads, or plate and grid leads in grounded-grid service. The charging current is a function of the rf plate voltage which, in turn, is a function of the dc plate voltage. Setting a limit on the dc voltage sets a limit on charging current without the difficult task of determining the current directly. This effect depends on frequency and is the reason for the upper frequency limit for maximum ratings.

average dc plate current

The fundamental limit on plate current is the available supply of electrons emitted by the filament or cathode of the tube. The maximum plate-current figure is intended to set a value which may be easily realized throughout the expected life of the tube. If operating conditions are chosen which require that the maximum plate-current limitation to be exceeded at the start of tube life, it may become increasingly difficult to maintain the desired value of plate current as the tube ages. There is a definite relationship between the *maximum instantaneous value* of plate and grid current and the *average dc* (meter reading) *plate current* which differs for each class of tube operation. In linear-amplifier service, for example, most transmitting tubes are run class AB_1, AB_2 (loosely termed class B).* In these cases, the peak plate current is about three times the indicated (average)

*Most class-B linear amplifiers are operated in class AB_2. Class-B operation is defined as cutoff operation with an 180° operating angle of plate current flow. Class AB_2 operation signifies less-than-cutoff condition with more than 180° operating angle.

dc plate current. For long life, the cathode emission should be great enough to provide two or three times the required peak value of plate, plus grid, plus screen current.

The user can quickly determine the allowable average dc plate current in linear service for thoriated tungsten filament-type tubes by merely multiplying the filament watts by a factor of about 5.5. This is a rule-of-thumb number that – over the years – has proven to give a conservative balance between allowable plate current and good tube life. For the 3-500Z, therefore, the filament power is 5 (volts) x 14.5 (amperes) = 72.5 watts. Therefore, allowable maximum average dc plate current for linear amplifier service is 72.5 x 5.5 = 400 milliamperes.

In the case of an indirectly heated cathode the rule-of-thumb is different. Emission from an indirectly heated cathode depends upon the emissive material and the active cathode area, assuming cathode temperature is the proper value. The rule-of-thumb in this case for oxide cathodes is that maximum average dc plate current is approximately 125 milliamperes for each square centimeter of cathode area. For example, the 4X150A tetrode has an active cathode area a little over 2.0 square centimeters and the average dc plate current rating is 250 milliamperes.

long pulse service

In pulse service where the "on" time is small compared to the "off" time, many transmitting tubes can be run to much higher peak power limits than are permissible in continuous service. In continuous service, the maximum voltage and current limitations are set with a safety factor in mind to consider average power dissipated on the tube electrodes. In pulse service, when the tube "rests" for an appreciable time, it is possible to set new guidelines on peak electrode dissipation and maximum ratings, provided the *average* electrode dissipation and maximum temperature ratings are not exceeded.

In pulse service (less than 0.1 second) a thoriated tungsten power tube may have an anode instantaneous peak dissipation capability as high as 100 times the average power capability, and the available filament emission may be as high as 80 milliamperes per watt of filament power. In some cases, the filament voltage has been boosted above normal to obtain emission levels as high as 150 milliamperes per watt with the penalty of greatly reduced tube life.

In the case of the oxide-coated cathode, the peak pulse current is not as clearly defined or as easily generalized as in the case of the thoriated filament tube. A figure of 500 milliamperes peak plate current per watt of heater power is often used for very short pulse service (less than 3 microseconds), and other numbers are available giving pulse plate current in terms of active cathode area.

In *long* pulse service (more than 0.1 second), the rise in temperature of the electrodes rather than the average power during the pulse often becomes the basic tube limitation, and twe maximum capability of the power tube is progressively derated as the pulse length increases. For a radiation cooled tube, a pulse length of 2.5 seconds is often considered equivalent to a continuous duty operation. In the case of an oxide-coated cathode, life tests indicate that a peak-to-average dc plate current ratio of 2.0 for long pulse (0.5 second) is not unrealistic. This corresponds to a duty factor of 0.5.

voice and cw operation

A shadow world exists between continuous duty (CCS) operation, and ICAS operation on the one hand and pulse operation. Amateur voice.and cw operations seem to fall into this shadow area. Cw operation may be compared to a form of pulse operation as it defines an "on" and "off" duty cycle wherein the two times are approximately equal. This would represent a duty cycle of fifty percent (0.5) and the pulse (cw) waveform would be nearly square.*

*Note quite true; waveshaping is necessary to some extent to reduce key clicks.

Voice operation, on the other hand, is a different and more complex problem. The voice waveform is not a square pulse; it has a large peak-to-average power ratio with irregular waveform. Normal speech, unclipped, uncompressed or otherwise altered, seems to have a peak-to-average ratio of about 14 dB.[3] Various compression and clipping techniques can reduce this ratio to 3 to 5 dB before severe distortion becomes apparent.[4] Thus, heavily clipped or compressed speech waveforms tend to resemble the cw duty cycle as far as the peak-to-average power ratio is concerned.

It is prudent to expect, therefore, that the power capability of a tube can be safely increased for *Intermittent Voice and CW Service* (IVS service) over the CCS rating provided the maximum element temperature of the electrodes is not exceeded and the cathode (or filament) emission is sufficient to satisfy plate and grid current peaks. In addition, the tube in question should not have an intolerable level of intermodulation distortion when operated in linear service under these enhanced conditions.

This type of intermittent operation is done everyday with the popular sweep tubes used in ssb equipment designed for amateur service. Small soft-glass envelope tubes (i. e., the 6LQ6) are run up to 250 or 300 watts PEP input with no apparent harm *provided* the maximum level of plate dissipation is held within reason, even though the average, long-term dissipation rating in tv service is only 30 watts or so. The user is taking advantage of the intermittent nature of amateur voice operation and the high peak-to-average ratio of the human voice to get more watts per dollar of tube investment. Many amateurs have found, to their regret, an overworked sweep tube tends to overheat and shows extremely short life when a voice clipper/compressor unit is used to bring up the average power of the equipment, or if extended cw operation is used. A moment's reflection upon the heating process in the tube will show the reason for this problem. The tube is being pushed so far that any margin of safety has vanished. Unfortunately, no one has yet been able to miniaturize the watt!

Thus, there's a limit beyond which pushing the transmitting tube becomes uneconomical. It may be well to push an inexpensive sweep tube to 300 watts PEP input, since a tuning error, or other maladjustment won't bankrupt the unlucky user. The owner of a more expensive transmitting tube, however, may well have second thoughts before he blasts his pet power tube. Obviously, some middle ground is called for where the peak-to-average power ratio of ssb and cw operation can afford new and conservative tube ratings more in line with today's usage.

intermittent voice service

In single sideband service, the two plate current values of significance are the single-tone plate current and the two-tone plate current. The ratio of single-tone to two-tone plate current may vary from 1.1/1 to 1.57/1, depending upon the class of operation. Two-tone plate current is useful as the magnitude of intermodulation distortion products may be specified as the reduction in decibels of one product from one tone of a two-equal-tone signal. Precedence exists, therefore, for providing typical operating data for linear amplifier service specifying the dc plate current under two-tone conditions of average plate current and plate current at the peak of the modulation envelope. Such data for the 8122 is shown in **table 1**.[2]

Based upon such data, extensive life tests have been run at the Eimac Division of Varian to determine if more meaningful operating conditions could be specified for either ssb or cw operating modes. As far as amateur operation is concerned, the limiting mode is cw, where the duty factor is about 0.5. The duty factor for single sideband transmissions with unprocessed speech runs about 0.05 for a 13-dB peak-to-average signal and could rise as high as 0.5 for high levels of speech compression or clipping. A duty factor of 0.5 (peak-to-average ratio of 2.0) for Intermittent Voice Service rating would therefore cover both table 2. Preliminary operating data for 8873 family of ceramic/metal, zero-bias power triodes.

Cathode: Oxide coated, unipotential	
Warm-up time	60 sec
Heater voltage	6.3 V
Heater current	3.2 A
Direct interelectrode capacitances, grounded-grid connection	
Input	19.5 pF
Output	7.00 pF
Feedback	0.04 pF
Maximum frequency ratings	450 MHz
Operating temperature, maximum, ceramic seals and anode core	250° C
Base	11-pin Special (JEDEC E11-81)
Radio-frequency linear amplifier, cathode driven, class AB_2	
Absolute maximum ratings, to 450 MHz	
Plate voltage	2200 Vdc
Plate current, continuous	250 mA
Plate dissipation	see note 1
Grid dissipation	5 W
Typical operation, Intermittent Voice Service[2] frequencies to 30 MHz	
Peak envelope or modulation crest conditions	
Peak voltage	2000 Vdc
Cathode voltage[3]	8.2 Vdc
Zero-signal plate current	22 mA
Single-tone IVS plate current	500 mA
Two-tone plate current (approximate)	312 mA
Average plate current	250 mA
Single-tone grid current	30 mA
Two-tone grid current	12 mA
Peak rf grid-cathode voltage	67 V
Peak driving power	26 W
Peak power input	1000 W
Single-tone IVS useful output power	587 W
Resonant load impedance	2140 ohms
Intermodulation distortion products:	
3rd Order	-35 dB
5th Order	-36 dB

notes. 1. 8873 is conduction cooled, and plate dissipation depends upon heat-sink cooling. 8874 plate dissipation is 400 watts, and tube has an axial-flow, forced-air cooled anode. 8875 plate dissipation is 300 watts, and tube has a transverse-flow, forced-air cooled anode.
2. Intermittent voice and cw ratings are based upon the maximum voltage and current ratings given for a signal having a peak-to-average power ratio of 2.0 or more. During short periods of adjustment (less than 30 seconds), the average plate current may be as high as the IVS value.
3. Cathode bias is obtained from a zener diode.

the cw and ssb speech-with-processing situations.

The derivation of a different rating from an existing rating may only take place after extensive life tests have been completed to make sure that tube life is not being shortened and that maximum temperature and dissipation limits are not

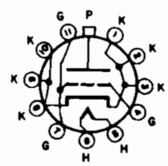

fig. 2. Tube base diagram for the 8873, 8874, 8875 family of zero-bias triodes. Multiple cathode leads keep cathode inductance to a minimum.

being exceeded. Any tube may be limited by grid or screen dissipation level and some may be limited by a plate voltage ceiling, or by available cathode emission. *Each tube type is an individual case, and to jump to conclusions or to interpret data from one type to another is risky and unfounded to say the least.* In all cases the total average current load on the oxide cathode will remain about the same for the new rating as for the average situation.

The *Intermittent Voice and CW (IVS)* rating may be defined as:

> That maximum voltage and current rating given for a signal having a maximum peak to average power ratio of 2.0 or more. During short periods of adjustment (less than 30 seconds), the average plate current may be as high as the IVS value.

In all cases, the IVS rating and "short period of adjustment" are limited by the maximum allowable temperature of the tube anode and seals.

using the ivs rating

The IVS rating is especially attractive to amateur operators as it outlines typical operating parameters for cw and ssb. How are the new ratings used? The following is an example of how an amateur operator can safely and properly tune up for an IVS operating condition with the aid of an inexpensive oscilloscope.

The oscilloscope is necessary for ssb adjustment at first since meter response to a voice waveform may vary from meter to meter and is, in any case, highly irregular and difficult to interpret.

1. The first step to achieve an IVS condition for either ssb or cw is to tune and load the linear amplifier with carrier (single tone) to an IVS rated value of dc plate current (as read on the plate meter), observing maximum "on" time. The amplifier is now ready for IVS CW operation. This could be called the "long-dash" tuning method. An electronic key is handy for this operation.

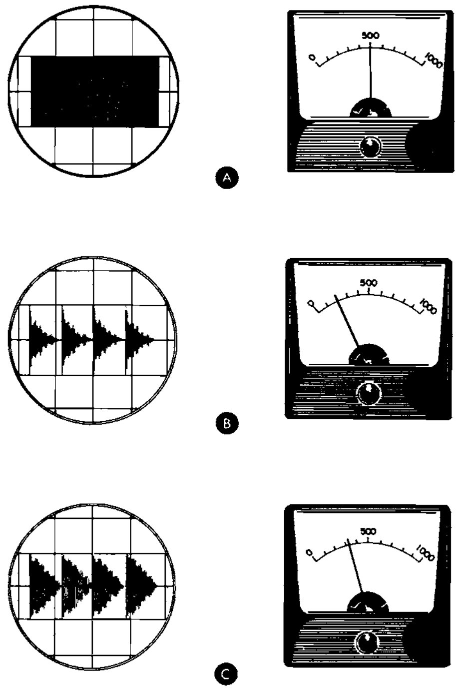

fig. 1. Typical meter readings for IVS operation.

A. With single tone (carrier), the linear amplifier is loaded to maximum IVS plate current (500 mA in this example). Oscilloscope shows carrier pattern. Pattern height is noted as 2 units. Observe short tuneup time.

B. Carrier is removed and voice modulation applied and gradually increased until voice peaks reach carrier height of 2 units as noted in pattern A. Plate meter kicks up to about 200 mA on voice peaks. No speech processing used.

C. Adding speech processing (clipping or compression). Note that plate meter now kicks up under voice peaks to about 325 mA, but that voice peaks on oscilloscope rise no higher than the single-tone limit of 2 units. However, area under the peaks is greatly enhanced, indicating greater average-to-peak ratio of voice signal. If oscilloscope peaks are greater than 2 units height, with or without voice processing, amplifier is being overdriven, with accompanying splatter and distortion.

2. For ssb observe the rf output pattern on the oscilloscope and note the amplitude for reference.

3. Remove the carrier. Insert audio and slowly increase audio gain so that the instantaneous rf peaks observed on the oscilloscope reach the same maxi-

table 1. Type 8122, linear rf power amplifier service (AB_1). Typical CCS operation at 30 MHz with two-tone modulation.

Plate voltage	2000 Vdc
Grid no. 2 voltage	400 Vdc
Grid no. 1 voltage	-35 Vdc
Zero-signal plate current	100 mA
Plate current	
Peak of envelope	335 mA
Average	250 mA
Grid no. 2 current	
Peak of envelope	10 mA
Average	7 mA
Average grid no. 1 current	0.05 mA
Effective rf load resistance	3050 ohms

mum level as obtained in *step 2* under carrier insertion. The amplifier is now working at the correct IVS level of peak input.

4. Observe the average current peaks on the plate meter for future reference.

In summary, the amplifier is tuned up to IVS condition with single-tone excitation to set the peak signal level. The single tone is removed, and audio is applied so the instantaneous signal peaks reach the same peak level as before, but the peak-to-average level of the intelligence may vary widely, depending upon voice characteristics, degree of speech processing, etc. This is summed up in **fig. 1**.

ivs ratings for the 8873 family of triodes

The new 8873 family of zero-bias triodes is the first to carry the new IVS rating. These ratings are based upon the original design concept of the tube, plus extended life tests where electrode temperatures, cathode emission and power output were carefully monitored. For example, the continuous plate current rating is 250 milliamperes. The cathode area of the 8873 is over 2 square centimeters; this corresponds quite closely to the 125 milliamperes per square centimeter rule-of-thumb stated earlier for an oxide-coated cathode. The life tests showed that a peak dc plate current rating of 500 milliamperes is reasonable at a duty cycle of 0.5 (peak-to-average power ratio of 2.0), corresponding to the IVS philosophy (**table 2**). The various input levels at a given plate voltage may now be established. At 2000 volts, for example, the average plate input is 2000 (volts) x 250 (milliamperes) = 500 watts. This corresponds to key-down service, such as RTTY. The two-tone rating (as in a short two-tone test) is 2000 (volts) x 312 (milliamperes) = 624 watts, average power. The IVS rating for ssb voice or cw is 2000 (volts) x 500 (milliamperes) = 1000 watts peak envelope power. In the case of voice and cw, the average current "load" on the cathode is the same.

Thus, today's power tube may be rated in two different and useful ways. Commonly, it bears the continuous duty (CCS) rating, and occasionally it bears the semi-obsolete ICAS rating. It is hoped that the new IVS rating will find favor in the future as it permits greater operating economy to be achieved in the use of all power-grid tubes.

what about . . .

The immediate question arises, "If this is so, what about the IVS ratings for the 3-500Z or the 8122 or the 4X150A, or whatever?" The present answer to this query is that *each tube type must be examined on its merits* and the outer limits established for any new rating, whether it be pulse, ICAS, or IVS. This is a continual process with most tube manufacturers, and more relevant date of this type will probably be forthcoming over the months.

Thanks to William McAulay, W6KM; Jack Quinn, W6MJG; and Robert Sutherland, W6UOV, for their help in the preparation of this article.

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ham radio

amateur service newsletter
W6SAI

modifying the Heath SB-200 amplifier for the new 8873 zero-bias triode

Simple modification of the SB-200 linear to provide increased power dissipation, better frequency stability, and lower drive. Two designs are featured — air cooled and conduction cooled

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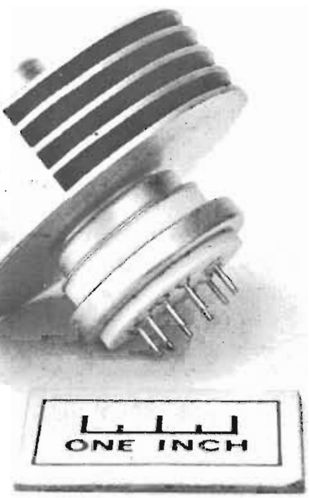
The high power capability, moderate cost and compact size of the new 8873 family of zero-bias, ceramic/metal power triodes make them well suited for new design, as well as for retrofit into popular amateur equipment that uses older tubes having restricted power capability and limited frequency range. The well-known Heath SB-200, a one-kilowatt PEP linear amplifier, is a likely candidate. This article covers the modification of this unit to use the new power tubes. The modification provides increased power dissipation,

The Eimac 8875 is a ceramic/metal zero-bias triode with a transverse cooler that provides 300 watts anode dissipation.



better high-frequency stability and lower drive requirements, and (in the case of the 8875) at a lower overall tube replacement cost then the original pair of tubes.

Based upon a study of the SB-200 circuit design, it was decided to try different modifications on two separate amplifiers. The first version uses the 8875, a high-mu power triode having 300 watts anode dissipation and capable of about 1200 watts peak input in Intermittent Voice Service (IVS).[1] The 8875 has five large, round, horizontal anode fins that may be adequately cooled with a small phono-motor fan, the type already in the Heath amplifier. This modification requires a minimum amount of disruption of the existing Heath circuitry.

The second amplifier has a more sophisticated and interesting modification: a conduction-cooled 8873 power triode (electrically equivalent to the 8875) is used with a finned heat sink for proper anode dissipation. The heat sink forms the vertical back wall of the rf enclosure.

This section discusses the first conversion, which provides full input level for the amplifier, with low intermodulation distortion and good tube life. The conduction-cooled version is described in the last part of the article.

the 8875 modification

The 8875 zero-bias power tube is shown in the photo and is capable of 1200 watts PEP input for ssb and 1000 watts when run in IVS service. The tube is about the size of a 4CX250B, has an 11-pin base and uses an inexpensive socket. Cathode and grid connections are brought out to multiple base pins, and, in addition, the grid is terminated in a low-inductance contact ring at the base of the tube which may be used for vhf operation. The anode is intended to be cooled by a horizontal air blast from a small fan. Dissipation is a function of cooling air, and a small phono-motor fan provides about 300 watts dissipation. For RTTY service, the power input level of the amplifier is dropped to about 600 watts. These levels are entirely compatible with the rating of the intermittent-duty power supply of the SB-200 amplifier.

The 8875 is mounted in a horizontal position in the approximate space previously occupied by the two glass tubes as shown in the chassis photo. The 11-pin tube socket is near the center of a small aluminum sub-chassis mounted to the rear wall of the amplifier enclosure. The existing cooling fan, mounted at the

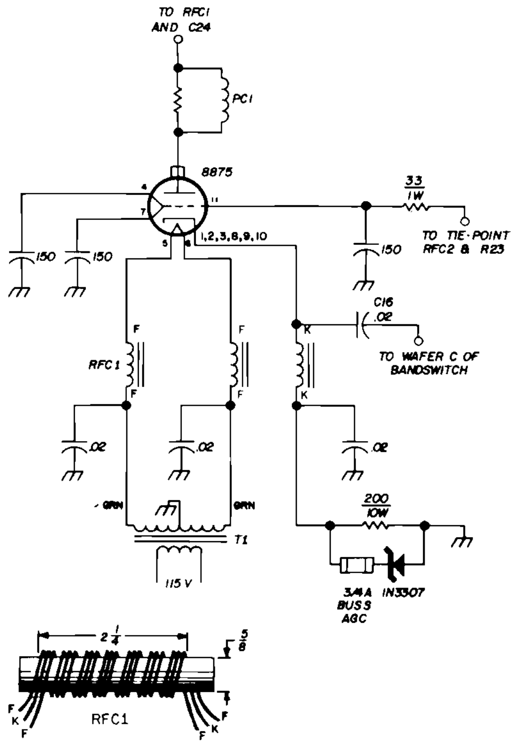


fig. 1. Revised schematic for the SB-200 shows the 8875 zero-bias triode. The 8875 requires 6.3 volts and the filament voltage in the SB-200 is on the high side because of reduced current drain. The original filament-choke windings are removed and three new windings put in their place. The filament windings are two 44" lengths of no. 20 enameled wire; the cathode winding is a 44" length of no. 26 insulated wire. Each winding has a 3" pigtail. Twenty trifilar turns are wound on the ferrite form. The ends are tied with twine and given a drop of epoxy to hold the windings in place. The socket for the 8875 is an E. F. Johnson 124-311-100. The 150-pF grid bypass capacitors are dipped mica units.

bottom of the enclosure, is moved to a new positon in respect to the anode of the 8875.

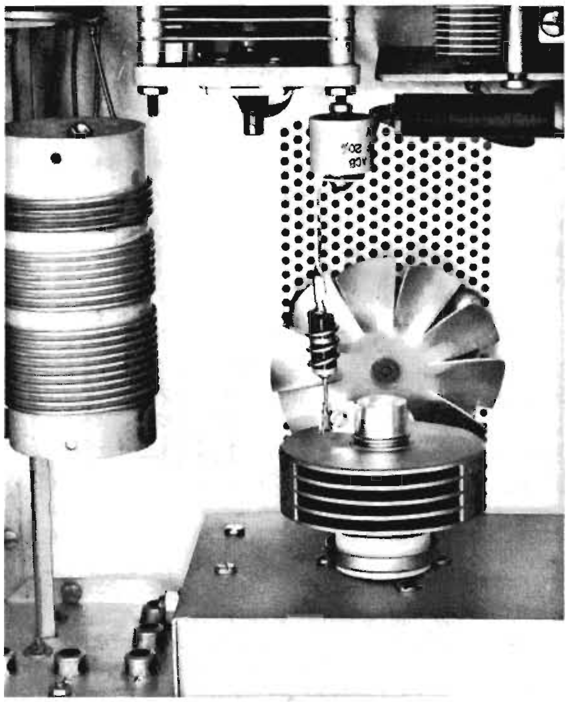
The revised circuit is shown in **fig. 1.** It uses most of the original components. A new filament choke or dropping resistor is required, as well as zener-diode bias for the 8875. All new components, with the exception of the zener fuse, are mounted within the sub-chassis, as shown in the rear-view photograph.

mechanical modifications

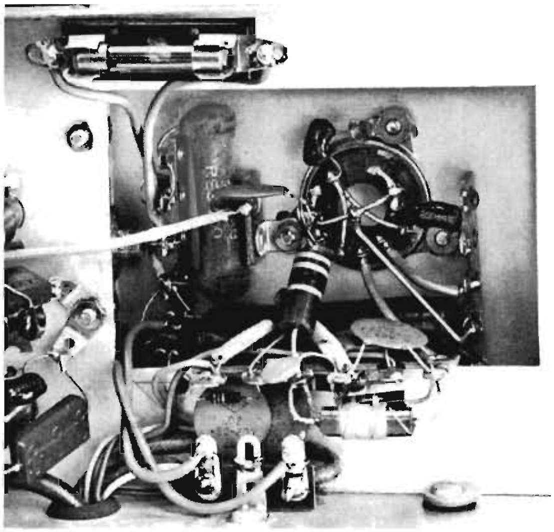
The first step is to remove the components around the existing 4-prong sockets, and then remove the sockets themselves.

Remove the filament choke from tie-point M. Unbolt tie-point AB, leaving the wiring connected. (See pictorial 11, page 39 of the Heath Instruction Manual. Original components are identified by

In the modified Heath SB-200 the 8875 is mounted horizontally in the space formerly occupied by the two glass tubes. Major plate circuit components remain unchanged.



their Heath nomenclature.) The next step is to cut out a rectangular hole on the rear wall of the enclosure as shown in the rear-view photo, with the dimensions



The rear panel of the SB-200 is cut out to allow access to the underside of the new sub-chassis. Untouched cathode coils are to left of the cutout. The zener fuse is at top of cutout with 200-ohm cathode resistor and zener diode mounted in sub-chassis. In this installation an extra cathode choke was used with the original SB-200 filament choke. The cathode choke has 15 μH inductance and 1000 mA current rating (J. W. Miller 4624).

shown in **fig. 2.** The new sub-chassis for the 8875 is placed over this hole. The sub-chassis is a 4 x 5 x 2-inch Bud AC-1404 chassis cut down to 1-¼" height and held in place with spade lugs and bolts. The tube socket is placed on the sub-chassis as shown in **fig. 2.**

The new sub-chassis interferes with various bolts holding the rf enclosure to the main chassis deck along the bottom rear edge, and it is necessary to provide clearance for these bolts. Proper clearance is provided by noting position of the bolts and drilling ¼-inch clearance holes in the sub-chassis at the points of interference.

Once the sub-chassis is in position, the tube is placed in the socket and the phono fan moved until the blades are positioned beneath the anode of the tube as shown in **fig. 3.**

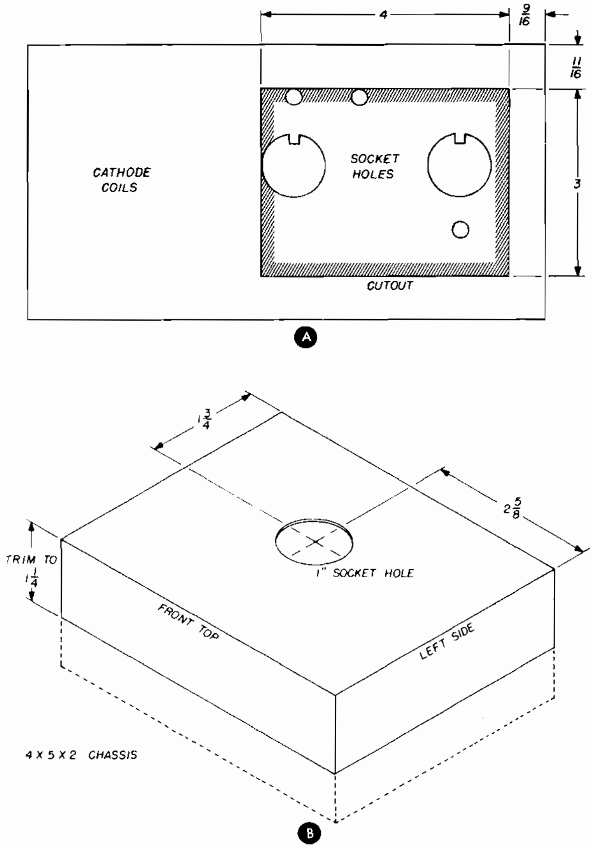


fig. 2. SB-200 chassis modifications to accomodate the 8875. The enclosure cutout for the sub-chassis is shown in A. The modified sub-chassis for the 8875 socket is shown in B.

electrical modifications

The portion of the original schematic of the SB-200 that is revised is shown in **fig. 1.** The 1N3307 8.2-volt zener diode is bolted firmly to the wall of the sub-chassis, using a thin coating of *Wakefield Thermal Compound* smeared on the zener stud to allow a good thermal bond. The new component layout in the sub-chassis is shown in the rear-view photograph.

Using the new tube, the existing filament voltage of the SB-200 is too high, and it is necessary to drop it slightly to avoid over-volting the tube filament; a 0.3-volt drop is necessary. This may be readily achieved by placing a 0.1-ohm wirewound resistor in series with one filament lead, or the filament choke may be rewound with the proper wire length and size to develop the required voltage drop. Since a cathode rf choke is required, the builder has the option of rewinding the present filament choke and adding a cathode winding as shown in **fig. 1** or using the existing choke and adding a cathode rf choke and filament dropping resistor. The latter was done for the first tests, and a new trifilar rf choke was substituted at a later date.

amplifier testing

When the modification is complete, all wiring should be checked and the resistance to ground from the anode clip should be checked. As in the original amplifier, before modification, the resistance should be about 180,000 ohms (the resistance of the filter bleeder resistor, $R_5 - R_{11}$). The amplifier should be connected to the exciter and to a dummy load. Before the amplifier is turned on, the exciter is tuned up, feeding through the unenergized antenna relay of the amplifier. The amplifier is now turned on, and the panel meter should read about +2400 volts in the HV position. Amplifier plate current is zero because of the cut-off bias voltage.

The amplifier controls are set as des-

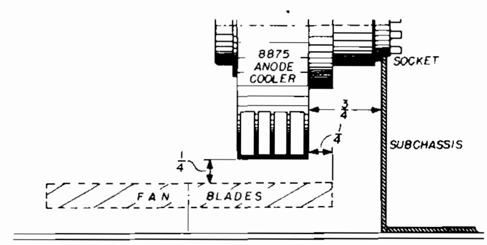
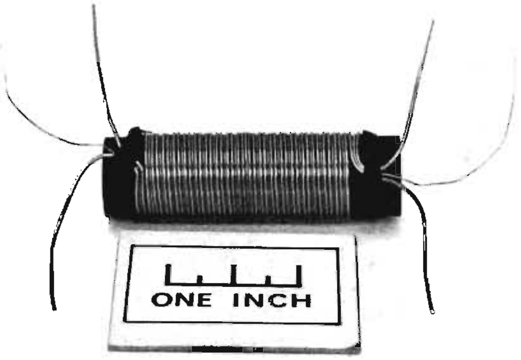


fig. 3. Phono-motor fan installation for the 8875. The motor must be moved so the shaft is in line with the center of the anode, and the tip of the fan blade clears the bottom of the tube by ¼ inch. The blade should also clear the edge of the anode by ¼ inch.

cribed in the *Operating Procedure* section of the Heath Manual (page 47), and the amplifier is again turned on. With no driver output, the meter of the amplifier should indicate an idling plate current of about 20 milliamperes. A ninety-second cathode warm up time should be observed before excitation is applied to the



New trifilar filament and cathode choke for the modified SB-200 uses ferrite rod from original unit. Winding details are shown in fig. 1.

8875. The drive level of the exciter is advanced until the plate current rises to about 200 milliamperes. Tuning and load controls are adjusted for maximum power output (minimum plate current) on the amplifier meter. Grid current should be about one division on the meter (25 milliamperes, or less). **Caution:** *The 8875 is easy to drive; watch out for excessive grid current.*

The amplifier may now be loaded for a maximum plate current of 500 milliamperes, using carrier injection from the exciter. Maximum grid current is 45 milliamperes. This corresponds to a drive level of 55 watts or less. Loading should be done quickly so as to not run excessive IVS plate current for more than 30 seconds or so.

When proper loading with carrier injection is achieved you will find that maximum power output occurs at this point along with the recommended values of plate and grid current. At maximum input the power output (measured with an accurate wattmeter) is between 520 watts (10 meters) and 630 watts (at the lower frequencies). Power gain is about 10 decibels. Under voice conditions, with no speech processing, voice peaks will run about 200 milliamperes on the meter.

Thanks to Merle Parten, K6DC, and Dick Rasor, WA6NXB, for their help and assistance in modifying the amplifier and making measurements on the completed version.

conduction-cooled linear

Modern power tubes such as the 8873-family have the capability of developing anode power dissipation densities (watts per square centimeter) comparable to the power densities in many jet and rocket engines. For this reason, effective cooling techniques are essential for long life and high tube reliability.

Conduction cooling is an efficient system of heat elimination, making use of the *heat source* (power tube or transistor), a heat transmission path (*thermal link*) and a *heat sink,* wherein the heat is removed. Many amateurs have seen transistors with tiny heat sinks on them; far fewer amateurs have observed heat-sink systems capable of dissipating several kilowatts of power. Such large systems exist, and the general design (suitably scaled down) may be adapted for use at amateur power levels. Although common in commercial and military gear, the heat-sink conduction-cooled system is just beginning to appear in amateur equipment (i. e., the Signal-One transceiver).

In the case of a power tube whose anode operates at a high voltage potential, the thermal link must have the dual properties of a thermal conductor and an electric insulator. One of the most practical materials for this task is *Beryllium Oxide* (BeO), an insulative ceramic (refractory) material which has the thermal conductive properties of aluminum.

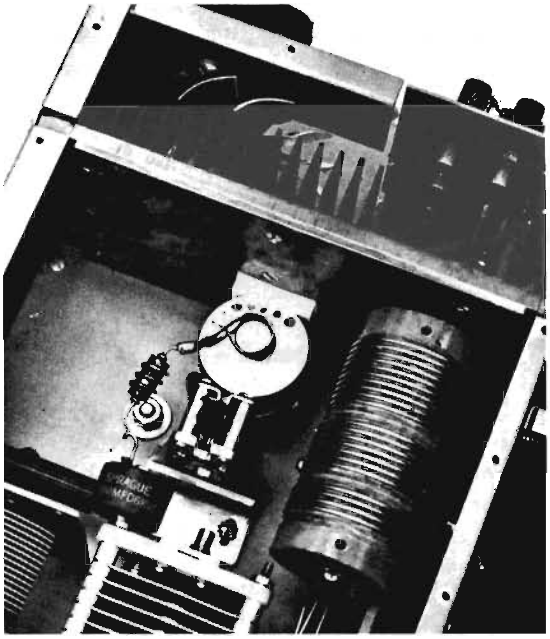
The 8873 zero-bias power triode makes use of a BeO thermal link and external heat sink. The link is detachable, providing mounting flexibility and reduced tube replacement cost. However, since two thermal interfaces occur with the detachable link (tube anode to link and link to heat sink), attention must be paid to ensure low thermal resistance at these two interfaces if optimum cooling performance is to be achieved.

The heat sink, receiving heat through the thermal link from the tube anode, emits energy in the form of radiant heat. The quantity of heat radiated depends upon the absolute temperature of the sink relative to the surrounding environment and the nature of its surface. A heat sink operating at an elevated temperature compared to its environment will transfer heat to the environment by radiation, convection and conduction, as is done in this case.

The added output capacitance of the tube supplied by the thermal link and heat sink is typically 6 to 10 pF, and this must be taken into account when vhf tank circuits are concerned.

The heat sink used with power tubes may be liquid or air cooled. In this case two or three hundred watts of anode dissipation are required so air cooling is feasible.

The 8873 conduction-cooled zero-bias triode.



Conduction-cooled 8873 zero-bias triode mounted in the Heath SB-200 linear. The 8873 is held in place with a toggle clamp that presses the anode of the tube against a beryllium-oxide thermal link and finned heat sink. The flat surface of the heat sink is covered with 1/8-inch copper sheet to distribute anode heat evenly. Under normal voice operation heat-sink dissipation provides sufficient cooling. For cw or long-winded voice operation a thermal switch turns on a small phono-motor fan to hold the temperature of the heat sink at a conservative level. When heat-sink temperature drops to normal value, the fan is automatically switched off.

the 8873

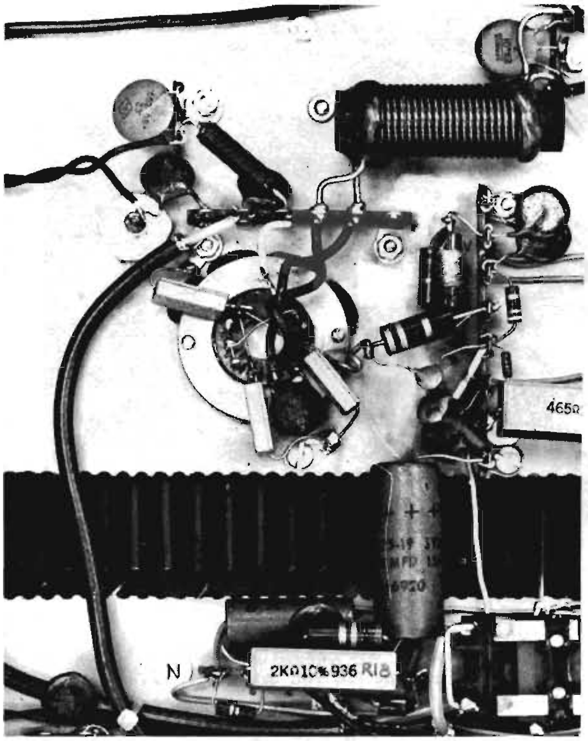
The 8873, like the 8875, is a ceramic/metal, zero-bias triode intended for hf and vhf service up to 450 MHz or so. No air cooling of the base is required if the socket is mounted on a chassis which has sufficiently low thermal resistance to drain the filament heat away from the stem of the tube.

The 8873 seemed a natural for retrofit in an existing Heath SB-200 amplifier. It was planned that the cabinet and power supply of the SB-200 could be used as a test bed for future experiments (such as a 50 MHz or 144 MHz amplifier) so a new, two-piece aluminum chassis was made. The power supply was rebuilt on one chassis and the amplifier section on another. Both units were then bolted together to resemble the original Heath chassis and shields. An amateur intending to modify his own SB-200 to this design probably would use the Heath metal work as-is.

heatsinking the SB-200 chassis

The 8873 anode is heat sunk to a finned radiator mounted at the rear of the amplifier enclosure. Generally speaking, the modification consists of removing the present tubes, sockets and auxiliary components and reworking the circuit electrically as described in the 8875 modification. Views of the heat sink installation are shown in the photographs. The heat sink measures 7-3/4 x 4-1/4 inches and is mounted to the chassis and side walls about 8-3/4 inches behind the front wall of the enclosure. The 8873 socket mounts in the center of the enclosure, with the center of the socket about 15/16-inch from the smooth surface of the heat sink.

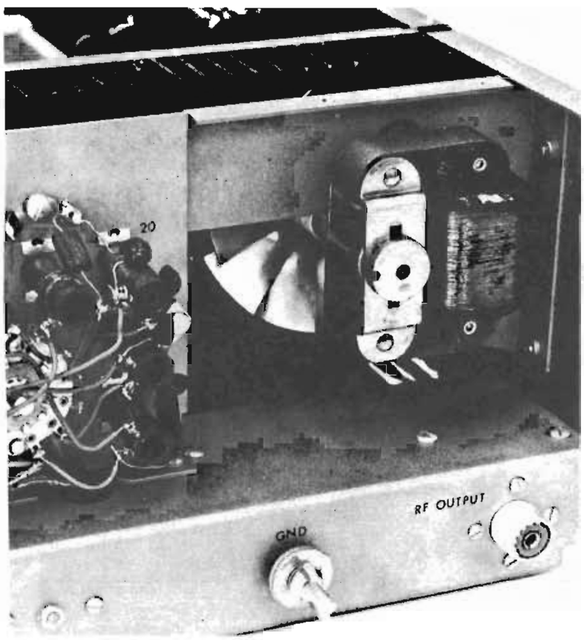
Anode heat flows from the 8873, through a BeO insulating block into the heat sink. Good bonding is essential between these three components in order to hold anode core and seal temperatures below the maximum permitted rating of 250°C. To hold the components firmly together, a DE-STA-CO toggle clamp is mounted in front of the tube. A small ½-inch ceramic insulator is substituted for the rubber nose of the clamp, which presses against the tube and heat sink. While the clamping action takes place, the tube and socket should be free to move. Accordingly, the socket is mounted in a clamp ring so that a slight amount of rotational and lateral movement can be accomplished. The rotational movement is required in order to align the flat surfaces of the tube, thermal link and



Bottom view of amplifier chassis shows sub-mounted tube socket. Chassis bolt holes are slotted so tube and socket may be moved slightly so the anode is properly seated against the heat sink. The rear of the amplifier chassis has been cut away below the heat sink so cooling air may pass through the fins. Filament choke is at upper right with zener diode mounted on chassis at upper left.

heat sink. The socket is then tightened in position after alignment and clamping takes place.

The heat sink provides about 160 watts of continuous anode dissipation when cooled by normal currents of 20°C air (room temperature). It is possible to raise the dissipation of the sink to about 200 continuous watts by passing cooling air across it from the small phono-motor fan which was a part of the original SB-200 assembly. The fan was included in this design, along with a thermal switch.



Rear view of conduction-cooled amplifier shows cooling fan and cathode tuned circuits. Chassis beneath the heat sink has been cut away for proper flow of cooling air.

When the temperature of the heat sink approaches a value that indicates high anode temperature, the fan is automatically switched on, increasing the capacity of the sink and protecting it from long-winded rag chewers and marathon talkers.

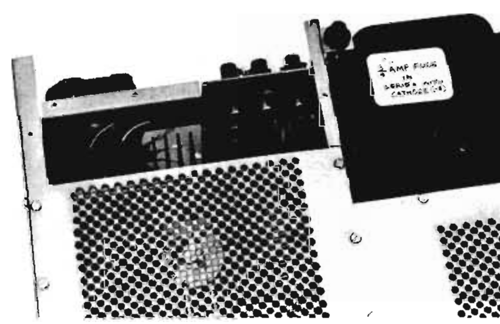
To determine the capacity of the heat sink system, temperature runs were made on the sink and tube anode, with various values of anode dissipation. Heat-sink and anode temperature were measured with temperature sensitive paint, and the thermal switch was moved about on the sink until it switched on when anode temperature reached about 180° C, well below the upper design limit of the tube.

To hold tube base temperature to a safe level, slots were cut in the chassis around the tube socket to allow cooling air from beneath the chassis to flow up and around the tube base (see under-chassis photo).

amplifier operation

Tuning and loading of the amplifier is normal, and follows the procedure outlined in the 8875 description. Under most operating conditions, the heat-sink temperature does not rise to the point at which the cooling fan is actuated, and amplifier operation is completely noiseless, a welcome "sound" these days!

While this unit is considered to be experimental, it points the way to the amplifier design of tomorrow: heat-sunk, noiseless, compact and highly efficient — quite a far cry from the old days of rack mounted gear, heavy, buzzing power supplies and black-crackle panels. How time flies!



Conduction-cooled amplifier with top shield in place. Shield has been trimmed at rear to allow air to flow over heat sink.

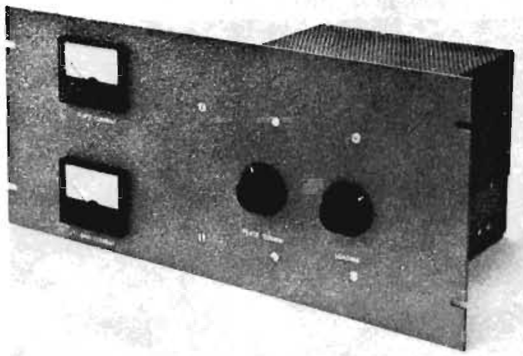
Again, thanks to K6DC and WA6NXB for their assistance in this experimental project.

reference

1. William I. Orr, W6SAI, "Intermittent Voice Operation of Power Tubes," *ham radio,* this issue.

ham radio

amateur service newsletter
W6SAI



two-kilowatt linear amplifier for six meters

This high performance six-meter linear features the new Eimac 8877 and provides excellent stability, good reliability and minimum harmonic output

Robert I. Sutherland, W6UOV, EIMAC Division of Varian, San Carlos, California 94070

The serious six-meter operator needs a high power amplifier that will function reliably over extended periods of time and have minimum harmonic radiation. Such amplifiers seem to be commonplace for the "dc bands" but are rather rare for 50 MHz and above. Many six-meter amplifier designs are cranky, hard to neutralize or otherwise unstable or tricky to adjust.

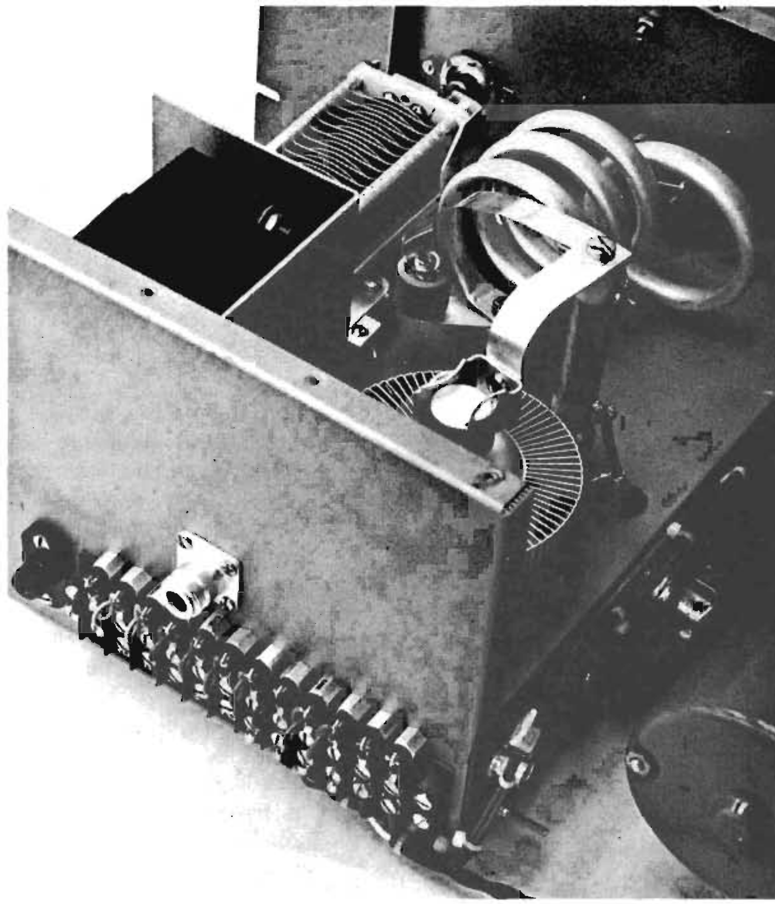
The amplifier described in this article has none of these undesirable attributes. It will run key-down on a 24-hour basis, if need be, and is stable and easy to adjust. I have used it over a period of months and it has proven to be a valuable

The new high-mu 8877/3CX1500A7 triode recently announced by EIMAC.

adjunct to the spread of six-meter equipment in my station.

This amplifier uses a grounded-grid circuit with a new high-mu triode just announced by Eimac: the 8877/3CX1500A7. This ceramic/metal triode is intended for linear service in the high-frequency and vhf range. The amplifier is intended for the maximum legal power input, 1000 watts dc, and can develop up to 2000 watts peak envelope power input during ssb operation. The amplifier requires a driver that can supply approximately 40 watts PEP at 50 MHz. Using a plate potential of 2600 volts and plate current of 750 milliamperes, power output will be about 1200 watts. This represents an amplifier efficiency of 61% and a power gain of 14.8 dB.

A schematic of the amplifier is shown in **fig. 1**. The control grid is operated at dc ground with a minimum of inductance between the tube and the chassis. The plate and grid currents are measured in the cathode return lead. A 12-volt 50-watt zener diode is placed in series with the cathode return lead to set the desired idling plate current. No special neutralization scheme is needed to attain completely stable operation.



Top view of the plate circuit of the linear amplifier showing the shorted-turn tuning scheme. The shorted-turn is hard-soldered to shaft coupler to allow front panel tuning. The "anti-inductance" strap can be seen connecting the top of the plate choke to the plate blocking capacitor. Note that the position of the plate blocking capacitor can be changed by loosening one screw and rotating the capacitor around the screw.

The plate circuit is a standard pi-network with tube output capacitance plus stray capacitance to the cabinet forming the input capacitance of the network (30 pF). The output loading capacitor is an air variable shunted by two fixed ceramic power. The input impedance of the tube is 54 ohms resistance in parallel with 26 pF capacitance. The match holds over the 1-MHz tuning range of the amplifier.

A 10,000-ohm 25-watt resistor in the cathode lead of the 8877/3CX1500A7 is

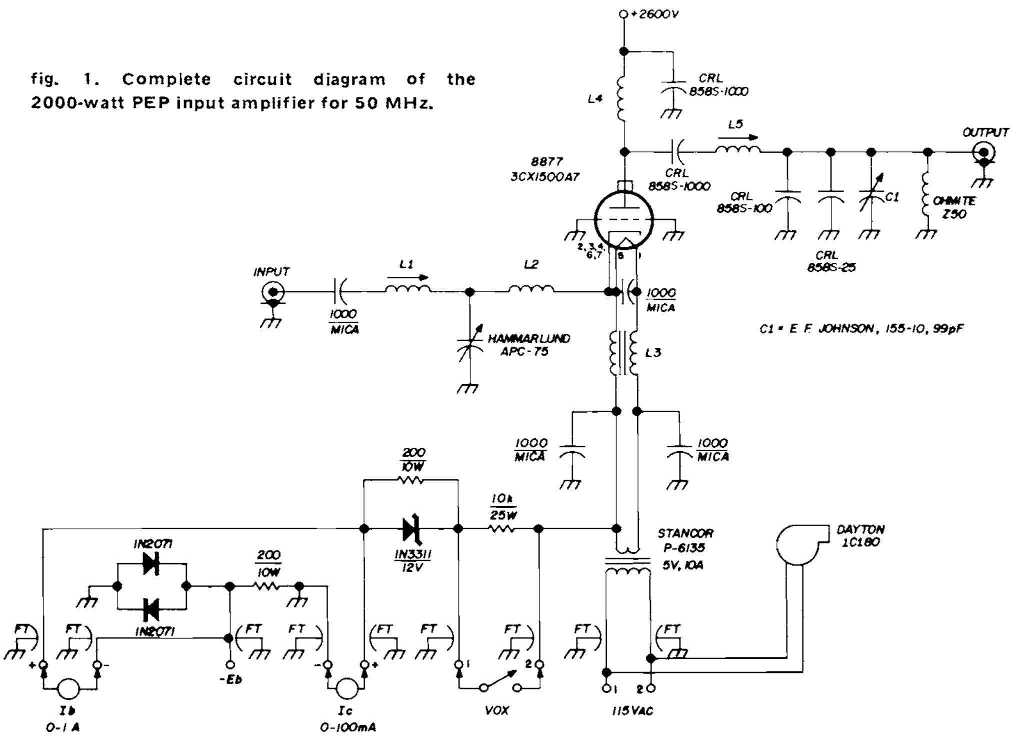


fig. 1. Complete circuit diagram of the 2000-watt PEP input amplifier for 50 MHz.

L1 6 turns no. 18 on a CTC 1538-4-3 form; coil length 7/8"

L2 6 turns no. 18, ½" diameter, 5/8" long, self-supporting

L3 Bifilar wound choke, ½" diameter core, 3" long, each coil 12 turns no. 10 Formvar; core is Indiana General CF-503

L4 54 turns no. 20 enameled on 1/2" diameter Teflon rod; winding length 1-13/16"

L5 3 turns 3/8" diameter copper tubing; inside diameter 1-7/8"; coil length 2-3/8"; shorted turn 2-¼" diameter 3/8" copper tubing ¼" from main coil

FT Erie 327 1000-pF feedthrough capacitors

capacitors. Amplifier tuning is accomplished by varying the inductance of the coil by adjusting the coupling between the coil and a shorted turn.

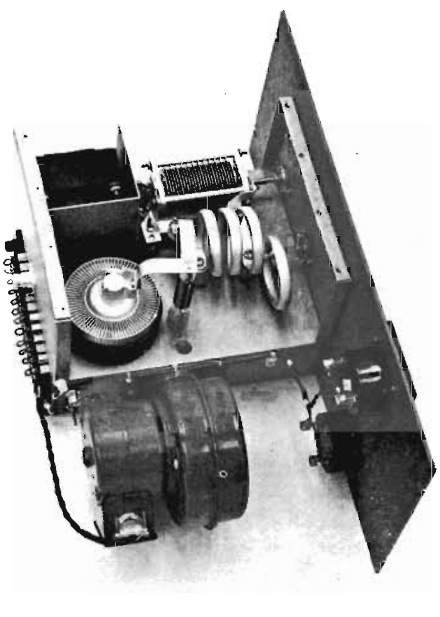
The cathode input circuit consists of a simple T-network. The network was calculated so that a 50-ohm cable from the driver would be matched to the input impedance of the 3CX1500A7 at full used to reduce standby current through the tube to a low value. When the exciter is turned on, a set of contacts on the vox relay (or other control relay) shorts out the 10,000-ohm resistor, causing the tube to operate at its normal idling plate current. The 200-ohm 10-watt resistor from the negative terminal of the plate supply to ground makes certain the negative terminal does not soar to the value of the plate voltage if the positive side of the power supply is accidentally shorted to ground.

The two 1N2071 diodes across the 200-ohm resistor limit any transient surges under the shorted condition which might cause insulation breakdown. Also, these diodes afford some transient protection of the two meters by providing a path around the meters. Additional protection could be obtained by putting two back-to-front parallel connected diodes across each meter. The 200-ohm resistor around the zener provides a load for the zener and prevents the cathode voltage from becoming quite high if the zener should burn open.

the plate circuit

Top views of the amplifier chassis are shown in the photographs. The closed ring near the front panel is the shorted turn used for tuning; it is made of 3/8-inch diameter tubing, hard soldered to a brass shaft coupler with copper-silver solder. Soft solder would not be advisable in this application because of the high circulating current in the shorted turn. The "anti-inductance" strap is used to set the tank circuit to the desired tuning range. This strap runs from the top of the plate rf choke to the plate blocking capacitor. The position of the blocking capacitor can be moved to allow the strap to be flexed and set to the proper position. Note that the current through the strap is going in the opposite direction from the current in the coil at any instant and therefore causes field cancellation.

Another view of the plate circuit. The air variable across the top edge of the chassis is the adjustable part of the loading capacitor. Two ceramic barrel capacitors are mounted in parallel with the air capacitor and can be seen at the end of the variable capacitor near the filament transformer shield.



To set the amplifier to the low-frequency end of the band, the shorted turn is completely decoupled and the position of the blocking capacitor and the anode strap adjusted to resonate the plate circuit to 50 MHz. As the shorted turn is coupled tighter, the total inductance in the plate tank circuit will be reduced, causing the resonant frequency to increase. When the shorted turn is fully coupled, the resonant frequency of the plate tank circuit will be about 51 MHz.

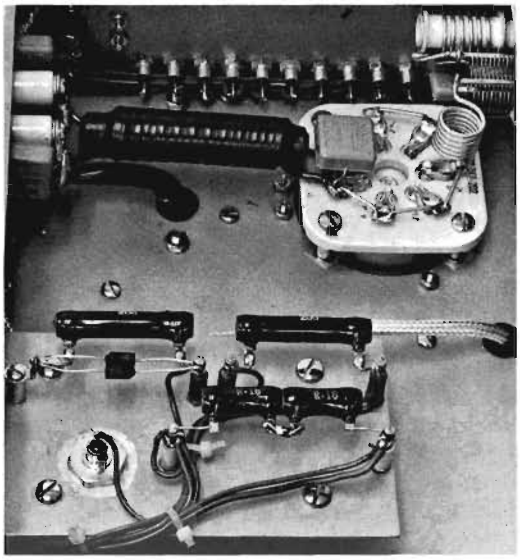
Amplifier loading is accomplished in the same manner as in a typical pi-network amplifier. The loading capacitor is the air variable along the top right edge of the chassis. The two ceramic fixed capacitors are at the left end of the air capacitor and at the end of the coaxial cable coming from the type-N coaxial receptacle mounted on the back panel.

The plate choke is made of 54 turns of no. 20 enameled wire closewound on a one-half inch diameter Teflon rod. The winding length of the coil is 1-13/16 inches. The choke is mounted on top of the ceramic capacitor which is used to by-pass the B-plus end of the choke.

Visible on the back of the front panel are the *Jackson* ball-drive assemblies. These handy devices provide a very smooth and slow "feel" to the tuning. The 5.0-volt 12-ampere filament transformer is visible inside its aluminum shield at the top left end of the chassis.

the input circuit

The input matching network is a standard T-design consisting of two series coils and one shunt capacitor. One coil and the shunt capacitor are variable. With these two adjustments it is possible to



View of the underside of the chassis showing the input circuit and the location of the zener diode and resistors. The T matching network is in the upper right hand side of the chassis. The heater-cathode choke is mounted between the socket and the ceramic stand-offs at the left side of the picture. Note that the socket is mounted below the chassis to allow passage of the cooling air. The straps grounding the grid to the chassis can also be seen under the threaded brass spacers used to sub-mount the socket.

cover a wide range of impedance transformations. The controls for the variable elements are brought out the left rear side of the chassis. Once the adjustments have been made, no tuning is required over the first megahertz of the band.

The input matching network can be seen in the top right corner of the under chassis photograph. The cathode-heater rf choke is near the tube socket. The choke is bifilar wound with twelve turns on each coil using no. 10 Formvar insulated wire. The core material is *Indiana General CF-503,* one-half inch in diameter.* The core permeability is a little high for this application, but the material was available and has not given any trouble. The Johnson 122-247-202 socket is mounted one-half inch below the chassis using threaded brass spacers. Four pieces of brass shim stock, or beryllium copper, are formed into an "L" shape to mount between the brass spacers and the chassis and make contact to the control grid ring.

*Available from Newark Electronics Corporation, 500 North Pulaski Road, Chicago, Illinois 60624. Order catalog number 59F1521

the tube

The 8877/3CX1500A7 is a new ceramic triode having good division between the plate current and the grid current. It has EIA base no. E7-2 which can be used with the standard septar sockets. The tube has a plate dissipation rating of 1500 watts, and has a mu of approximately 200. The cathode is indirectly heated, and the filament requirements are 5.0 volts at 10 amperes.

performance data

Many different operating conditions were tried with this amplifier. The conditions most suitable for amateur ssb operation at 2000 watts PEP input are:

Plate voltage	2600 Vdc
Plate current (single-tone)	750 mA
Plate current (idling)	40 mA
Grid voltages	-12 Vdc
Grid current (single-tone)	58 mA
Power input	1950 W
Power output	1200 W
Efficiency (apparent)	61 %
Drive power	40 W
Power gain	14.8 dB

The intermodulation distortion products at full peak envelope power input under the above operating conditions are:

3rd order	-44 dB
5th order	-37 dB
7th order	-64 dB
9th order	-68 dB

ham radio

rating tubes

for linear amplifier service

Peak envelope power and intermodulation distortion are important parameters when selecting tubes for linear amplifiers — here's what they mean and how they are measured

Robert I. Sutherland, W6UOV, William I. Orr, W6SAI, Eimac division of Varian

The power-handling capability of a given tube in single-sideband service depends upon the nature of the signal being transmitted and the tube's power dissipating capability. The method of establishing single-sideband service ratings should be such that relatively simple test equipment can be used to determine whether or not a tube is operating within its maximum ratings.

It is impractical to establish a rating based on voice-signal modulation because of the irregular waveforms and the varying ratios of peak-to-average signal power found in different voices. The most convenient rating method, and probably most practical, uses a single-tone audio signal to modulate the ssb transmitter. By using this test signal at its full modulation capability, the amplifier will operate under steady, maximum-signal conditions which are easily duplicated and observed.

When a single sine-wave tone modulates a single-sideband transmitter the rf output appears as a steady, unmodulated signal on an oscilloscope (see **fig. 1A**). This is because the output is a continuous signal having a frequency removed from that of the carrier by the modulating frequency, as shown in **fig. 1B.**

two-tone tests

Consequently, the operation of a linear amplifier under single-tone modulation is comparable to that of a cw transmitter under key-down conditions. As such, the performance of the power-amplifier stage at maximum signal (or peak) conditions can be determined from meter readings. However, this simple test lacks information on the linearity of the

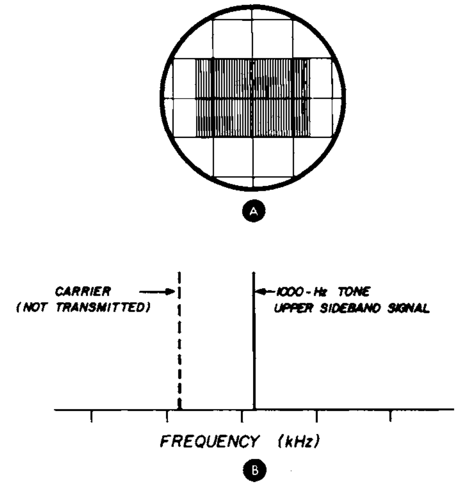
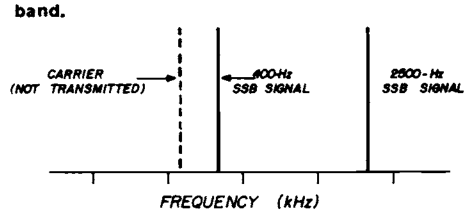


fig. 1. Rf output of ssb transmitter with single-tone modulation. Oscilloscope pattern is shown in A; spectrum is shown in B.

stage. To study linearity by observing amplifier output, some means must be provided to vary the output signal level from zero to maximum with a regular pattern that can be easily interpreted. A simple means is to use two equal-amplitude audio tones to modulate the ssb transmitter. This is termed a *two-tone* test. With this procedure the transmitter emits two steady signals separated by the frequency difference of the audio tones (**fig. 2**).

In some ssb generators, the two-tone signal is obtained by impressing a single tone at the audio input and injecting the carrier (by unbalancing the balanced modulator) to provide the second equal amplitude rf signal (**fig. 3**). The resultant beat between the two rf signals produces a scope pattern which has the appearance of a carrier 100 per cent amplitude-modulated by a series of half sine waves as shown in **fig. 4.**

fig. 2. Spectrum of ssb transmitter modulated by a two-tone test signal containing 400- and 2500-Hz tones and transmitting upper sideband.



When using the two-tone technique to measure the distortion of a linear rf amplifier it is sometimes more expedient to use two rf signal sources (separated in frequency by the desired number of cycles) and to combine them in a manner which will minimize the interaction between them. The two rf signals represent the two equivalent sideband frequencies generated by the two-audio-tone system and produce exactly the same scope pattern.

A linear amplifier is usually rated at peak envelope input or output power level. *Peak envelope power* (PEP) is the root-mean-square (rms) power generated at the peak of the modulation envelope. With two-tone or single-tone test signals the approximate relationships between single-and two-tone meter readings, peak envelope power and average power (class B or AB operation) can be determined from the formulas shown in **appendix 1.** Although the equations for average power output are different for the two tests, the PEP formulas are identical.

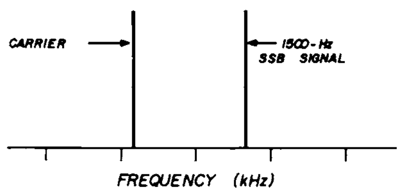


fig. 3. Spectrum of ssb transmitter modulated by 1500-Hz tone and injecting carrier to obtain second rf signal equal in amplitude to the tone.

multitone relationships

The approximate equations given in **appendix 1** are for single- and two-tone conditions, the most common test situations. However, in some multi-channel transmitter applications many more tones are used. The following method can be used to determine the peak-envelope-power to average-power ratio. (For the purposes of this explanation it is assumed that all the tones are equal.)

The following examples demonstrate two important relationships between

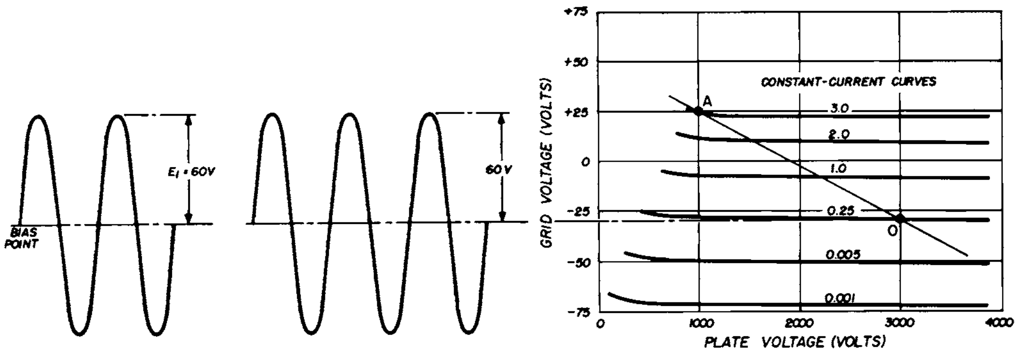


fig. 5. Single-tone condition.

single and multitone signals amplified by a linear system.

Assume the amplifier is set up for a single-tone driving signal and a Point "A" on the operating line is established (see **fig. 5**). A definite PEP output is developed under this condition. To drive this linear amplifier to the same PEP output with a multitone signal, the drive signal voltage for each tone must be 1/nth (n = number of tones) the amplitude of the single-tone signal.

By assuming a perfectly linear amplifier where the input waveshape is exactly reproduced in the output load, these grid waveshapes can be used to demonstrate the relationship of PEP to Average Power.

For the single-tone case, PEP = Average Power; for the two-tone case, PEP = twice Average Power. However, in the two-tone case the average power is one-half that of the single-tone case, so the resultant peak envelope power ratings are identical.*

The two test frequencies (f_1 and f_2) are equal in amplitude but slightly different in frequency. As a result, when they are exactly in phase the two crest voltages add directly to produce the crest of the two-tone envelope. When the two frequencies are exactly out of phase the

*This is best illustrated with two practical examples.

single-tone

$$\text{Average power} = \frac{E_1(\text{rms})^2}{R_L} = \frac{\left(\frac{60}{\sqrt{2}}\right)^2}{R_L} = \frac{1800}{R_L}\ \text{W}$$

$$\text{PEP} = \frac{E_1(\text{rms})^2}{R_L} = \frac{\left(\frac{60}{\sqrt{2}}\right)^2}{R_L} = \frac{1800}{R_L}\ \text{W}$$

Therefore, PEP = average power

two-tone:

$$\text{Average power} = P_{1avg} + P_{2avg} = \frac{E_1(\text{rms})^2}{R_L} + \frac{E_2(\text{rms})^2}{R_L}$$

$$= \frac{\left(\frac{30}{\sqrt{2}}\right)^2}{R_L} + \frac{\left(\frac{30}{\sqrt{2}}\right)^2}{R_L} = \frac{450}{R_L} + \frac{450}{R_L} = \frac{900}{R_L}\ \text{W}$$

$$\text{PEP} = \frac{(E_{1rms} + E_{2rms})^2}{R_L} = \frac{\left(\frac{30}{\sqrt{2}} + \frac{30}{\sqrt{2}}\right)^2}{R_L}$$

$$= \frac{\left(\frac{60}{\sqrt{2}}\right)^2}{R_L} = \frac{1800}{R_L}\ \text{W}$$

Therefore, PEP = $2 \times P_{avg}$

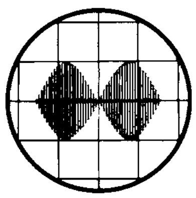


fig. 4. Scope pattern of ssb transmitter modulated by two-tone test signal.

cusp of the two-tone envelope results (see **fig. 6**).

Note that the voltage amplitude at the crest of the resultant two-tone envelope is equal to that of the single-tone envelope

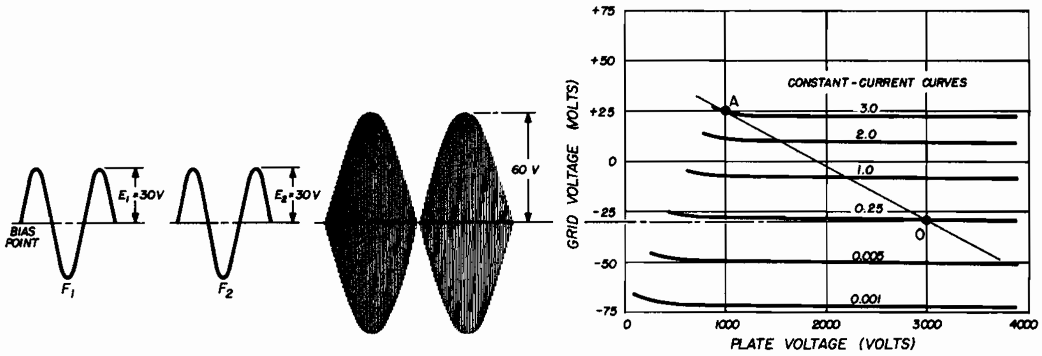


fig. 6. Two-tone condition.

and therefore the tube is driven to the same point on the operating line in each case. If the tube is driven to the same peak plate current and the same peak plate voltage swing by different excitation signals, then the peak envelope power output for both signals is the same.

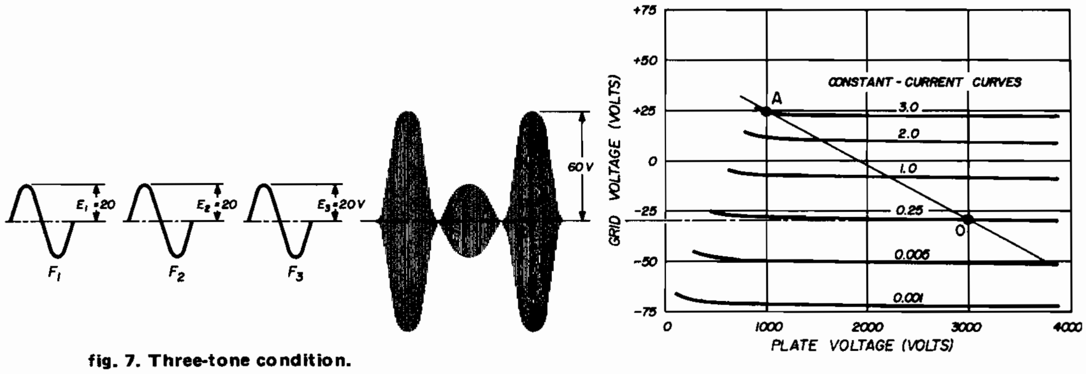


fig. 7. Three-tone condition.

The same holds true for a three-tone test signal. Note that the sum of the three individual tone-crest exciting voltages add in phase to drive the tube to the same peak current and peak plate voltage swing as that of the single-tone case (see **fig. 7**) so the PEP output is the same as the single- and two-tone examples.

These results (equal amplitude tones) may be summarized by the following expressions:

$$PEP = n\ P_{avg}$$

$$PEP = n^2\ P_t$$

Where P_{avg} is the average power of the composite signal, P_t is the average power in each tone, and n is the number of tones.

example

An fm repeater is to be designed to simultaneously rebroadcast one to eight channels. Each channel must have an average power output of 100 watts. How much peak envelope power must the linear amplifier deliver?

Each channel can be considered to be a single-tone signal. Therefore, the PEP of each channel is equal to the average power of each channel. The *maximum power output* requirement of the amplifier will be under the 8-tone condition. The average power output for the composite 8-tone signal will be 8 times the 100 watts-per channel power. Therefore, the linear amplifier must be capable of 800 watts of average power output.

The peak envelope power will be eight times the average power of the composite signal ($PEP = nP_{avg}$) or 6400 watts. A tube must be selected to deliver this peak-envelope and average power at an intermodulation distortion level compatible with the degree of interchannel crosstalk that can be tolerated.

measurement standards

To describe adequately the performance of a tube in single-sideband linear service, it is necessary to determine many parameters. The normal electrode voltages and currents must be specified as well as the two-tone currents, the operating point, peak envelope power and the magnitude of the intermodulation-distortion products. These parameters are defined as follows:

Peak envelope power is the root-mean-square power at the crest of the envelope. This term is usually shortened to PEP.

Idling plate current determined by the operating point is called the *zero-signal*

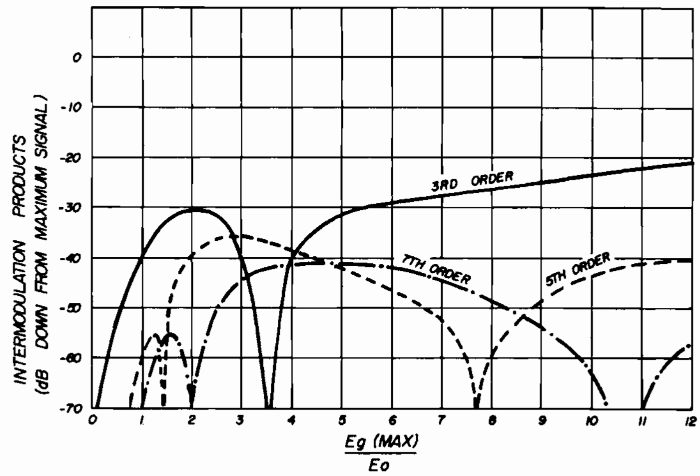


fig. 8. Graph showing intermodulation distortion products. As drive is increased, the various IMD products pass through maxima and minima. Misleading conclusions can be drawn if the equipment is tested near a cusp on the IMD curve where a particular IMD product drops to an extremely low level.

appendix 1

Approximate relationships between meter readings, peak envelope power and average power for class B or AB operation with one- and two-tone tests.

parameter	single-tone	two-tone
dc plate current	$I_b = \frac{i_{pm}}{\pi}$	$I_b = \frac{2i_{pm}}{\pi^2}$
plate input (watts)	$P_{in} = \frac{i_{pm}E_b}{\pi}$	$P_{in} = \frac{2i_{pm}e_p}{\pi^2}$
average output (watts)	$P_o = \frac{i_{pm}e_p}{4}$	$P_o = \frac{i_{pm}e_p}{8}$
PEP (watts)	$P_o = \frac{i_{pm}e_p}{4}$	$P_o = \frac{i_{pm}e_p}{4}$
plate efficiency	$N_p = \frac{\pi e_p}{4E_b}$	$N_p = \left(\frac{\pi}{4}\right)^2 \frac{e_p}{E_b}$

definition of symbols:

i_{pm} = peak of the plate current pulse (plate current pulse is not sinusoidal)

e_p =peak value of plate swing, assumed to be sinusoidal when tank-circuit has sufficiently high Q.

E_b = dc plate supply voltage

plate current and is designed I_{bo}.

The other two plate current values of significance are the *single-tone plate current* and the *two-tone plate current*. The ratio of single- to two-tone current is 1.57:1 in a true class B amplifier (180° plate conduction angle). For other classes of linear operation and for different zero-signal plate currents, this ratio varies from 1.1 to 1.57:1.

The standard method of specifying the magnitude of the distortion products is to specify the reduction in decibels of one product from one tone of a two-equal-tone signal.

For example, assume that a particular tube under a given set of operating conditions has third-order distortion products of −35 dB and fifth-order distortion products of −50 dB. This means that the third-order product has an amplitude of 35 dB below one of the two test tones and the fifth-order product has an amplitude 50 dB below one of the two test tones. (It is also correct to add the amplitudes of the two third-order products and compare them to the *sum* of the two tones. The decibel ratio is still the same as the example.)

It is *not* correct to compare one distortion product to the sum of the two tones; that is to say, the PEP value of the signal. The resulting distortion figure would be 6 dB better than the correct example (−41 dB rather than −35 dB and −56 dB rather than −50 dB).

Normally the tube under test is adjusted to the full drive condition, and all the pertinent parameters are measured. The drive signal is then reduced. At each test point, all the parameters are measured again. The resulting data can then be plotted as a function of drive voltage.

It should be noted that maximum intermodulation distortion does not necessarily occur at maximum drive level, and it can be shown mathematically that an intermodulation characteristic like **fig. 8** can be expected. In practice there is very good correlation between mathematical prediction and actual test results.

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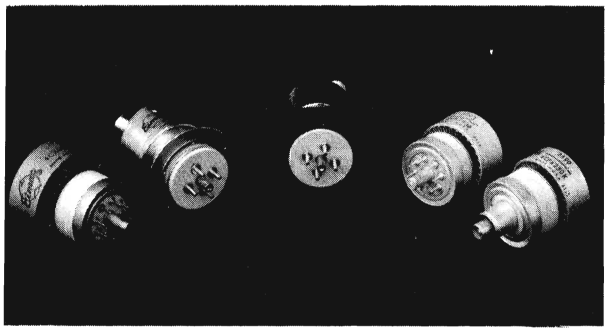

vhf/uhf effects in gridded tubes

Here is a rundown on the characteristics you should understand to get the most out of gridded tubes at uhf

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The technically-minded amateur has known for some time that a vacuum tube becomes progressively less effective as the frequency of operation is increased. Amplifiers require greater driving power, output power drops off, and the manufacturer may be obliged to derate the tube at the high-frequency end of the operable range. If the frequency is raised high enough, the gain of the amplifier will drop to an unusable level. Upon further increase in frequency, the gain will drop to unity or less. At the same time this is happening, the input impedance of the amplifier drops as does the maximum impedance realizable in the plate circuit.

Numerous factors contributing to the reduction of amplifier output at vhf and uhf can be listed and divided roughly into three groups:

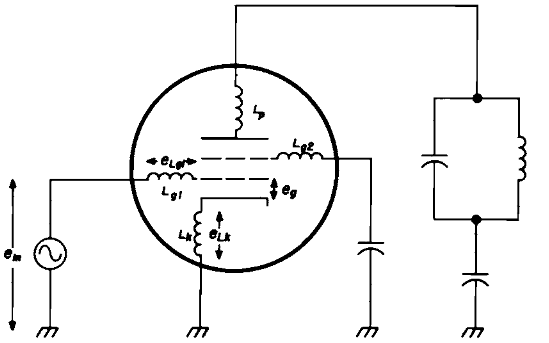
1. Circuit-reactance limitations
2. Circuit and tube loss limitations
3. Electron transit-time limitations.

circuit-reactance limitations

At the very-high and ultra-high frequencies there exists a situation which is quite different from that which exists at low frequencies. At low frequencies the electrical circuits and the tube are quite distinct. As frequency increases, this ceases to be true, and it is found that part of the resonant circuit exists within the tube (**fig. 1**). The electrode leads, while they are normally short and have large surface area in modern tubes, have a small but finite inductance. As the frequency of operation is raised, the reactance of the **lead inductance** will become quite appreciable, often reaching undesirable proportions in the vhf-uhf regions. The effect of lead inductance is to create a voltage drop such that the applied voltage across the external terminals of the tube will not entirely appear across the electrodes. Driving voltage is thus lost across the lead inductance.

In addition, while the interelectrode capacitances may be small, at ultra-high frequencies they approach a large fraction of the capacitance required to establish resonace in an external circuit. As such, the interelectrode capacitances represent a limitation in terms of actual operation as the external tank circuit may "disappear" within the tube. The combination of the electrode-lead-inductance and the interelectrode capacitance may cause an **internal resonance** in the uhf region.

fig. 1. The total input voltage does not act upon the grid-to-cathode electron stream because of grid and cathode lead inductance—the voltage across them subtracts from the input voltage.



A typical internal resonance experienced in a large power tube is the circuit consisting of the control-grid cage and mounting cone, and the screen-grid and cone. These two assemblies can form a quarter wavelength long tuned-line circuit shorted at the tube envelope by the capacitor consisting of the two contact rings and the dielectric material used in the envelope. The internal resonant frequency could be in the range of 1400 MHz for a five- to ten-kilowatt tube. This could lead to a 1400-MHz parasitic oscillation.

The smaller the tube, the higher the resonant frequency. All tubes will have internal resonances and the designer must move them out of the normal usable frequency range or load the circuit in such a way as not to degrade the performance within the rated frequency range. Even if resonances do not occur, the combination of reactances within the tube may constitute an undesired network that creates a mismatch between the driver and the tube electrodes.

magnitude of lead inductance

The most important reactance encountered in a vacuum tube is that associated with the lead inductance. An estimate of the magnitude of lead inductance may be made using the following equation:

$$L = 1.00508\ell \left(2.303 \log_{10} \frac{4\ell}{d} - 1 + \frac{d}{2\ell}\right)$$

where L = inductance (microhenries)
ℓ = wire length (inches)
d = wire diameter (inches)

This equation assumes no mutual inductance with some other nearby lead or wire.

As an example of how great lead inductance can be, consider the case of a lead that is 0.1-inch in diameter and one-inch long. This lead will have an inductance of 0.015 μH and an inductive reactance of 47 ohms at 500 MHz. An inductive reactance of this magnitude, for example, between the screen element and the screen by-pass capacitor outside the tube can cause stability problems. A voltage drop exists across the screen reactance caused by the current flowing through it which is the vhf current charging the output capacitance of the tube. This voltage is impressed upon the screen, which thereby is removed from ground potential, disturbing the grid-plate isolation normally afforded by this element.

This source of positive input resistance can be better understood by realizing that it results from a signal appearing across the cathode lead inductance, driving the cathode a small amount, as in a cathode-driven amplifier. (In a cathode-driven amplifier, the alternating current component of the plate current has to flow through the cathode lead inductance to reach the cathode. Since the driving signal is in series with the output load through the cathode-to-plate resistance of the tube, some of the power in the load is supplied by the driver.) Unwanted feed-through power is thus supplied by cathode lead inductance in a grid-driven stage.

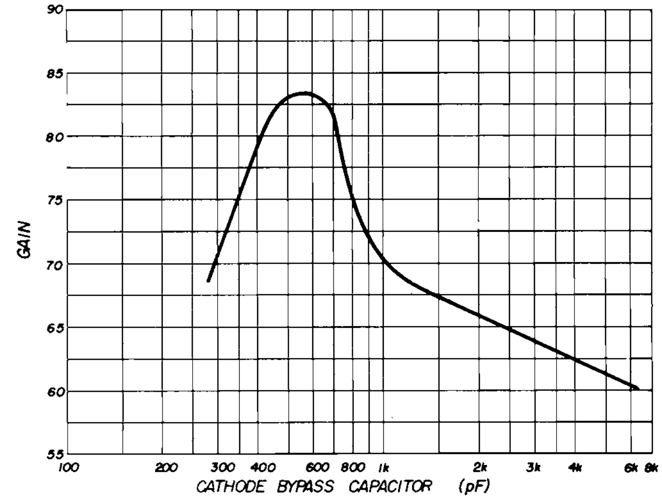


fig. 2. Gain of a 5CX1500A 100-MHz class-B amplifier with different values of cathode bypass capacitance. Gain peak occurs when bypass capacitor is series resonant with tube, socket and cathode lead inductance.

cathode lead inductance and input loading

Tube gain will also be adversely affected due to a reduction in input resistance because of the cathode-lead inductance. A small amount of cathode-lead inductance in conjunction with the grid-to-cathode capacitance of a vacuum tube will cause a resistive load to appear across the input of the tube. The magnitude of this added shunt input resistance is:

$$R \cong \frac{1}{\omega^2 L_k C_{gk} g_m}$$

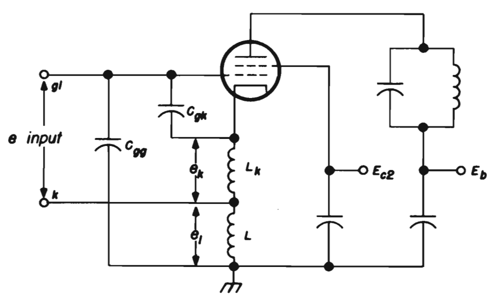
R = input resistance due to cathode-lead inductance
$\omega = 2\pi f$ where f is frequency (hertz)
L_k = cathode lead inductance (henries)
C_{gk} = Grid-to-cathode capacitance (farads)
g_m = transconductance (mhos)

The above equation shows that as the frequency is increased, the grid-loading due to cathode-lead inductance increases. For example, if the frequency of the amplifier is tripled from 144 to 432 MHz, the input resistance of a particular tube at 432 MHz would be one-ninth the resistance at 144 MHz. If it is desired to drive the 432 MHz tube to the same maximum grid voltage swing, then nine times the power is required just to cover the input loading due to cathode-lead inductance.

There will be other losses such as skin

effect, dielectric, radiation, transit time and the circuit reactance effects shown in **fig. 1.** The amount of grid loading due to cathode-lead inductance may be reduced somewhat by separating the input and output circuit paths back to the cathode. Some miniature tubes and low power transmitting tubes, such as the 4CX250B, 6146 and others, have multiple cathode leads to minimize the cathode-

fig. 3. Cathode lead neutralization.



lead inductance effects. Transistors also have this problem, only it is called emitter-lead inductance instead of cathode-lead inductance.

vhf/uhf cathode-lead inductance neutralization techniques

It is possible to neutralize the effects of the cathode-lead inductance by choosing the value of cathode by-pass capacitor so it will be **approximately** series resonant with the total lead inductance (tube, socket and circuit inductance). This technique is particularly effective in low noise stages of vhf/uhf receivers.

Fig. 2 illustrates stage gain as a function of cathode by-pass capacitance value in a high power vhf amplifier. This particular graph was obtained from an experimental 100-MHz, 5CX1500A amplifier running in class B. The 5CX1500A cathode lead, socket and circuit lead inductance was measured to be just over four nanohenries. A capacitance of 637 picofarads was calculated to be necessary for cathode lead inductance neutralization. The graph agrees fairly well with the calculated data.

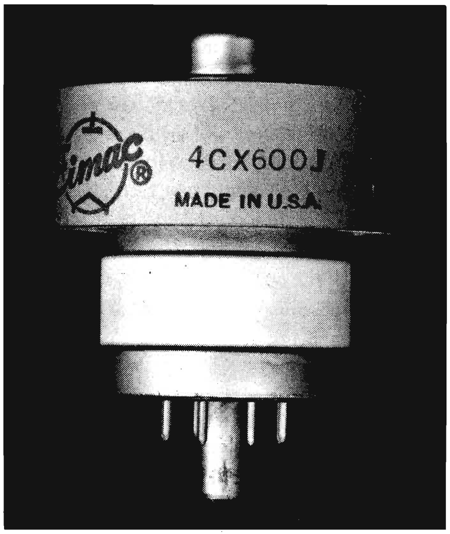
A neutralization technique described in the October 1939 issue of **Electronics** is of interest (**fig. 3**). I have had no personal experience with this technique, but it does appear to have merit. The voltage drop e_L across L, caused by the cathode current, is reversed in polarity with respect to e_k in the sense that C_{gg} and L are in series between grid 1 and the cathode. Thus, the current flowing back to the grid through C_{gg} is 180 degrees out of phase with the applied voltage, e_{input}. For a certain value of L the currents through C_{gk} and C_{gg} will be equal as well as opposite in phase; thus, the conductance between grid 1 and the cathode is zero. For neutralization of the cathode lead inductance the following ratio must be met:

$$\frac{L}{L_k} = \frac{C_{gk}}{C_{gg}}$$

For the 5CX1500A 100-MHz amplifier component values for neutralization of the cathode lead inductance could be:

$L_k = 4$ nanohenries
$C_{gk} = 35.8$ picofarads
$C_{gg} = 10$ picofarads
$L = 14$ nanohenries

The well known 4CX250B—rated to 500 MHz.



This tube has twice the cathode area of the 4CX250B and provides better than —40 dB intermodulation distortion.

L_k and C_{gk} are typical characteristics of the 5CX1500A and SK840 socket while C_{gg} and L are added components.

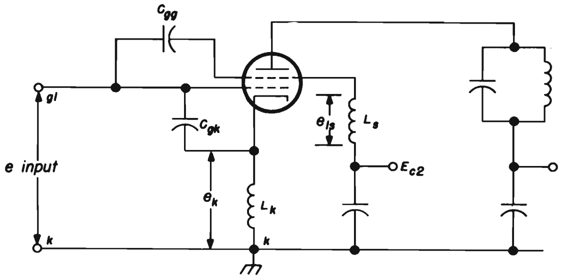
A very similar method of neutralizing that part of input resistance caused by the cathode lead inductance uses an inductance in the screen-grid lead and has been often used in 6146-type gear at 144 MHz (**fig. 4**). Basically, this circuit is the same as **fig. 3.** There are two differences: the point K (cathode pin) is now at ground potential and only screen current flows through the inductance L_s.

Other circuits may be used to minimize the effects of the cathode-lead inductance. The grounded-grid, cathode-driven (or probably more correctly called the "grid-separation" circuit), is often used (**fig. 5**). In this case the cathode is driven while the grid in a triode, or the grid and screen grid in a tetrode, are operated at some low rf potential. The grid structures then act like a shield between the input and output circuits. The main advantage, as far as cathode-lead inductance is concerned, is that this inductance is now just another inductor in series with, and therefore a part of, the input tuned circuit. There are other advantages that can be credited to the "grid separation" amplifier which will be discussed in the section on interelectrode capacitances.

intermodulation distortion and input loading

An appreciable amount of input loading can increase intermodulation distortion. Any tube plate-characteristic non-linearity will cause a variation in this input loading with signal level and thereby present a varying load to the driver, thus causing increased distortion in the drive voltage.

fig. 4. Neutralizing cathode lead inductance with an inductor in the screen lead on the tube side of the screen bypass capacitor.



screen lead inductance

The screen lead inductance between the screen element and the screen by-pass capacitor may help or hinder the operation of an amplifier. Below the **self-neutralizing frequency**[1,2] of the tube, the screen lead inductance is usually detrimental to the stability of the amplifier, as the rf current flowing through this inductance will cause an unwanted rf voltage to be developed. The point where the screen bypass is connected to the screen terminal may very well be at rf ground potential, but the potential of the screen itself may be varying above and below ground by the magnitude of voltage developed across the screen lead inductance, eL_s (**fig. 4**).

The magnitude of the developed voltage depends on the inductance of the screen lead and the frequency of operation. The higher the frequency, the greater the inductive reactance and the greater the rf current through this inductor. The current is greater because the capacitive reactance of the output capacitance of the tube will be smaller, and the capacitor will be charging to the same rf plate voltage swing each cycle.

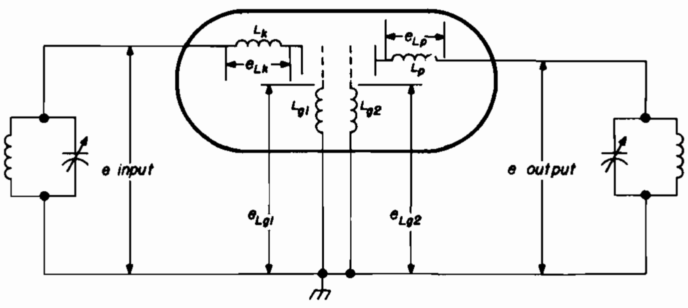
At operating frequencies below the self-neutralizing frequency of a tetrode, screen inductance may be added in order to neutralize the amplifier. That is, the self-neutralizing frequency is lowered to the operating frequency.[3] At operating frequencies above the self-neutralizing frequency of the tube, a series capacitor is sometimes added to move the self-neutralizing frequency up to the operating frequency (**fig. 6**). As was briefly touched upon previously, sometimes it is desirable to add a certain amount of screen lead inductance to neutralize the number of leads to use to provide the inductance he needs for his design.

In a cathode-driven (grounded-grid) amplifier, control-grid inductance is very important. Just as in the case of the screen-lead inductance in a grid-driven tetrode amplifier, the control-grid inductance in a cathode-driven amplifier may aid or hinder the designer. The control-grid inductance may cause instability, a loss in drive voltage due to the voltage divider effect (**fig. 1**) or it may be used to provide a method of neutralizing the amplifier.[2]

plate lead inductance

The plate in modern tubes used for vhf/uhf operation is usually designed with a

fig. 5. Cathode-driven or grid-separation amplifier.



cathode-lead inductance portion of the input resistance.

control-grid lead inductance

The **control-grid lead inductance** in a grid-driven amplifier is usually not of much concern as the relative magnitude of the tube lead inductance as compared to the external inductance added to it to attain resonance is very small. The control-grid lead inductance is wholly a part of the input resonant circuit with no current being induced from the output circuit, and it becomes part of the input tuned circuit. An exception to this is in the case of a vhf/uhf **distributed amplifier.** In this application an active filter is designed using all of the tube lead inductances and interelectrode capacitances. The control-grid inductance is important in this case, and modern distributed amplifier tubes are made with four grid leads available to the equipment designer. He then has the choice of the

massive anode structure in order to dissipate the heat that is generated in this element of the tube. For this reason, plate lead inductance is usually low enough so it is not of any great concern. If normal good engineering practice is followed in designing vhf/uhf circuitry, the plate lead inductance becomes inconsequential.

interelectrode capacitance

In addition to lead inductance, interelectrode capacitance plays an important role in the operation of tubes in the vhf/uhf regions. Interelectrode capacitances due to active parts of the tube structure are incapable of reduction beyond a certain point. However, in many tubes the interelectrode capacitance results largely from capacitance between leads in areas of the tube where electrons do not flow. It is the job of the tube designer to reduce this unnecessary capacitance to a minimum.

input capacitance

The input capacitance of a grid-driven tube is the sum of the grid-to-cathode and grid-to-screen capacitances. The larger the input capacitance, the greater the drive power must be. This can be explained by the very large increase in input charging current necessary to charge the input capacitance. As the frequency increases the reactance of the input circuit becomes smaller, and for the same peak grid voltage, the charging current must increase. As power is proportional to the square of the current, the power lost in the input circuit necessarily increases as the charging current rises. The driver must supply this extra power.

To reduce this loss, the circuit designer must keep the input capacitance down and may limit the magnitude of the peak grid voltage. Peak grid voltage can also be minimized by operating with less bias. Quite often in certain amplifiers the class-B mode is more desirable than class-C operation. Reducing the peak grid voltage and the charging current will reduce the amount of power that must be dissipated by the control grid. Radio frequency power dissipated by the control grid unfortunately cannot be measured by the dc meters on the front panel of the amplifier, so the operator has no means of knowing if charging currents cause excessive grid temperature. High input capacitance also limits the bandwidth of the input circuit. For those applications requiring large instantaneous bandwidth, great care must be taken in the design of the equipment.

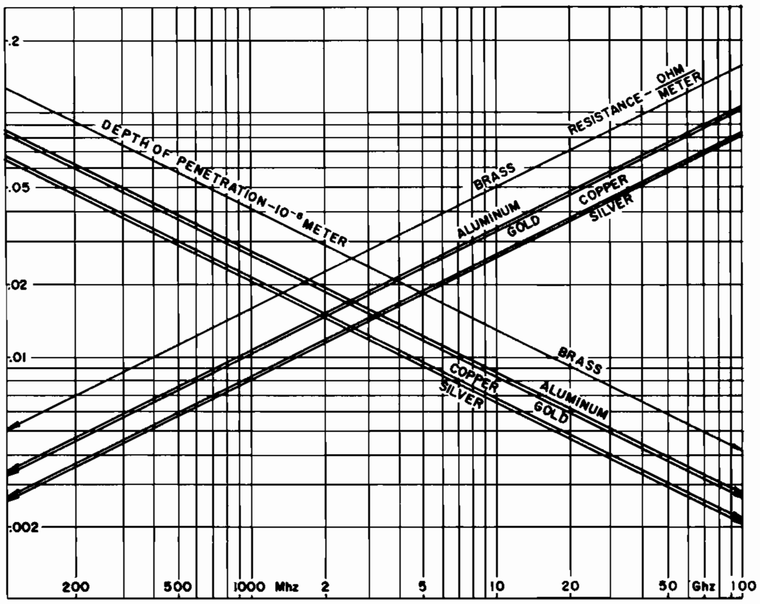
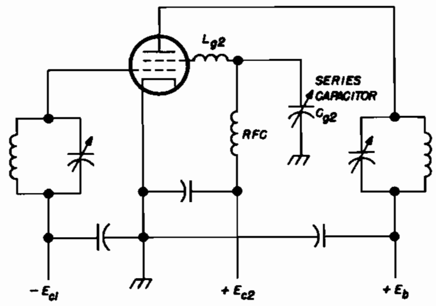


fig. 7. Chart for computing resistivity and depth of penetration for metallic conductors between 100 MHz and 100 GHz.

fig. 6. Screen neutralization at operating frequencies above self-neutralizing frequency of tube.



The grid-separation or cathode-driven amplifier offers quite an advantage as far as input capacitance is concerned. The input capacitance consists only of the cathode-to-grid capacitance. For the same tube in the cathode-driven configuration, the input capacitance will be roughly half that value in the grid-driven circuit. This is quite an advantage for applications requiring wide bandwidths.

output capacitance

The output capacitance of a power tube is an important factor in determining what plate-load resistance can be used. This in turn determines the stage gain and power output that is available. The equivalent shunt resistance (plate-load) of a parallel resonant circuit can be written as $R_L = Q/2\pi fC$ or $R_L = 2\pi fLQ$ where R_L is the plate-load resistance, Q is the loaded Q of the resonant circuit and f is the resonant frequency. For

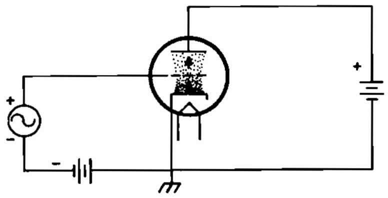


fig. 8. The effect of transit time on grid losses. At the instant shown, the grid potential is increasing in a positive direction and there is a consequent disproportionate number of electrons between grid and cathode so there is electron flow from the grid even though it may be negatively biased.

operation at a given frequency, to **increase** the shunt resistance it is necessary to **decrease** the shunt capacitance. This can be done to a point by reducing the circuit capacitance and increasing the tank coil inductance, maintaining the same frequency of resonance. Eventually, this process is limited by the fact that the capacitance external to the tube has been reduced to zero; the shunt resistance is finally determined by the tube interelectrode capacitance. The larger the interelectrode capacitance, the smaller the shunt resistance that can be realized. Accordingly, power output tends to drop off as the load resistance, or as the square of frequency, as frequency increases.

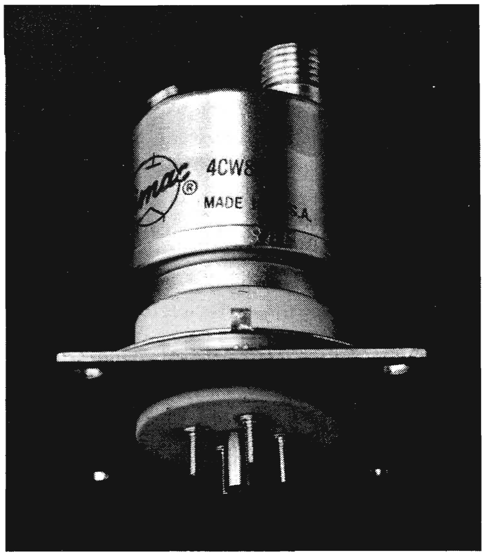
There are other problems brought about by the effect of output capacitance. The output capacitance must be charged and discharged during each cycle of the radio frequency. Again, as the frequency increases, the reactance of the output capacitance decreases. Therefore, with the same value of peak rf plate voltage, the current flowing through the output capacitance must increase as the frequency increases. The output capacitance of a tetrode is made up of the screen grid structure, the plate structure and the tube envelope.

The charging currents must flow over the surface of these components of the tube, all of which have varying degrees of rf resistance. It is possible for the charging currents to exceed the dissipation rating on the screen grid even though the dc meters indicate all is within ratings. It is advisable in vhf/uhf circuits, therefore, to try to achieve the lowest usable value of peak rf plate voltage by using the lowest plate load resistance and drawing the highest plate current consistent with desired output power and efficiency. Running the tube in this manner will lighten the rf stress on the tube seals and reduce the rf current in the screen grid structure thus providing for a potentially longer tube life.

feedback capacitance

The feedback capacitance in a grid-driven amplifier is the capacitance from anode to the control grid. The higher the frequency of operation, the greater the chance for instability due to rf feedback from the output circuit through feedback capacitance to the input circuit. In the vhf/uhf region this capacitance and other tube capacitances and inductance must be adjusted to provide for neutralization by added circuit components.[4]

The grid-separation amplifier helps minimize the effects of the feedback capacitance. The feedback capacitance in this configuration is very much less than the grid-driven case since it is the capacitance from anode to cathode with the grid, or grids, shielding the output from the input. In some applications no neutralization will be required. Other applications may require quite extensive neutralization.[3,5]



Distributed-amplifier tube discussed in text. By choosing the number of control-grid leads, the designer can control lead inductance. The tube in the right-hand photo has a screen by-pass capacitor installed. Center pin is one of the heater pins; the square tab on the capacitor is the other heater connection and the cathode. The threaded pins are control-grid leads.

Tube designed for large phased-array radar system. Two of these tubes were used in the 1000-watt 432-MHz amplifier used by WA6LET with a 150-foot dish for moonbounce contacts in 1965.

circuit and tube-loss limitations

The power losses associated with a tube and circuit all tend to increase with frequency. In the vhf/uhf region, almost all radio-frequency current flows in the surface layers of a conductor because of skin effect (**fig. 7**). The resistance and rf losses in a conductor increase with the square root of frequency because the layer in which the current flows decreases in thickness as the frequency increases. Insulating supports in the tube and external circuit have losses associated with the molecular movements produced by the electric fields. These dielectric losses will usually vary directly with frequency. Also, there will be additional losses due to the radiation of energy from the wires and leads carrying rf current. The power radiated from a short length of wire carrying current increases as the square of the frequency.

All these factors contribute to a general reduction in tube and circuit efficiency as operating frequency is increased. In the manufacture of power tubes the resistance losses are reduced by increasing conductor surface area and by proper choice of lead materials. Dielectric losses are reduced by selection of envelope and insulating materials. Support insulators are positioned, when possible, out of high voltage fields. Radiation losses are reduced by constructing vhf and uhf tubes and circuits so as to be totally shielded. At times it is prudent to use concentric line construction techniques so that tube and circuit fields are entirely confined.

transit-time limitations

Electron transit-time effects can contribute to reduced tube output in many ways. Transit time is the finite time an electron takes in going from the cathode to the grid. If the transit time (a function of grid-to-cathode distance and grid-to-cathode voltage) is an appreciable fraction of one ultra-high-frequency cycle, then an electron in transit in the grid-to-cathode region might be still heading for what was once a more positive location than the cathode surface it just left, but now finds the grid may be less positive or perhaps even negative (**fig. 8**). As a result of transit-time effect in the cathode-grid region, there will be a dispersal of "out of step" electrons. Because of this dispersion of the electron stream, the plate-current pulses are not as sharp as the current pulses liberated from the cathode.

In addition, energy is required to accelerate the electron towards the anode, and this energy is supplied by the driver. As the operating frequency is raised, more energy is required because the grid-input resistance due to transit time varies inversely as the square of the frequency. That is, if the frequency is doubled, the input resistance due to transit time effects will be one-fourth that at the lower frequency. The extra power required to overcome transit-time loss due to grid-input resistance is supplied by the driver and appears as lost drive-power—required but put to no practical use other than to heat the tube seals and waste precious exciter output.

Paradoxically, transit-time loss and cathode-lead inductance loss (both of which cause input loading) are not all evil because they often tend to stabilize a "wild" stage. Cure of the trouble may lead to higher stage gain with the possibility of oscillation and instability!

Circuit loading can be used to achieve stability and this is often done at the lower frequencies by adding resistors across i-f transformers, for example. The circuit designer, therefore, finds that a tradeoff exists between loading and stage gain that will work to his advantage or disadvantage, depending upon his ability to analyze the circuit. There is often more than one successful path to a proper design and the good circuit engineer has several alternative paths (in his head, if not on paper) to choose from. A candid realization of uhf/vhf effects will help the circuit designer obtain maximum power, efficiency and reliability from his equipment.

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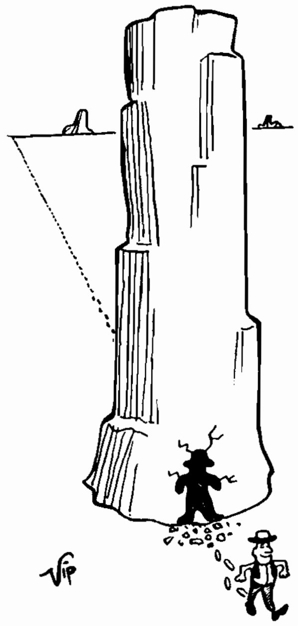
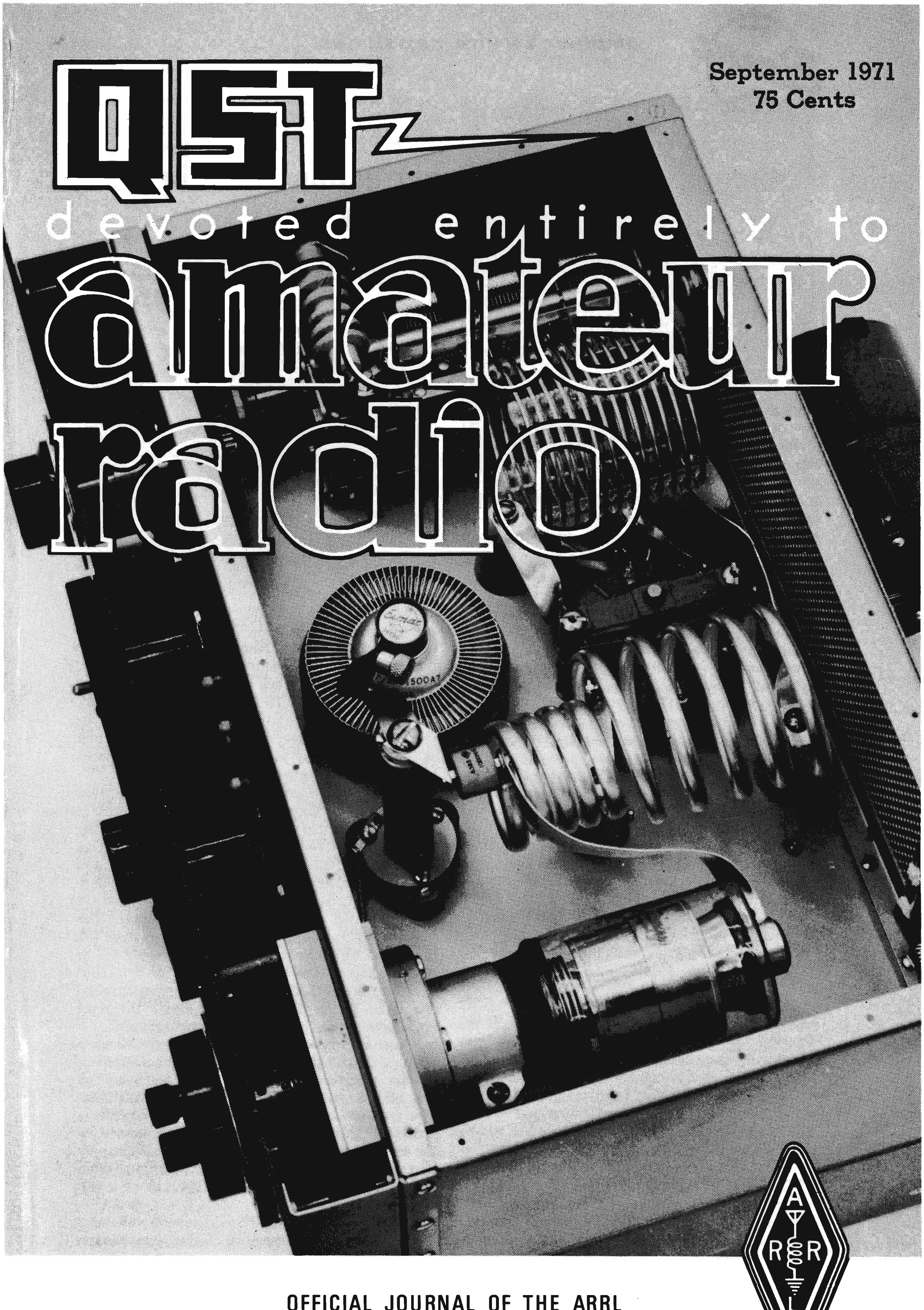


High-power linearity with Eimac power-grid tubes.



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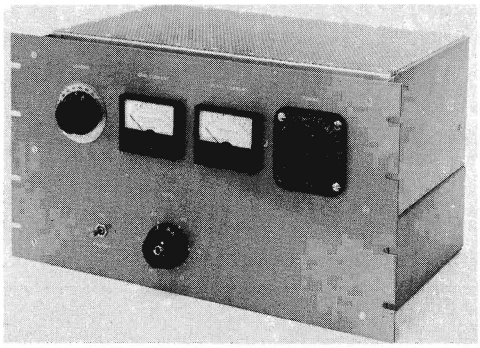
W6SAI

Custom Design and Construction Techniques for Linear Amplifiers Using the 8877

BY MERLE B. PARTEN,* K6DC



TEST RESULTS at 50, 144, and 220 MHz using the new 8877 tube were so gratifying that I decided to try it in a 3.5- to 28-MHz grounded-grid amplifier. Since so many articles about construction have been written in the past, I decided to compile into one text enough data and design variations to allow a home constructor to build an 8877 amplifier without having to bolt-for-bolt copy this design.

Design Advantages

The 8877 is a big brother to the new 8873 series of ceramic/metal power tubes. It is a zero-biased high-mu triode having an oxide-coated cathode. The plate dissipation is 1500 watts. Heater-to-cathode capacitance is low eliminating the need for filament chokes when operated below 30 MHz. An inexpensive 7-pin socket may be used reducing the overall cost. The grid connection is near the chassis level and permits low-inductance grounding. Average IMD products for the 8877 in linear service run 38 dB below one tone of a two-tone test signal for 3rd order products, and 44.5 dB for 5th order products.

Design Considerations

In building anything, whether it is a new home or an amplifier, there is always something you would like to change after the job is completed. Think the project through before picking up the hammer!

Professional designers are not immune to mistakes. Some manufacturers have chosen a symmetrical knob placement and size pleasing to the eye. The band switch and multimeter-switch may be located side-by-side, using identical knobs. Even the knowledgeable operator may close the key, look at the meter, reach for the meter switch (he thinks), and grab the band switch instead. (It only happens once per band!) Different knob sizes or placement could prevent a catastrophic error such as this. Don't worry about beauty in a front-panel layout where performance might suffer.

One simple design error can cause several more severe errors to show up as building progresses. A mock-up assembly was made to determine the space required for the coils and capacitors. This rough layout allowed positioning the tube and loading capacitor. A point overlooked was where the loading capacitor shaft terminated on the front panel. It was too close to the left edge. The mistake was solved by using a set of gears from the junk box. An alternate layout would have been to mount the coil and switch assembly on the front shield surface with the switch shaft passing through the front panel. This would eliminate the right angle drive.

Any cost-conscious builder will first review his on-hand supplies. A suitable plate-tuning capacitor, mounted vertically, might free enough space on top of the chassis to mount a plate transformer. The rectifier diodes, capacitors, and bleeder resistors could be located below the chassis, cooled by the airflow from the blower. Suitable power-supply parts may be purchased as replacement items from some amateur equipment manufacturers at a reasonable price.

Convenient location of controls should be a factor in the layout design. A right-handed person

* EIMAC Division of Varian, San Carlos, CA.

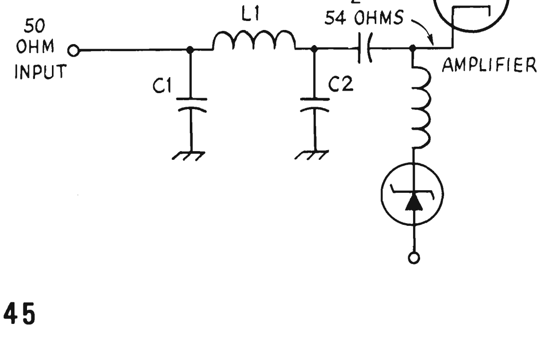


Fig. 1 — Simplified diagram of the input circuit. Component values are given in Table 1.

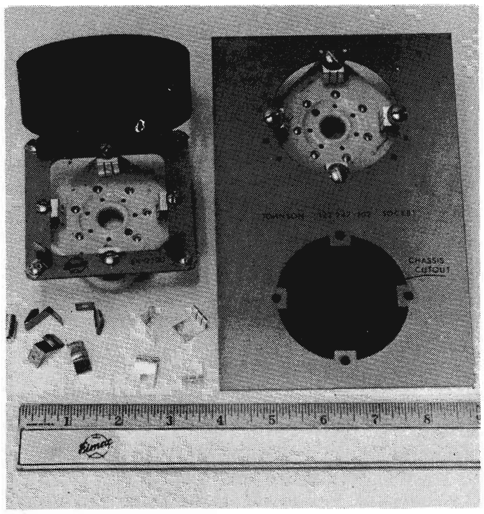


Fig. 2 — The builder may construct his own socket from a combination of components. This is one area in which the amateur can reduce the overall cost of the project.

usually finds it easy to adjust the loading control with his left hand as he adjusts the plate tuning control with his right hand.

When building uncomplicated equipment such as a linear amplifier, the home constructor has a cost advantage over the manufacturer. Labor and engineering amount to nearly 60 percent of the cost of a commercially-made unit. Think of what that saving would buy in new or surplus parts! A home-built unit may not match one commercially made in appearance, because of differences in shop tools, but it should be as reliable electrically.

A word of caution about buying used vacuum-variable capacitors. When there are no apparent cracks or flaws, a "leaker" or defective unit is hard to detect. If the lead screw moves the bellows too easily, the capacitor *could* be a dud. Have it "high-potted" at the rated voltage, or make sure you have a return guarantee so you can check the unit yourself.

Metal Fabrication

Quality workmanship does not depend entirely on the use of a metal brake and shear. Avoid use of tin snips when cutting materials. Sawing along a line causes less mechanical distortion. If you *must* use the snips, cut outside of the mark on the first slice, then approach the line with several thin slices.

Neat and accurate bending can be accomplished with two pieces of 1 1/2-inch angle iron about two feet long. The material to be bent is clamped between the angle irons, using C-clamps and a vise. Use a piece of flat wood or phenolic as a buffer between the hammer and the material to be bent, to avoid hammer marks. A test bend using a scrap piece of aluminum is sometimes helpful where a critical fit is needed. Use common sense to figure whether the bend will add or subtract metal thickness to marked length.

Tips on Shielding and Isolation

Shielding between input and output circuits of a grounded-grid amplifier reduces the possibility of parasitic oscillations, even when fundamental oscillation is not a problem. Generally, either type of oscillation can be detected by observing the grid and plate meter readings as the plate tuning control is rotated through resonance. If the system is free of parasitic oscillations, maximum grid current and minimum plate current will occur at approximately the same dial setting.

Long plate leads usually encourage the vhf parasitic type of oscillation. In this particular amplifier, however, there was no evidence of this.

Fundamental oscillation in a high-mu triode will not occur unless there is a feedback path between the input and output circuits. A path could be established, either by a control shaft not properly grounded, or through a wire not sufficiently bypassed. A metal shaft should be grounded where it passes through an open hole. Shaft bushings of the C-clamp variety do an adequate grounding job. If mechanical strength is not a factor, a nonmetallic shaft will do a good job of preventing undesired coupling. When using shaft couplings, replace the slotted screws with Allen screws. Nothing is more frustrating than a slipping shaft. This applies to knobs, too.

Wires passing through shields require the usual decoupling precautions, using ceramic bypass capacitors and perhaps small rf chokes or ferrite beads. This is especially important in treating the point of exit of the high-voltage lead.

If the high-voltage wire is kept in the plate compartment and not passed into the input area of the amplifier chassis, it will require less decoupling and be less prone to feed back. Shielded wire such as RG-8A/U cable, may be used between the plate choke and the exit point.

Regarding TVI and RFI leakage through an opening in a shielding surface, I recall a demonstration by Phil Rand, W1DBM, and Lew McCoy, W1ICP, at a radio club in Cleveland, just after WW2. A TV set was placed a few feet from a shielded box containing a 29-MHz transmitter. The top of the box had a four-inch row of 1/4-inch holes. Near the row of holes was a 1/4 x 4-inch slot. One or both openings could be closed with a shield plate. When the 4-inch slot was exposed, the TV set displayed severe interference. With only the row of holes exposed, no TVI was evident.

From this we may conclude that the mating surfaces of an rf enclosure should be free of wrinkles and slots. Securing clean surfaces every 2 inches seems to do the job. Some aluminum material is anodized, making it non-conductive. To avoid slots due to the insulation, check the surface with an ohmmeter. If the surface appears to be nonconductive, it must be cleaned.

TABLE 1

VALUES OF CIRCUIT Q OF 1

MHz	C1 Opt.	pF (USE)	C2 Opt.	pF (USE)	L µH
3.5	839	(820)	842	(820)	2.36
4.0	734	(750)	737	(750)	2.07
7.0	420	(430)	421	(430)	1.18
14.0	210	(220)	211	(220)	0.59
21.0	140	(150)	140	(150)	0.39
28.0	105	(100)	105	(100)	0.30

Practical capacitor values in parentheses

TABLE 2

L1 COIL WINDING DATA

BAND MHz	NO. TURNS	WIRE SIZE	INDUCTANCE RANGE IN µH	F* MHz
3/8-inch Diameter Forms				
3.5	14	24	1.64 - 4.58	5.05
4.0	14	24	1.64 - 4.58	5.8
7.0	10	24	0.96 - 2.32	10.1
14.0	7	16	0.44 - .74	19.5
21.0	5	16	0.28 - 0.52	29.2
28.0	4	16	0.17 - 0.34	40
1/2-inch Diameter Forms				
3.5	15	20	1.975 - 3.67	5.05
4.0	13	20	1.584 - 3.045	5.8
7.0	10	16	0.76 - 1.21	10.1
14.0	6	16	0.43 - 0.736	19.5
21.0	5	16	0.35 - 0.55	29.2
28.0	4	16	0.26 - 0.39	40

Coil winding data is given to allow a choice of either 3/8-inch or 1/2-inch slug-tuned forms. The winding should be close spaced and start at the top of the form.

* A grid dip meter should be used to assure that the inductor resonates at the indicated frequency. These adjustments should be made with capacitors C1 and C2 out of the circuit.

Air Cooling

The opening in a shield surface where blower air enters the chassis may be a source of rf leakage. In this amplifier, brass-wire screen is mounted in the air stream to minimize this leakage. Tiny globs of solder at several crossover points assure positive connection on the screen. The disadvantage of this method is the eventual collection of dust, restricting air flow. It requires periodic cleaning.

The question so often asked is, "Do I actually need this much air?" Remember, heat is what destroys a tube! If the blower noise is too great, place the fan elsewhere and duct the air to the amplifier. Only a slight hiss will remain as the air passes through the anode cooler.

The Input Circuit

The cathode impedance of an 8877/3CX1500A7 is about 54 ohms. Direct coupling from the exciter to the cathode without the use of a cathode-tuned circuit *will* work, but performance will be degraded. The reduced-drive requirements and improved distortion products make the small effort of putting a "flywheel" in the input circuit worthwhile.

The input pi-network circuit for each band is set and forgotten. Final adjustment of the slug-tuned coils is made with the amplifier operating and will be discussed later.

Fig. 1 shows the basic input circuit. A computer program for 50 to 54 ohms and a selection of three input circuit Q figures produced the most practical values of capacitance. Q values of 3, 2, and 1 in the computer run indicated that at 3.5 MHz the required network capacitors would be 2500, 1700, and 850 pF, respectively (values rounded out for illustration). A circuit Q of 1 was chosen based on price, physical size, and nearness to stock values. I accurately measured the value of more than 50 5-percent mica capacitors and found that about 90 percent of them were on the low side of the marked value. Keep this in mind when making a selection!

The Q of the input circuit is so low that any of the polyiron or ferrite-core materials are satisfactory. The slugs used in this amplifier were coded red (1-20 MHz).

Miller No. 4400 ceramic forms are one source of 3/8-inch cores. Cambion and Millen are other sources. The Millen No. 69046 is a good choice for 1/2-inch forms.

The Socket

For grounded-grid application, a reasonably priced socket is available from Eimac. The socket is part number SK-2206. If you prefer to build one yourself, a 7-pin septar socket, made by E. F. Johnson (part number 122-247-202) may be used. Fig. 2 shows construction details. Proper standoff spacing is obtained with 3/8-inch metal spacers, plus two metal washers. The finger-stock clips may be ordered from Eimac (part number 149-842).

Output Circuit Considerations

In any amplifier, the plate-to-grid capacitance of the tube adds to the stray capacitance, making it difficult to achieve the desired values of capacitance in a pi-network circuit at 28 MHz. A high value of input capacitance results in a high plate-circuit Q, and high circulating current. The higher Q is an advantage in attenuating harmonics, but the efficiency will be reduced because of high circulating current and heat loss. Since some compromise in circuit Q between 3.5 and 28 MHz is necessary, the best place to "cheat" is at the 10-meter end of the operating range. The 28-MHz coil can be wound with tubing or flat strap which aids in dissipating the heat.

If the 10-meter trade-off is chosen, then consider using an air-spaced capacitor for plate tuning. For contest operating or fast band changing it allows retuning quickly. If a capacitor such as the Johnson 152-1 is used, it will require extra capacitance to be switched in for 3.5 MHz.

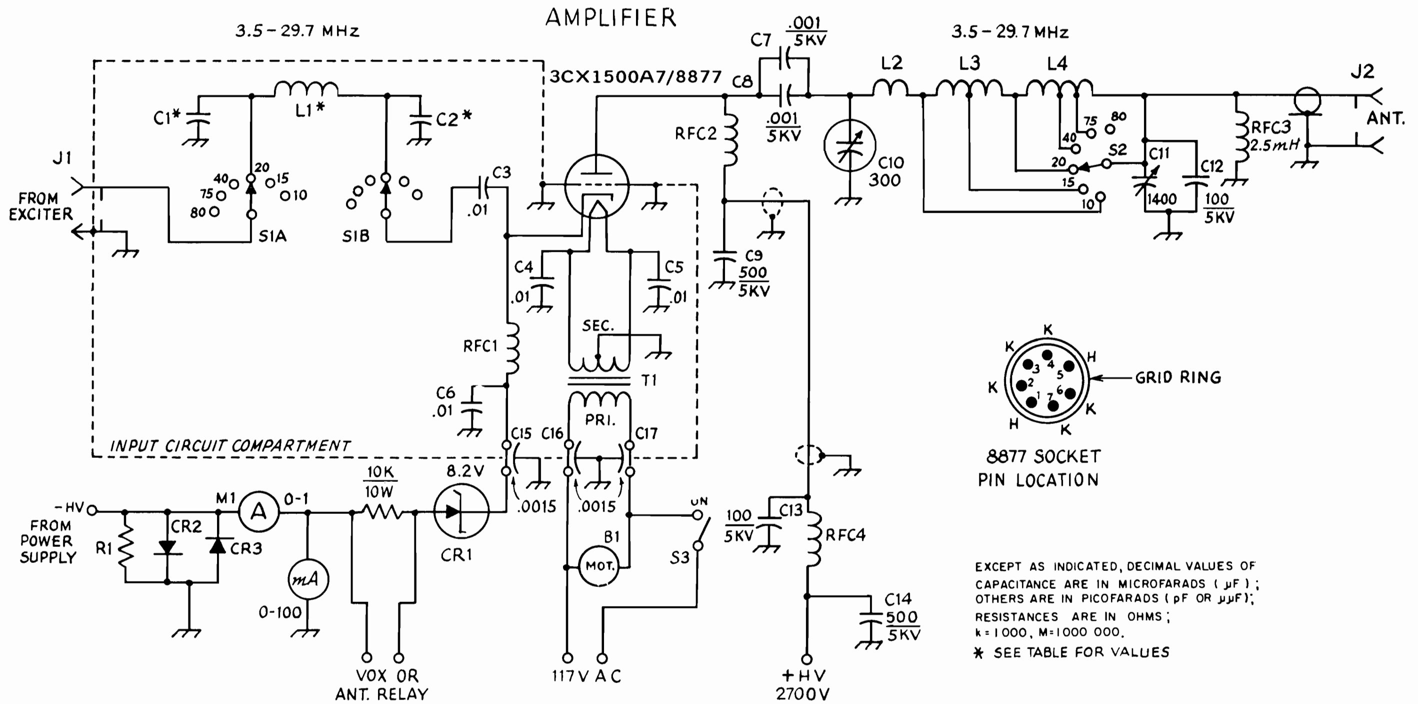
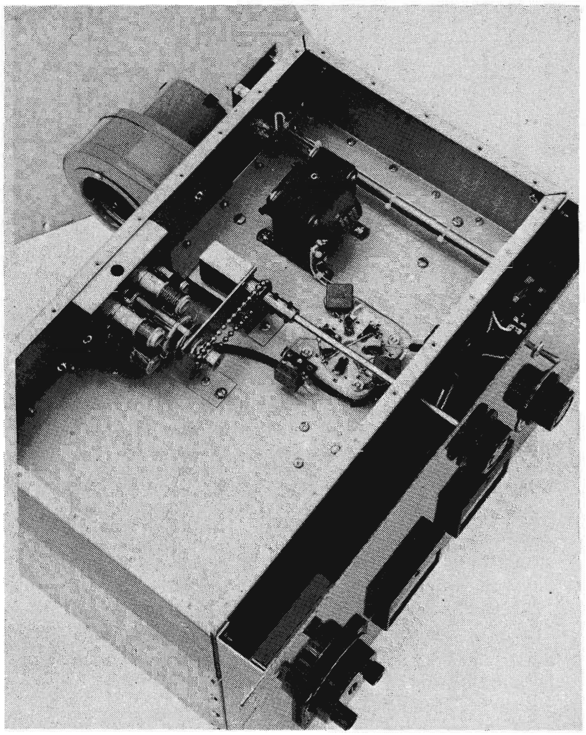


Fig. 3 — Circuit diagram of the 2-kW amplifier. Component designations not listed below are for text reference. The values for C1, C2, and L1 are given in Table 1.

B1 — Blower. (Dayton 4C012).
C3-C6, incl. — .01-μF, 600-volt disk ceramic.
C7,C8 — .001-μF, 5-kV (Centralab 858S).
C9 — 1500-pF, 5-kV (3 parallel 500-pF Centralab 858S).
C10 — Vacuum variable, 5-300 pF.
C11 — 4-section broadcast variable, 365 pF per section. All sections parallel-connected. (J.W. Miller 2104).
C12,C13 — 100-pF, 5-kV (Centralab 850S).
C14 — 500-pF, 5-kV (Centralab 850S).
C15,C16,C17 — Feedthrough, .0015-pF, 400-V.
CR1 — Zener diode, 8.2 V, 50 W (Motorola 1N3307).
J1 — BNC, chassis mount (Amphenol UG-1094/U).
J2 — SO-239 chassis connector.
L2 — 10-meter coil (see text).
L3 — 15- and 20-meter coil (see text).
L4 — 80- and 40-meter coil (see text).
M1 — 0-1 A dc meter.
M2 — 0-100 mA dc meter.
RFC1 — 15-μH, 1-A choke (Miller 4624).
RFC2 — 160 turns, No. 24 Formvar, wound on a 3/4-inch dia ceramic insulator, 4 inches long.
RFC3 — 2.5 mH, 300 mA.
RFC4 — 10 turns, No. 14 wire, 1/4-inch ID, 1-inch long.
S1 — Ceramic rotary switch, 2 pole, 6 position (Centralab PA-2045).
S2 — 1 pole, 6 position (Millen 51001).
T1 — 5-V, 10-A filament transformer.



The input circuit is mounted on a small bracket to keep the adjustments inside the chassis. S1 and S2 are coupled with a chain drive. A separate shaft arrangement with a knob could have been used for the input circuit eliminating the need for the chain drive.

The above points are illustrated in Fig. 4. The values given for a Q of 12 are the ones used in the amplifier described. All are for an input resistance of 2000 ohms.

A construction detail sometimes overlooked is the junction marked (A) in Fig. 4. If the amplifier is to operate properly at 28 MHz, the connection between capacitor C2 and the 50-ohm output terminal should be very short. A 5-inch lead, for example, represents quite an inductance. If a long lead is necessary, use RG-8A/U cable from the 50-ohm connector to the junction of L and C2.

Coil and Switch Mock-up

The actual inductance of the plate coil must be close to the design value. This is simple if the author's data and layout are duplicated. A drastic change of layout might be desired, so the following mock-up procedure for the coil and switch assembly will assure optimum operation. Fig. 5 shows the points referenced below.

1) Cut a piece of sheet metal or aluminum slightly larger than the space selected for mounting the coil and switch assembly.

2) Mount the band switch on the sheet in the position it will occupy in the final setup.

3) Measure the distance from the switch arm to the capacitor (C2, if it were mounted) and along the chassis to the tube (if it were mounted). See item M, Fig. 5.

Cut a 3/8-inch-wide copper strap to the length measured. Attach one end to the switch arm and add a jumper to the 80-meter switch position. The free end will be connected to Cx, at (M), later.

4) Measure and cut a strap for the path between the tube position (A) and the 28-MHz coil (D). This path is via the chassis at (A), through the position of C1, then to a point near the connection of the 28-MHz coil, blocking capacitor and rf choke (B) (C) to (K).

Don't let the length of the leads scare you. They actually represent circuit inductance not wound into the coils. Each strap will be about a foot or so long. Now we must find the added amount of inductance needed for each band in the form of a coil.

A pi network is a resonant circuit using a coil and two capacitors in series. For this mock-up, a fixed value of capacitance is used to represent the *effective* value of capacitors C1 and C2 in Fig. 4. The value changes with each band and can be made with stock values or parallel combinations.

Assuming the use of design values in Fig. 4(A) for a Q of 12, the effective series capacitance of C1 and C2 used for Cx is:

MHz	*Cx*
3.5	230 pF
4.0	202 pF
7.0	115 pF
14.0	57 pF
21.0	38 pF
28.0	28.7 pF[1]

5) Place the band switch in the 10-meter position. Connect Cx (10-meter value) to the two strap ends (K) and (M). Wind a coil (D) of 1/4-inch copper tubing having 5 to 8 turns about 2 inches in diameter. Space the turns about 1/8 inch apart.

Cone-and-pillar insulators (E) (F) (H) (J) are used to make stable mountings for the coils. Connect the (10) and (20) taps to the proper insulators and install straps (G) between (F) and (H). Securely mount one end of the 10-meter coil to insulator (E). Position the free end as shown for connection to the junction of the blocking capacitor, rf choke and C1 (B) (C). The insulator at (C) represents the height and location of the plate rf choke.

The free end of the lead (B) attached to (C) is then moved from turn to turn until the circuit resonates at 28 MHz. Remove excess turns. This coil stays in the circuit on all bands. Its position affects the inductance of the remaining coils, so mount it securely and connect the lead (B) from (C).

6) The 20-meter coil can be determined next. It is mounted between (E) and (F). Wind the coil in the same direction as the 10-meter coil (D). Normally 1/4- or 3/16-inch copper tubing is used for the 15- and 20-meter coils. This is because of the larger surface and better heat-handling capability.

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[1] If the plate-tuning capacitor has a high minimum capacitance, use 35 to 40 pF for this band.

Change the band switch and Cx to 20 meters. The coil diameter and turns are juggled along with the turn spacing to arrive at 14-MHz resonance. The coil must fit the space between insulators (E) and (F). Five turns about 3 inches in diameter should work. The turns may be squeezed or spread to adjust the inductance.

7) Connect a lead to the 15-meter tap on band switch (15). Change band switch and Cx to 15 meters. The 10-meter coil plus 2 turns of the 20-meter coil should be about the correct tap point for 15 meters. If it is, solder the lead to the coil. The 20-meter position should be rechecked. Don't forget to change Cx.

8) For 80 meters, inductance is added to the 10- and 20-meter coil previously wound. Air-Dux 2404T coil stock may be used. It is 3 inches in diameter, has 4 turns per inch, and is of No. 10 wire. This coil is connected between (H) and (J). Leads for the 40- and 75-meter taps are attached to the appropriate band-switch lugs. The 80-meter band must be adjusted first. Place the proper value for Cx in position, rotate the band switch to 80 meters, and trim the coil to resonance. The 40- and 75-meter taps are determined in the same manner as the 15-meter tap was found. Be sure to recheck each band when all of the taps are in place.

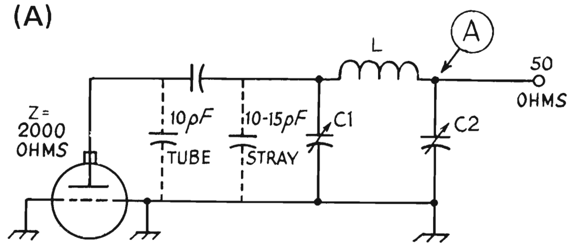
9) The switch, coil, and insulators are removed from the test plate *as a unit*. The test plate serves as a template to lay out the holes on the amplifier chassis. The tank circuit then is installed in the amplifier in one piece and the leads are attached to the variable capacitors. All leads and straps should be made as short as possible.

Proving the Mock-up Procedure

I used two methods to determine that the value of inductance was correct on each band. The amplifier tube was placed in its socket and connected to the tank circuit. Then, with a 2000-ohm load resistor shunted across the plate circuit, an Rx meter was used to check the impedance at the output connector. When the tuning capacitors were adjusted to the calculated value, the Rx meter indicated 50 ohms.

The second method of testing the amplifier was under full-power operating conditions into a

(A)



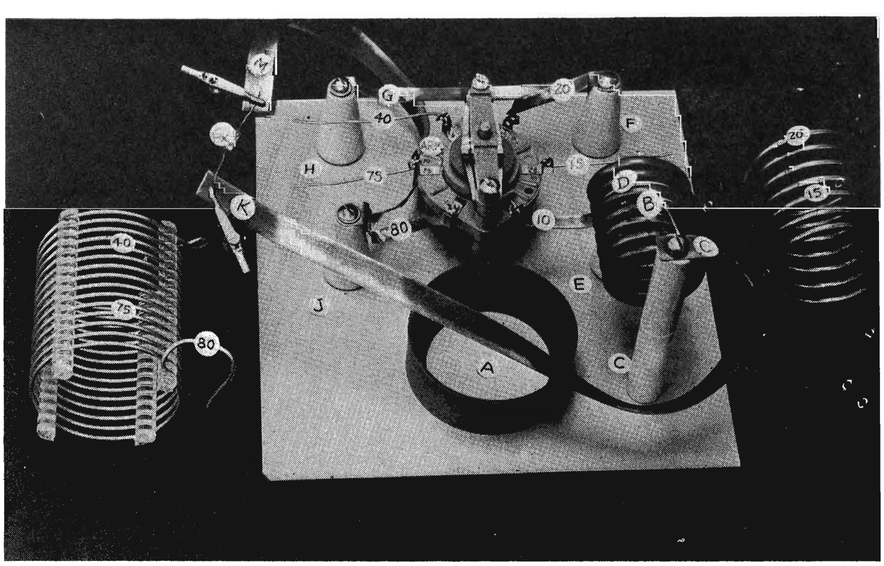
(B)

	Freq. MHz	*Input* C1 pF*	*L µH*	*Output C2 pF*	*Notes*
For *Q* of 12	3.5	273	8.54	1473	
	4.0	239	7.47	1289	
	7.0	136	4.27	737	A good set of
	14.0	68	2.14	368	compromise
	21.0	45	1.42	246	values.
	28.0	34	1.07	184	
For *Q* of 15	3.5	341	6.90	1961	Note high input
	4.0	298	6.04	1716	C at 28 MHz, and
	7.0	171	3.45	918	high output C at
	14.0	85	1.73	490	3.5 MHz.
	21.0	57	1.15	327	
	28.0	43	0.86	245	
For *Q* of 10	3.5	227	10.12	1123	
	4.0	199	8.85	983	All practical
	7.0	114	5.06	562	values except
	14.0	57	2.53	281	for 28 MHz input
	21.0	38	1.69	187	C. Large coils
	28.0	28	1.26	140	would be required.

*Value indicated is total of stray, tube, and tuning capacitor capacitance.

Fig. 4 — Simplified diagram of the output circuit. (B) lists the values for C1, C2, and *L*.

Fig. 5 — The mock-up used by the author for determining the correct tap points for the output circuit. The letters and numbers are for text reference. ▼



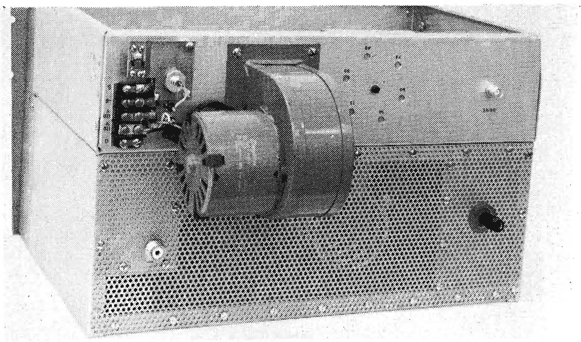


Fig. 6 — Rear view (inverted) of the amplifier. The diodes are mounted on a piece of circuit board.

dummy load. The settings of C1 and C2 were the same as determined earlier.

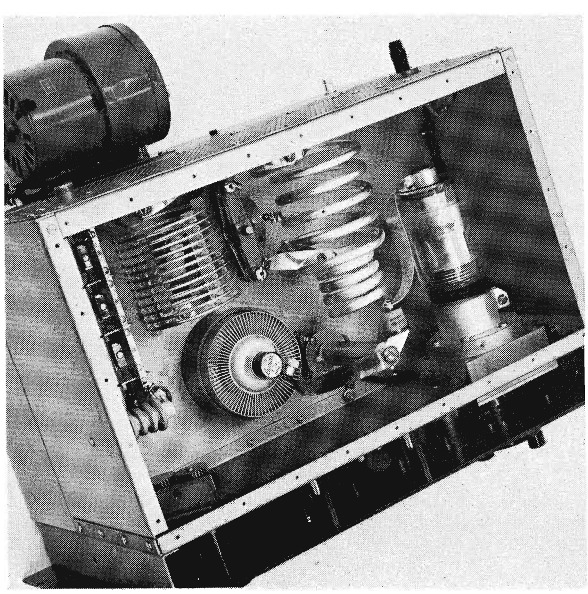
Control Circuits

There are two choices of methods for controlling standby-transmit conditions. First, the plate voltage may be turned off during standby, in which case no added grid bias is required. The second method allows the plate voltage to be left on at all times, but calls for additional bias during standby. The extra bias is required to cut the tube off completely during this period.

The second method is perhaps more commonly used, and only requires a fixed value of resistance to be placed in series with the biasing Zener diode during standby. Exciter or antenna relay contacts can be used to short out the resistor during transmit periods. This method is preferred when using solid-state rectifiers and high-capacitance filters in a power supply.

For cw operation, an additional resistor may be used to bias the tube close to cutoff, and can be switched out for ssb. A fuse in series with the Zener diode is a feature to be considered. It may save a tube, Zener, or meter from damage if an arc occurs, causing a high current surge. Use a 1-ampere fuse.

A piece of printed-circuit board may be used to mount the meter-protecting diodes. The Zener diode is also mounted on this board. Fig. 6 shows the meter diodes on the left side of the board. Space was allowed to mount a second Zener diode for added bias, switchable from the front panel.



Mechanical Assembly

Wires from the front panel to the rear of the chassis are routed through a duct made of 3/8-inch tubing, threaded at each end to accept nuts. The small blower in the photograph (Fig. 6) provides sufficient air for a 2-kW PEP operation.

To the left of the blower are the six access holes for tuning the input coils. A round disk with matching holes is mounted inside the chassis, held by a center screw. Once the adjustments are made, the disk is rotated enough to cover the holes. Then the screw is tightened to hold the disk in place. Metal buttons will work just as well.

Also shown in the photograph is a method of securely mounting the connector to a plate that is independent of the removable perforated screen. On the upper left is the high-voltage connection.

Metering

Be sure to protect the meters from damage with a pair of back-to-back diodes. Any of the inexpensive silicon diodes connected in parallel, but with anodes in opposite directions across the meter terminals, will conduct if the voltage exceeds approximately 0.6 volt. An extreme surge may even short one of the diodes. It is therefore advisable to place them where they may be tested or replaced easily. A shorted diode may shunt the meter sufficiently to give a false meter reading, and calibration should be checked if a surge is ever experienced. Only two diodes are needed in this amplifier to protect both meters.

Negative-lead metering is preferred both for safety and simplicity. This method requires all grounds to be removed from the negative points in the high-voltage power supply. The negative points are then connected to a common negative bus which is grounded through a resistor. A separate wire is used to connect the negative bus to the plate meter and back to the tube cathode through the meter and Zener diode, completing the high-voltage path. Since the tube is connected in the grounded-grid configuration, metering for grid current is placed in series with the grid (ground) and cathode. Thus both meter movements are only a few ohms above ground.

The Power Supply

If a power supply is modified for negative-lead metering it is a good idea to use a grounding resistor in the power supply as well as in the amplifier. This limits any voltage difference between the negative bus and the chassis to a very low value. The resistance must be high enough to prevent shunting the meters, but low enough to provide a low voltage difference between -HV and ground. Any value of *wirewound* resistor between 25 and 500 ohms, rated 10 to 25 watts will suffice.

The top view of the amplifier shows the vacuum-variable capacitor mounted to a subpanel. A small blower mounted on the rear of the cabinet provides sufficient air to cool the tube during full-power operation.

With safety of life as a factor, the writer prefers to use parallel resistor combinations.

Final Testing

Laboratory tests at Eimac indicate best performance to be at an anode potential of 2700 to 3000 volts. The efficiency runs between 60 and 65 percent.

Plate impedance figures are based on a 2 kW PEP input using 2700 volts at 740 milliamperes. The grid current for the 8877 runs about 15 percent of the plate current. At full power input, the grid current should be about 110 mA.

When plate voltage is applied, the zero-signal plate current should be about 95 mA. Drive should be applied through a directional coupler. On each band, after fully loading the amplifier to the above conditions, tune the input coil for minimum reflected power. No further adjustment is required and the directional coupler can be removed.

As the plate tuning control is "rocked" through resonance note the action of the grid and plate current. If maximum grid current and minimum plate current occur at approximately the same point, the amplifier is probably stable.

For maximum efficiency from *any* amplifier at *any* power level, it is important to have proper drive and loading. Changing from high to low power (2 kW to 1 kW, for instance) is *not* just a matter of reducing the driving power. Using the 15-percent ratio of grid-to-plate current, the ratio would be 110 to 740 mA for 2 kW input, and 55 to 370 mA for 1 kW input. If the anode voltage of 2700 is maintained, a change from 2 kW to 1 kW will double the plate load impedance. Therefore, the tuning and loading controls *must* be readjusted. Additionally, the drive must be reduced. If the dummy load survives tune-up logging of dial settings, you are finished.

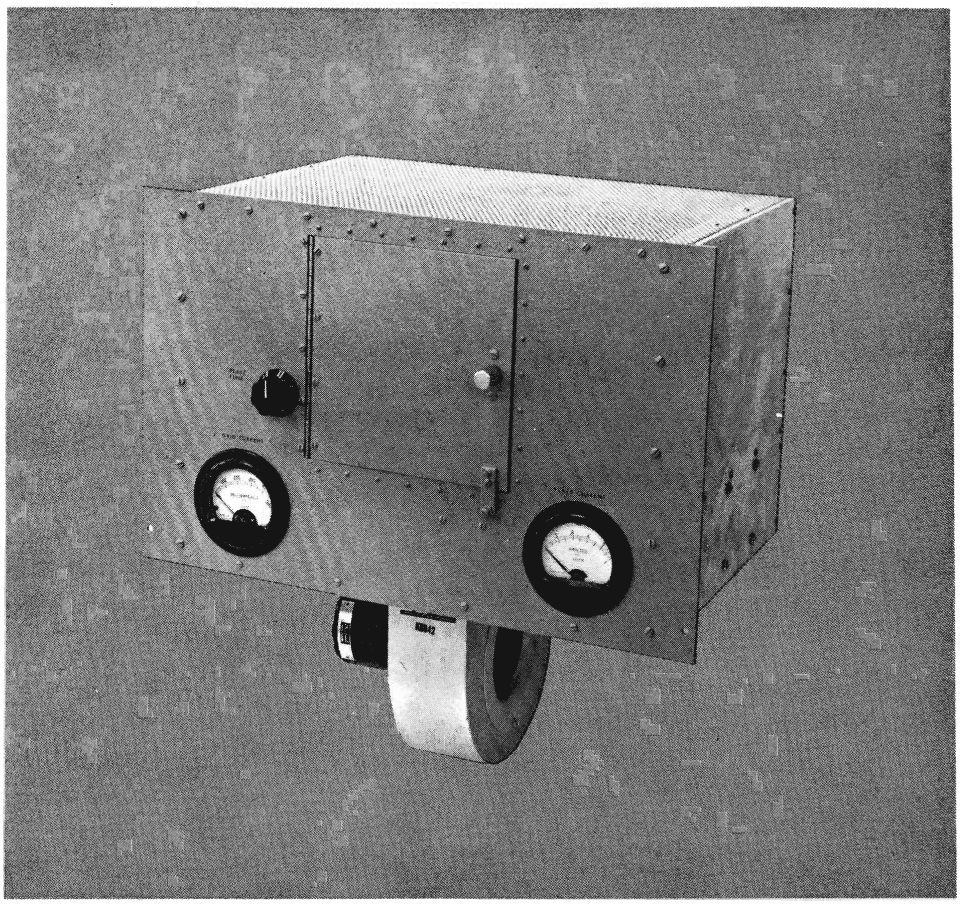
A note of thanks goes to Bob Sutherland, W6UOV, and Ray Rinaudo, W6ZO, for their encouragement in preparing this article and to Bill Orr, W6SAI, for his editorial assistance. QST

[EDITOR'S NOTE: A constructional technique used by the author to eliminate rf leakage through large holes in the chassis will be discussed in a subsequent issue of *QST*. It is called a "waveguide-beyond-cutoff" and can be used to duct air from a blower to the amplifier chassis. No screens are required in the air path, yet the chassis remains rf tight.]

SOCKET AND CHIMNEY COMBINATIONS FOR THE 3CX1500A7/8877

Two Air System Sockets and Chimneys are available for this tube:

(Grounded Grid operation with grid grounded to chassis)	SK-2210
(Grounded Cathode operation)	SK-2200
Teflon Air Chimney (for VHF use)	SK-2216



design data for a two-kilowatt vhf linear

The input matching network and plate circuit are unique in this strip-line amplifier— self neutralization is also featured

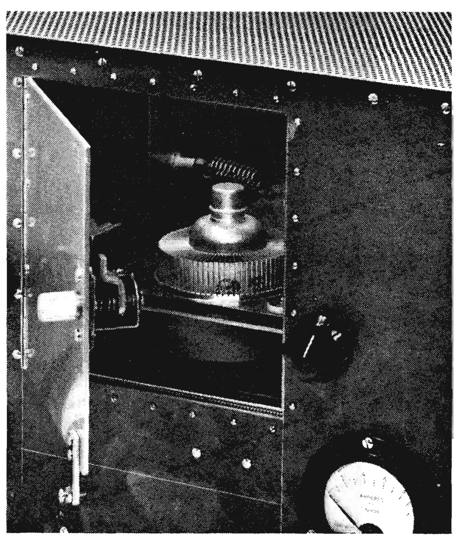
Robert I. Sutherland, W6UOV, Power Grid Development Lab,
Eimac Division of Varian, San Carlos, California 94070

This article describes the design of a 150-MHz, 2000-watt PEP input, grounded-grid linear amplifier. I built this amplifier to conduct life tests and determine the performance characteristics of the new Eimac 3CX1000A7 high-mu ceramic triode at 150 MHz. Many of the design techniques and performance figures at 150 MHz will be of interest to the serious vhf and uhf experimenter. By changing the cathode matching network and the plate line length, this amplifier would be excellent for the 144-148 MHz band.

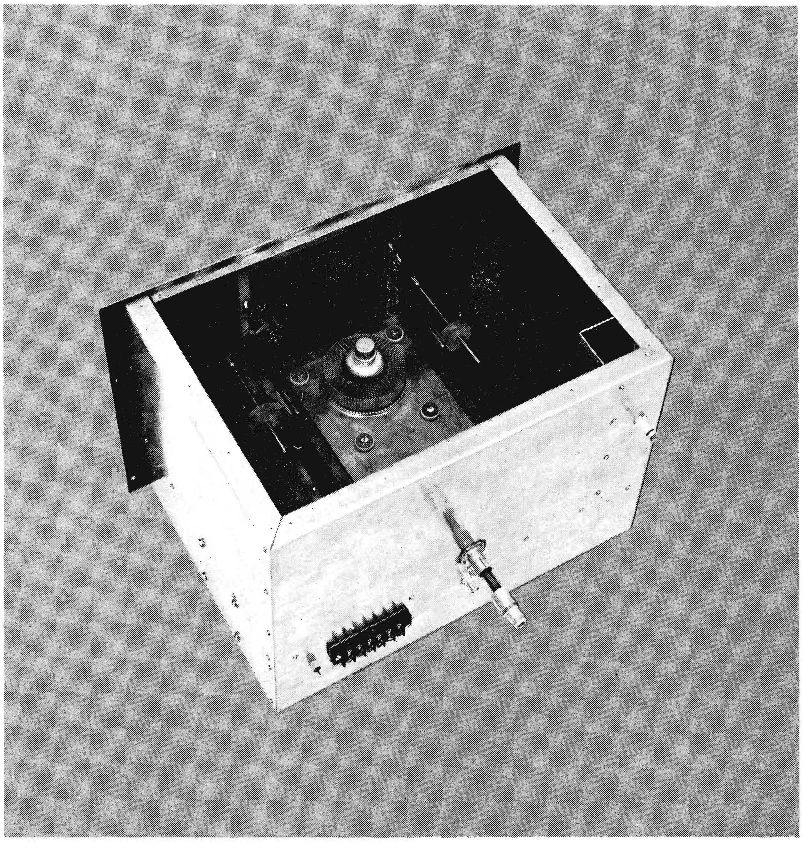
Several propagation modes, in which the maximum legal input power can be justified, are possible in the two-meter band. Meteor scatter, moonbounce, forward scatter and tropospheric bending are but a few.

Running the full, legal one-kilowatt aver-

Amplifier with the access door open; the 3CX1000A7 can be seen mounted in the plate line.



age power input to the final amplifier in the cw and ssb mode has been fairly easy in the region between 2 and 30 MHz with the large selection of power tubes and circuits available. At these frequencies, many small tubes can be paralleled to develop also use four such tubes to develop an average plate input power of 1 kW for ssb radiotelephone service, with reasonably low intermodulation distortion. A single 4CX1000K can be used to give 1 kW average input power, under voice conditions,



Top view of the amplifier with the shield removed. The variable tuning capacitor on the left was changed in the final model to a fixed capacitor. The output coupler coax may be seen protruding from the back panel.

power-handling capability. At 144 MHz, however, tubes that will handle 1-kW average power with good efficiency, stability and reliability are few, and large arrays of television sweep tubes are impractical.

choice of tubes

Many successful amateur designs have been built for 144-MHz, 1-kW continuous duty (single-tone) using the 4X150A, 4CX250A and 4CX300A tetrodes. Generally, two tubes are used in push-pull. You can in a 144-MHz amplifier[1] with good efficency, stability and low drive.

Many popular high-frequency amplifiers, using zero-bias glass triodes such as the 3-400Z, 3-500Z, and 3-1000Z, are now being used in grounded grid. The 3-400Z and 3-1000Z have been used successfully in 150-MHz grounded-grid amplifiers, but neutralization is difficult.[2] The 3CX1000A7, a zero-bias, ceramic, high-mu triode is now available, which simplifies these problems. It is well-suited for vhf.

The 3CX1000A7 is electrically similar to the 3-1000Z. The amplifier described in the following paragraphs uses this metal-ceramic tube in a circuit for 150 MHz, and it has all the advantages of zero bias and grounded grid. Bias and screen-voltage supplies aren't required. Stage gain is high, and neutralization is simple and effective.

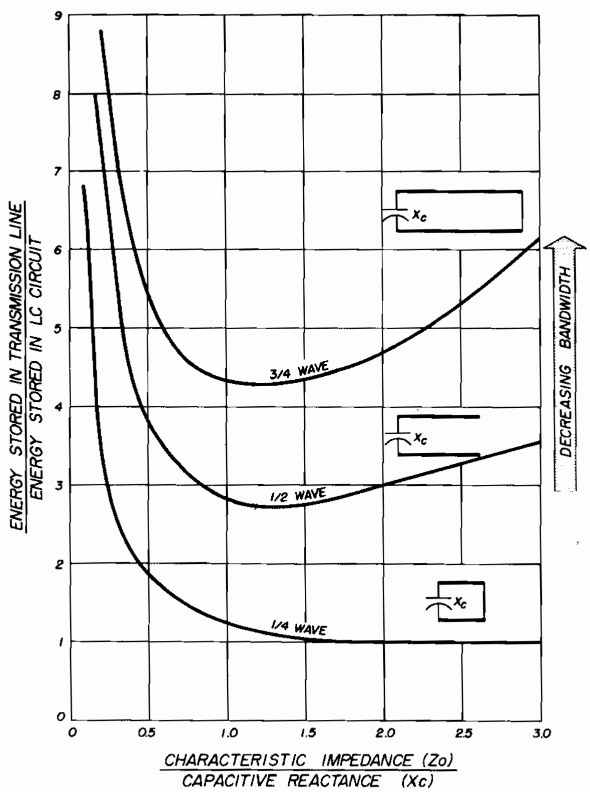


fig. 1. Frequency dependence of tuned transmission-line segments as compared to an LC circuit. Zo is the characteristic impedance of the line and Xc is the capacitive reactance acting on the sending end of the line.

the plate circuit

The choice of the plate circuit is important in the design of a vhf amplifier. I used a strip-line amplifier because it requires less machining.[3] A brake, shear and drill are all that are required. These tools are found in any well-equipped metal shop. You could do the job yourself with a hacksaw, file, drill and some perseverance.

The plate tank has a characteristic impedance, Z_o, of 60 ohms. The higher the impedance of this part of the circuit, the less will be the Q in the line; consequently, you will have a greater bandwidth. A wide bandwith isn't necessary to pass the modulation frequencies, but it is desirable if you want to eliminate a trimming device. A large amount of heat must be dissipated by the plate circuit, and if the circuit Q is low, the tank circuit will not detune as rapidly as it would with a high-Q circuit.

There's a practical limit as to how high you can make the plate strip-line impedance. Tube output capacity, plus strays, will limit the impedance and length of plate line:

$$X_c = Z_o \tan \iota$$

where:

X_c = tube output reactance.
Z_o = characteristic impedance of plate line.
ι = line length in electrical degrees.

For a quarter-wave line, $Z_o/X_c = 2$ is adequate (**fig. 1**).

The plate strip-line is also useful for making a sandwich-type blocking capac-

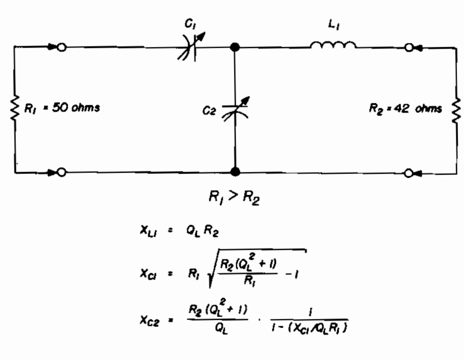


fig. 2. The schematic and equations used to calculate the input matching network. For this amplifier, R1 = 50 ohms, R2 = 42 ohms and assigned loaded Q is 2.

itor. In this strip-line assembly, a 10-mil-thick piece of **Isomica** was used for the dielectric. Mylar or teflon could be substituted with good results.

Output coupling is by a sliding contact on the inside surface of the plate line. As the sliding contact moves toward the anode, the area of the pickup loop increases, thereby increasing the coupling.

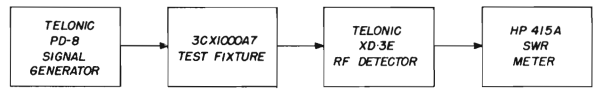
the input circuit

The input, or cathode tuned circuit, is a T network designed to match a 50-ohm coaxial line to the 42-ohm cathode driving impedance of the 3CX1000A7 (**fig. 2**).

The EIMAC SK-870 socket mounts directly on the chassis with the control grid contacts in contact with the chassis; that is, the control grid is at dc ground. The grid current is metered in the cathode return lead. The SK-870 socket is modified, as described later, to accomplish neutralization.

The driving impedance of the 3CX1000A7, in grounded-grid operation, is 42 ohms in parallel with about 20 pF of input capacitance. In order to simplify the design of the input circuit, it is assumed that the input to the 3CX1000A7 is resistive only. By making the matching network components variable, it's possible to resonate the input network, including the 20 pF of tube capacitance, while accomplishing the desired match.

fig. 3. Block diagram of the test set-up to determine the self-neutralizing frequency of the 3CX1000A7 triode and SK870 socket.



The input matching network in relation to the tube socket at the input coax line. The coil between the socket and the variable capacitor assembly is L1. The coil in the lower left-hand corner is the bifilar filament choke.

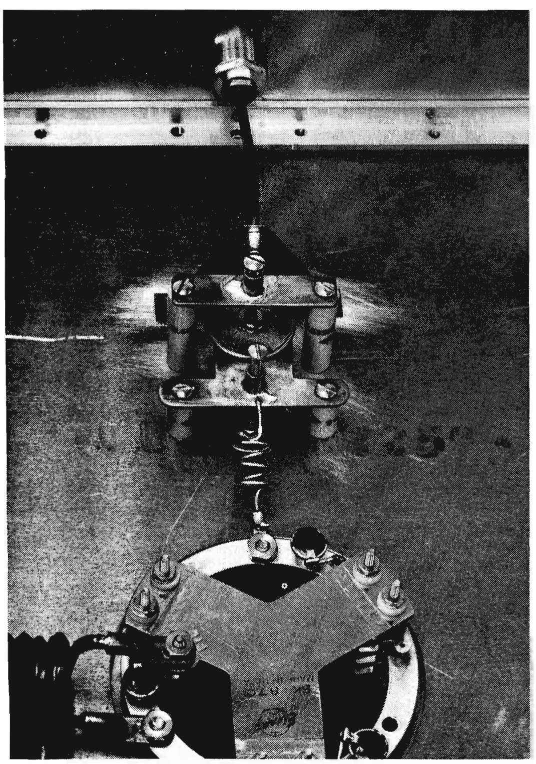


Fig. 2 shows the matching network and the design equations used. Under full plate-current load, the input match is adjusted to a 1:1 vswr, because only under full-load conditions will the tube input resistance be at the design value of 42 ohms. It's possible, however, to adjust the input network without the amplifier operating. Measure the resistance of a good-grade 42-ohm carbon resistor on an impedance bridge, then place the resistor between grid and cathode. The filament and all other voltages on the 3CX1000A7 must be off for this adjustment. A small amount of drive is applied through a standing-wave bridge. The input network is then adjusted for a 1:1 vswr at the operating frequency.

The "cold" adjustment won't be exactly correct, but the technique does allow the network to be tested and closely adjusted before application of power. The final adjustment is done with all voltages applied to the amplifier and tube with the 42-ohm carbon resistor removed.

self-neutralizing frequency

The 3CX1000A7 and SK-870 were tested in a special test fixture that treats the tube and socket as a four-terminal network. The anode was driven by a high power signal generator. The feedthrough signal was then detected on the cathode (see **fig. 3**). The control grid was grounded during this test. A curve was then plotted showing the relative magnitude of the feedthrough signal as a function of the signal-generator frequency.

The radio-frequency signal amplitude on the anode must be kept the same at all frequencies. A frequency will be found at which there is a minimum feedthrough signal; this point is called the **self-neutralizing frequency** of the tube and socket (**fig. 4**). The frequency thus obtained may be used as a guide in the design of a neutralizing circuit for the amplifier.

The graph of **fig. 4** indicates a self-neutralizing frequency of 105 MHz; this is lower than the operating frequency (150 MHz). The amplifier was unstable at 150 MHz before neutralization was applied. To raise the self-neutralizing frequency closer to 150 MHz, it was decided to lower the control-grid lead inductance. The SK-870 socket was modified by removing the six grounding spacers (see photo). These are metal spacers approximately 1/8-inch thick on each of the six SK-870 mounting screws. This modification allows the socket to be mounted 1/8-inch closer to the chassis.

The self-neutralization curve was again run with the same fixture and this modified

The Eimac SK870 socket. The six 1/8-inch metal spacers must be removed to neutralize the amplifier.

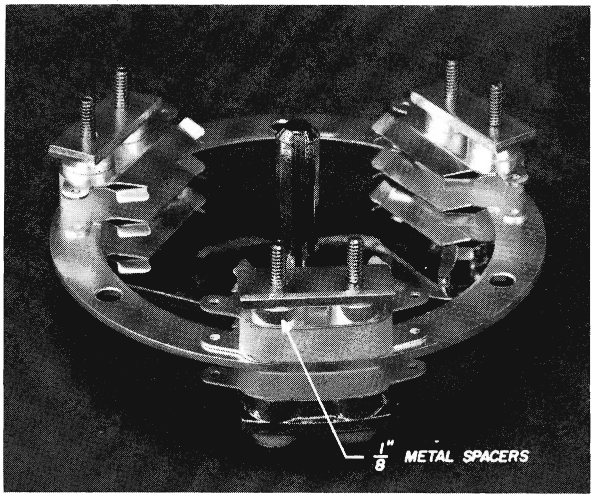


fig. 4. Relative input-to-output circuit isolation as a function of frequency before the socket modification. The self-neutralizing frequency is just below 105 MHz.

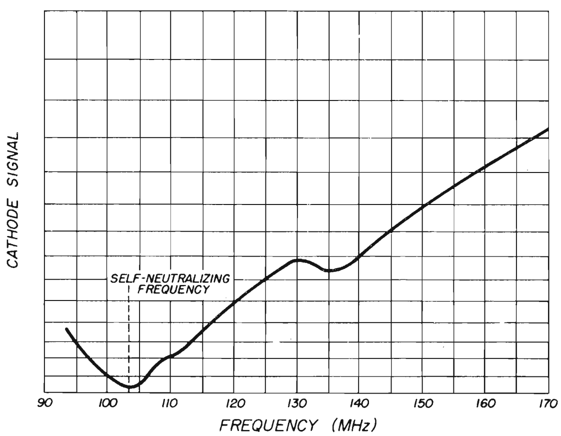


fig. 5. Relative input-to-output circuit isolation as a function of frequency after the socket modification. The self-neutralization frequency has moved up to 136 MHz.

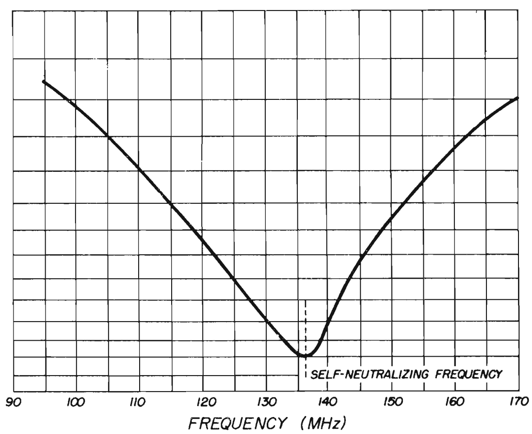
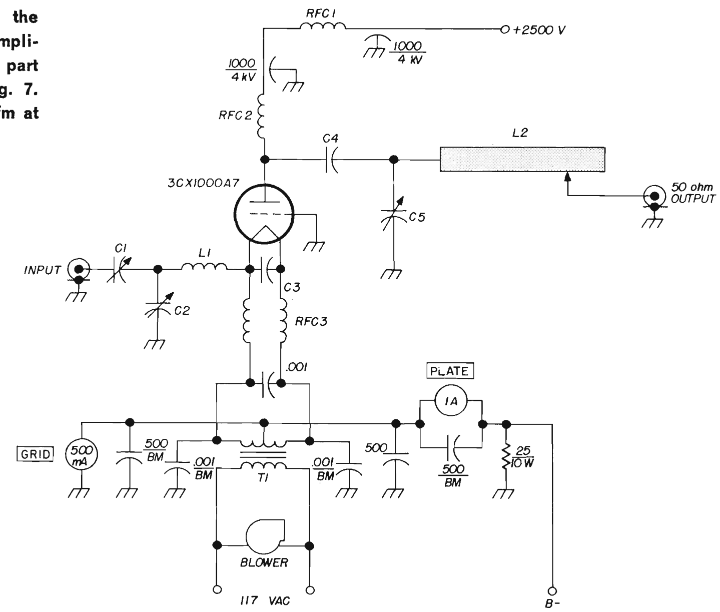


fig. 6. Schematic of the grounded-grid 150-MHz amplifier. C4, C5 and L2 are part of the plate line; see fig. 7. The blower delivers 37 cfm at 0.4 inch of water.



C1	11.9 pF (see matching network photo)
C2	1.06 pF (see matching network photo)
C3	1500 pF. Three 500-pF, 500-V stud-mounted button micas, spaced uniformly around the filament ring
L1	0.089 μH. 3 turns no. 14 wire, 1/16" diameter, 5/8" long
RFC1, 2	10 turns no. 16 tinned, 0.650" OD, 1-1/2" long
RFC3	bifilar coil no. 10 wire, each coil 5 turns, 3/4" ID, 1-5/8" long
T1	5 V, 33 ampere filament transformer

socket. **Fig. 5** shows the results: the self-neutralization frequency has now been moved to 136 MHz. The socket was again mounted in the amplifier, which was then retested for stability. The amplifier was stable, and the power output peak coincided with the plate current dip. Either arrangement of the socket grid return leads raised the self-neutralization frequency sufficiently to achieve stability or the neutralization point is sufficiently broad to include 150 MHz. It wasn't necessary to improve the neutralization further. Undoubtedly this simple socket modification is necessary to achieve optimum stability in the two-meter band.

amplifier assembly

The schematic of the amplifier is shown in **fig. 6**. Note that the circuit is much the same as that of an hf amplifier. Only the

Closeup of the output load coupler. The teflon block, with the fingerstock mounted on top, forms an adjustable loop as it slides back and forth to adjust the loading.

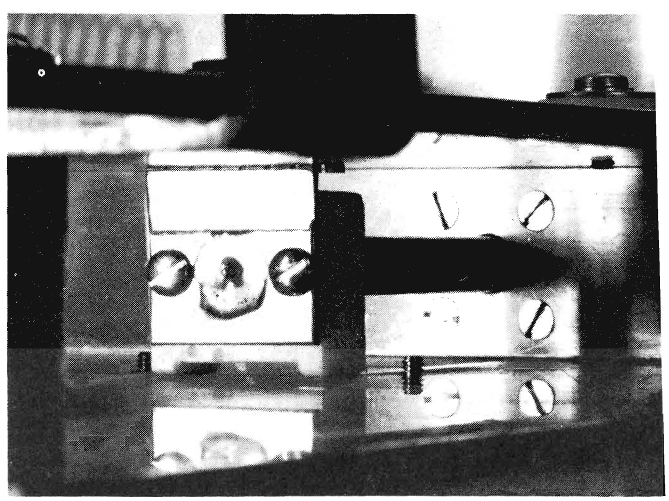


plate tank circuit and the input matching network can be considered unique. The metering of the grid and plate currents is in the negative return leads. The grid is metered in this manner to allow a very low impedance radio-frequency connection between the control grid and the chassis.

As noted previously, this grid circuit is part of the amplifier neutralization scheme, and the use of a grid by-passing arrangement could have complicated the amplifier neutralization. The plate current is metered in the negative return lead in the interest of safety. It is always dangerous to have a plate current meter mounted on the front panel if the meter is operating at a high potential with respect to ground.

A 25-ohm, 10-watt safety resistor **must** be connected between the negative terminal of the plate power supply and ground. If either grid or plate meter should become open circuited, and the plate side of the power supply shorts to ground, the negative side of the plate supply could assume a potential equal to the supply voltage. The 25-ohm resistor will prevent the negative terminal voltage from soaring and will load the power supply sufficiently to cause the fuses, or overloads, to function.

Fig. 7 is a pictorial of the plate line and the dimensions that were used in the 150-MHz amplifier.

operating results

The amplifier has been operating at 150-MHz many hours under the following conditions:

Plate voltage	2500 volts
Plate current	800 milliamperes
Plate input power	2000 watts
Grid voltage	0 volts
Grid current	240 milliamperes
Output power	1175 watts
Drive power	50 watts
Filament voltage	5 volts
Filament current	33 amperes

Note the output power and the required drive power; the power gain is 23.5 times or 13.7 dB.

No intermodulation distortion tests were made at 150 MHz although tests have been run at 2 MHz to determine the tube characteristics. Under the same operating conditions given for 150 MHz, the third-order intermodulation products were no less than 32 dB down, and the fifth-order products were better than 38 dB down from one tone of a two-equal-tone test signal.

references

1. Barber et al, "Modern Circuit Design for VHF Transmitters," CQ, November & December, 1965.
2. Orr and Sayer, "Semi- and Super-Cathode Driven Amplifiers, *QST*, June & July, 1967.
3. William I. Orr, W6SAI, and John T. Chambers, W6NLZ, "Strip-line Kilowatt for 432 MHz," *ham radio*, September, 1968.

ham radio

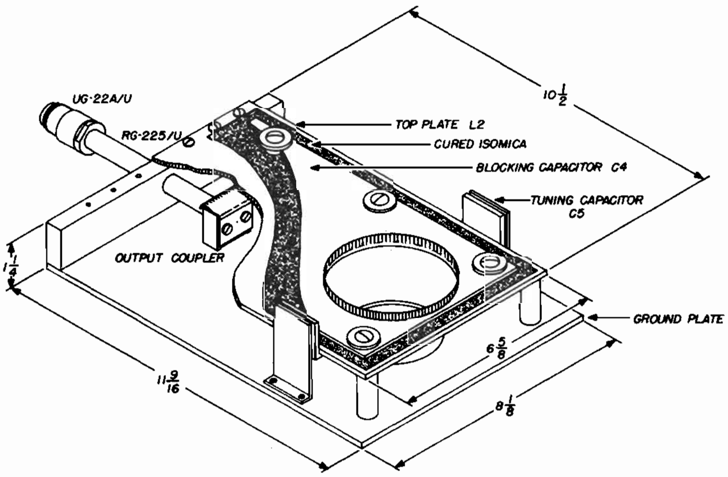
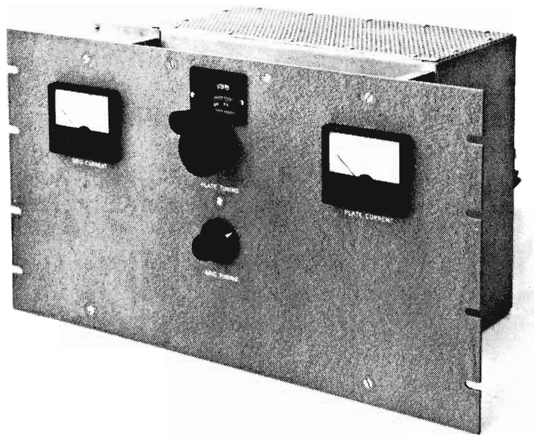


fig. 7. Cutaway view of the 3CX1000A7 plate line. The isomica is the dielectric for tthe plate-blocking capacitor C4. The two support pillars and six insulating shoulder washers are made of teflon as is the sliding output coupler block. The finger contact strip for the anode of the 3CX1000A7 is Instrument Specialties Company no. 97-133. A more detailed drawing of the plate line is available from the author.



high performance 144-MHz power amplifier

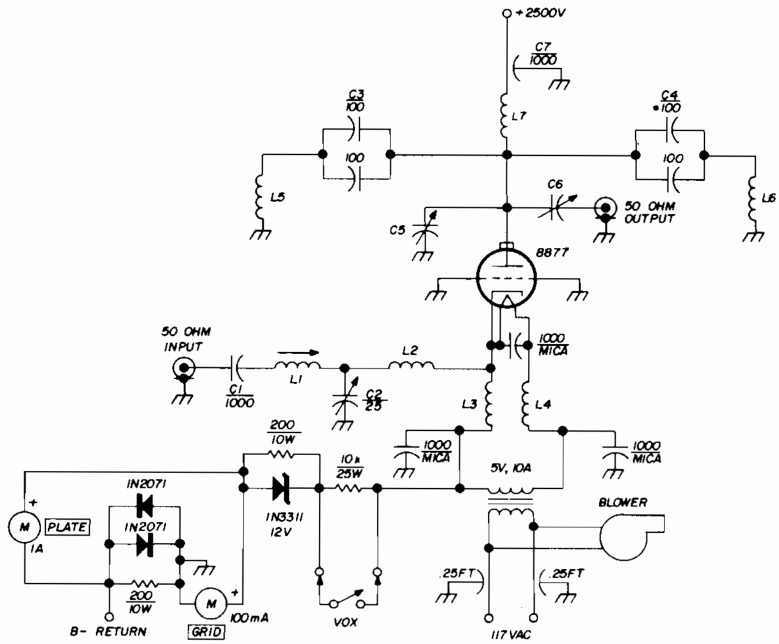
This efficient easy-to-tune grounded-grid 8877 amplifier can be run at 2000 watts PEP ssb or 1000 watts cw

Robert I. Sutherland, W6UOV, EIMAC Division of Varian

The new Eimac 8877 is a ceramic/metal high-mu triode rated for use up to 250 MHz. Operation of this tube at 50 MHz proved to be so satisfactory[1] that other 8877 amplifiers have been designed and built for frequencies up to 350 MHz. Two of these amplifiers are of interest to the serious vhf operator. One amplifier is designed for the amateur 2-meter band and is described here. The other amplifier covers the range from 150 to 230 MHz, and is well suited for use on the amateur 220-MHz band; it will be described later.

The 8877 triode has good division between plate and grid current and low intermodulation distortion. It has a plate dissipation rating of 1500 watts and mu of approximately 200. The cathode is indirectly heated; filament requirements are 5.0 volts at 10 amperes. The tube base mates with a standard septar socket.

This 144-MHz 8877 linear amplifier is designed for the serious DXer who demands reliable service combined with good linearity and efficiency. The compact grounded-grid design presented here uses a half-wave plate line[2] and a lumped T-network input circuit. The amplifier reserve. For operation at 2000 watts PEP the plate voltage should be between 2500 and 3000 volts; under these conditions the amplifier will deliver 1240 watts



C1	1000 pF (Centralab 858S-1000)	L1	5 turns no. 14, 3/4" long on 1/2" diameter form with white tuning slug (CTC 1538-4-3)
C2	25 pF variable (Hammarlund HFA-25B)	L2	4 turns no. 14, air wound, 3/4" long
C3,C4	each consists of two parallel-connected 100-pF, 5000-V capacitors (Centralab 850S-100)	L3,L4	10 turns no. 12 enameled, bifilar wound, 5/8" diameter
C5	plate-tuning capacitor (see fig. 3)	L5,L6	plate resonators (see fig. 4)
C6	output loading capacitor (see fig. 2)	L7	7 turns no. 14 wire, 5/8" diameter, 1-3/8" long
C7	1000 pF, 4 kV feedthrough (Erie 2498)		

fig. 1. Schematic for the grounded-grid two-meter triode amplifier. Operating bias for the 8877 is supplied by a 12-volt zener diode in the cathode lead.

requires no neutralization, is completely stable and free of parasitics, and is very easy to operate.

This amplifier is designed for continuous duty operation at the 1000-watt dc input level, and can develop 2000 watts PEP input for ssb operation with ample output. With the higher plate-voltage supply, up to 13.8-dB gain can be obtained with an amplifier efficiency of 62%.

the circuit

In the amplifier circuit in **fig. 1** the 8877 grid is operated at dc ground. The grid ring at the base of the tube provides a low-inductance path between the grid element and the chassis. Plate and grid currents are measured in the cathode-return lead; a 12-volt, 50-watt zener diode in series with the negative return ly burn open.

input circuit

The cathode input matching circuit is a T-network which matches a 50 ohm termination to the input impedance of

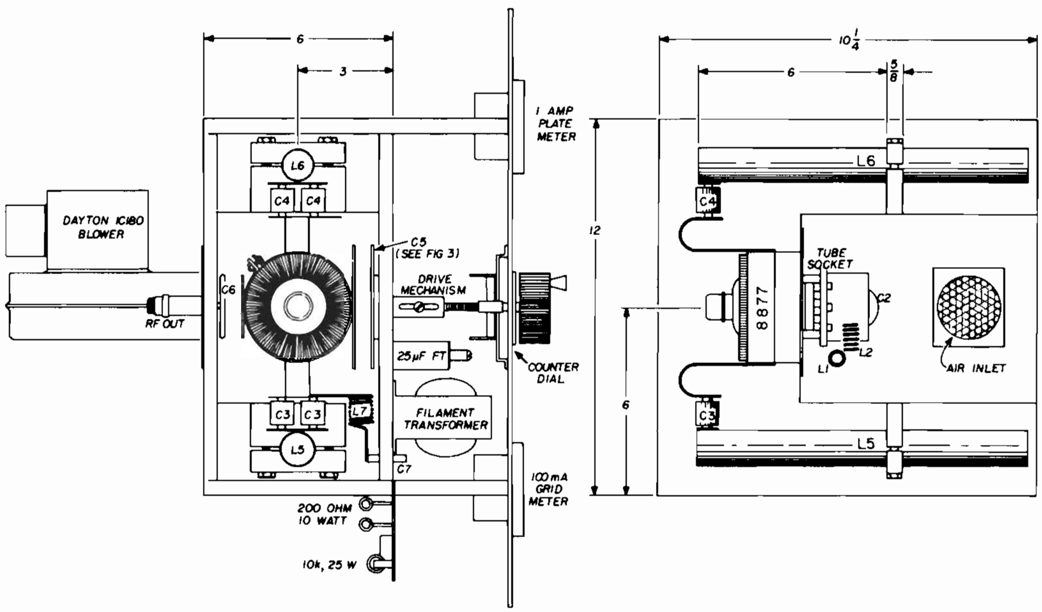


fig. 2. Structural details of the amplifier show the relative size and position of the various components. Assembly is made of aluminum panels.

sets the desired value of idling current. Two additional diodes are shunted across the meter circuit to protect the instruments.

Standby plate current of the 8877 is reduced to a very low value by the 10,000-ohm cathode resistor; this resistor is shorted out when the vox circuit is energized, permitting the tube to operate in normal fashion.

A 200-ohm safety resistor insures that the negative power circuit of the amplifier does not rise above ground potential if the positive side of the plate-voltage supply is accidentally grounded. A second safety resistor across the 1N3311 zener diode prevents the cathode potential from rising if the zener should accidental-

the tube (about 54 ohms in parallel with 26 pF). The network consists of two series-connected inductors and a shunt capacitor. One inductor and the capacitor are variable so the network is able to cover a wide range of impedance transformations.

The variable inductor (L1) is mounted on the rear wall of the chassis and may be adjusted from the rear of the amplifier. The input tuning capacitor (C2) is adjustable from the front panel. When the network has been properly tuned no adjustment is required over the 4-MHz range of the 2-meter band.

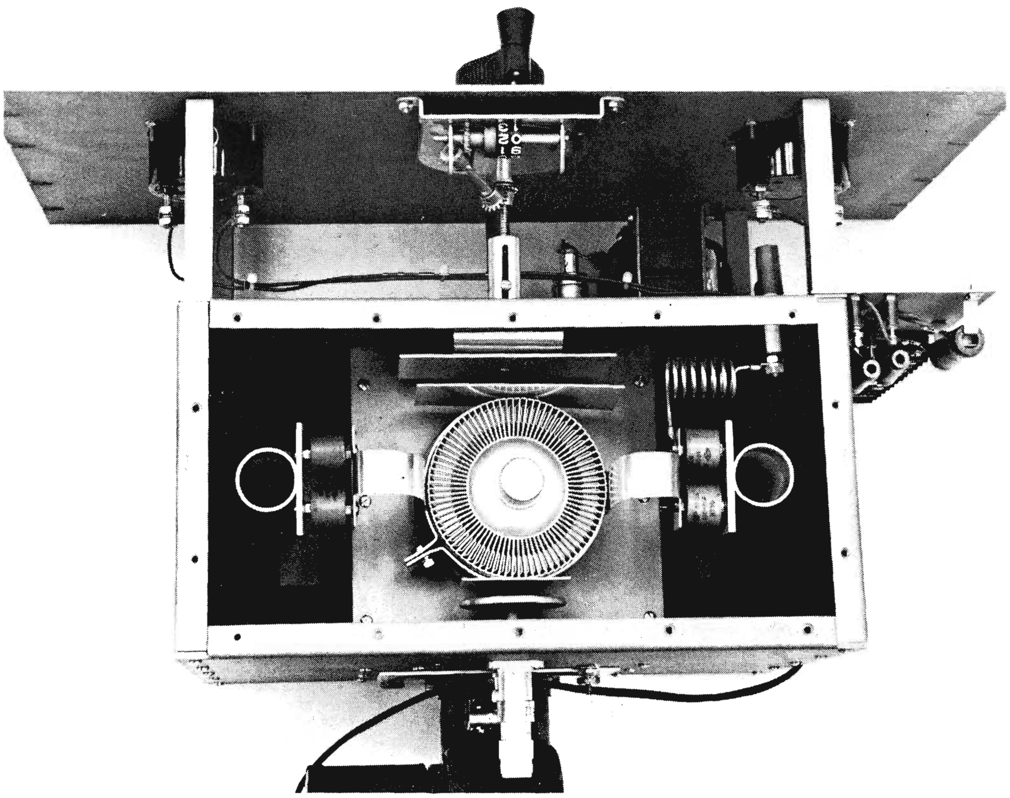
Underchassis layout of components is shown in the photograph. The cathode input circuit is in center compartment.

The slug-tuned coil in the input matching circuit is mounted on the rear wall. Air-wound filament chokes are placed in front of the socket. The cathode-heater rf choke is near the top edge of the enclosure. All of the cathode leads of the

mechanism are placed in the area between inclosure and panel.

plate circuit

The plate circuit of the amplifier is a transmission-line type resonator. The line



Top view of amplifier showing plate compartment. 8877 tube is at center with plate lines on each side.

socket, plus one heater pin (pin 5) are connected in parallel and driven by the input matching network.

The ceramic socket for the 8877 is mounted one-half inch below chassis level by spacers. Four pieces of brass shim stock (or beryllium copper) are formed into grounding clips to make contact to the control grid ring. The clips are mounted between the spacers and the chassis. The aluminum clamps holding ends of plate lines are visible in the side compartments. The filament transformer and dial

(L5 plus L6) is one half-wavelength long with the tube placed at the center (**fig. 2**). This type of tuned circuit has several advantages. A quarter-wave circuit would normally be preferred because of its greater bandwith, but I wanted to use easily obtainable standard copper water pipe as the center conductor of the transmissionline tank circuit. The resulting high-impedance transmission line would make a quarter-wave plate tank circuit physically short and difficult to handle.

In addition, the heavy rf current that flows on the tube seals and control grid would, in the process of charging up the output capacitance to the plate voltage swing, tend to concentrate on one side of the tube if a single-ended quarter-wave circuit were used. This current concentration would cause localized heating of the tube. The best tuned circuit configuration to minimize this effect is a symmetrical cylindrical coaxial cavity. Unfortunately, this type of cavity is complex and difficult to build.

A practical compromise is to use two quarter-wave lines connecting to opposite sides of the tube. It is interesting to note that each of the two quarter-wave lines is physically longer than if only one quarter-wave line were used. This is because only one-half of the tube output capacitance loads each of the two lines.

Resonance is established by a moving

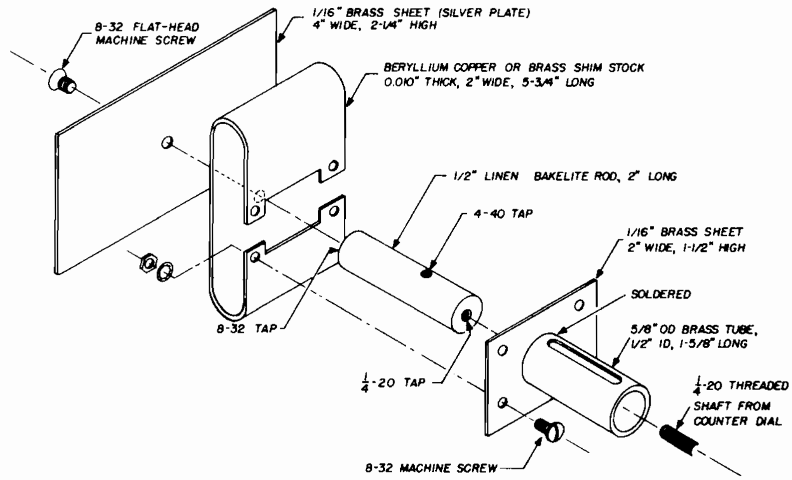


fig. 3. Variable plate portion of plate-tuning capacitor C3. This arrangement permits the capacitor to be adjusted under full power without "jumpy" tuning as there are no moving or sliding contacts which carry heavy rf current.

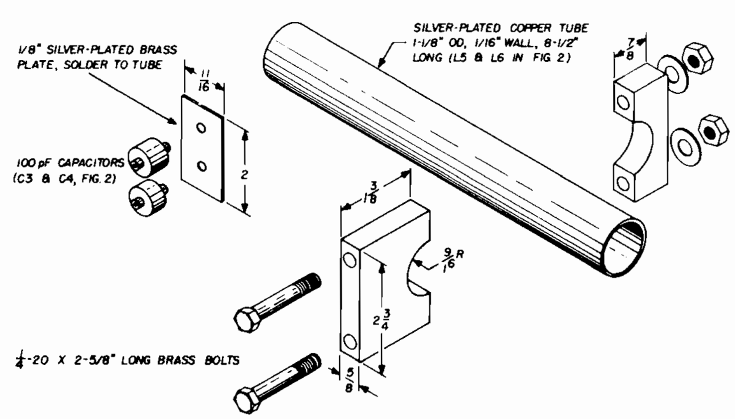


fig. 4. Details of plate lines L5 and L6. Copper tubes are standard copper water pipe.

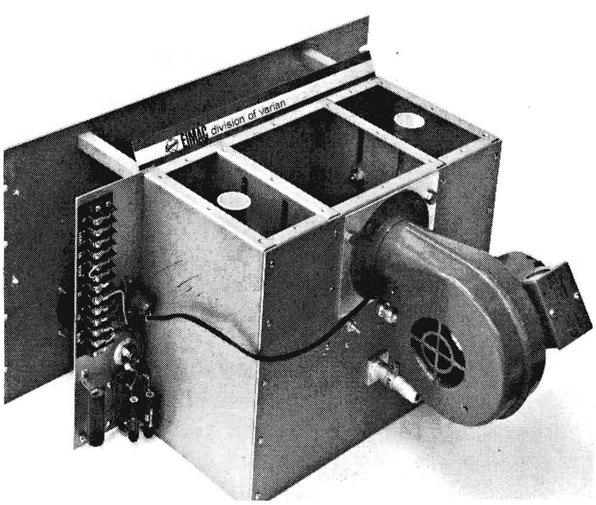
plate capacitor (C5); antenna loading is accomplished by a second capacitor (C6) placed at the anode of the 8877. Output power is coupled from the plate circuit through the series capacitor into a 50-ohm output. In the top-view photo tuning capacitor C_5 is at the front of the compartment; variable loading capacitor C_6 is at the rear. The plate choke is visible in the front corner.

construction

The two-meter amplifier is built in an enclosure measuring 10¼ x 12 x 6¼ inches. The 8877 socket is centered on a 6 x 6 subchassis plate. A centrifugal blower forces cooling air into the under-chassis area; the air escapes through the 2-5/8-inch diameter socket hole.

The plate tuning mechanism is shown in **fig. 3.** This simple apparatus will operate with any variable plate capacitor, providing a back and forth movement of about one inch. It is driven by a counter

Rear of amplifier showing blower and coaxial output connector. Amplifier is upside down in this photograph.



dial and provides a quick inexpensive and easy means of driving a vhf capacitor. The ground return path for the grounded capacitor plate is through a wide low-inductance beryllium-copper or brass shim stock which provides spring tension for the drive mechanism.

The variable output coupling capacitor

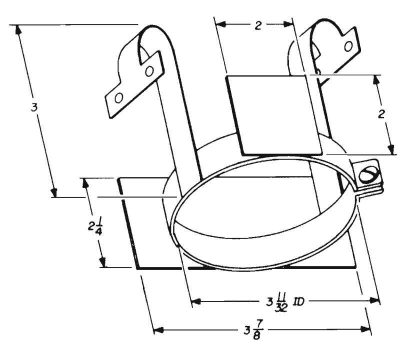


fig. 5. Anode clamp assembly for the two-meter linear amplifier.

is located at the side of the 8877 anode. The type-N coaxial fitting is connected to the moveable plate of the coupling capacitor. The fitting is centered in a special tubular assembly which allows the whole connector to slide in and out of the chassis, allowing the variable plate of the coupling capacitor to move with respect to the fixed plate mounted on the tube anode clamp. When the final loading adjustment has been set the sliding fitting is clamped by an arrangement similar to the slider on a variable wire-wound resistor.

The length of the plate-line inductors (L5 and L6) is adjusted by means of dural blocks placed at the shorted end of the line (**fig. 4**). The position of the blocks is determined by setting capacitor C5 at its lowest value and adjusting line lengths so that that plate circuit resonates at 148 MHz with the 8877 in the socket.

The plate rf choke is mounted between the junction of one plate strap and a pair of the dual blocking capacitors; the high-voltage feed-through capacitor is mounted to the front wall of the plate compartment. The blocking capacitors are rated for rf service, and inexpensive tv-type capacitors are not recommended for this amplifier. A short chimney to direct cooling air from the socket through the anode of the 8877 is made from Teflon and clamped between the chassis deck and the anode strap.*

operation

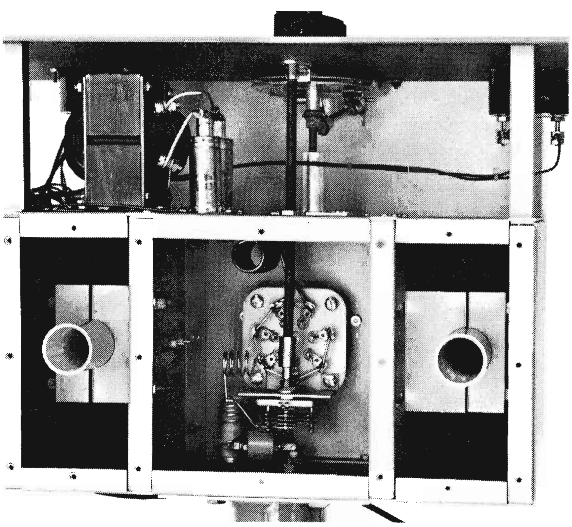
Amplifier operation is completely stable with no parasitics. The unit tunes up exactly as if it were on the "dc bands." As with all grounded-grid amplifiers excitation should never be applied when plate voltage is removed from the amplifier.

The first step is to grid-dip the input and output circuits to near-resonance with the 8877 in the socket. A swr meter should be placed in series with the input line so the input network may be adjusted for lowest swr.

Tuning and loading follows the same sequence as any standard grounded-grid amplifier. Connect a swr indicator at the output and apply a small amount of rf drive. Quickly tune the plate circuit to resonance. The cathode circuit should now be resonated. The swr between the exciter and the amplifier will not necessarily be optimum. Final adjustment of the cathode circuit for minimum swr should be done at full power because the input impedance of a cathode-driven amplifier is a function of the plate current of the tube.

table 1. Performance data for the 144-MHz power amplifier under the conditions most suitable for amateur ssb (2000 watts PEP) and cw (1000 watts).

Plate voltage	3000 V	2500 V	2500 V
Plate current (single tone)	667 mA	800 mA	400 mA
Plate current (idling)	54 mA	44 mA	44 mA
Grid voltage	-12 V	-12 V	-12 V
Grid current (single tone)	46 mA	50 MA	28 mA
Power input	2000 W	2000 W	1000 W
Power output	1240 W	1230 W	680 W
Efficiency (apparent)	62 %	62 %	68 %
Drive power	47 W	67 W	19 W
Power gain	13.8 dB	12.6 dB	15.5 dB



Underchassis view of the two-meter amplifier. The cathode input circuit is in the center compartment. Plate lines are visible in the side compartments.

Increase the rf drive in small increments along with output coupling until the desired power level is reached. By adjusting the drive and loading together it will be possible to attain the operating conditions given in the performance chart in **table 1.** Always tune for maximum plate efficiency: maximum output power for minimum input power. It is quite easy to load heavily and underdrive to get the desired power input but power output will be down if this is done.

I would like to thank K6DC for his help in adjusting and determining the operating conditions for this two-meter amplifier.

*Detailed drawings of the anode clamp, plate resonator and blocking capacitor assembly, and variable plate tuning capacitor (C5) are available from R. Sutherland, EIMAC Division of Varian, 301 Industrial Way, San Carlos, California 94070. Ask for drawing numbers 168658, 168648 and 168647.

references

1. R. Sutherland, W6UOV, "Two Kilowatt Linear Amplifier for Six Meters," *ham radio,* February, 1971, page 16.
2. R. Barber, R. Rinaudo, W. Orr, R. Sutherland, "Modern Circuit Design for VHF Transmitters," *CQ,* November, December, 1965.

ham radio

A 144-MHz Amplifier Using the 8874

BY RAYMOND F. RINAUDO,* W6ZO

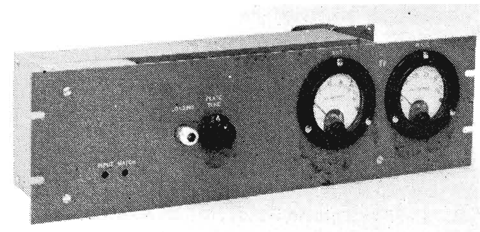


Fig. 1 – Front view of the 144-MHz amplifier. The grid-current meter is on the left and plate meter on the right. Plate-circuit loading is increased by pushing in on the coax connector. Input circuit tuning is done with an insulated screwdriver through the two holes at the left side of the panel labeled INPUT and MATCH.

THIS 144-MHz amplifier is an inverted ultra-audion, grounded-grid, or a cathode-driven amplifier, depending upon what point in electronics history you choose to speak from. The first description has now been dropped and is probably recognized only by the real old timers, or students of radio history.[1] The second, grounded grid, is still widely used but somehow fails to describe how an amplifier works; it also implies no grid bias, but bias is often used. The third, by its words, cathode driven, tells you how the amplifier operates, without being unduly restrictive as to operating voltages.

The Cathode-Driven Amplifier and Tube

The cathode-driven amplifier has become very popular with amateurs using the high-frequency bands. This popularity has developed primarily in the past fifteen years. One of the principal reasons is the availability of single-sideband exciters with a PEP output of one hundred or more watts. With exciters of that power capability, the natural step in increasing power level is to go to an amplifier which will absorb all, or almost all, of the exciter power – the cathode-driven amplifier. Also in the past few years, new modern-design tubes have been introduced which were developed for this service, having a very high mutual conductance and operating with very little or no grid bias. The best of these tubes operate satisfactorily at the lower vhf range, but are difficult to handle in the upper part of that range.

Now a new family of tubes has been developed which performs very well at hf, vhf, and well into the ultra-high-frequency range. These tubes are the 8873, 8874, and 8875. The tubes differ from each other only in the anode cooler construction. The 8873 is intended for heat-sink cooling, the 8874 for forced-air cooling, and the 8875 for cooling by large volume but very low-pressure air, such as is supplied by a fan. The 8874 is used in the amplifier which is to be described.

The requirements for a good cathode-driven tube will be briefly reviewed.

1) In order to have acceptable gain, the tube should have high mutual conductance.

2) The tube should have low grid interception of electrons. All other things being equal, the tube having the lowest grid interception is the easiest to drive.

3) The tube should have the least possible inductance between the grid in the tube and the external grid connection. Inductance in the grid lead causes degeneration which, in turn, means that more drive power has to be supplied to drive the tube to a particular plate current.

Incidentally, the long wire grid lead in the old glass high-mu triodes is the main reason for lack of vhf capability of those tubes. It is of interest to note that while the cathode-driven tubes require low grid-lead inductance, the grid-driven tube requires low cathode-lead inductance; lack of attention to this detail results in a tube which is hard to drive, and, for the same reason in both cases, degeneration.

The 4X150A/4CX250B tetrode tubes are still considered very good performers in the vhf range. However, the 8873 triode family, in cathode-driven service, gives power gains which approach that of the older tetrode types in a grid-driven arrangement and does that without the necessity for a screen supply or neutralization; and at the lower

* Eimac Division of Varian, 301 Industrial Way, San Carlos, CA 94070.

[1] It is interesting to note that a cathode-driven amplifier was described in *QST* almost forty years ago. See *QST* for September, 1933, "The Inverted Ultraudion Amplifier," by Hugo Romander, W2NB (now W6CH).

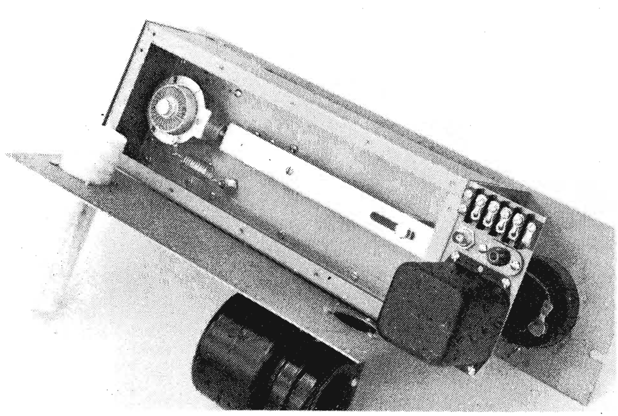


Fig. 2 – Looking into the box containing the plate-circuit strip line. The slot in the plate line is for a rough adjustment of frequency. Once set properly, all of the 2-meter band can be covered using only the front-panel controls. The Teflon chimney and wave-guide-beyond-cutoff vent pipe are attached to the box cover.

plate voltages the tube can be operated at zero bias. If fixed bias is needed, it can be obtained from a comparatively inexpensive Zener diode.

The 2-meter amplifier to be described uses one 8874 in a cathode-driven circuit and with a strip-line plate circuit. It is capable of 1100-watts PEP input for ssb suppressed carrier and 550 watts for cw. As an a-m linear amplifier it will run at 500-watts input.

Construction

The amplifier is built so as to fit behind a standard 5 1/4 × 19-in. panel as shown in Fig. 1. The plate circuit enclosure is 13 × 5 × 3 in. and made of aluminum (see Fig. 2). This one happens to be of the hand-made variety, but a standard chassis of this size could have been used and would have saved some time. The cathode input circuit is in a 5 × 3 3/8 × 1-in. aluminum box. A standard 4 1/2 × 3 1/2 × 1-in. chassis would have served as well, as the box is not crowded. Two end brackets space the rf unit 1 3/4 in. behind the panel to allow room for the meters.

The tube socket is centered between the two sides of the plate-circuit enclosure and is 1 1/2 in. from one end. The grid is connected directly to the chassis by a grid collet. The grid collet was made by soldering a grid contact ring, Eimac part No. 882931, to a 1/16-in.-thick brass ring. The brass ring has three No. 6-32 stud bolts attached which match the location of the three mounting holes of the Johnson 124-311-100 socket. There are other ways to make a good grid collet, and ingenuity of the individual builder can assert itself. Certainly the multiple contacts that finger stock gives is the type of thing needed. Never depend on the control-grid socket connections (pins 4, 7, and 11) to be good enough for vhf or uhf service.

Contact to the anode of the tube is made by a plate collet. The collet was made by sandwiching an Eimac plate contact ring, part No. 008294, between two 1/16-in.-thick brass pieces. The upper piece is circular with a 1-in.-wide tab on one side, the tab having a 3/8-in. lip bent at 90 degrees for the plate blocking-capacitor mounting. The lower brass piece is also circular, but without the tab. The plate collet is shown in Fig. 3.

The plate line is made of copper, 1/8-in. thick and 1-in. wide. A 3/8-in. lip is bent at the tube end of the line for connecting to the plate blocking capacitor. The far end of the line stops about 1/4-in. short of the enclosure wall. The plate line is supported near the far end by a 1 × 1 × 1/2-in. block of copper, and by a 1-in.-high ceramic insulator 5 1/4-in. from the tube center. A 1/4-in. bolt passes through the plate-line slot, the support block and a slot in the enclosure wall. The exact location of the ceramic insulator is not critical, except that it must not interfere with the location of the plate tuning and coupling capacitors which are between it and the tube! Both copper and soft aluminum have been used for the plate line and the 1 × 1 × 1/2-in. support block. Very careful power measurements showed no difference in performance, whichever material was used. Brass might be satisfactory, if it is silver plated. For those willing to experiment, 1/16-in.-thick material might be used instead of 1/8-in., as in this amplifier. The line would then have to be slightly shorter to tune to resonance.

Rf power is taken out of the amplifier by capacitance coupling to the plate line. A 1-in.-dia disk is positioned near the tube end of the line and coupling is varied by a sliding arrangement. Details of this coupling device are shown by Fig. 6.

Plate tuning is done with a 1-in.-dia disk on a 1/4-in. threaded shaft which positions the disk relative to the plate line. The shaft turns in a threaded bearing mounted on the enclosure wall. When the plate-line support block is properly set, it is possible to tune the amplifier from 144 to 148 MHz without changing the position of the block. The plate tuning capacitor is spring loaded to prevent a variable ground-return path, sometimes encountered in this kind of device. The spring loader is made by soldering a threaded nut of the correct size to a piece of thin metal; brass or steel will do. This is then threaded on the turning shaft until it rests on two pillars mounted on the outside enclosure wall. Just enough compression to prevent play in the tuning shaft gives the best results. This device can be seen just to the right of the output coaxial connector in Fig. 4.

The tube anode is cooled by a blower mounted on the plate-circuit enclosure cover. The only escape for the air is through the tube anode cooler, the chimney, and then the vent pipe. The chimney used in this amplifier was made from sheet Teflon approximately .050-in. thick, formed into a cylinder and then taped to keep that shape. Since this material is not readily available, a suitable arrangement can be worked out using the standard ceramic chimney for the 4X150 or 4CX250, such as the Eimac SK-606. The idea is to prevent the air from going directly to the vent pipe without going through the tube anode.

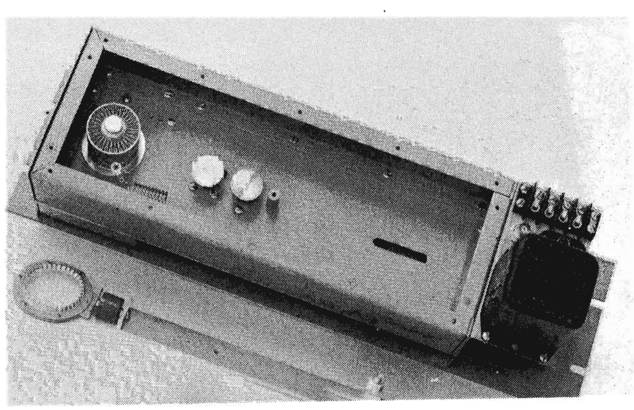


Fig. 3 — Plate-line box with the strip line removed. The output coupling capacitor is the disk nearer the tube. The other disk is the plate-tuning capacitor. A one-inch-high ceramic pillar supports the plate line and is located next to the tuning capacitor.

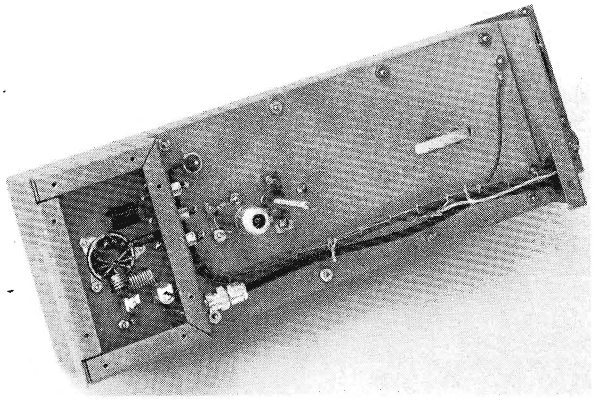


Fig. 4 — The front of the amplifier with the panel removed. The input-circuit enclosure which is to the left is normally covered with a piece of perforated aluminum. The spring-compression device which prevents erratic tuning of the plate-circuit resonating capacitor is shown just to the right of the type N output coax connector.

Circuitry

In cathode-driven service the average input impedance of the 8874 is approximately 95 ohms. An input circuit, therefore, is needed for two reasons: (1) the input to the amplifier should look like 50 ohms to accommodate the transmission lines and the driver output impedance; this calls for an impedance transformer; and (2) the 95-ohm average input impedance varies tremendously during the rf cycle. For example, if the tube is operated Class B, the plate current is zero during half the rf cycle and the impedance is very high, but during the part of the cycle that the tube peak currents occur, the impedance is lower than 95 ohms. Since drivers don't like to have the load impedance varying wildly, a storage reservoir should be provided. A tuned circuit with some Q serves this function and also transforms the impedance.

The cathode input circuit is an L-pi network using lumped constants. It was designed to have a loaded Q of 3. Knobs were not provided for adjusting the input circuit because of its broadband nature. Instead, screwdriver slots in the capacitor shafts permit adjustment of the capacitors from outside the box. The "screwdriver" should be nonmetallic. A short 1/4-in.-dia Bakelite shaft filed in the shape of a screwdriver on one end and with a knob on the other end works very well. If the input circuit is matched at 146 MHz, changing frequency to either 144 or 148 MHz without touching the input adjustments will give an SWR of less than 1.7 to 1. At either 145 or 147 MHz the SWR is less than 1.35 to 1. Of course the input can be matched at any frequency within the band, if desired.

Fig. 5 shows tha layout of parts in the input-circuit compartment. All of the six cathode terminals of the socket are connected together. The bifilar heater choke is wound on a small length of insulating rod and was made bifilar for the convenience of using only one form instead of two. Sharp-eyed readers who like to count turns will note that the outside layer of this coil has one less turn than the one on the inside. Again, it was a matter of convenience, not a requirement.

Although the pi-L or L-pi and T networks have been around a long time, the T network has not been used much. The T is useful where low-impedance transformations are necessary. The solid-state designers have this sort of problem in their work and have developed tables of solutions for a wide variety of impedance transformations.[4] A copy of these tables is very useful, indeed.

All of the power leads into the rf compartments enter via feedthrough bypass capacitors. They help keep the rf where it belongs. Three are low-voltage types, which feed the cathode and filament circuits. The fourth is a high-voltage unit for the plate voltage supplied to the tube.

Cooling the Tube

The cooling of the tube is done somewhat differently than is usually the case. The tube data sheet, for 400-watts plate dissipation, specifies a minimum air flow of 8.6 ft^3/min for 50 degrees C at sea level (you can use a little less if the air is cooler than 50 degrees C). The pressure required to move that quantity of air through the tube cooler is 0.37 inches of water, assuming no back pressure on the discharge side of the anode. This cooling air is provided by a Dayton model 2C782 blower.[5] The blower forces air into the box containing the tube plate circuit. A Teflon chimney connects the top of the tube anode to the vent pipe fastened on the box cover. All other holes in the plate-circuit box are sealed, so that the air blown into the box

[4] *Matching Network Designs with Computer Solutions*, Application Note AN-267, Motorola, Box 20912, Phoenix, AZ 85036.

[5] Available from W. W. Grainger, Inc., 2750 W. Fulton St., Chicago, IL 60612.

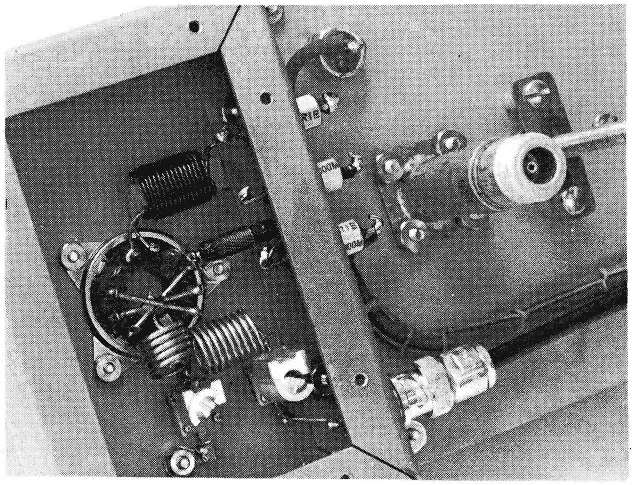


Fig. 5 — Close-up view of the input-circuit box. The bifilar-wound filament choke is above. The cathode choke is to the right and the input T-network coils and capacitors are below the tube socket. The slide mounting of the type N output coupling capacitor is shown to the right of the input-circuit enclosure.

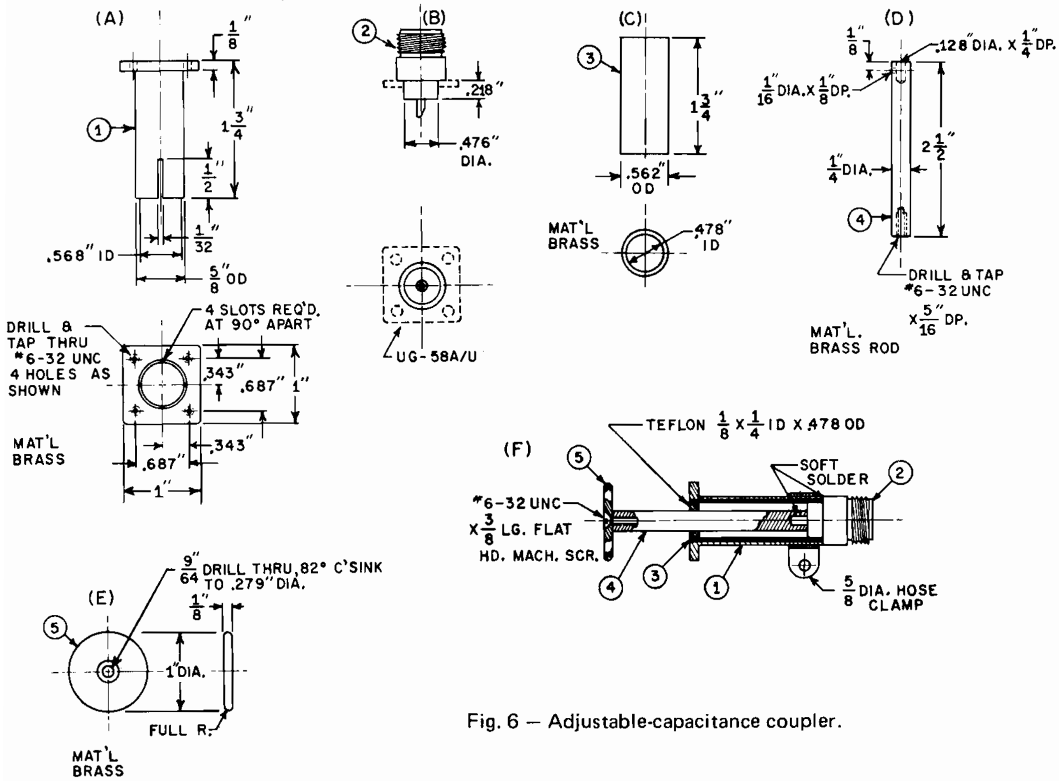


Fig. 6 — Adjustable-capacitance coupler.

can only escape by going through the tube anode cooling fins and then to the outside world. The vent pipe has practically no air restriction, yet acts as a very effective choke at these radio frequencies.

Equipment built for amateur service has not often taken advantage of the particular principle of microwave radio by which the vent pipe works – electronically, that is. In this case it is used for what it will *not* do, rather than for what it *will* do; that is, it will not transmit a radio signal of any frequency below about 4600 MHz. The pipe has a diameter of 1 1/2 in. For frequencies above 4600 MHz this would be an excellent transmission line. However, for frequencies much below 4600 MHz, the cutoff frequency, it becomes a choke. Its effectiveness as a choke depends upon how long it is. The usual description of this phenomenon is "wave guide beyond cutoff"; in this case "beyond" means below! The attenuation of this device is given by the formula:[6]

$$\text{Aa} = \text{Aperture attenuation (dB)} = 32\frac{D}{d}$$

D = length of pipe

d = inside diameter of pipe

The vent pipe used in this amplifier is 4 1/2-in. long and 1 1/2-in. ID. When these figures are put into the formula, this gives an attenuation of 96 dB! This attenuation is for frequencies well below the cutoff frequency, not for those close by. However, the 144-MHz output of the amplifier and its harmonics up to about the 20th can be considered well below cutoff.

6 *Electrical Design News*, October, 1963.

Tune-up Procedure

Tuning a cathode-driven amplifier is not much different than tuning one which is grid driven. There is one precaution that must be observed, though. Never run drive power into a cathode-driven amplifier unless the plate voltage is on. Running normal drive power with no plate voltage produces high grid dissipation and will quickly destroy the tube.

The 8874 has an indirectly heated cathode. Always allow at least 90 seconds for the heater to warm up before the tube is required to draw plate current.

When tuning a new amplifier for the first time, it is very helpful to start with reduced plate voltage. If the final operating conditions with this amplifier are to be 2000 V at 500 mA, then apply 1000 V to the plate and only enough drive to be able to tune the plate circuit to resonance. Then increase drive and adjust plate loading and tuning to maximize power output, while watching the grid meter to make sure that the current is not excessive, indicating high grid dissipation. When the drive, plate loading and tuning are optimized for maximum power output at 1000 V and 250-mA plate current, then the loading adjustments will be very nearly correct for the 2000-V 500-mA condition. Before going to the higher plate voltage, the cathode input circuit can be adjusted for a low input SWR. Do not waste time matching the input circuit exactly, as the input match will change somewhat at the higher power level.

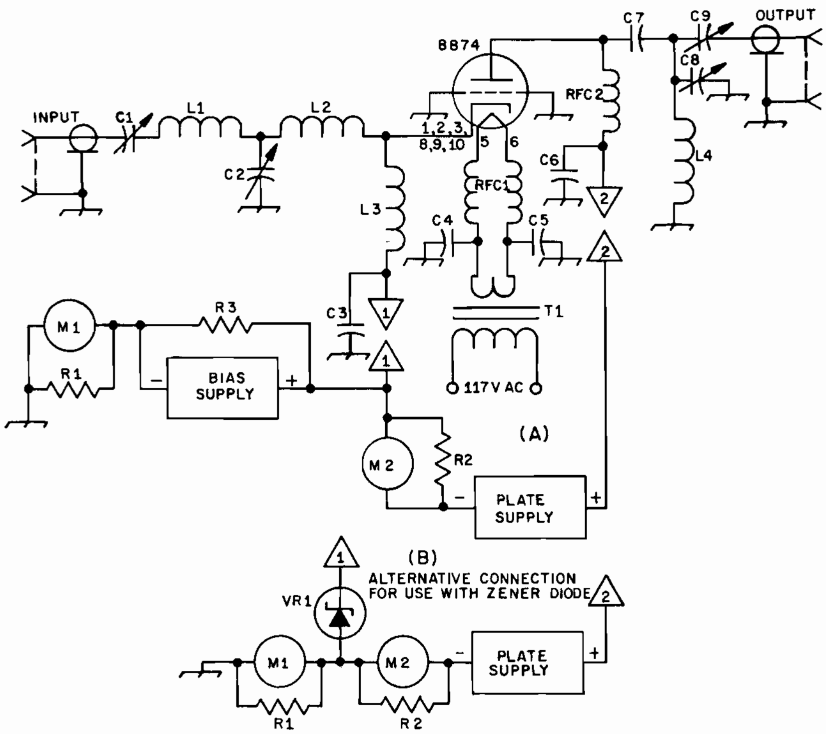


Fig. 7 — 144- to 148-MHz 8874 amplifier.

C1 — 2.4 to 24.5 pF air variable (E. F. Johnson 189-509-4).
C2 — 1.7 to 14.1 pF air variable (E. F. Johnson 189-505-4).
C3, C4, C5 — 1000-pF feedthrough (Erie 327-005-X5UO-102M or equiv.).
C6 — 1000-pF feedthrough, 2.5 kV (Erie 1270-10 or equiv.).
C7 — 100 pF, 5 kV (Centralab 8505-100N or equiv.).
C8, C9 — See text.
L1 — 9 turns No. 16 wire, 3/8-in. dia, 5/8-in. long.
L2 — 5 turns No. 16 wire, 3/8-in. dia, 3/8-in. long.
L3 — 0.47-μH choke, 300 mA (Miller 4588 or equiv.).
L4 — Plate line; 1 X 1/8-in. copper or soft aluminum, 8 1/2-in. long.
M1 — 50-mA dc meter.
M2 — 500-mA dc meter.
R1, R2 — 50 ohms, 10 W.
R3 — Bias-supply bleeder; select for approx. 100 mA current and appropriate wattage.
RFC1 — Bifilar wound, each winding 15 turns on 3/8-in.-dia form. See text and photos.
RFC2 — 10 turns No. 16 wire, 1/2-in. dia, 1 1/8-in. long.
T1 — Filament trans., 6.3-V 4-A secondary (Triad F-43X or equiv.).
VR1 — See text.

Now the plate voltage can be raised to 2000 V and drive power increased to give about 500-mA plate current. Again, plate loading and tuning and drive power are adjusted for maximum power output at the desired power level. And of course the grid-current meter is watched to avoid excessive grid dissipation. When the plate-circuit adjustments have been completed, then the input circuit can be adjusted for minimum SWR. If the drive power is free of harmonics or other spurious signals, an SWR of better than 1.1 to 1 is easily obtained on the drive line.

Of course it is not necessary to tune up for the first time with reduced plate voltage. However, if you start off at 2000 volts, then a great deal more care must be used. Unwanted or unexpected effects happen a lot faster and with more violence at 2000 volts than at 1000!

The preceding tune-up procedure applies to either cw or ssb linear operation. For a-m, the method would be the same provided that loading is adjusted for the most output at the crest of the modulation cycle when the drive signal is being 100-percent modulated. A peak-voltage-responding indicator is needed on the output of the amplifier to do that. An rf vacuum-tube voltmeter is one of the instruments that can be used.

A second method that can be used is to adjust the amplifier loading and tuning for minimum distortion when the drive signal is being 100-percent modulated. Still another method that will produce the correct loading adjustment is to simulate the conditions that exist at the crest of the modulation cycle, but by using a cw signal only. For example, if a driver with a 5-watt carrier is to be used, it will have a peak output of 20 watts

Table I

Typical Performance								
Plate Voltage	*No - Signal Plate Current*	*Grid Bias*	*Plate Current*	*Grid Current*	*Drive Power*	*Power Output*	*Efficy.*	*Gain*
1000	10 mA	–4.6V	250 mA	24 mA	7.2W	146W	58%	13.1 dB
1500	10	–7	250	25	8.4	238	63	14.5
2000	10	–9.8	250	22	9.4	320	64	15.3
1000	92	0	300	26	7.0	171	57	13.9
1500	140	0	400	27	11.4	362	60	15
2000	150	–1.7	500	29	18.3	636	64	15.4

at the crest of the modulation cycle. Therefore a cw signal of 20 watts can be used as a driver temporarily while the linear amplifier is being adjusted for maximum power output, just as it is for cw or ssb operation. After the adjustments are completed the 5-watt driver can replace the 20-watt unit, and the amplifier is ready to go.

The important point to remember about an a-m linear amplifier is that the amplifier must be adjusted to accommodate the peak power conditions. One of the most difficult facts to accept is that amplifier settings which give the best and most readable signal are not the same as those which give the strongest carrier. After the amplifier has been correctly tuned, one is always tempted to touch up the tuning to get more carrier output. When the temptation is great, reread the paragraphs on a-m linear-amplifier adjustment!

You should now have amplifier performance which is very nearly like that shown in Table I. Because of variations from tube to tube, the grid current at a particular operating level may vary considerably from that shown in the table, possibly as much as 2 to 1. Drive power requirements may also vary, but to a smaller degree. This may be lesser or greater by 10 or 20 percent.

Performance

The amplifier can be used for cw or as a linear amplifier for either ssb or a-m work. The mode of operation will determine the maximum input power that can be used. Table I shows the measured performance of the amplifier with different plate voltages and idling plate currents.

Table II

Type 8874 Tube Data	
Heater Voltage	6.3 V
Heater Current	3.2 A
Maximum Plate Voltage	2200 V
Maximum dc Plate Current (key down)	250 mA
Maximum dc Plate Current (modulation crest)	500 mA
Maximum Plate Dissipation	400 W
Maximum Grid Dissipation	5 W

Some conclusions can be reached from the data shown in Table I. The table shows that higher efficiencies and gain are possible with higher plate voltage and, of course, higher power output. Also, the lower the idling plate current, the lower the gain, but efficiency goes up.

For linear-amplifier service, good results can be obtained if the tube bias is set for an idling plate current of 25 to 50 mA. A Zener diode in the cathode return is probably the simplest way to get the needed bias. However, a variable low-voltage supply can be used and has the advantage of flexibility if much experimenting is to be done. A suitable circuit for such a supply is shown in the 1971 ARRL *Handbook.*

Other construction techniques such as coaxial resonant cavities can be used to build a good amplifier. What has been described is a fairly simple, easy-to-build unit. The 8873 family of tubes will do a good job at vhf and uhf in a well-designed amplifier. QST

The Simulmonitor

(Continued from page 23)

A commercial transistorized amplifier suitable for Z1 is available at very nominal cost. It has a diode and filter capacitor, for rectification and filtering, so it is necessary to power the unit with only the 6.3 volts ac from the filament supply. This amplifier provides the local audio, and drives a local speaker for monitoring purposes during maintenance. Amplifier Z2 is similar. It provides the needed line level from the tone oscillator and the microphone.

A jumper shown as a broken line between Terminals 5 and 9 of K2 determines whether a tone burst is received only when the primary receiver is operative, or when either primary or secondary receiver is operative. This is the first step toward a future control system.

This, then, is the Simulmonitor, as conceived and utilized by CFAR. As channel occupancy increases and more satellite receiving setups come into the system, the approach becomes more sophisticated, and the Simulmonitor may change almost beyond recognition. In other areas the technique is used under other names and with other features, but the basic concept will most likely remain, and provide many benefits. The net result will be further advancement in the story of amateur radio repeaters. The story is just now starting to be told. QST

RADIO FREQUENCY

INTERFERENCE TASK GROUP

DIVISION OF VARIAN
301 Industrial Way
San Carlos, California

AS-50

4163

THE AMERICAN RADIO RELAY LEAGUE, INC.

ADMINISTRATIVE HEADQUARTERS NEWINGTON, CONNECTICUT, U. S. A. 06111

RADIO FREQUENCY INTERFERENCE TASK GROUP

Dear Radio Amateur,

This letter and its enclosures are supplied in response to your request for information on radio-frequency interference (RFI). Assembled by the American Radio Relay League (ARRL) RFI Task Group, the material is intended to assist you in providing your neighbors with factual information on RFI, and to suggest steps you and your neighbors can take to resolve RFI problems.

In 1973 the Federal Communications Commission (FCC) received 42,000 RFI complaints. Of these, 38,000 involved interference to home-entertainment equipment. And perhaps most important, 34,000 of these 38,000 complaints would never have come to the FCC's attention if the manufacturers had corrected design deficiencies in their home-entertainment products at the time of manufacture! Miriam Ottenberg's article in the Washington Star-News (Enclosure 1) provides a good overview of the RFI problem, and will make interesting reading for both you and your neighbors.

The FCC's experience with RFI, of course, goes back many years, and as seen in Enclosure 2, they have found over the years that 90% of all television interference (TVI) problems can be cured only at the television receiver. Certainly, if your own television receiver experiences no interference while you are on the air, it is most likely that interference to a more distant television receiver is not the fault of your transmitter. Further, when it comes to audio equipment (phonographs, Hi-Fi's, electronic organs, intercoms, etc.), the only cure for RFI is by treatment of the audio device experiencing the interference (Enclosure 3). After all, there's nothing a radio operator can do to his transmitter which will stop a neighbor's phonograph from acting like a short-wave receiver.

It is clear, therefore, that almost all RFI problems experienced with home-entertainment equipment result from basic design deficiencies. The few small components or filters which would prevent RFI are often left out by the manufacturers in their attempts to reduce costs, and hence, to reduce the prices on equipment. Manufacturers have asserted that "since only 1% of those who purchase their products will experience RFI, it is not necessary to design equipment so that it will be less susceptible to interference."

That only 1% of the population experiences RFI, however, is clearly an underestimation. The tremendous growth in radio communications over the past decade has brought increasing numbers of transmitters in close proximity to home-entertainment equipment. An example of this growth is the more than one million transmitters now operating on frequencies assigned to the Citizen's Radio Service. Most of these stations are in homes, and thus, are close to television receivers, radios and audio equipment used for home entertainment. As an example of the high density of radio transmitters to be found in urban and suburban areas today, take a look at the enclosed map of Arlington, Va. (Enclosure 4). With situations like this prevailing in many parts of the country, it is apparent that the interference potential to unprotected electronic equipment is considerable, and it is growing every day.

Given the present situation, what can we as radio amateurs do to help the consumer resolve RFI problems he is experiencing? The approach we suggest is one of encouraging the consumer to contact the manufacturers of his equipment, to request that they provide those components and services necessary to eliminate RFI which is a source of annoyance to both of you. This is not to say, however, that we should not assist in resolving RFI problems. On the contrary, we, by virtue of our technical competence, have an obligation to help the consumer. He should be made aware that we will make every effort to cooperate in resolving the problems which exist (for example, by running tests, assisting in writing letters to the various manufacturers involved, etc.), but that ultimately, <u>it is the manufacturers' responsibility to correct those deficiencies which led to the interference being experienced.</u>

Many responsible manufacturers already supply filters for reducing television interference when such cases are brought to their attention (Enclosure 5). A simple letter to the manufacture involved, stating the relevant facts on the interference problem, is all that is usually required to obtain the necessary high-pass filter. If a given manufacture is not listed in Enclosure 5, it is still possible that he will supply a filter; as such, send a letter either directly to him or to the Electronic Industries Association (EIA):

Electronic Industries Association
2001 Eye St., N.W.
Washington, D.C. 20006

In the case of interference to stereo receivers, AM-FM radios, phonographs, Hi-Fi's, electronic organs, and other home-entertainment equipment, it is suggested that a letter be sent to the manufacture or to the EIA. An example of such a letter is shown in Enclosure 6.

Regardless of the problem experienced, please fill out the enclosed form (Enclosure 7) so that the ARRL can accumulate statistics on RFI which will be of considerable use in our contacts with the FCC, EIA, and manufacturers of equipment used for home entertainment.

It may be of interest to know that the interference problem has come to the attention of the Congress of the United States. Further, the Congress is becoming increasingly aware of the manufacturers' responsibility to provide equipment which is not susceptible to RFI.

It has been the intention of this letter, and its enclosures, to provide you with factual information on RFI, and to suggest steps which can be taken to resolve RFI problems. It must be remembered, however, that to be effective, the suggestions supplied must be carried out in an atmosphere of cooperation between all parties concerned...the consumer, the radio amateur, and the manufacturer. Clear presentations of RFI problems, together with respectful requests for help, will be much more effective in obtaining a manufacturer's assistance than will letters demanding help. We, as radio amateurs are uniquely qualified to assist the public in understanding and correcting RFI problems, and we should, of course, continue to put forth our best efforts to resolve amicably those problems which occur in our own neighborhoods.

The ARRL RFI Task Group would appreciate hearing from you, and having your ideas and suggestions regarding the matter of RFI. Those who wish to help in the resolution of the RFI problem will be placed on a special list for future mailings by the Task Group.

Vy 73,

The ARRL RFI Task Group
W4UMF, W1ICP, W4KFC

Enclosures (7)

Washington Star-News

Strange Sounds, Bad Pictures, FCC Headache

By Miriam Ottenberg
Star-News Staff Writer

Fairfax County neighbors of an amateur radio operator threw fish at his door, scrawled obscenities on his wall and snubbed his wife because they blamed him for the strange sounds coming out of their television sets.

The cause of that neighborhood squabble is reflected in more than 42,000 complaints to the Federal Communications Commission last year from people whose television sets, hi-fis, phonographs, stereos, electronic organs, tape recorders, intercoms and radios have delivered unwanted voices or distorted pictures.

IN THE LAST three years, according to the FCC, the number of complaints has doubled as the number of licensed citizen band operators has grown to 810,000 using three to five million transmitters as a growing number of unlicensed operators have flooded the airwaves and as more homes have acquired second TV sets and a wide range of other electronic entertainment devices.

Across the nation, about two-thirds of the complaints deal with interference to TV, but to the Washington area, more complaints center on interference to audio systems such as amplifiers, electric guitars and organs and, especially, hi-fis.

"On several occasions," reported Harold R. Richman, engineer in charge of enforcement in FCC's Washington are a district office, "pastors have reported that mobile units calling their bases have come over the amplifier right in the midst of the sermon.

"**IT'S VERY** embarrassing. The man doesn't know he's in church and what he says can be heard over what the minister is saying."

Although complainants regularly blame the ham operators or CBers for the interference that tears up their pictures and obliterates their sound, the FCC says that in nine out of 10 cases, the fault is in the receiver and not the sender.

"Those ham operators and the three to five million citizen band radios would not cause interference to television and other home entertainment devices if companies had considered interference problems when they manufactured the sets," said James C. McKinney, chief of FCC's enforcement division.

"I don't know of any sets being built with filtering devices adequate to screen out unwanted signals. Our present rules don't require manufacturers to do anything along these lines," he said.

"**WE TELL** them that the man next door is doing nothing wrong," McKinney said. "We gave him a license to transmit and he's doing it properly. We advise them to have a high-pass filter installed on their set to filter out the unwanted signals. But they argue that they don't see why they have to spend money to fix their set when it was all right until their neighbor started transmitting.

"We're left in the middle of an argument we can't win."

As the problems and the protests have mounted, ham operators — weary of being unfairly accused — have urged legislation to require TV sets to be built with adequate filters to reject unwanted signals. Thousands of supporting letters from across the country reached Rep. Charles M. Teague, R-Calif., after he introduced legislation to require that radio and television receivers meet FCC standards for filtering out interference.

AMONG enthusiastic supporters of the measure was the Prince Georges County Council, which had been told about interference problems during its hearings on antenna heights.

In a resolution sent to Teague and Maryland Congressmen, the council urged passage of the Teague measure to benefit TV owners by eliminating the possibility of interference and aiding the general public "by allowing the full benefits of amateur radio to be realized in an atmosphere free of contention over alleged interference."

"It's just a coil device that could be installed easily in the sets," commented Council Chairman Francis B. Francois. "I think if people knew the truth, they'd demand that manufacturers design their sets to eliminate interference. Amateur operators are being unfairly blamed. The real cause is improper manufacture of the sets."

Teague died in January before his bill reached the hearing stage. Now, Rep. Torbert H. Macdonald, D-Mass., chairman of the House Commerce subcommittee on power and communications, has told the Star-News that there is considerable opinion that the FCC can require the manufacturers to put the necessary filters into television sets without additional legislation.

HE SAID he would explore the problem with the new FCC chairman Richard E. Wiley, and "if he says he needs legislation, I have no objection to putting it in although it's simpler and speedier if it can be done without Congress ordering it."

Asked the opinion of the industry, a spokesman for the Electronic Industries Assn. at first said that the association recommended that manufacturers not add interference rejection devices to television sets where less than one percent of customers have problems.

The association said it recommended that manufacturers make filters and information available on a local, individual basis.

Later, after learning that the interference problem is rising rapidly, Eugene J. Koschella, assistant vice president of the association's consumer electronic group, said the association has asked the FCC to send representatives to its service committee meeting this week to recommend possible solutions to the more frequent problems.

"Apparently the wide growth in over-the-air broadcasting, CB, ham, two-way radio, increased power AM and FM stations, etc., warrant a fresh look into our positions," Koschella acknowledged.

ALTHOUGH Koschella said major manufacturers provide free filters for complaining TV owners, those seeking built-in filters argue that even free filters require a $15 service call for a technician to install them.

And in some cases, manufacturers have warned that if a TV owner installs the filter himself, he'll lose his warranty.

Dr. Theodore J. Cohen, a senior research scientist and project manager for an electronics group and a licensed amateur radio operator since he was 12, has spearheaded the campaign for legislation to guard against unwanted signals.

He has proposed legislation that goes beyond the Teague bill to give the FCC authority to regulate all audio and radio-frequency receiving devices so their susceptibility to interference is reduced.

Without that legislation, the FCC has no authority to deal with interference problems plaguing such home entertainment devices as stereos, electronic organs, hi-fis and tape recorders.

Currently, the FCC is trying to cope with interference problems by telling complainants, that their own receivers are the source of their problems and supplying information they can give their TV or audio serviceman to get rid of interference.

"WE SEEK ways to resolve these problems without sending out technicians we don't have," McKinney explained.

He pointed out that the FCC's enforcement force today numbers 460 people — five less than in 1948 — before television grew to its present 119 million sets and before the citizen band was created.

"I'm not making a plea for more men but for an end to what is probably our biggest problem — interference to home entertainment devices, unnecessary interference," he said.

FEDERAL COMMUNICATIONS COMMISSION
FIELD ENGINEERING BUREAU

STATEMENT TO TELEVISION RECEIVER OWNER

The Federal Communications Commission has received your complaint concerning interception of unwanted radio signals from another radio station and you have requested investigation of the condition. The FCC does not have staff and facilities to investigate all cases involving interception of unwanted radio signals which affect television reception. Television receivers are not designed by the manufacturers to work in all cases in the presence of strong signals from other radio stations. Under these circumstances additional treatment to your television receiver will probably be required to correct the condition. Commission experience shows that 90 percent of the cases of this nature require treatment of the receivers rather than treating the transmitting equipment at the transmitting station. Interception of unwanted signals which disrupt television reception can be very annoying. In order to restore good reception as quickly as possible, it is important to recognize that the solution of the problem will, in most cases, depend on corrective treatment to your television receiver.

Your television serviceman should be able to resolve your problem with a signal rejection filter or other technical treatment to your receiver. We urge that the matter be settled locally without assistance of the FCC. It is suggested that you furnish your serviceman with the attached sheet.

Enclosure (2)

NOTE TO SERVICEMAN

The data received from the complainant indicates that the interference problem which is being experienced is coming from a nearby station below 54Mc/s or an FM broadcast station in the 88 to 108 Mc band. If the station is below 54 Mc/s or you are unable to identify the station, try a good quality high pass filter across the antenna terminals of the television set. If the interference is from an FM broadcast station, try an FM band reject filter. High pass filters and FM band reject filters, with good rejection characteristics, are listed below.

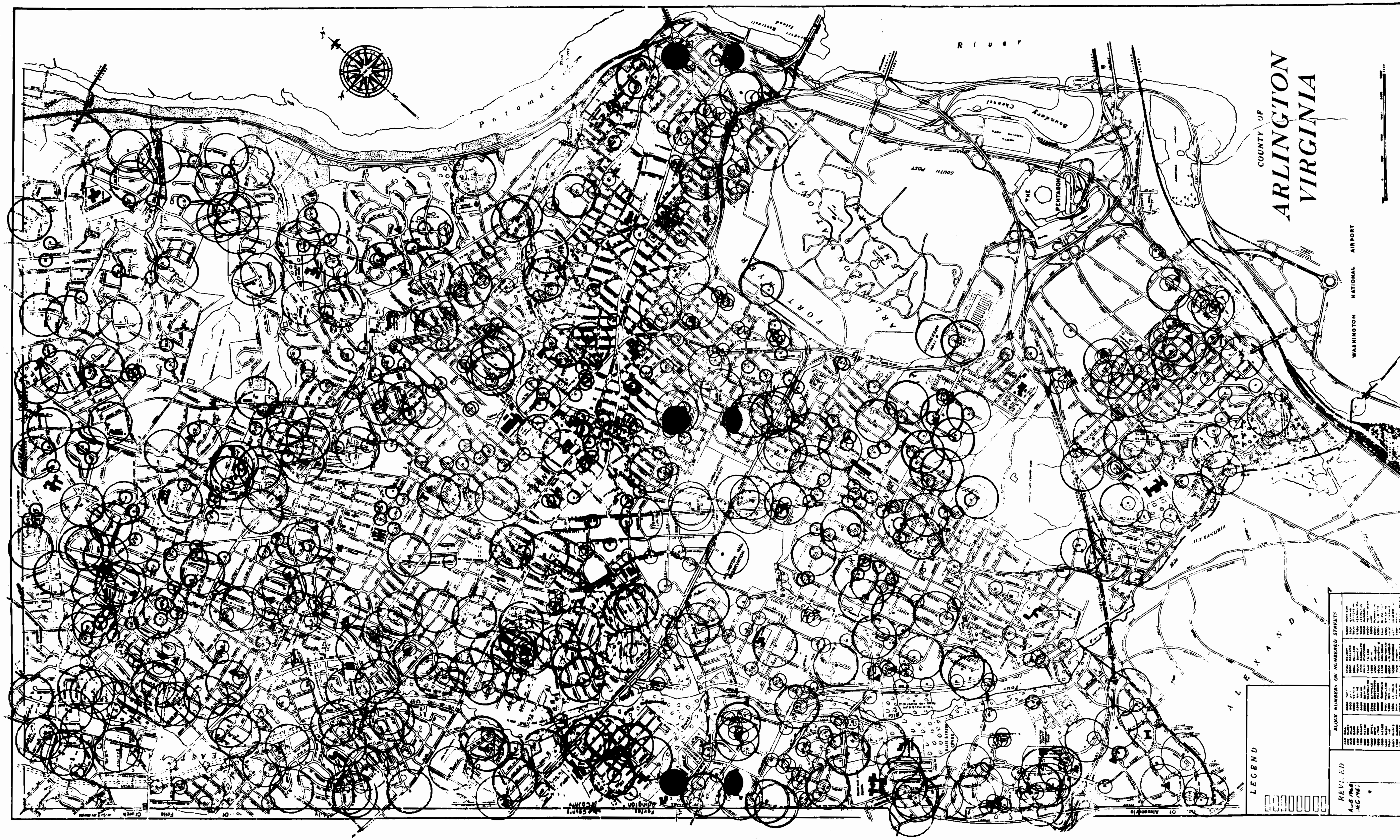
High Pass Filters

Make	Model	Remarks
R.L. Drake	TV-300-HP	
Finco	3013	

FM Band Reject Filters

Make	Model	Remarks
JFD	TR-FM	General use, 10 db loss on Channel 6
Drake	300 FMT	" " " " " " " "
Finco	3006	" " " " " " " "
Drake	300 FMS	Especially useful for interference from stations in the low end of FM band where reception of Channel 6 is desired.

In about 50 percent of the cases, a filter on the antenna terminals of the TV receiver will not fully remove the interference. There are a number of other possibilities: (1) the filter is not effective at the antenna terminals, place it at the tuner input terminals; (2) the filter may have insufficient rejection ability to reject the interfering signals sufficiently; two filters in series may be needed; (3) cross modulation may be occurring in an antenna signal booster amplifier generating a signal that can't be removed by the filter at the receiver; place the filter in front of the booster or remove the booster; (4) the antenna and/or lead in cable may have corroded joints causing cross modulation and thus generating signals that can't be removed by the filter; repair the antenna system or install new components as needed; (5) the receiver may be subject to audio rectification in the input of the first audio stage; a bypass condenser of .005 uf from the grid of the first audio stage to ground will usually solve this problem. (Treatment applies to tube type receivers only; consult the manufacturer concerning solid state receivers.)


COUNTY OF
ARLINGTON
VIRGINIA
Potomac
River
THE PENTAGON
SOUTH POST
FORT MYER
WASHINGTON NATIONAL AIRPORT
LEGEND

Comments on Map of Arlington, Virginia

The attached map of Arlington, Virginia, A Washington, D.C., suburb, helps to convey a better appreciation of the burgeoning RFI problem. Prepared from FCC licensing records approximately a year ago, the map shows amateur stations as 200-foot circles, and Citizen's Radio Service stations as 50-foot circles. That RFI can be experienced well outside these zones is well known to amateurs, and the areas represented by the circles are for comparison purposes only. The triangles depict fixed stations in the Land Mobile Service, although not all of them are indicated. Not shown are military radio installations, microwave relay stations, commercial repeaters, broadcast stations, experimental stations, or mobile radio stations, which number in the hundreds in this area.

It is readily apparent that no electronic device in this sample area is immune to the likelihood of strong near-field radiation from a radio transmitter of some kind. Yet Arlington, Virginia, is but an example of a situation which is becoming increasingly serious in every part of our country, and indeed, the world.

The great and growing density of radio transmitting stations appears either to have been thus far underestimated or largely disregarded by the manufacturers of RF-vulnerable electronic equipment. Even today the argument is offered that only a small percentage of such equipment will ever be subject to strong RF fields, and this presumption offered as justification for short-cut equipment designs.

Enclosure 4

FEDERAL COMMUNICATIONS COMMISSION
FIELD ENGINEERING BUREAU
Washington, D. C. 20554

AUDIO DEVICES - INTERCEPTION OF RADIO SIGNALS

This bulletin is intended to assist you with the problem of interception of radio signals by a record player, Hi-Fi or stereo amplifier, tape recorder, public address system, telephone or the audio section of a television or radio receiver where the condition is observed on all channels or all across the dial. In the following discussion these units are referred to as "audio devices."

Audio devices are designed to amplify audio signals such as music or speech and are not intentionally designed or intended to function as receivers of radio signals. The Commission cannot give any protection to audio devices which respond to signals from a nearby radio transmitting station. The problem is not caused by the improper operation or by technical deficiencies of the radio transmitter. Strong radio signal energy gains entry into the audio circuitry, "overloads" the amplifier, is "rectified" and amplified, and appears at the loudspeaker as undesired sound. The only "cure" is by treatment of the audio device. You should therefore, contact a qualified technician, the dealer, or the manufacturer of your audio device (or telephone company for telephone problems) for assistance.

You may very reasonably ask, "Why do I have to do something to my audio device? It works fine except when the radio station is transmitting. Why is it my problem and not the responsibility of the operator of the radio station?" The answer lies in policies concerning the economics of the design and sale of these devices, in a highly competitive market. The device has two objectives: 1) to reproduce a desired audio signal and 2) to reject unwanted signals which may degrade the overall performance of the device, at a reasonable cost.

The state of the electronics art is such that it is possible to manufacture "custom built" audio devices; that is, to install in them complete shielding and special circuits to reject nearly all types of undesired signals. The cost of special designs and circuitry would necessitate an increase in the price of the device.

Perhaps less than one percent of the total number of audio devices in use today will be located near the strong field of a radio transmitting facility whereby the device will respond to radio signals which it is not designed or intended to receive. If it does, it requires the addition of filtering or shielding, or both.

Manufacturers believe it is unfair and unnecessary to burden the mass consumer market with the additional costs of special circuits and designs inasmuch as the number of devices affected is relatively small. Many manufacturers, dealers, and servicemen have devised procedures and processes to improve the radio signal rejection capability of audio devices.

The conditions at a particular location are not necessarily stable. Two situations may arise:

1) The user may live in an area where the average audio device has no difficulty in rejecting unwanted signals present. The situation in the area may be changed by the installation of a new station nearby, or by the appreciable strengthening of existing signals by the increase in power of a given station. Under these conditions, the situation has changed from that of a location in the ninety-nine percent category where no special treatment of the unit is required to that of the less than one percent category where special treatment becomes necessary.

2) The user may reside in an area that is in the ninety-nine percent category having no strong unwanted signals and later move into an area where strong radio signals are prevalent. The situation has then changed from that of being located in the ninety-nine percent category where special treatment of the unit is not necessary, to the less than one percent category where special treatment is needed.

Information concerning the foregoing is not necessarily widely distributed or fully recognized by all manufacturers, dealers, and servicemen. The statement of a salesman or dealer that he sells a good quality device, or that there is nothing wrong with the unit, is not enough and avoids the issue. Persons in possession of an audio device not capable of rejecting radio signals must fully recognize the situation involved. An audio device may perform the task for which it was designed with excellence and still require special treatment to improve its capability to reject strong radio signals in some localities.

<u>You are urged to bring this information to the attention of your serviceman, dealer, or manufacturer of your equipment.</u>

High-Pass Filters Available

Manufacturer	Where to Write	What Filter Supplied	Notes
Motorola	Consumer Products Div. Consumer Relations Dept. 9401 West Grand Ave. Franklin Park, IL 60131		
Heath	Benton Harbor, MI 49022	1. Drake TV-300-HP, or 2. reimbursement for locally-procured filter, upon receipt of invoice	
Olympic International	89-89 Union Turnpike Glendale, NY 11227	Drake TV-300-HP	
Westinghouse Electric Corp.	Portable Products Div. P. O. Box 333 Edison, NH 08817		Westinghouse no longer makes TV sets
RCA Sales Corp.	600 North Sherman Drive Indianapolis, IN 46201	Drake	
TMA Company	1020 Noel Avenue Wheeling, IL 60090		Muntz TV TMA TV Howard Stereo
Magnavox	7 Regional Service Centers East Rutherford, NJ; Atlanta, GA; Westlake, OH; Skokie, IL; Dallas, TX; Torrance, CA; So. San Francisco, CA	Drake	
Emerson	Emerson Television Sales Corp. Jersey City, NJ 07302		

1234 Someplace St.
Some City, USA 98765
23 June 1977

Stereo Manufacturers, Inc.
7654 Speaker Street
Anywhere, USA 34567

ATTN: Technical Service Department

Dear Sir:

I recently purchased a Stereo Manufacturers Model A234 stereo (Serial Number Y67-76) with J-55 speakers. I am pleased with the general performance of the unit except that it is highly susceptible to radio-frequency interference. The signal from a neighboring short-wave radio station causes strong interference regardless of the source selected (phono, FM, AM). The signal masks the radio's entire dial range, and varying the volume control has no effect whatsoever on the intensity of the interference.

The owner of the short-wave station has been cooperative, but he says that the problem is inherent in the basic design of the stereo's circuitry. As such, there is nothing the radio operator can do, it would appear, to correct the situation.

I understand from reading a bulletin published by the FCC (FE Bulletin No. 25) that many manufacturers have procedures to improve the radio signal rejection capability of their products, and that the consumer is urged to bring interference problems to the attention of the manufacturer. As such, I ask for your assistance in eliminating this most aggravating problem.

Sincerely yours,

John Q. Public

cc: Federal Communications Commission
1919 M St., N.W.
Washington, D.C. 20554

Electronic Industries Association
2001 Eye St., N.W.
Washington, D.C. 20006

American Radio Relay League
225 Main St.
Newington, Conn. 06111
(ATTN: RFI Task Group)

Enclosure 6

Report of Radio Frequency Interference (RFI)
(Please Print or Type)

The purpose of this form is to assist the American Radio Relay League (ARRL) in collecting information on radio-frequency interference (RFI). Such information, and the statistics to be derived therefrom, will be of considerable help to the League in its contacts with the Federal Communications Commission, and with various representatives of the electronic's industry... contacts which we hope will lead to the elimination of RFI as a source of annoyance to both the consumer and the radio operator.

Type device______________________________(Television, Radio, Hi-Fi, Tape Recorder, Phonograph, Electronic Organ, Intercom, Other (Specify)).

Manufacture______________________________

Model No.______________________ Serial No.______________________

Description of Interference (picture, Sound, Channels affected, Effect of varying volume control, etc.):

Other devices in the same location experiencing interference:

Other devices in the same location which are <u>not</u> experiencing interference:

Distance to radio station's antenna________________________

Description of radio station:

Transmitter_____________________watts Antenna________________

Frequencies of operation for which interference is observed:

Consumer's name and address:___________________________

____

____

Add any remarks you may have on the reverse side of this form.

Mail form to: ARRL RFI Task Group, ARRL, 225 Main St., Newington, Conn. 06111

Enclosure 7

Comments on
Radio-Frequency Interference (RFI)

Many interference problems could be eliminated with adequate shielding and better engineering of the product, but this is not being done. (Thus), passage of (legislation) would assure the consumer of being able to purchase electronic devices that would meel certain standards which would make it free of many present-day interference problems.

Ray Clark, K5ZMS/5

All the PR in the world won't do us a bit of good until the RFI problems are borught under control. Today if you tell someone you're a ham, the comment is, "Oh, you're one of the guys that tear up TV." If the image of amateur radio is to be improved, then the first priority is to get the RFI problems under control. The consumer is entitled to properly designed entertainment apparatus and the use of this apparatus free of interference problems... likewise, the ham or other operator of communications equipment is entitled to the use of this equipment free of interference complaints.

Dan Rasmussen, WA9UBI

An alternative to outright legislation compelling manufacturers to certify performance tests of interference rejection of their product, would be to compel manufacturers to place a disclaimer on their product which has not been tested such as:

"This device is not guaranteed to reject interference when used in the vicinity of strong radio sources."

Katashi Nose, KH6IJ

We hear a lot about what more the Government should be doing to protect the consumer. Usually these proposals involve the creation of new or expanded bureaucracies, at great cost to the taxpayers. The potential of the electronics of radio, it seems to me, has barely been tapped, and I believe that radio amateurs have a tremendous capacity to open up new frontiers, and in the process, to improve the quality of components being merchandised, not only to your group, but also to the general public.

Congressman J. Kenneth Robinson

What is needed is Congressional or FCC action to require all manufacturers of TV sets, stereos, and a-m receivers to build interference suppression into their designs. Some lead bypassing and narrow filters at the input would go a long way toward solving the present problem.

Jim Fisk, W1DTY

Electronic proliferation in the forms of cable TV and non-RF electronic devices that are interference-prone make it imperative that legislation be extended to devices other than the usual "home entertainment" devices. I am enclosing a copy of an Electronic News article (on rf interference to pacemakers) which may be of interest.

Edward Erickson, W2CVW

ALMOST EVERYTHING YOU WANT TO KNOW ABOUT MOON BOUNCE

DIVISION OF VARIAN
301 Industrial Way
San Carlos, California

AS-49

4139

Moonbounce Notes

During the last year there has been an upsurge of interest in amateur communication via reflection from the moon. All the bands from 50 MHz through 2400 MHz have been involved.

This activity has created more interest in moonbounce and each neophyte "moonbouncer" has had many questions concerning just how to get started:-

-Which band should be used?

-How much power is needed?

-How good should the receiver be?

- What kind of antenna should be used, and how big should it be?

In the process of determining antenna parameters, it is necessary to know how to find and track the moon. In addition, the type of antenna mount, the aiming system and the physical location of the antenna on the available plot of land are a function of the path of the moon.

Many articles have been published which will answer these questions and help the beginner and old time "moonbouncer" alike. Much other data are available which have not been published. It is the intent of this compilation of moonbounce notes and articles to reproduce in one place the literature necessary to allow the potential "moonbouncer" to make the basic decisions necessary to start his project.

As time goes on, additional notes will be added.

Contributions from those interested in EME (earth-moon-earth) communication will be gratefully received.

Thanks to the American Radio Relay League for permission to reprint certain articles from QST magazine.

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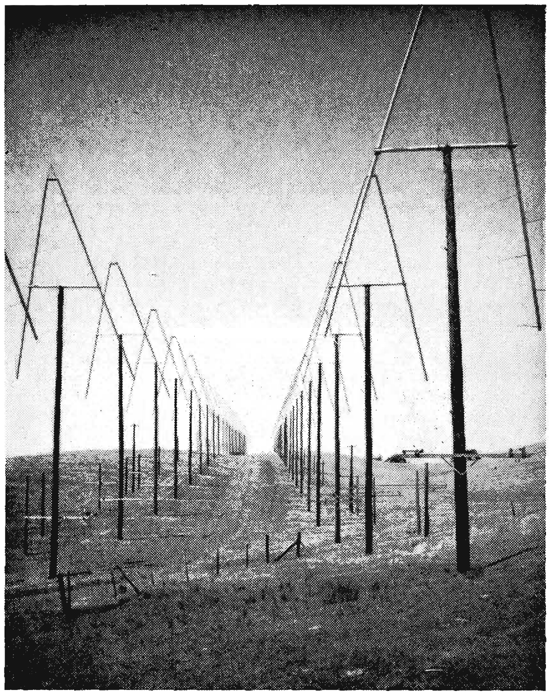

This field full of phased log-periodic antennas was used by the author to obtain reflections from the moon in the 10-meter band. In case you'd like to duplicate the feat, the array is 1200 feet long and 75 feet wide. It has a gain of 27 db.!

The Moonbounce Problem, 28 Mc. and Up

Basic Facts for Determining Equipment and Antennas Needed for Lunar Communication

BY H. T. HOWARD,* W6UGL

The purpose of this article is to stimulate amateur interest in moonbounce communication, by presenting the basic parts of the problem, such as noise figure, path loss, and antenna gain, in familiar terms. Once these basic factors are understood, they can then be applied to equipment and antenna design for communication via the moon or man-made satellites.

Moonbounce was accomplished on ten meters several months ago at this station with about 1 kw. p.e.p. single sideband, using the array of 48 log periodics shown in the first photograph. The array is 1200 feet long by 75 feet wide, and it has a gain of 27 db., over the range of 20 to 65 Mc.! The beam produced is approximately 1½ degree thick by 30 degrees in azimuth and can be placed to intercept the moon or sun track for about two hours each day. Power is distributed in the array with open-wire line, and tapered sections to maintain the wide bandwidth, and in the usual operation *each antenna* handles from 5 to 10 kilowatts. A circuit diagram of the array would look like a corporation organization chart; that is, it starts with one feed line and progressively branches down to the individual antennas which are specially designed and built log periodics, each having a pair of 40-foot booms and a total of 48 elements.

At each power division point there is a movable tap arranged so that the relative phase between antennas is completely adjustable. In practice, the phasing is changed each day to follow the moon's elevation. It takes two men with wrenches and a jeep about two hours to move all of the taps. The array is normally used with a 300-kw. (600-kw. p.e.p.) c.w. transmitter for radar studies of the solar corona and the ionized regions between the earth and the moon.[1]

The selection of ten meters for the moonbounce experiment mentioned above avoids controversy

* Radioscience Laboratory, Stanford University, Stanford, Calif.

[1] Research supported by the Electronics Research Directorate of the Air Force Cambridge Research Laboratories, Bedford, Mass., under contracts with Stanford University, Stanford, Calif.

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over the use of large nonprivate antennas for v.h.f. records. Six or possibly fifteen meters might yield similar results if tried. The whole idea, though, is to demonstrate that the absolute minimum antennas for h.f. and lower v.h.f. moonbounce are ridiculously large for individual construction.

Since the array is linearly polarized, Faraday fading is a very important consideration,[2] and the unrecommended expedient of whistling into the microphone was used, until the signal faded up to a usable strength. Then the call was signed in voice and, with the help of some imagination, was received 2½ seconds later. The use of circular polarization will reduce fading, and is certainly required for any serious v.h.f. lunar-communication attempt.

The trick of ten-meter moonbounce points out several facts that will become obvious as you read further. First, station equipment needed for moonbounce on our lower bands is a minimum, and commercially available, but the antenna required is gigantic. Second, cosmic noise and ionospheric effects play a large role below about 100 Mc. With increasing frequency, the antenna becomes physically smaller, but the receiver and transmitter must be the best that amateur ingenuity can produce.

The average loss in decibels for the earth moon-earth path, assuming 500 watts of r.f. power at the antenna terminals and a moon reflectivity of about 7 per cent, is given in Fig. 1. Path loss will vary approximately ± 1 db. during each month as range to the moon changes.[3] Moon reflectivity is currently the subject of several scientific investigations, and while reflectivity appears to be higher at frequencies below 450 Mc., and is perhaps lower above that frequency, the figure given should be accurate enough for a first approach to the problem. If the transmitter power at the antenna terminals is less than 500 watts due to feed-line losses or other practical considerations (such as money) this path-loss number should be increased by the number of db. difference.

The next problem is that of receiver noise figure and sensitivity. Fig. 2 is a plot of cosmic noise *vs.* frequency, presented to give the equipment designer an idea of what is needed for a front end. The *min* and *max* lines show the sky temperatures and minimum usable noise figure that can be expected when the antenna is directed toward the coldest and hottest portions of the sky, respectively. This variation is easily observable even with simple equipment and is a good method of checking antenna and system performance.[4] Fortunately for the communications problem, larger areas of the sky are cold than are hot.

Below about 1000 Mc., cosmic noise is the dominant factor and varies with the portion of the galaxy observed. It can be seen that being cosmic-noise-limited, that is, having the feed-line loss and receiver-noise contribution less than the cosmic noise, at all times, is an engineering feat nearly impossible at 220 Mc. and higher, with the present state of the art.

Before going further, it is necessary to clear up some confusion concerning receiver sensitivity and noise figure that has arisen because of improper use of the relation:

Ideal receiver sensitivity = kTB

where

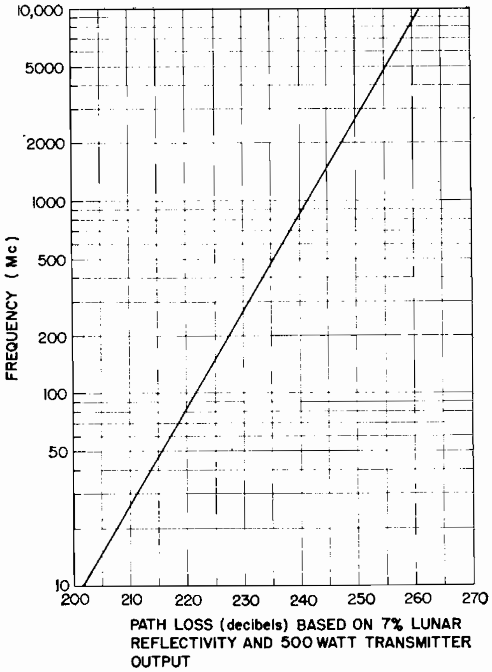


Fig. 1—Moonbounce-path loss vs. frequency.

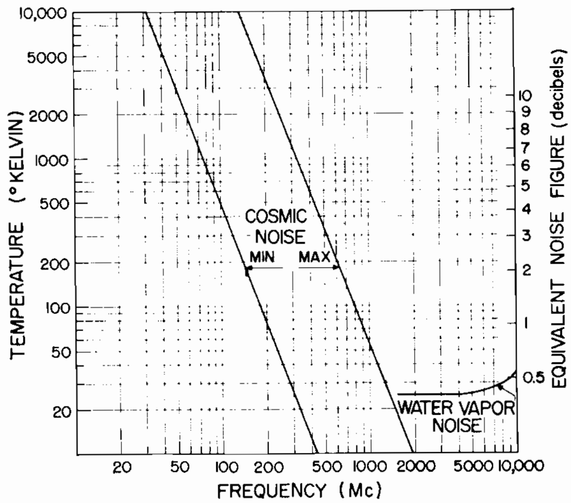


Fig. 2—Cosmic and water-vapor noise limits vs. frequency.

[2] Dyce, "The Appearance of the Moon at Radio Frequencies," *QST*, May, 1961.

[3] Pettengill, "Lunar Studies." Lecture notes presented at course on Radar Astronomy, summer session 1961, Massachusetts Institute of Technology, Cambridge, Mass.

[4] Downes, "A Simple Radio Telescope," *Sky and Telescope*, August, 1962.

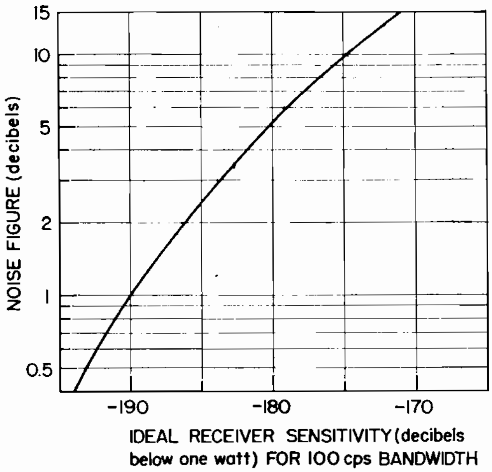


Fig. 3—Ideal receiver sensitivity vs. noise figure.

k is Boltzmann's constant, 1.38×10^{-23} joule/°K

T is temperature in degrees K

B is bandwidth in c.p.s.

If one uses room temperature of 290 degrees K, then it can be shown that:

Ideal receiver sensitivity (−dbw.) =

$$204 \text{ db.} - 10 \log B - \text{db. noise figure.}$$

This relation is correct for systems with noise figures greater than 3 db. (system temperature greater than 290 degrees K), but needs revision to be correct for present-day low-noise amplifiers. By using an equivalent system temperature for T instead of 290 degrees K, we can still satisfy the IRE definition for noise figure and be consistent with present practice. All of this is simply saying that it is possible for a directive antenna and receiver at u.h.f. to look at a portion of the sky that is colder than 290 degrees K.

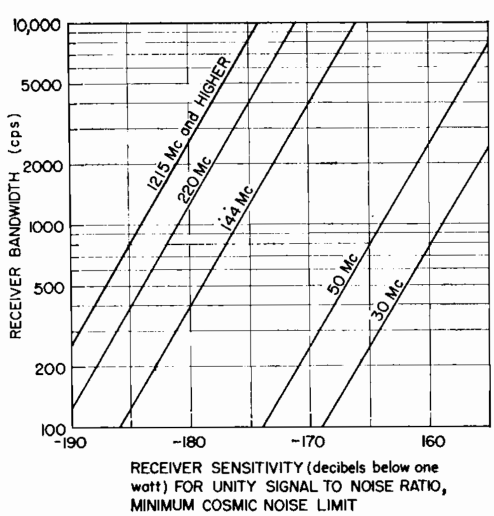


Fig. 4—Receiver sensitivity vs. bandwidth for the various amateur bands.

A plot of the above interpretation for a receiver with a 100-cycle bandwidth is shown in Fig. 3. There are some surprises in this graph that arise from proper use of noise figure. For instance, an improvement in receiver noise figure from 3 db. to 2 db. improves receiver sensitivity not just 1 db., but nearly 3. Going from a 10-db. crystal mixer to a 2-db. paramp yields a sensitivity improvement of 12 db. This makes it pretty obvious that the best possible noise figure and the lowest possible line losses are all important, at frequencies where system noise is greater than cosmic noise.

Fig. 4 uses the information of Fig. 2, and assumes that the system performs to the lower cosmic-noise limit. It shows what receiver sensitivity to expect in each case, for unity signal-to-noise ratio with various bandwidths. If the system is not cosmic noise-limited, the number obtained from Fig. 4 should be decreased by the number of db. difference between the ideal case of Fig. 4 and the actual system. Again, both noise figure and transmission-line loss enter here. The number from Fig. 4, as modified by reality, is the receiver sensitivity in decibels below 1 watt, and can be added algebraically to the path loss of Fig. 1 to obtain the two-way antenna gain necessary.

For example, select 1296 Mc. and assume a parametric-amplifier front end with a 2-db. noise figure [5] and 2 db. of feed-line losses. From Fig. 1 the total path loss is 244 db. and from Fig. 2 the system is definitely not cosmic noise-limited.

Example:

Fig. 1:	Total path loss for 500 watts power output	244
	Feed-line loss	2
		246 db.
Fig. 4:	Cosmic-noise-limited receiver sensitivity (500 c.p.s. bandwidth)	−187 dbw.
Fig. 2:	Receiver n.f. = 2 db. = 170° K	
	Line loss = 2 db. = 170° K	
	340° K = 3.4 db.	
	Cosmic noise = 0.5 db.	
Fig. 3:	Difference between 0.5 db. cosmic noise and 3.4 db. actual receiver system	+ 10 db.
		− 177 dbw.
		69 db.

$$\frac{69}{2} = 34.5 \text{ db.}$$

This is the antenna gain required at each station for unity signal-to-noise ratio in a 500-c.p.s. bandwidth, but as W1FZJ has pointed out,[6] the ear can be a narrower filter if properly trained.

[5] Troetschel and Heuer, "A Parametric Amplifier for 1296 Mc.," *QST*, January, 1961.

[6] Harris, "The World Above 50 Mc.," *QST*, June, 1961.

(Continued on page 6)

very little keeping you from beginning in e.m.e. work except the antenna.

The bare minimum gain required from the 144-Mc. antenna is 20 db. over a dipole. This does not mean that echoes are not possible with slightly less gain, but for any hope of reliability through the moon's cycle, 20 db. is the line when using "normal" receiving systems.

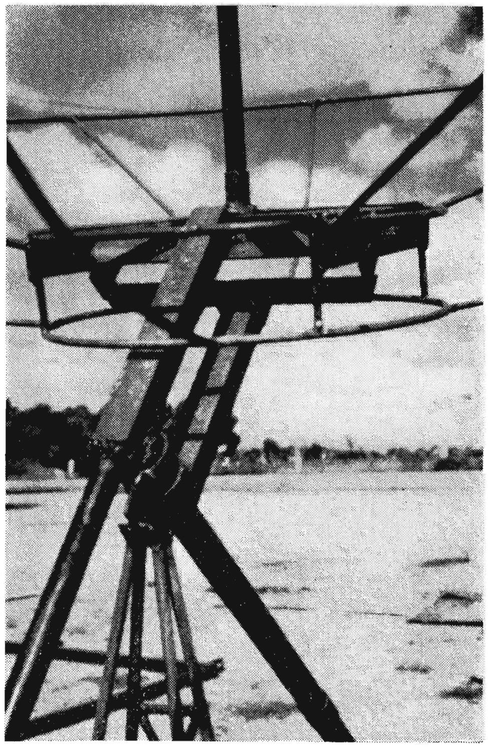
Now about the antenna. To my knowledge, no one has yet been satisfied with the performance of Yagis on an e.m.e. circuit. W6DNG used them but has since changed to an extended expanded collinear which he says is the best of more than 50 e.m.e. antennas he has tried. F8DO has a Yagi array, but doesn't feel it is performing as well as it should. However, short Yagis of 4 or 6 elements may be the answer if you must try them. VE3BZS/VE2 has an array of sixteen, 4-element Yagis and is now doubling that number. He's had some success in hearing his own echoes. K6HCP, using two 26-foot boom Yagis, ran several hours of tests over a period of days with K6MYC with completely negative results. Transmitting at K6HCP and listening at K6MYC produced nothing. The opposite was also tried without success although K6MYC could hear his own echoes.

The antenna at K6MYC is a 160-element collinear which I believe is producing close to the theoretical 24 db. gain. Echoes can be received almost anytime during the moon's 28-day cycle, assuming the Faraday rotation (polarization rotation in the ionosphere) is correct. We will discuss Faraday rotation later as well as the 28-day cycle, which is related to sky temperature (cosmic noise) at various "look" angles. F8DO, VE3BZS/VE2, ZL1TFE, ZL1AZR, WB6DEX, and of course VK3ATN have all heard K6MYC on e.m.e. WB6KAP has an antenna almost identical to K6MYC's and has had equally good results.

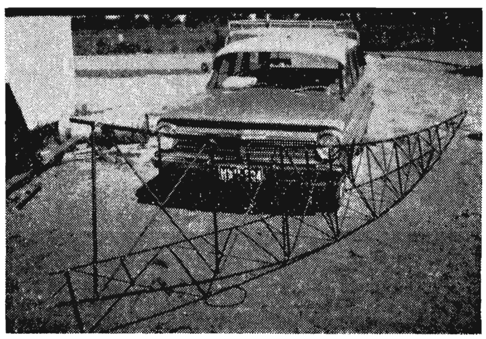
The cubical quad looks good since it fits into the low-Q class with collinear types. ZL1TFE heard signals with four 5-element quads patterned after those by W1CER and modified by W7FS. Don't rule out expanded quads as they should be quite practical. Size and weight seems to be their chief drawback.

Rhombics, the king of the h.f. antennas, seem to have a place in v.h.f. circles as well. This antenna does not allow much moon time each month, but the gains achieved can be extremely high, at much less cost than most other arrays. VK3ATN uses four rhombics stacked one above the other with six-foot spacing between for his 2-meter e.m.e. antenna. The antenna is 342 feet long per leg and has an apex angle of about 10 degrees. The gain is calculated at 33 db., over perfect ground, but actual gain is probably closer to 27 to 30 db. VK3ATN has been very successful using this antenna and 150 watts input. The LaPort rhombic is being tried and seems to have possibilities. ZL1AZR has a single-layer one and has copied K6MYC and possibly VK3ATN. More layers or a side-by-side configuration may be in order. The antenna is only 70 feet long. The disadvantages of rhombics are immobility and low elevation angles.

All of the antennas thus far discussed have been linearly polarized. Now let's consider some sort of circularly-polarized antenna. First a definition of circular polarization is in order; let us use the helix to simplify the explanation. Since a helix has no linear element, it theoretically radiates equally in all planes and the wave is launched in the direction of the spiral. Depending on whether the helix is wound clockwise or counterclockwise, the antenna would be called right- or left-hand



Shown is the hub assembly of the VK3ATN dish. The declination and hour angle drive motors are yet to be mounted as is a 16-inch diameter bearing.



This is one of the more than 20 steel tubing trusses being used in the 50-foot dish at VK3ATN. The 20-foot long trusses weigh 30 pounds each and are within ⅛-inch tolerance of a parabolic curve.

circularly-polarized respectively. For point-to point communication using helices, both antennas should be wound the same direction. When listening for one's own echoes, a right-hand signal radiated at the moon will return left-hand. This means that to hear your own echoes the direction of circularity must be switched. Circular polarization can also be achieved with properly phased crossed dipoles orthagonally mounted.

WB6DEX currently uses nine 20-element crossed Yagis, but runs only the horizontal 90 elements when testing with another station using horizontal polarization. Otherwise he would lose about 3 db. by putting half of his power into the vertical elements. However, if both stations used 180 elements circularly-polarized arrays, 3 db. would be gain on both ends — obviously very worthwhile.

Also the problem of long term fading due to Faraday rotation would be eliminated. This polarization mismatch can cost as much as 20 db. when using linear-polarized antennas. Helix antennas have not yet been successfully used for a two-way amateur e.m.e. contact as far as I know, but maybe W8JK's helix will intrigue some of you who need a new challenge. (See W1CER's article on page 20, November, 1965 *QST*.)

Certainly one antenna that should not be overlooked is the parabolic reflector, be it circular or cylindrical. However, dishes of a useful size at 144 Mc. are impractical for the average amateur, but at 432 and higher the picture brightens. (K2UYH described a homebuilt dish in the August, 1966 *CQ*.)

To summarize on antennas, my personal experience tells me circularly-polarized antennas for 432 and above, if at all possible, and below 432 shoot for maximum gain in a low-*Q*, linear polarized array.

Now let's look at a smaller but still important component in the e.m.e. station, the preamp. It may not be entirely necessary if your converter has a noise figure of 3 db. or less, but if located near the antenna the preamp can reduce feedline losses on receiving and possibly lower the system noise figure a bit. The noise figure to aim for at 2 meters is 2 db. You can try for less, but don't expect a noticeable increase in sensitivity, because the lowest sky temperature encountered at 144 Mc. is about 1.9 db. At 432 and above cosmic noise is less and very low-noise devices become more useful. It is doubtful that your system will be cosmic-noise limited. On 144 and 432 transistors appear to be the way to go, and more specifically, FETs or the steadily improving MOS dual-gate FET. Many types and brands are available for under $2. Many good preamp circuits have been published, but most lack protection for the transistor. A pair of diodes, typically 1N100s, back-to-back at the input to ground will save much grief. If you insist on using regular bipolar transistors, be sure to build a good stripline filter to help eliminate overloading of the transistor by strong local stations in the broadcast band and higher. Normally a filter is not needed ahead of a FET.

Little need be said about the balance of the converter except that crossmodulation (overload) of the mixer stage can sometimes be a problem. The use of FETs as mixers is a current solution. Recently RCA began marketing a dual-gate MOS FET pair that look ideal for converters, a 3N140 front end and a 3N141 mixer. Both are under $2 and may be the best yet for 144 and 220.

Next month we'll look at methods used during e.m.e. tests and pass along some time-saving hints. Also a thorough examination of the problems encountered is in order, as is a discussion of antenna mounts and drive mechanisms. In the meantime, you should read W6UGL's article, "The Moonbounce Problem, 28 Mc. and up," on page 20, September, 1963, *QST*.

A Layman's Look at E.m.e.—Part II

K6MYC continues his discussion this month of propagation problems effecting e.m.e. communications and what the amateur can do to alleviate some of them.

Although there are electrons everywhere in our atmosphere and beyond, those in the ionosphere have the greatest effect on v.h.f. and u.h.f. signals leaving this planet. This cloud of electrons is in a constant state of flux, their number either increasing or decreasing, or moving about to form clouds or blobs, much the same as vapor clouds. For our discussion, however, think of the ionosphere as a homogeneous layer with no irregularities. A plane-polarized 2-meter signal entering this layer is gradually rotated and may go through several rotations before passing through the ionosphere and into space. If electron content is high, as it normally would be during daylight hours, the signal may rotate many more times than it would during early morning hours. This phenomenon is known as *Faraday rotation.*

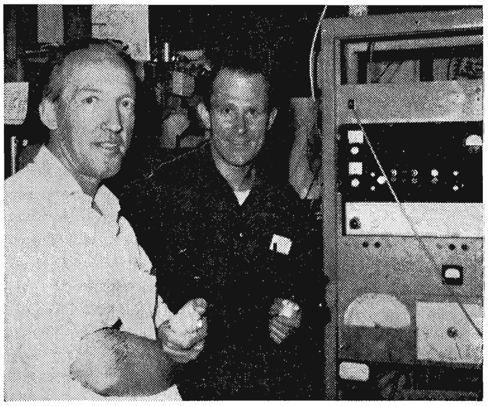
Regardless of the plane of polarization originally, the wave may come through the layer in any plane until it strikes the moon. As an example, consider the direction of rotation to be clockwise. When the signal strikes the moon and is reflected, it maintains its plane of polarization until beginning to re-enter the ionosphere, where again it begins to rotate, still in a clockwise direction, until returning to the antenna from which it was transmitted. An originally horizontal signal may have rotated six times plus 45 degrees leaving the ionosphere, and another six times plus 45 degrees upon re-entry, adding up to a net 90-degree rotation change, or vertical polarization. The signal received on a horizontal antenna may suffer a 20- to 30-db. loss from polarization shift alone.

The problem of Faraday rotation is further complicated when contact with another station is attempted. The transmitted signal must pass through two probably-different ionospheric sections before arriving at the other antenna. The polarization of the arriving signal may match the plane of one of the two antennas, but not necessarily both, or either. To put it simply, your own echoes may be coming back well, but the other station may not hear anything. But if transmissions are continued for an hour or so, chances are your own echoes will fade and the other station may start hearing you.

(A demonstration of this occurred on Dec. 20, when K6MYC and VK3ATN had another e.m.e. QSO. During the entire QSO, 1302 to 1310 GMT, neither was able to hear his own echoes. VK3ATN also heard W6YK for 8 minutes following. — EDITOR)

Another interesting fact about Faraday rotation is the relation to the hemispheres involved. A plane-polarized signal leaving the northern hemisphere twisting clockwise will return in the southern hemisphere counter-clockwise. It is possible that the effects of Faraday rotation can be nullified if the electron content of the ionosphere were the same for both paths. The shift related to hemispheres does not occur if both stations are in the same hemisphere. Schedule times should be chosen when both stations can use approximately the same antenna elevation angle (the moon the same distance above the horizon at both station) as there are usually two to three times as many electrons in the horizon path to the moon as in the path at a 45-degree elevation angle. The best time for ionospheric stability is between 2200 and 0600 local time at both stations.

Another factor entering into echo quality is *scintillation,* which cannot be corrected with circular polarization. An uneveness of electron density forms in the ionosphere and acts on a signal much like a lens on light. These "blobs" can have a focusing or defocusing affect on a signal producing unrealistically strong echoes, or no echoes at all. Scintillation, from my observations, is more apparent at frequencies below 144 Mc.

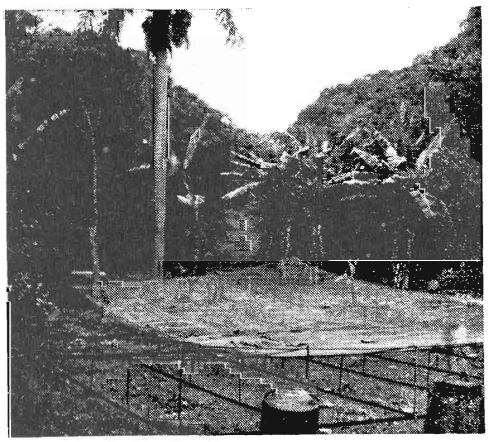


New Zealand, e.m.e. buff Ralph Carter, ZL1TFE, (left) recently visited K6MYC in Saratoga, California. Ralph is actively working towards e.m.e. contacts on 144 and 432 from his home in New Zealand.

Reprinted from February, 1968, QST

Libration fading caused by the rocking motion of the moon also effects echoes. For short periods the path loss can be reduced by as much as 6 to 10 db. The moon is a rough surface and acts like many reflectors. Sometimes they add up in phase, while on the average they give a seven per cent πr^2 reflectivity. Libration spread is more troublesome at frequencies higher than 144 Mc.

Another factor having a large bearing on whether or not contacts can be made with marginal systems is *cosmic noise*. On 432 and above this should not cause much concern, but at 144 it is a different story. The minimum cosmic noise at 144 Mc. is about 1.9 db., which is quite easily heard with modern transistors. Cosmic noise is greatest is the direction of the Milky Way, or the galactic center. From my experience cosmic noise can make a 2-db. receiving system perform like a 6-db. system, or worse, when the moon is near the galactic center. There is usually a period of five to seven days each month when the moon is at its lowest declination angles. These days should be avoided if success depends upon



W1FZJ/KP4 is building this 50 foot square "dish" for 432-Mc. e.m.e. tests. The dish will later be expanded to 150 feet for use on 144. A movable feed will be mounted atop a 60-foot tower in the center of the dish

optimum receiving capabilities. Even the period as the moon is increasing its declination to its peak of approximately 27 degrees is not especially quiet. In my opinion the ten-day period after the moon has reached its declination peak is the best.

The following suggestions are offered as possible solutions to the problems just discussed:

1) Faraday rotation can be handled with circular polarization, or in part by carefully planned schedule times.

2) Larger than minimum antennas help overcome scintilation, libration fading and cosmic-noise effects.

3) An effective method of reporting and confirming signal reports helps in completing information exchanges. Avoid using code characters requiring dots, such as the letters I, E, S and H, and the numbers 2 through 7. The following system is currently being used by those scheduling VK3ATN:

T — signals detected
M — letters or portions of calls copied
O — Both calls and report copied
MT — nearly solid copy
5 — solid copy, no need for code

By this system an O plus both calls received at both ends and confirmed with RRR establishes a contact. Had this sytem been in use for my November 22nd test with VK3ATN we probably would have made another contact. However, by the old system VK3ATN was sending 3s represented by the letter E. Es are easily lost to fading and are sometimes not discernible from noise pips ringing in narrow-bandwidth audio filters. Especially after many hours of listening for weak signals, dashes are much easier to detect.

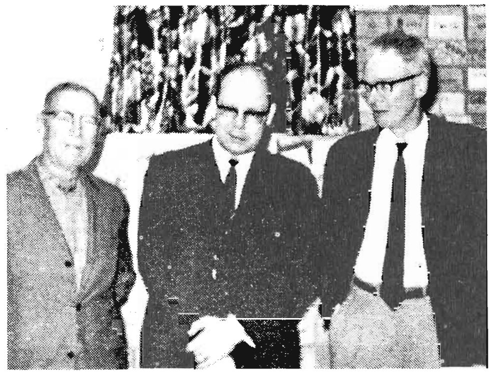
4) Receiving system modifications such as post detection, phase lock, noise blanking and cancellation all can help find signals in the noise. F.s.k. should offer a 3 db. signal-to-noise improvement and is an area for experimentation.

5) Keep transmitting and receiving periods short. I prefer 1 to 2-minute periods, particularly in daylight hours when Faraday rotation is rapid; 90 degrees every 15 to 30 minutes. Echoes can appear, peak and fade in 5 minutes or less. Five-minute periods are used by many, since some detection schemes require 3 to 5-w.p.m. c.w. speed for proper integration time.

6) Use relatively slow-speed c.w., under 10 w.p.m. When testing with VK3ATN my transmission periods are two minutes long. During the first minute each call is sent 2 or 3 times, and the report is sent the second.

7) Be sure of your frequencies, times and calling sequences. Frequencies must be within one kilocycle.

8) Keep your antenna as close as possible on the moon. If your antenna has a 5-degree beamwidth at the 3 db. points, you probably can't afford to be 5 degrees off. It is worthless to build a good antenna system and then waste it with poor



Willis Brown, W3HB, Bethesda, Maryland, recently hosted Andy Kalt, DL8PK, Wahn, West Germany (center), and Bill Smith, W3GKP, of early moonbounce fame. DL8PK is active on 2 meters in Germany. By the way, Massachusetts meteor jockey W1JSM is the son of W3HB.

aiming. This has been the principal cause of many e.m.e. failures.

9) Don't start listening for echoes in a narrow bandwidth (under 500 cycles) unless you are experienced or have a receiving system that requires it. I prefer an 800 to 1000-cycle bandwidth but most of my receiving is done in a 2.1-kc. bandwidth, with the ear providing the "selectivity."

10) When searching for weak echoes, continuously sweep the 500 to 1000-cycle portion of the band where the signal should be. I've found I can detect signals this way that might otherwise be lost in the noise. The ear can detect pitch changes easier than a steady note.

11) Doppler shift on two meters is not much of a problem. I've never heard an echo shift more than 500 cycles at 144 Mc. If the moon is rising, the signal will appear high in frequency. As the moon passes due south there will be little or no shift; then as the moon begins to set, the echo will appear lower in frequency. When listening for your echo from a rising moon, set the receiver so the transmitted signal produces a 200 to 300-cycle note. The echo will then produce a 500 to 700-cycle note. The opposite is true of a setting moon. Doppler on 432 and higher is of more concern and will produce a 1-kc. shift or more, except when the moon is due south of your antenna.

Next month's concluding discussion of this series will cover antenna mounts, drives and readout systems.

E.M.E. for the Layman — Conclusion

This month we conclude a three-part discussion of e.m.e. (earth-moon-earth) principles by Mike Staal, K6MYC. The final section covers antenna mounts, drive systems and readout mechanisms.

First the prospective moonbouncer must decide if he is going to use his antenna system for anything other than e.m.e. experiments. This decision governs the selection of an appropriate mount and drive system. A very simple mount can be constructed if the antenna is to be used only for e.m.e. and thus be aimed at a specific point in space. This may be a logical place to begin, but you will probably soon become frustrated at being limited to perhaps 5 or 6 hours each month when the moon passes through the antenna's pattern. I suggest at least a partially-steerable array.

If only e.m.e. is contemplated, a polar (or equatorial) mount would be a wise selection as it requires only one drive mechanism for tracking and some form of manually tilting the array slightly from day to day to set the *declination*[1]. To accomplish this, your antenna mast or tower must be mounted parallel to the axis of the earth. Thus, if your station location is at 35° north, the mast would be fixed at an angle of 35° from the earth's surface at such location, oriented in a north-south direction (see fig. 1.). The declination (manually-tilted axis) changes from day to day. Information may be found in *The American Ephemeris and Nautical Almanac*, 1968, available through the Superintendent of Documents, U.S. Government Printing Office, Washington, D.C. for a nominal price. All that is necessary now is that your drive mechanism rotate the antenna at a rate of 15° per hour to track the moon.

This is all fine and dandy for e.m.e., but if you want to use your array for satellites, meteor scatter, aurora or something similar, a polar mount is not much good. A drive system permitting the array to be fully steerable in both azimuth and elevation (az-el) is the answer.

The array at K6MYC is mounted atop a homemade 12½-foot tower. The four legs of the tower are fastened to a platform which in turn is bolted to the roof of the garage directly above the operating position. A large unmodified prop pitch motor is mounted inside the top of the tower. A husky steel plate is welded to the rotating gear and another plate is attached to the first with ordinary door hinges, see the photographs. These hinges are employed in the elevating mechanism. To this plate a 3-inch aluminum channel is attached and the main boom of the array is clamped in this channel. A jack screw with right-hand left-hand square threads starting from the center out raises and lowers the array. At the lower end of the jack screw is a 20-to-1 gear reduction box giving a zero to 90° elevation time of three minutes. With the plates together the array is pointing straight up. The entire elevation drive rotates with the array.

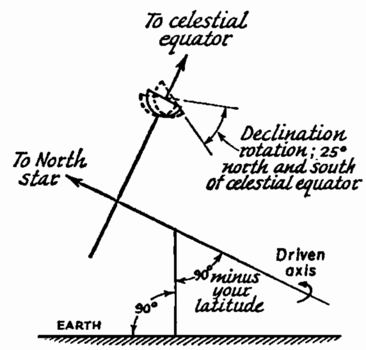
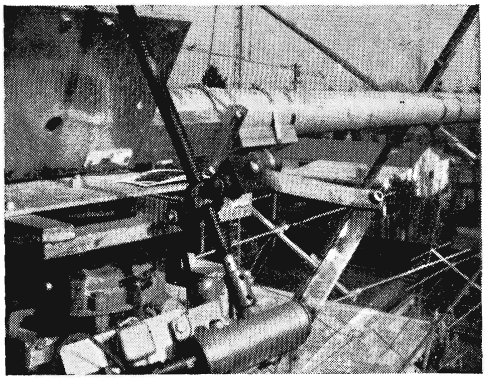


Fig. 1.

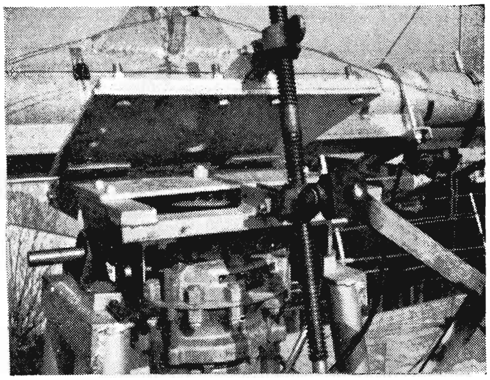
Selsyn hookups are used for direction readout and may be varied to suit the particular builder. I'll let you work out your own azimuth system, but my elevation selsyn mount is quite simple. The selsyn is attached to the main array boom and aligned with it. A weight was tightly affixed to the selsyn shaft and, of course, the weight always hangs straight down regardless of the position of the array. The mates to both selsyns are mounted on a panel in the shack. Crude, perhaps, but it gives *one-degree accuracy*, and in e.m.e. you can't afford less!

A handy item for telling if your array is pointing at the moon is the RCA SQ2520 photo-cell costing about $2, or its equivalent. This device is sensitive enough to detect the light of even a small sliver of moon. When placed at the end of a 20-inch long one-inch diameter tube and the leads connected to an ohm meter, it is an

Reprinted from March, 1968 QST



Mounted on the lower end of the jack screw is the 20-to-1 reduction system. Note the collinear elements and main boom.



The elevation selsyn is mounted on the boom to the right of the mount. Note the jack screw, elevation plates and channeling holding the main boom on the mount assembly.

accurate indicator of proper aiming. Obviously it must be mounted so to be aimed along the exact plane of your array. It is useful only at night when the moon is visible.

As can be seen, the problems of mounting, steering and controlling an e.m.e. array are mostly mechanical and must be left to the ingenuity of the builder. Following the basic principles given here on locating the moon the builder may develop his own system.

It has been a pleasure to present these notes on e.m.e. problems, and it is my hope that many of you will become interested in building your own e.m.e. system. — *K6MYC*

The World Above 50 Mc.

CONDUCTED BY BILL SMITH,* K4AYO

Beginning Moonbounce—101

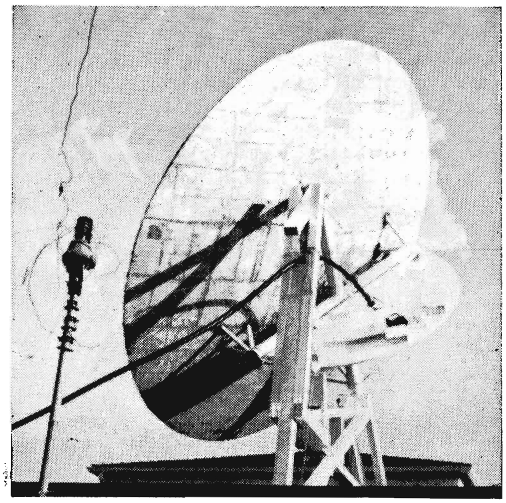
Each month we receive letters from prospective moonbouncers inquiring for reference material, and hints how to begin their e.m.e. project. In this column for January, February and March 1968 appeared a three-part series by K6MYC, designed especially to answer the most common questions. For those who do not have these issues, we'll paraphrase some of the highlights this month, but suggest you obtain the originals from a friend, or ARRL Headquarters at the nominal fee of 75 cents each.

Basically, this is what is required: 500 watts or more of transmitter output, the best possible receiver front end, a bigger antenna than most of us will ever erect, the means of aiming the array at the moon, and much perseverance. All, but the latter, may be store-bought, if you're so inclined.

Lets look at each. The transmitter power is easily acquired at 144 MHz., the most popular e.m.e. band, 220 MHz., where apparently there is no active e.m.e. work, and at 432 MHz. 1296 and up are progressively more difficult. There are numerous transistors capable of achieving the necessary noise figure at 144, many in the one dollar price range. The picture doesn't change too much at 432; at 1296 the device will cost 10 dollars or more.

The antenna, its type, size and aiming, may be considered together. Success has been had with collinears, Yagis, rhombics and dishes, or parabolic reflectors. The most popular, because it is tolerant of less-than-optimum amateur construction techniques, is the collinear. K6MYC designed, and later discussed in the April, 1967 edition of this column, a modification of a commercially available collinear. The modified version of that antenna is now on sale. At 2 meters, it is probably the best available commercial antenna.

Both SM7BAE and ZL1AZR, who together hold the world's 144-MHz. e.m.e. record, use multiple-Yagi arrays. Another promising 2-meter Yagi array was described by Oliver Swan at the recent West Coast V.H.F. Conference. In tests at K6MYC, a four-bay array of these Yagis, spaced 80 inches both horizontal and vertical, recovered the same amount of e.m.e. signal from KØMQS as did a 40-element collinear array. Physically the collinear array is about three times as large as the Yagi array. Details of this antenna will appear soon in *QST*.



Mounted on the roof of his Los Angeles home, this is the homemade dish of WB6IOM. He used this dish to successfully work G3LTF and establish a new 1296-MHz. moonbounce record. The 16-foot diameter dish consumed 450 square feet of sheet aluminum and 70 pounds of epoxy to bond the aluminum sheets. (WB6IOM photo)

Rhombics, used with much success by VK3-ATN and KØMQS, are capable of developing gain in excess of 30 db. over isotropic. Their disadvantages are physical size (several hundred feet in length) and fixed direction, except in the case of VK3ATN who varied the direction a few degrees by a pully and track arrangement. Rhombics are not feasible at the average city amateur location.

The parabolic reflector, more commonly known as a "dish," is essentially a low-efficiency antenna, something in the order of 35 percent. In addition, because of its physical size, especially at 144 and 432, it is not practical for the back-yard e.m.e. enthusiast. However, at 1296 and higher, good gain can be developed from a modest size dish. A picture of WB6IOM's 16-foot dish, used in establishing the world's e.m.e. distance record on 1296, appears elsewhere in this column.

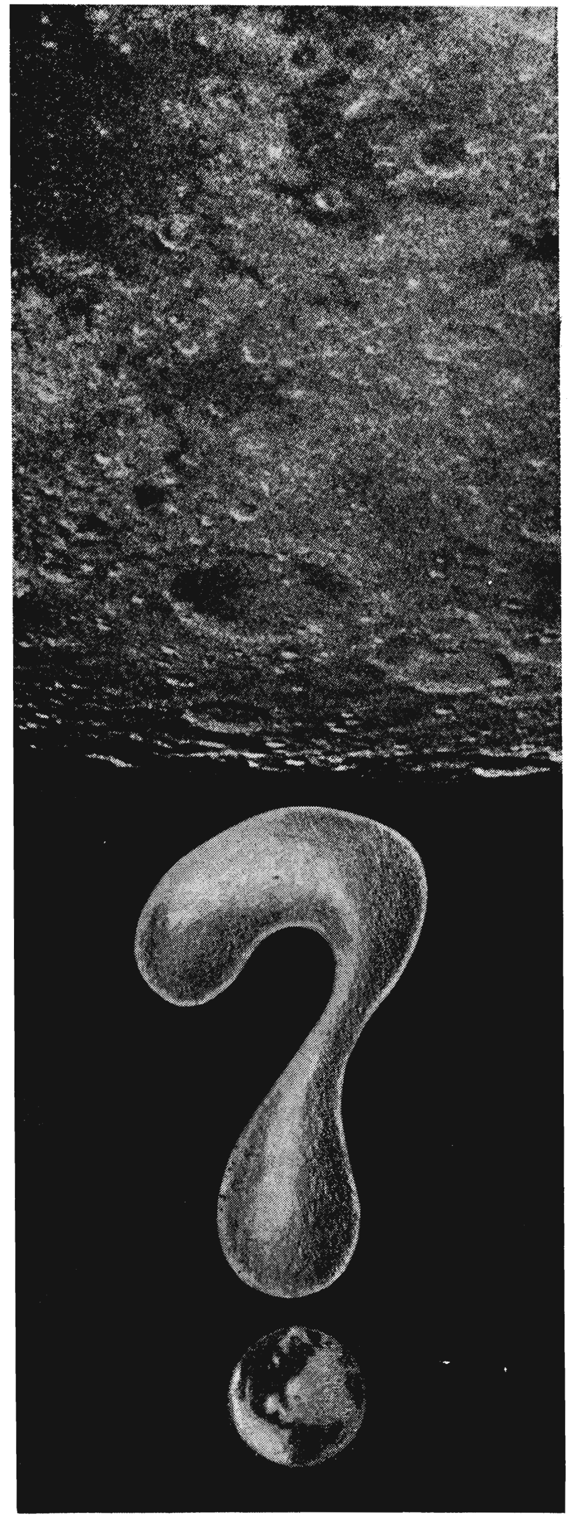
Even more important is how you aim the array. It matters not how much gain the array has if it can not be aimed at the moon. Three systems are available; fixed position, partially steerable (polar mount), and fully steerable. A fixed-position array is the simplest to build. You have only to determine the place in space where the moon will travel through the array's pattern at a given time, and fix the array in that position. This method, however, limits the time each month the moon will pass through the antenna's pattern, and who you can work because of matching the "window." The window is a mutual place in space where antennas at both stations are pointed at the moon simultaneously.

The partially steerable, or polar mount, antenna is especially suitable for e.m.e. work. It needs to be set only once daily for declination (the angle in degrees north or south of the celestial Equator, or elevation angle) and then rotated in azimuth (horizontal plane) to track the moon. The moon travels across the sky at approximately 15 degrees per hour.

A fully steerable array, in both azimuth and elevation (az-el mount), is more flexible for use on other propagation modes, but is difficult mechanically to construct and calibrate for e.m.e. purposes. This is the most desirable type for satellite work.

All right, we've thrown out some facts; what do they boil down to? For the e.m.e. neophyte I'd suggest the following, and you e.m.e. greybeards may sit back and stroke them. Try 144 MHz., there is more activity, and technically 2 meters is more easily achievable. Construct a collinear array of at least 160 elements. That puts you into the 20-db. gain e.m.e. ballpark. Mount the array in a fixed position, taking into consideration who you wish to schedule. The mount may be modified at a later date to a polar configuration, after you become more familiar with e.m.e. techniques.

Much of this discussion may be directly applied to satellite programs, hopefully to soon again grace the amateur horizon through the Amsat and Nastar projects. E.m.e. and satellite work is within the grasp of many of us. As KØMQS recently said, "if I can work e.m.e., anyone can." What Dick said is true — *if* you have the perseverance to put the system together, and stay with it until it works. You still can't buy that!



How High the Moon

BY DON LUND,* WAØIQN

Reprinted from July, 1965 QST

The year 1964 will long be remembered by v.h.f. moonbounce enthusiasts. First, the patient work of Bill Conkel, W6DNG, and Lenna Suominen, OH1NL, paid off with the first two-meter moonbounce contact, and then KP4BPZ really showed the possibilities of such work. Postmortems on the week end of June 13 and 14 were held wherever v.h.f. men gathered, but one aspect of our participation seems clear. Many groups and individuals depended upon their ability to visually align their antennas on the moon. Cloudy weather meant failure: partly cloudy weather meant disastrous breaks in tracking the moon.

Getting around this trouble is really pretty easy. First, you need to know where your antenna is pointed. If you're using good rotators, the indicators tell you. If you're using an "Armstrong" system, attaching "setting circles," which are circular dials with 360-degree markings, will tell you. Now the only thing you need to know is where in the heck the moon is!

The purpose of this article is to show two ways of calculating where the moon will be in the sky at a given time on a given date. The first way is quick and dirty. With no mathematics and no references other than this article, it will predict the moon's position to an accuracy of 5 degrees or so for observers within the United States. Since an antenna with an honest 20-db. gain will have a half-power beam width of about 13 degrees, 5-degree accuracy should be acceptable for most applications. If this isn't good enough, a second way is described. It is both accurate and tedious. To use it, one needs a table of trigonometric functions and one reference book. Either of these methods will help you aim your antenna at the moon in fair weather or foul.

All of this discussion will be in terms of elevation and azimuth coordinates. Elevation is the height in degrees of the center of the moon above the horizon. Azimuth is the bearing of the moon, measured clockwise from North. For example, the elevation of the horizon is 0 degrees, and the elevation of a point directly overhead is 90 degrees. The azimuth of the eastern horizon is 90 degrees, while the azimuth of the southern horizon is 180, and so on. We are going to stick to "az-el" coordinates because this is the simplest type of mounting for an amateur to build and align. Also, because of the moon's rapid motion in declination (declination is the same, in celestial coordinates, as latitude in geographical coordinates), other types of mountings do not offer the advantage for the moon that they do for heavenly bodies with fixed declinations.

The Moon's Position; Quick Way

If we watch the moon's path across the sky for a month or so, we see that it shows a cyclic variation. The moon might, on the first night, rise quite high in the sky. The next night it would not rise quite as high, and the next night it would be even lower. After about 13 days it would be lowest in the sky, and the next night it would be

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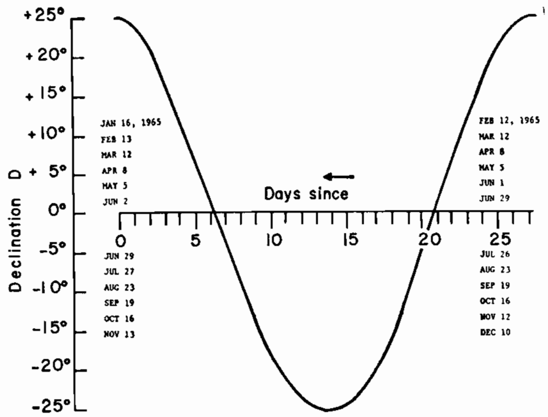

Fig. 1—The average declination of the moon during 1965.

higher again, until after about 27 days, it would be at its highest again. This is because the moon wobbles in declination. The wobble is almost sinusoidal, with a period of about 27 days, as shown in Fig. 1. The dates given are the starting dates for the oscillation. Since the period isn't exactly 27 days, it is necessary to slip a day every so often, as on September 18–19. The maximum amplitude changes about 1½ degrees during 1965, so the curve shows the mean declination, D, during 1965. Thus on July 26, 1965, the moon's declination is + 25 degrees. On August 2, seven days later, the declination is 0, while on August 3, D = − 4 degrees.

Knowing the moon's declination, we may compute its path across the sky (see Fig. 2). We see that when the moon's declination is most positive, it passes highest in the sky; when the declination is most negative, it is lowest in the sky. At some time, call it T, the moon is due South. At $T - 1$, that is one hour before T, the moon is on a solid line corresponding to the declination from Fig. 1, where it crosses the dashed line marked "$T - 1$." An hour and a half later, the moon is still on the same solid line, and is where the dashed line marked "$T + ½$" crosses it. At $T - 1$ and $T + ½$, we can read the moon's azimuth and elevation off the bottom and side scales. One word of caution about Fig. 2: It has been computed for an observer whose latitude is 40 degrees North, which is on a line passing through San Francisco, Indianapolis and Philadelphia. For observers north and south of this line, the elevation scale is squeezed or stretched. However, for the kind of accuracy we need, the curves will produce acceptable results over most of the continental United States, except Texas, Florida and Maine.

All that is needed now is to find the time, T, at which the moon is due South. This is shown in Fig. 3. Again, the dates are the starting times of the periods, which are about 29 days long. The time can then be read directly in local standard time. For example, the moon is due South at midnight on July 12, 1965. On August 3, 22 days later, the moon should be due South at about 4:40 P.M. local standard time. As before, Fig. 3 represents an average curve for 1965, computed for an observer at the middle of the United States. East and West Coast times may be off by several minutes.

In summary, the complete procedure is:

a) Given the date, find D from Fig. **1**.

b) Given the date, find T from Fig. 3.

c) Knowing D and T, enter Fig. 2, reading off azimuths and elevations at hourly intervals before and after T. For illustration, let's say we want the azimuth and elevation of the moon on August 3, 1965. From Fig. 1, $D = -4$ degrees, and from Fig. 3, $T = $ 4:40 P.M. In Fig. 2, the $D = -4$ degrees curve must lie a sixth of the way down from $D = 0$ degrees to $D = -25$ degrees. Pencilling a curve like that in, at $T - 3$, that is at 1:40 P.M., the azimuth is 127 degrees and elevation is 29 degrees. At 2:40 P.M., the azimuth is 141 degrees, and elevation 32 degrees. Following along, we can find elevation and azimuth every hour. Sounds a little complicated at first, but with some practice, it becomes quick and easy.

The Moon's Position: Exact Way

For the man who has everything — a 300-foot dish and an IBM computer — the easy way may be neither satisfying nor accurate enough. For the man with such excellent capabilities, we offer a cookbook which shows one way of computing the moon's elevation and azimuth. We won't define things like hour-angle, for these definitions would constitute a full course in astronomy. Rather, we will just tell you how to compute, and let you study the references if you wish.

The first step is to compute the local sidereal time, which we call T_s. Pick a Greenwich Mean Time, T_g, for which we want to compute the moon's position in the sky. At this point we must refer to *The American Ephemeris and Nautical Almanac, 1965* (or whatever year you wish) which is available from the Superintendent of Documents, U. S. Government Printing Office, Washington, D. C. Copies are often available at nearby observatories, and occasionally at nearby universities. In the *Ephemeris*, under the section

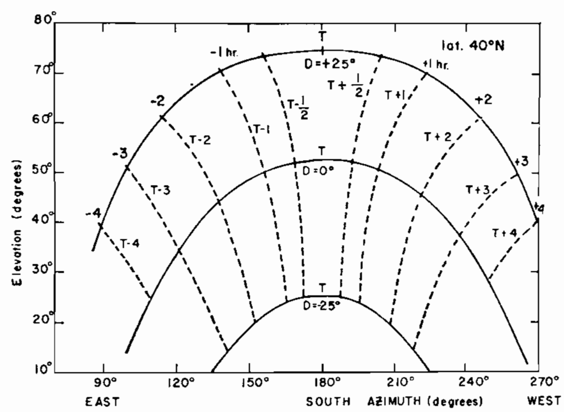

Fig. 2—Azimuth and elevation of the moon.

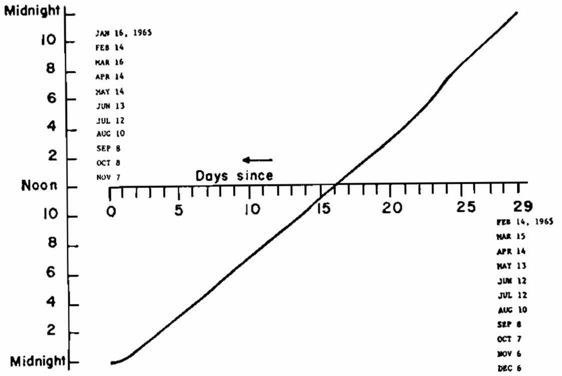

Fig. 3—The average local standard time at which the moon appears due south.

titled "Universal and Sidereal Times, 1965" there is a column called "Sidereal Time, Hour Angle of the First Point of Aries — Apparent." One entry is given for each day of the year. Look up the value for the date you wish, and call the value T_a. Then the local sidereal time may be computed from

$$T_s = 1.002778 T_g + T_a - 24 \frac{l_1}{360}$$

where l_1 is your longitude in degrees, west of Greenwich. Next, compute the hour angle of the moon; call it *h.a.m.* To do this, in the *Ephemeris*, in a section titled "Moon, 1965, For Each Hour of Ephemeris Time," for each date there is a column showing "Apparent Right Ascension" for each hour of time. Look up the Apparent Right Ascension for the date and time of interest; call it *r.a.m.* Then the hour angle of the moon, in degrees is:

$$h.a.m. = (T_s - r.a.m.) \frac{360}{24}$$

Next, we compute the elevation of the moon; call this angle E. This is computed from

$$\sin E = (\sin D \times \sin l_2) + (\cos D \times \cos l_2 \times \cos h.a.m.)$$

where l_2 is the latitude of the observer and D is obtained from the column "Apparent Declination" which is just to the right of the "Apparent Right Ascension" column in the *Ephemeris*. Having found E from the tables of trigonometric functions, look up $\cos E$. Then the azimuth, A, can be computed from:

$$\cos A = \frac{\sin D - \sin l_2 \times \sin E}{\cos E \times \cos l_2} \quad \text{and}$$

$$\sin A = \frac{\cos D \times \sin h.a.m.}{\cos E}$$

From the trigonometric tables, we can then look up A.

For the person who needs this accuracy, and has access to an IBM computer, a Fortran program for the above may be obtained by writing the author.

Summary

To permit aiming antennas at the moon through cloudy skies, we have shown two ways of computing the position of the moon in the sky. The first way is as simple as we know how to make it. Its accuracy is poor by astronomical standards, but should be sufficient for most amateur applications. The second way is more accurate, but involves tedious computations. We comment that we have ignored certain fine points in the second method, such as the difference between Ephemeris and Greenwich Mean Time and the fact that the Ephemeris values of right ascension and declination are as seen from the center of the earth. Such refinements can be introduced if the need for ultimate accuracy arises.

References

For a general reference which provides an excellent introduction to the terms and ideas used here, we would recommend *Astronomy*, by R. H. Baker (D. Van Nostrand Co., Inc., 1960). For more detailed information, which includes the derivation of expressions like those which we have used in the Exact Method, we could recommend *Elementary Mathematical Astronomy*, by C. W. C. Barlow and G. H. Bryan, as revised by H. S. Jones (University Tutorial Press, Ltd., 1961). Tables in Figs. 1, 2 and 3 were supplied by the High Altitude Observatory, Boulder, Colorado. QST

Tracking the Moon—In Simple English

Practical Ideas for Designing and Aligning a Polar Mount

BY VICTOR A. MICHAEL,* W3SDZ

A MAJOR pitfall facing the prospective moonbouncer is the antenna mount and tracking system. Even a 50-foot dish is of no value in lunar communication, if it cannot be pointed at the moon and kept there. When we began our moonbounce efforts, many hours were spent pouring over astronomy texts. It was determined rather quickly that a whole new language would have to be learned for a proper understanding of the moon-tracking problem. Gathered together here are some of the essentials involved.

Earth-Space Relationship

Understanding the earth-moon relationship in space is the first step in solving the moon-tracking problem. This relationship is best illustrated with a polar mount, as in Fig. 1. A polar mount is simply an elevation-azimuth mount with its azimuth axis parallel to the axis of the earth. Thus a polar mount at the equator would have its axis parallel to the earth (horizontal, to the viewer on the ground), while at the North Pole the axis would be vertical, or at a 90-degree angle to the plane of the earth. Your latitude determines the position of the polar axis with respect to the earth's surface, as illustrated in Fig. 1. Once this is determined, we can proceed to a few other terms.

Celestial Equator. An extension of the earth's equator; the circle that would be formed at a right angle around the polar axis.

Meridian. The north-south line directly overhead.

Hour Angle. The angle in degrees to the right of the meridian. (Degrees can also be transferred into time: 15 degrees equals 1 hour; 1 degree equals 4 minutes.)

Declination. Angle in degrees north or south of the celestial equator.

Using the Nautical Almanac

This is the most important tool you will use in setting up, calibrating, and using your moonbounce antenna. It is available from the U. S. Government Printing Office for \$2.00. Be sure you get the right book; there is a similar publication from the same source titled *The American Ephemeris and Empirical Nautical Almanac.* This is more expensive, and harder to use for amateur applications.

On page 39 is a portion of the tables found in the *Nautical Almanac.* It will be noted that the position of the moon is plotted for each hour of GMT. As an example, at 1200 GMT Jan. 1, 1965, the GHA (Greenwich Hour Angle) is 15 degrees 16.5 minutes. This means that the moon has passed overhead at Greenwich, and is now 15 degrees 16.5 minutes, or just over one hour, to the right of the meridian, as the observer faces south. The declination is given as S 23 degrees 39.5 minutes, which means that the moon is at this position south of the celestial equator.

Once you know where the moon is at Greenwich, a simple formula may be applied to determine its position with respect to your own location. The declination is always the same, no matter where you live. The only factor that changes is the hour angle. The Local Hour Angle (LHA) can be obtained by the formula

$$\text{LHA} = \text{GHA} \begin{matrix} -\text{ west} \\ +\text{ east} \end{matrix} \text{ longitude.}$$

Getting back to our example, suppose you live at 75 degrees west longitude. We find that the moon would have an LHA of 300 degrees 16.5 minutes, or approximately 4 hours before meridian.

Mount Design Considerations

After you examine your almanac you will discover a few facts about the moon's habits that will help you to design a mount. First of all, the moon spends about two weeks above the celestial equator and about two weeks below. The maximum declination is about 25 degrees

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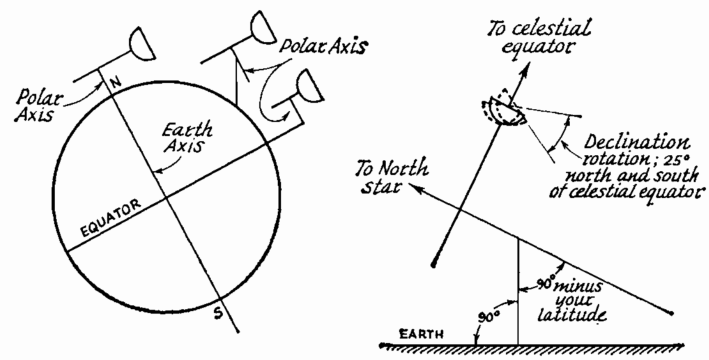

Fig. 1—Principles of the polar mount for moon tracking. At the left it can be seen that the polar axis is always parallel to the axis of the earth. Its position with respect to the earth's surface depends on the latitude of the observer. Two planes of rotation are required; the declination, which may be varied a small amount from day to day as required; and hour-angle, which should be controlled with a clock drive to follow the moon.

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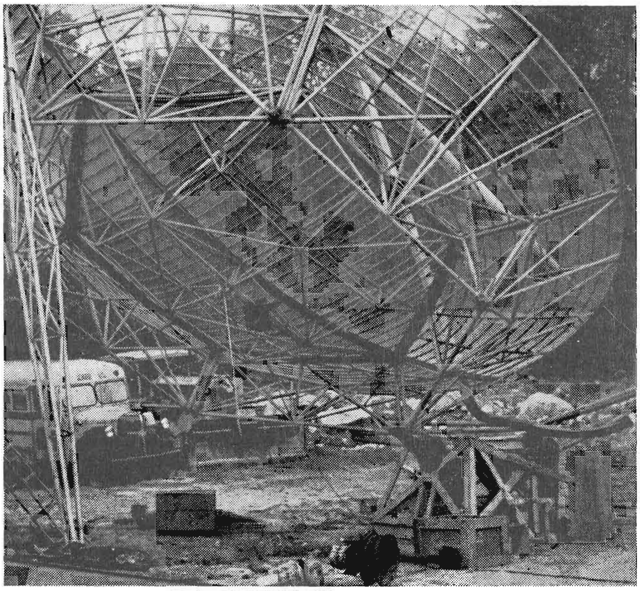

Fig. 2—Simplified polar mount and 28-foot dish at W1BU. Principles of the mount and its clock drive are explained in the text.

north or south of the celestial equator. Tracking ability for about 3 to 5 hours of hour angle each side of local meridian should be satisfactory.

At this point it is possible to make some compromises in order to simplify the mount in favor of a larger antenna. For instance, at W1BU, Sam gave up two weeks out of a lunar month in order to use the 28-foot dish of Fig. 2 in recent moonbounce tests. He can elevate the antenna above, but not below, the celestial equator. The high edge (upper left in the picture) is elevated to the desired position, while the lower edge rests on the pedestal at the lower right. The hour angle is controlled by a clock drive, just visible at the lower center. Though complex enough, this is far simpler than the true polar mounts used on the 18-foot dish at W1BU, or the mount and drive for the 256-element collinear array at W3SDZ, Fig. 3.

Calibration of the Mount

Obviously, if you are going to use the information in the *Nautical Almanac* with your mount, there must be some system of readout. There are

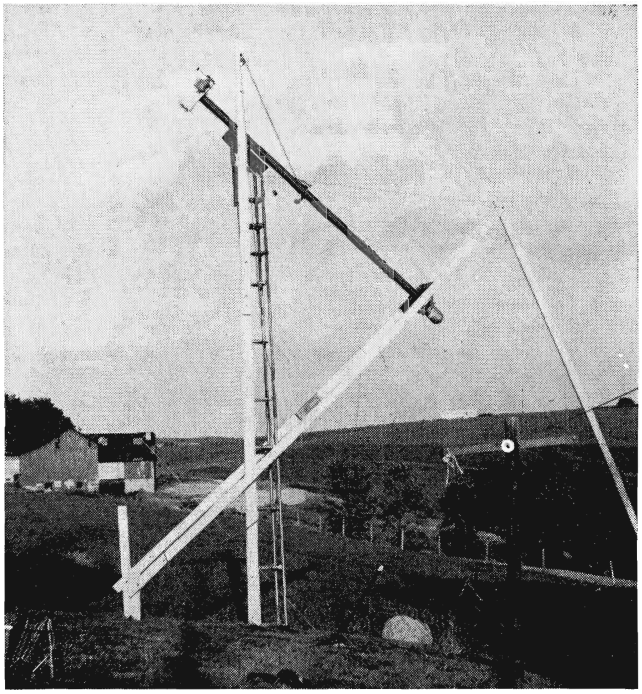

Fig. 3—Polar mount at W3SDZ, before the 256-element 432-Mc. collinear array was in place. The complete array is pictured in November 1964 QST, page 75.

many possibilities, and many different systems will evolve. As a starting point, a few ideas will be discussed here, and then "to each his own."

As a practical matter, the declination need be set no more than once per day, for it changes less than 2 degrees in 24 hours. For a few hours of moonbouncing effort each day, less than 1-degree variation is involved. Unless your antenna pattern is much sharper than the best amateur efforts to date on frequencies below 1300 Mc., this error is no problem. Declination readout can be rather simple: a calibrated scale on the antenna mount, a selsyn readout, or even a good 10-turn pot geared to the declination axis, and connected to a mercury battery and a meter. Anything accurate to plus-or-minus 1 degree should be all right.

Hour-angle readout and automatic tracking are the chief problems in moon tracking. The moon appears to move across the sky at slightly less than 15 degrees per hour. Actually, it is the earth that is moving at 15 degrees per hour. The moon is also moving, but at less than 1 degree per hour. Thus our basic problem is to drive the polar or hour-angle axis at 15 degrees per hour with a clock. The simple procedure of turning off the hour-angle drive for about 3 minutes once each hour, until the moon catches up, keeps things more than accurate enough for antenna tracking.

Now a "clock" doesn't necessarily have to look like a clock. For instance, a large synchronous 60-cycle motor driving a gear train at 1 revolution per day, coupled directly to the hour-angle axis, will work. The W1BU system is shown schematically in Fig. 4. Actual readout can be by any method that will develop plus-or-minus 1-degree accuracy.

When the mount is made, the antenna mounted, and the readout devices reading, the next question will be where is the antenna *really* pointing? This may sound simple, but most would-be moonbouncers have had trouble with this problem. Fortunately, nature has provided us an almost constant radio signal that permits a rather accurate calibration of the antenna system to be made. That signal comes from the sun. It will be seen that the almanac gives identical information on GHA and declination for the sun; thus, by listening to solar noise you can calibrate your polar mount in the same terms of reference as you will use in moon tracking. Some time spent reading K2LMG's "Antenna Patterns from the Sun," *QST* for July, 1960, will be well spent at this point.

What you have just read covers some of the essentials. It is hoped that enough information has been given to enable the prospective moonbounce enthusiast to determine his requirements for mounting and tracking. If any serious experimenter in this field needs help at this point, the author will be glad to try to be of assistance.

QST

G.M.T.		SUN		MOON				
		G.H.A.	Dec.	G.H.A.	v	Dec.	d	H.P.
d	h	° ′	° ′	° ′	′	° ′	′	′
1	00	179 09·2	S23 02·3	201 09·9	11·7	S22 48·3	4·9	54·0
	01	194 08·9	02·1	215 40·6	11·7	22 53·2	4·7	54·0
	02	209 08·6	01·9	230 11·3	11·7	22 57·9	4·7	54·0
	03	224 08·3	· · 01·7	244 42·0	11·6	23 02·6	4·5	54·0
	04	239 08·0	01·5	259 12·6	11·6	23 07·1	4·4	54·0
	05	254 07·7	01·3	273 43·2	11·5	23 11·5	4·4	54·0
	06	269 07·4	S23 01·1	288 13·7	11·6	S23 15·9	4·2	54·0
	07	284 07·1	00·9	302 44·3	11·5	23 20·1	4·1	54·0
	08	299 06·8	00·7	317 14·8	11·4	23 24·2	4·0	54·0
F	09	314 06·5	· · 00·5	331 45·2	11·5	23 28·2	3·8	54·0
R	10	329 06·2	00·3	346 15·7	11·4	23 32·0	3·8	54·0
I	11	344 05·9	23 00·1	0 46·1	11·4	23 35·8	3·7	54·0
D	12	359 05·6	S22 59·9	15 16·5	11·3	S23 39·5	3·5	54·0
A	13	14 05·3	59·7	29 46·8	11·4	23 43·0	3·5	54·0
Y	14	29 05·0	59·5	44 17·2	11·3	23 46·5	3·3	54·0
	15	44 04·8	· · 59·3	58 47·5	11·3	23 49·8	3·2	54·0
	16	59 04·5	59·1	73 17·8	11·2	23 53·0	3·1	54·0
	17	74 04·2	58·8	87 48·0	11·3	23 56·1	3·0	54·0
	18	89 03·9	S22 58·6	102 18·3	11·2	S23 59·1	2·9	54·0
	19	104 03·6	58·4	116 48·5	11·2	24 02·0	2·7	54·0
	20	119 03·3	58·2	131 18·7	11·2	24 04·7	2·7	54·0
	21	134 03·0	· · 58·0	145 48·9	11·1	24 07·4	2·5	54·0
	22	149 02·7	57·8	160 19·0	11·2	24 09·9	2·5	54·0
	23	164 02·4	57·6	174 49·2	11·1	24 12·4	2·3	54·0

Table I—Section of a page from the *Nautical Almanac,* showing solar and lunar data for each hour of January 1, 1965, at Greenwich.

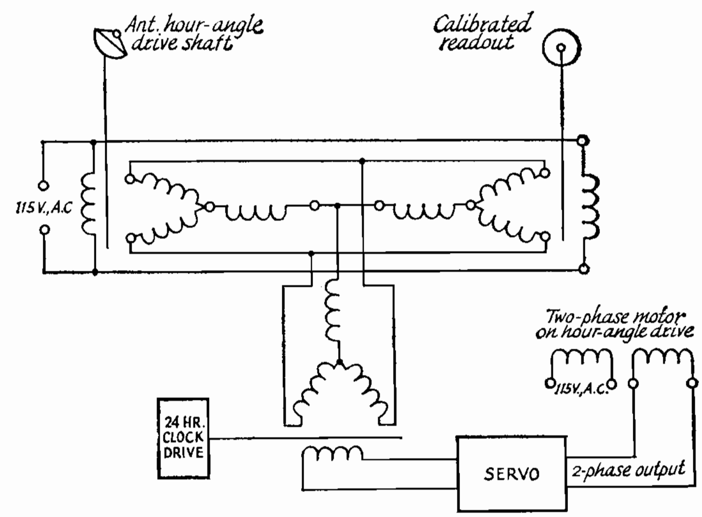

Fig. 4—Schematic diagram of the clock drive and readout system at W1BU.

USING SUN NOISE

BY DON LUND,* WAØIQN

One may hear the question "How many db's of sun noise do you get?" asked among serious v.h.f. men. Checking system performance of an advanced v.h.f. station by measuring the amount of noise received from the sun can be most useful, but there are some pitfalls that must be avoided. Our aim here is to set out the relationship between the power radiated by the sun, the antenna characteristics, and the receiver performance. If we know any two of these sets of parameters, we can measure the third. Finally, we'll explore some of the pitfalls inherent in talking about "db's of sun noise."

Solar Temperature, Antennas and Receivers

Twenty years or so ago, astrophysicists were arguing over whether the outer atmosphere of the sun was hotter than the visible surface. Radio astronomy provided some of the evidence that the outer atmosphere was much hotter than most astrophysicists had previously imagined. The result was that the "apparent temperature" of the sun increased with wavelength, at all wavelengths longer than a centimeter or so. (Apparent temperature comes in because the size of the sun is different at different wavelengths. So the sun is taken to be the same size as the optical sun, and apparent temperature is the temperature it would have, to radiate the measured power, at given wavelength, from this size of disc.)

What happens if we point an antenna at the sun? If the beamwidth of the antenna is just exactly the size of the sun, the antenna temperature will be the same as the temperature of the sun at this wavelength. Antenna temperature doesn't mean that we could burn a finger on the antenna; it means that the antenna is delivering the same amount of power to the transmission line that would be delivered by a resistor heated to the antenna temperature. This means that if we took a 50-ohm resistor, and heated it to 400,000°K, it would generate the same amount of noise at 432 Mc. as would be delivered to a 50-ohm resistor by an antenna with a ½-degree beamwidth pointed at the sun.

Antennas used by hams are not that sharp. If an antenna with a 10-degree beamwidth were pointed at the sun, its gain would be less, and it would deliver less power to the transmission line than a ½-degree antenna. Said differently, the antenna temperature would be lower for the broader antenna. In equation form

$$T_a = T_s \frac{\Omega_s}{\Omega_a}$$

where T_a is the antenna temperature, T_s the apparent temperature of the sun, Ω_s is the solid angle subtended by the visual sun (7×10^{-5} steradians), and Ω_a is the solid angle corresponding to the half-power beamwidths of the antenna[1]. If θ_H and θ_V are the half-angles to half-power beamwidths in the horizontal and vertical planes in degrees, then

$$\Omega_a = \frac{\pi}{4} \frac{\theta_H}{57.3} \frac{\theta_V}{57.3}, \text{ approximately.}$$

For illustration, an antenna which was 15° to the −3 db. points in the horizontal plane and 10° in the vertical plane would "see" a solid angle

$$\Omega_a = \frac{\pi}{4}\left(\frac{7.5}{57.3}\right)\left(\frac{5.0}{57.3}\right) = 8.99 \times 10^{-3} \text{ steradians}$$

The antenna is connected to a feed line which has some loss. If we call the feed line loss, when expressed as a ratio, A, we have

$$T_b = A\, T_s \frac{\Omega_s}{\Omega_a} + (1 - A)\, T_o$$

for T_b, the temperature at the receiver terminals due to the power received from the sun. T_o is the earth's temperature, usually taken as 290°K.

With no signal input, the receiver temperature is

$$T_R = (N - 1)\, T_o$$

where N is the noise factor of the receiver (noise factor is related to noise figure in the following way: if we express noise figure as a ratio, and add 1, we have the noise factor. A 6-db. noise figure corresponds to a noise factor of 5.).

If the sun noise at the output of the receiver is d decibels above the receiver noise, and if we converted to a ratio, call it D, then $d = 10 \log_{10} D$, and combining all the above, we have:

$$A\, T_s \frac{\Omega_s}{\Omega_a} + (1 - A)\, T_o = D\,(N - 1)\, T_o$$

The answer to "how many db.'s of sun noise" then is

$$D = \frac{1}{N-1}\left[A \frac{T_s\, \Omega_s}{T_o\, \Omega_a} + (1 - A)\right]$$

An equation much like this has appeared here before[2]; perhaps this presentation, which shows where such an equation comes from, will help in understanding what will be said later.

Let's work an example, showing how practical results may be predicted. If the receiver has a noise figure of 5 db., then $N = 4.16$. If the feed-line loss is 2 db., then $A = 0.631$, and if we are interested in 432 Mc. the apparent solar temperature is about 500,000°K for a condition when the sunspot number is 50 (see below). T_o is 290°K

[1] For further discussion, see Pawsey and Bracewell, *Radio Astronomy*, Oxford University Press, Oxford, England, 1955, p. 21. Steradian: The solid angle subtended at the center of a sphere by a portion of the surface whose area is equal to the square of the radius of the sphere.

[2] See Bray and Kirchner, "Antenna Patterns from the Sun," *QST*, July 1960.

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(about room temperature) and $\Omega_s = 7 \times 10^{-5}$. If the antenna beamwidth is 10° by 10° to the half-power points, its half-beamwidth to half-power points is 5° by 5°, and $\Omega_a = 6.0 \times 10^{-3}$. Then

$$D = \frac{1}{4.16 - 1}\left[0.631 \frac{5 \times 10^5}{2.9 \times 10^2}\frac{7 \times 10^{-5}}{6.0 \times 10^{-3}} + (1 - 0.631)\right] = 4.14$$

Converting this back to decibels, the sun noise should be almost 6.2 db. above the receiver noise for this system. There is one problem with this calculation: The sun radiates noise of both vertical and horizontal polarization (usually equal amounts) while most antennas accept only one polarization. If this is the case, the antenna only accepts half the incident radiation, and we must subtract 3 db. for polarization loss. In such a case, the sun noise would be 3.2 db. above receiver noise.

Making Measurements

The radio astronomer would measure N, A, and the antenna parameters, and then knowing these would measure T_s daily by measuring daily values of D. As hams, we are probably more interested in measuring the antenna parameters, or in monitoring our receiving system to make sure everything is working the way it should. This way leads to some trouble, simply because we don't know enough about T_s. At frequencies below about 1000 Mc., the apparent solar temperature isn't very well known for several reasons. The first is that not too many solar observatories have measured solar temperatures daily over a long period of time in this frequency range. While Potsdam, Ottawa, and Toyokawa, among others, measure daily solar temperatures between 1,000 and 10,000 Mc., and have over most of a sunspot cycle by now, not very many protracted measurements are available for the frequencies we are talking about. The second reason is that the solar temperature varies from day to day. Radiation at these frequencies comes from high in the solar atmosphere, and there is still much to be learned about this region of the sun. Therefore, solar temperatures often show little correlation with sunspot number, which is really a measure of activity in the lower part of the sun's atmosphere. The best guess that can be made as to solar temperature as a function of frequency, and the amount it increases for a Wolf Sunspot Number of 100, is shown in Table 1.

TABLE 1

Frequency	*Temperature (Sunspot No. = 0)*	*Percentage Increase (Sunspot No. = 100)*
144 Mc.	1,100,000°K	10%
220	1,100,000	12%
432	400,000	50
1296	150,000	100

These values have been obtained by comparing the reported results of Allen[3] with the daily values reported by the Toyokawa Observatory of the Research Institute for Atmospherics of Nagoya University. The accuracy of these values is not very good.

With this caution in mind, some good information can be obtained from monitoring solar temperature. One thing that can be done is to find, experimentally, what the beamwidth of an antenna is. If D turns out to be more than 2 (that is, 3 db. above receiver noise), we can find the half-power beamwidth ($2\theta_H$ and $2\theta_V$) by pointing the antenna at a point in the sky that the sun will cross, and letting the sun slowly drift through the antenna pattern. When the sun is in the center of the antenna pattern, put a 3-db. attenuator *between the antenna and transmission line* (not between the converter and i.f. strip). Such an attenuator is easily made from coaxial cable (about 29 feet of RG-58/U for 432 Mc.). Clock the times at which the receiver output from sun alone is the same as with the sun at the center of beam and the 3-db. attenuator in line. Since the sun drifts one degree every four minutes, dividing the minutes (between calibrated −3 db. points) by 4 gives the half-power beamwidth in degrees (2θ). Turning the antenna on its side and repeating will measure the beamwidth in the other plane. If the antenna does not give more than 3 db. of sun noise, you will have to use a signal generator, and rotate the antenna to measure these beamwidths.

Knowing the beamwidth and the feed-line loss, one can measure the receiver noise figure (assuming a value for T_s). This can be compared with the noise figure measured by using a noise generator. If by measuring solar temperature, using the values you think are correct for your system, you come close to the values shown in Table 1, then you can be sure that your system is performing properly. By measuring these things daily, you can check the performance of your total system.

Summary

In the preceding sections, we have discussed how to measure receiving system parameters, and how to monitor system performance to guard against deterioration. Should you suddenly get 1 db. of solar noise, when you have been getting 3 db., you know that your system needs some checking. Finally, we discussed some of the reasons why this is not an exact measurement, but rather should be taken as an indicator of system performance. QST

[3] Allen, "The Variation of Decimetre-Wave Radiation with Solar Activity," *Monthly Notices of the Royal Astronomical Society*, p. 174 (1957).

when D is negative, sin D is also negative.

For a time M minutes before sunset (or after sunrise) find an angle X degrees by dividing M by 4. That is,

$$X^\circ = \frac{M}{4}$$

Now find the elevation angle ϕ from the equation

$$\sin \phi = A \sin X + B\,(1 - \cos X)$$

The procedure described above gives the elevation angle of the sun at any time it is above your horizon, to about 1 degree accuracy. To get a better picture of the fine structure of your antenna pattern, especially at the low angles which are most important, better accuracy is needed. About 0.2 degree can be achieved by careful calculation and by applying certain corrections.

Read your latitude and longitude to 0.1 degree or better, and find the sunset (or sunrise) time to the nearest minute. Now adjust this time slightly by $\frac{3.3}{A}$ minutes. Add this to the sunrise time, or subtract it from the sunset time. Use this adjusted time to calculate the elevation angles as described above, and then apply a final correction by adding the amount shown in Fig. 3 to the calculated values. These corrections take

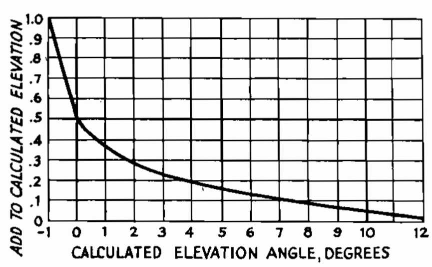

Fig. 3—Chart showing elevation angle corrections to be applied for results of high accuracy.

into account the refraction of the signal (and the light) by the atmosphere, and the difference in size between the radio sun and the visible sun.

Finding the Sun's Azimuth

If you do the job on a sunny day, the simplest way is to have a friend keep the beam pointed toward the sun in azimuth, lining it up by eye. Alternately, calculate the sun's azimuth in advance and rotate the antenna from time to time as required. When readings are taken, the beam should be within about one-fifth of a beamwidth of the sun's azimuth.

Azimuth is found from the formula

$$\cos \Theta = \frac{\sin D - \sin L \sin \phi}{\cos L \cos \phi}$$

ϕ is the elevation angle already calculated. Again, remember that when D is negative, sin D is also negative. The azimuth, Θ, is measured eastward from north in the morning, and westward from north in the afternoon. When cos Θ comes out negative, Θ is larger than 90 degrees and the sun is more south than north.

Plotting the Results

Now that the angle of the sun and the signal-strength readings have been obtained, the antenna pattern can be plotted. Fig. 4 is a curved-earth grid with elevation angles plotted on it. Taking the readings that were made as the sun ran its course, divide the signal voltage from the sun (E_S) by the signal voltage from the resistor (E_R). Do this for each reading taken. Now square each of these values of (E_S/E_R) to obtain the value of $(E_S/E_R)^2$. The next step is to compute the value Y, using the equation

$$Y = \sqrt{\frac{\left(\frac{E_S}{E_R}\right)^2}{\left(\frac{E_S}{E_R}\right)^2_{\min}} - 1}$$

where $(E_S/E_R)^2$ is each of the readings that were taken as the sun crossed your antenna and $(E_S/E_R)^2_{\min}$ is the value of the reading after the sun is below the zero-degree elevation angle by 5 minutes or more. Then find the greatest value of Y. At this reading, calling it $Y_{\max}$, assign an arbitrary value of slant range — 500 miles. This is then one point on the plot: 500 miles and the angle to the sun at that time. Now take 500 miles and divide it by $Y_{\max}$ and multiply all of the other Y values by this amount. Plot on Fig. 4 the angle for each signal-strength reading and distance just found. Drawing the curve, you now have your antenna pattern in the vertical plane.

There is one caution. The sun is not really a point source of radio waves. It can be represented as a ring of about 1-degree angular diameter on the outside and about one-half degree on the inside. Because of this, the nulls in the antenna pattern will not appear to be sharp. For this reason, a sample antenna pattern is shown in order to guide you in your plot. When the curve shows a dip, it probably is a very deep null as indicated by the dotted lines on the same curve, Fig. 5. Because the depth of the nulls cannot be determined, the antenna pattern taken by this method would probably not satisfy an exacting scientist, but in practice the signals that are received on such an antenna, amateur or otherwise, are not from point sources either. Thus the antenna pattern taken by this method is truly an operational pattern.

For those interested in meteor scatter an estimate of optimum range can be made. The memeteor trails will be most prevalent at a height of 50 miles. From your antenna pattern note the range at which the elevation angle line through the peak of your lowest lobe intercepts the 50-mile height. Multiplying this number by 2 will yield your approximate optimum meteor scatter range. In the example shown in Fig. 5, this would be about 1000 miles.

Noise Figure and Antenna Gain Check

There is another interesting sidelight to this

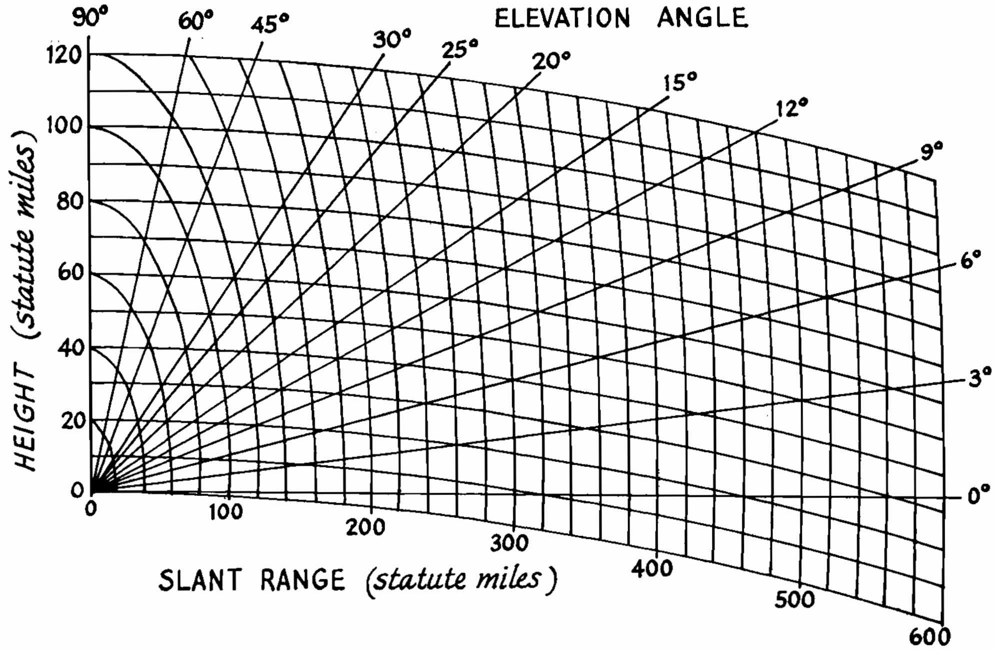

Fig. 4—Curved-earth grid for plotting results obtained from solar noise readings.

subject. The "quiet" sun is a more or less calibrated source of radio energy. Thus by a few simple calculations you can get an idea of your antenna gain or noise figure for actual received signals. Because the amount of energy that is received is a function of both the noise figure and antenna gain, you can start with one of the known values and find the other. The equation which applies is

$$\frac{G_P}{F_P} = 290\,\frac{L_P}{K}\left[\left(\frac{E_S}{E_R}\right)^2_{\max} - \left(\frac{E_S}{E_R}\right)^2_{\min}\right]$$

where G_P = the power gain on your antenna
F_P = the noise figure expressed as a power ratio
L_P = the transmission line loss for your cable and your length
K = a constant dependent upon the frequency band

and

$$\left(\frac{E_S}{E_R}\right)^2_{\max}$$

is the maximum signal ratio from the antenna pattern data taken above. This value will occur at the peak of the first vertical lobe. $(E_S/E_R)^2$ min is the signal ratio at the time the sun was a few minutes below the horizon.

This formula will only apply when the sun is quiet. If the answers are out of line the test should be repeated until a quiet day is found. A quiet sun radiates the lowest amount of energy; all other conditions produce greater received power.

Your antenna gain is probably the least well-known number of your radio system.

TABLE II

Frequency Band	*Value of K*
144 Mc.	2.9
220 Mc.	2.8
432 Mc	2.7
1296 Mc	0.65

To calculate the antenna gain:

1) Estimate the noise figure of your converter by taking the manufacturer's noise figure, or from tube data if it is a home-brew model. This value will probably be expressed in db. Convert the db. noise figure to a power ratio by the common db. formula, F_P = antilog $F/10$ where F is the noise figure in db. from above. This conversion can also be made using the decibel chart in the *ARRL Handbook*.

2) The factor K is listed in Table II for the various amateur bands above 50 Mc. The 6-meter band has been omitted because of the strong background of radio energy in this frequency range in large areas of the sky, which could adversely influence the results. For the higher bands the background radiation is much less. It is possible that one of the bright radio stars could be near the sun when the measurement is being taken, and would therefore influence the readings on the higher frequencies, also, but the chances of this are remote.

3) L is the line loss. This figure is easily estimated by looking up the transmission-line manufacturer's data for your frequency. It is usually expressed in db. loss per hundred feet. Thus, calculate the db. value for your length and convert the db. loss to a power ratio as you did above for the noise figure.

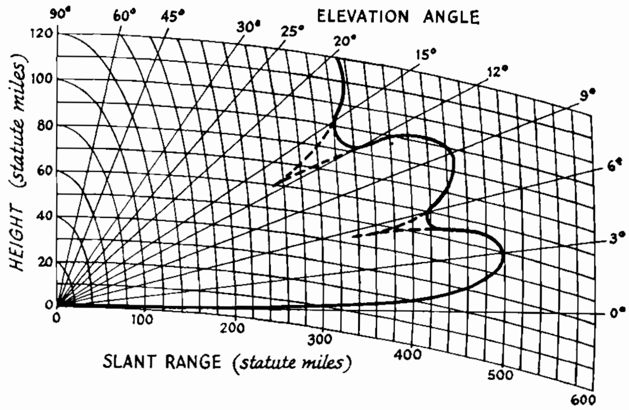

Fig. 5—Representative vertical antenna pattern. Dips in the heavy line represent nulls in the pattern which are actually much deeper than data will indicate. This is due in part to the fact that the sun is not a true point source for radio noise. Antenna pattern may be more like that shown in dotted lines.

Substitution of the values in the formula will yield the power gain G_P of the antenna. This can be converted to db. gain by the common formula

$$G = 10 \log G_P,$$

or by using the *Handbook* table.

What we have really been talking about here is a practical use of radio astronomy. The methods used here also apply to the detection of radio stars. Many interesting experiments can be performed. For those who are interested, take a radio look at Cygnus A or the center of our galaxy in Sagittarius, when they are rising or setting. Both are good strong noise sources, and real DX!

References

American Ephemeris and Nautical Almanac, issued yearly by U. S. Naval Observatory. Consult it at library.

Astronomical Phenomena for the Year, published annually. A reprint of selected pages from the above, 25 cents from Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C.

The Telephone Almanac, issued annually. Free from Bell System Telephone Company business offices.

Information Please Almanac, published by Macmillan Company, New York City; sold at newsstands and bookstores

MORE 50-MC. MOONBOUNCE EXPERIMENTS

Aylmer
Quebec
Canada

Technical Editor, *QST:*

The purpose of this letter is to give corrections and additions to a previous one,[1] and to describe further 50-Mc. moonbounce experiments at VE3BZS/2.

In the formula for the Doppler shift, the transmitter, frequency, *f*, should have been expressed in cycles, not megacycles. Also, it was mentioned that the antenna was usually aimed optically. However, the following formulae were referred to when the moon was obscured by clouds:

$$\sin E_T = \sin L_T \sin D + \cos L_T \cos D \sin H_T$$

$$\sin A_T = \frac{\cos D \cos H_T}{\cos E_T}$$

where E_T is the elevation of the transmitting antenna
A_T is the azimuth from true north
L_T is the latitude of the transmitter
H_T is the hour angle of the moon, and is approximately $\frac{360}{(transit)} \times t$, where t is the time in hours after local mean time of moonrise at the equator, and (*transit*) is the time in hours between ephemeris transits of the moon (approx. 25 hours).

Similarly for the receiver.

The formulae are approximations, since E_T is the elevation of the moon at the earth's center, not at the station. However, the difference is less than one degree, at most. Also, the local hour angle definition may not be standard.

The moonbounce experiments were continued with different antenna polarizations to see if improvement could be obtained. It was pointed out by Soifer[2] that crossed Yagi antennas transmit and receive the same sense of elliptically-polarized radiation. Thus, theoretically, assuming specular reflection of the radio wave at the moon's surface and therefore reversal of the sense of the polarization, the antenna used in the experiments previously described should not have received the transmitted echoes. Neglecting effects of the ionosphere and lunar surface, the fact that some echoes were received is probably attributable to ground reflection effects and/or transmission of elliptically-polarized waves due to mismatch in the antenna system.

[1] "50-Mc. Moonbounce Experiments," Technical Correspondence, *QST*, May, 1962, p. 49.

[2] Soifer, "Research, Tracking and Reporting, Project Echo A-12," *QST*, June, 1962.

A trial was made recently using two of the crossed Yagis transmitting radiation polarized in one sense and the other two receiving in the opposite sense. Stronger and more frequent echoes were recorded, even though the system gain (neglecting the problem of reverse polarization) was only one-fourth that previously used.

Another trial was then made with the four antenna units transmitting vertically. This was chosen over horizontal to have the major lobes of the upper and lower bays as nearly coincident as possible, in the event of ground reflection. During this trial the moon was obscured, and some power-line interference was present, but results were the best so far.

The 50-Mc. trials were brief and incomplete, but results seem to indicate that the echo amplitude varies widely; that the average signal-to-noise ratio of the echoes is less than that given by the formula in the previous letter; that, when using circular polarization, improved performance results if the transmitting and receiving antennas have opposite polarizations; and that, with the system parameters used, no distinct advantage between circular and linear polarization was noticed.

— *Alan Goodacre, VE3BZS/2*

SUN NOISE

Technical Editor, *QST:*

Some questions have arisen over the definitions used in the article on sun noise (April 1968 *QST*, page 42), and perhaps a note clarifying these is in order. To conform to common usage, as stated in the *IEEE Standards*, the relation between noise figure and noise factor should be:

Noise Figure = 10 log (Noise Factor).

Thus, ". . . add 1 . . ." is incorrect, and a noise figure of 6 db. corresponds to a noise factor of 4. Also, as used in the equations, A is the fractional transmission of the feed line, rather than its fractional loss. Fractional transmission equals one minus fractional loss, resulting in A being correctly given in the example. However, the redefinition of N, as above, changes the results of the example so that the sun noise, taking one linearly-polarized component, should be 4.8 db. above the receiver noise, for the other constants assumed in the example.

One last word of caution: the whole presentation was based on the ratio signal/noise rather than signal + noise/noise. For systems where the sun is only slightly above the noise, reduction to signal/noise ratio, as described in January 1968 *QST*, page 34, may be required. — *Don Lund, WAØIQN, P.O. Box 1664, Boulder, Colorado 80301.*

50-MC. MOONBOUNCE EXPERIMENTS

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Technical Editor, *QST:*

Communication by moon-reflected radio waves offers amateurs the opportunity for making v.h.f. contacts anywhere in the world, for periods of from a few minutes to several hours each day, if the considerable technical problems can be solved. The following describes some attempts to obtain moon echoes on 50 Mc.

Previous work with the reception of 50-Mc. transmissions by W7RDY at VE7AIZ gave evidence of perhaps two or three consecutive weak echoes during each of half a dozen trials.[1] Lack of more consistent results was assumed to have been due to Faraday rotation of the plane of polarization in the ionosphere, causing loss of signal. Antennas at both ends were horizontally-polarized Yagis. It seemed worth while to make another attempt at VE3BZS, using circular polarization, in a manner similar to that used by K1HMU [2] to overcome the Faraday-effect problem.

Four Yagis, each with 5 elements in a horizontal plane and 5 in a vertical plane, were arranged in box configuration approximately 20 feet square. The vertical driven elements were fed 90 degrees out of phase with the horizontal ones. The antenna could be rotated only in azimuth, and was usually optically aimed. The transmitter used a heterodyne exciter for maximum stability. An external tunable oscillator at 1 Mc., with 50 times frequency multiplication, gave receiver injection at 50 Mc., plus the Doppler shift, plus or minus the audio filter frequency of 940 cycles. This beating signal and the returned signal, if any, are fed into the regular 50-Mc. converter, and then into the station receiver, set for 600-cycle bandwidth. Then follows the 940-cycle audio filter, with a bandwidth of 20 cycles, and a tape recorder.

An important receiver point is that the gain of the receiver should be set so that the noise at the output of the audio filter disappears when the external injection is turned off. Under this condition the effective predetection (i.f.) bandwidth of the receiving system is determined by the audio filter. The heterodyne system for the transmitter allows the oscillators to run continuously, permitting better frequency stability than when turning the oscillator on and off. Heterodyning also reduces the drift at the signal frequency, for a given amount of oscillator drift, compared to a conventional oscillator-multiplier system. Absolute frequency stability was not extremely good, due to lack of temperature control of crystals and transistors, but relative drift to the rece ver was from one to two cycles per minute. This is good enough to permit audio filter selectivity of 10 to 20 cycles to be used.

Because of this narrow bandwidth the Doppler shift had to be calculated. The approximate formula used was:

$$\Delta f = f\left[\frac{37.04}{(\text{transit})}(\cos L_T \cos H_T + \cos L_R \cos H_R)\cos D \times 10^{-6} + 5.54\left(\frac{1}{(\text{semi})_1} - \frac{1}{(\text{semi})_2}\right) \times 10^{-2}\right]$$

where Δf = Doppler shift in cycles ;
- f is = transmitter frequency in megacycles
- $(\text{semi})_1$ = semidiameter of the moon expressed in seconds of arc
- $(\text{semi})_2$ = semidiameter of the moon expressed in seconds of arc 12 hours later for the day concerned
- (transit) = the time in hours between ephemeris transits of the moon (approx. 25 hours)
- L_T = latitude of the transmitter
- H_T = hour angle of the moon and is approximately $\frac{360}{(\text{transit})} \times t$ where t = time in hours after local mean time of moonrise at the equator
- L_R, H_R similarly for the receiver
- D = apparent declination of the moon

[1] *The VHF Amateur*, February, 1961, pp. 13–16.
[2] See photos in *QST*, November, 1961, p. 89.

The necessary information for the calculation may be obtained from a current *American Ephemeris and Nautical Almanac*. The first term in the square brackets is usually dominant and at moonset at 45 degrees latitude amounts to about − 110 cycles at 50 Mc.

Three trials were made and only one or two weak but identifiable echoes were received. Signal-to-noise power ratios were of the order of 1:1, or less. This means that little or nothing can be heard of the return signal by ear, but a visual presentation shows evidence of a return. The advantage of visual methods in detection of very weak signals increases with very narrow receiver bandwidth, since signal and noise tend to sound the same under these conditions.

The average signal-to-noise ratio at the output of the audio filter was calculated using the following formula, which neglects fading effects produced by the motion of the moon's surface, Faraday rotation and ground reflection:

$$\left(\frac{S}{N}\right)_{\text{POWER}} = \frac{1.6 \times 10^{-26}\, G_R \lambda^2\, G_T P_T\, 10^{\frac{-2KL}{10}}}{4.1 \times 10^{-21}\left(.22\lambda^{2.4}\;\; 10^{\frac{-KL}{10}} + F - 1\right) B}$$

where P_T = transmitter power output in watts
- K = attenuation of transmission line in db./100 ft.
- L = transmission line length in units of 100 ft.
- λ = wavelength in meters
- G_R = gain of receiver antenna over isotropic radiator
- G_T = similarly for transmitter
- F = noise figure of receiver at wavelength λ
- B = effective noise bandwidth of receiver in c.p.s.

It should be noted that P_T, K, and F vary with λ for given components. Frequency stability problems make minimum B vary with λ also. For given conditions there is an optimum λ to produce maximum average signal-to-noise ratio. The words "average signal-to-noise ratio" are used, since the instantaneous noise power may deviate from the average value, but the actual signal-to-noise ratio should be within a factor of two of the average about 50 per cent of the time.

The calculated signal-to-noise power ratio using:
- P_T — 400 watts
- K = 3 db./100 ft.
- l = 100 ft.
- λ = 6 meters
- G_R = 64
- G_T = 64
- F = 2
- B = 20

gives $\left(\frac{S}{N}\right)$ = 1:1.

A possible explanation for lack of consistent (although weak) echoes may be that the image antenna produced by ground reflection (assumed perfect for sake of argument) is causing cancellation and reinforcement [3] of the circularly polarized radiation in such a manner that alternate zones of radiation are produced where the polarization changes from being completely vertical to being completely horizontal. When the moon is in a zone where the radiation is predominantly linearly polarized, Faraday rotation may cause loss of signal, when transmitting with circular polarization.

A comparative test between VE3BZS/2 and another local station, with distant stations using horizontally polarized antennas, gave signals several db. lower than expected. Probably this was due, in part at least, to ground reflection producing predominantly vertically polarized radiation at low angles, when using circular polarization at VE3BZS/2.

The results of these 50-Mc. tests seem to show that 50 Mc. is not too practical under present conditions for amateur moonbounce work. Also circularly polarized antennas may suffer a loss in efficiency under conditions of good ground reflection in combating Faraday-rotation effects.

The narrow-band methods used in the receiver and transmitter should be adaptable for use on higher frequency amateur bands to allow use of existing equipment with little modification and cost. — *Alan Goodacre, VE3BZS*

[3] *The A.R.R.L. Antenna Book*, pp. 46–48.

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Moonbounce Notes

During the last year there has been an upsurge of interest in amateur communication via reflection from the moon. All the bands from 50 MHz through 2400 MHz have been involved.

This activity has created more interest in moonbounce and each neophyte "moonbouncer" has had many questions concerning just how to get started:-

-Which band should be used?

-How much power is needed?

-How good should the receiver be?

- What kind of antenna should be used, and how big should it be?

In the process of determining antenna parameters, it is necessary to know how to find and track the moon. In addition, the type of antenna mount, the aiming system and the physical location of the antenna on the available plot of land are a function of the path of the moon.

Many articles have been published which will answer these questions and help the beginner and old time "moonbouncer" alike. Much other data are available which have not been published. It is the intent of this compilation of moonbounce notes and articles to reproduce in one place the literature necessary to allow the potential "moonbouncer" to make the basic decisions necessary to start his project.

As time goes on, additional notes will be added.

Contributions from those interested in EME (earth-moon-earth) communication will be gratefully received.

Thanks to the American Radio Relay League for permission to reprint certain articles from QST magazine.

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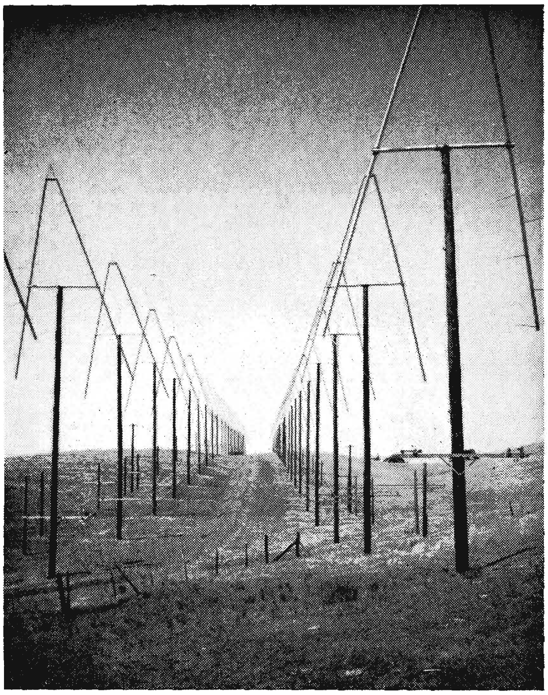

This field full of phased log-periodic antennas was used by the author to obtain reflections from the moon in the 10-meter band. In case you'd like to duplicate the feat, the array is 1200 feet long and 75 feet wide. It has a gain of 27 db.!

The Moonbounce Problem, 28 Mc. and Up

Basic Facts for Determining Equipment and Antennas Needed for Lunar Communication

BY H. T. HOWARD,* W6UGL

THE purpose of this article is to stimulate amateur interest in moonbounce communication, by presenting the basic parts of the problem, such as noise figure, path loss, and antenna gain, in familiar terms. Once these basic factors are understood, they can then be applied to equipment and antenna design for communication via the moon or man-made satellites.

Moonbounce was accomplished on ten meters several months ago at this station with about 1 kw. p.e.p. single sideband, using the array of 48 log periodics shown in the first photograph. The array is 1200 feet long by 75 feet wide, and it has a gain of 27 db., over the range of 20 to 65 Mc.! The beam produced is approximately 1½ degree thick by 30 degrees in azimuth and can be placed to intercept the moon or sun track for about two hours each day. Power is distributed in the array with open-wire line, and tapered sections to maintain the wide bandwidth, and in the usual operation *each antenna* handles from 5 to 10 kilowatts. A circuit diagram of the array would look like a corporation organization chart; that is, it starts with one feed line and progressively branches down to the individual antennas which are specially designed and built log periodics, each having a pair of 40-foot booms and a total of 48 elements.

At each power division point there is a movable tap arranged so that the relative phase between antennas is completely adjustable. In practice, the phasing is changed each day to follow the moon's elevation. It takes two men with wrenches and a jeep about two hours to move all of the taps. The array is normally used with a 300-kw. (600-kw. p.e.p.) c.w. transmitter for radar studies of the solar corona and the ionized regions between the earth and the moon.[1]

The selection of ten meters for the moonbounce experiment mentioned above avoids controversy

* Radioscience Laboratory, Stanford University, Stanford, Calif.

[1] Research supported by the Electronics Research Directorate of the Air Force Cambridge Research Laboratories, Bedford, Mass., under contracts with Stanford University, Stanford, Calif.

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over the use of large nonprivate antennas for v.h.f. records. Six or possibly fifteen meters might yield similar results if tried. The whole idea, though, is to demonstrate that the absolute minimum antennas for h.f. and lower v.h.f. moonbounce are ridiculously large for individual construction.

Since the array is linearly polarized, Faraday fading is a very important consideration,[2] and the unrecommended expedient of whistling into the microphone was used, until the signal faded up to a usable strength. Then the call was signed in voice and, with the help of some imagination, was received 2½ seconds later. The use of circular polarization will reduce fading, and is certainly required for any serious v.h.f. lunar-communication attempt.

The trick of ten-meter moonbounce points out several facts that will become obvious as you read further. First, station equipment needed for moonbounce on our lower bands is a minimum, and commercially available, but the antenna required is gigantic. Second, cosmic noise and ionospheric effects play a large role below about 100 Mc. With increasing frequency, the antenna becomes physically smaller, but the receiver and transmitter must be the best that amateur ingenuity can produce.

The average loss in decibels for the earth moon–earth path, assuming 500 watts of r.f. power at the antenna terminals and a moon reflectivity of about 7 per cent, is given in Fig. 1. Path loss will vary approximately ± 1 db. during each month as range to the moon changes.[3] Moon reflectivity is currently the subject of several scientific investigations, and while reflectivity appears to be higher at frequencies below 450 Mc., and is perhaps lower above that frequency, the figure given should be accurate enough for a first approach to the problem. If the transmitter power at the antenna terminals is less than 500 watts due to feed-line losses or other practical considerations (such as money) this path-loss number should be increased by the number of db. difference.

The next problem is that of receiver noise figure and sensitivity. Fig. 2 is a plot of cosmic noise *vs.* frequency, presented to give the equipment designer an idea of what is needed for a front end. The *min* and *max* lines show the sky temperatures and minimum usable noise figure that can be expected when the antenna is directed toward the coldest and hottest portions of the sky, respectively. This variation is easily observable even with simple equipment and is a good method of checking antenna and system performance.[4] Fortunately for the communications problem, larger areas of the sky are cold than are hot.

Below about 1000 Mc., cosmic noise is the dominant factor and varies with the portion of the galaxy observed. It can be seen that being cosmic-noise-limited, that is, having the feed-line loss and receiver-noise contribution less than the cosmic noise, at all times, is an engineering feat nearly impossible at 220 Mc. and higher, with the present state of the art.

Before going further, it is necessary to clear up some confusion concerning receiver sensitivity and noise figure that has arisen because of improper use of the relation:

$$\text{Ideal receiver sensitivity} = kTB$$

where

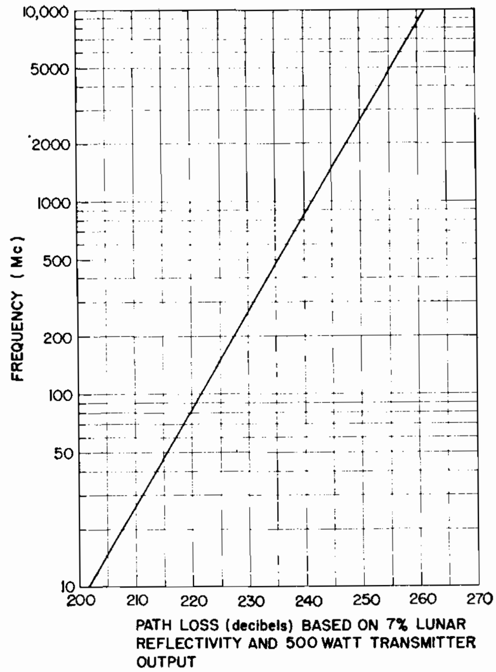


Fig. 1—Moonbounce-path loss vs. frequency.

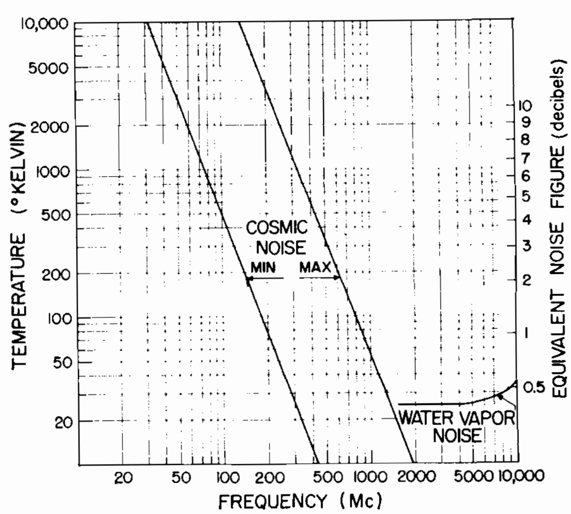


Fig. 2—Cosmic and water-vapor noise limits *vs.* frequency.

[2] Dyce, "The Appearance of the Moon at Radio Frequencies," *QST*, May, 1961.

[3] Pettengill, "Lunar Studies." Lecture notes presented at course on Radar Astronomy, summer session 1961, Massachusetts Institute of Technology, Cambridge, Mass.

[4] Downes, "A Simple Radio Telescope," *Sky and Telescope*, August, 1962.

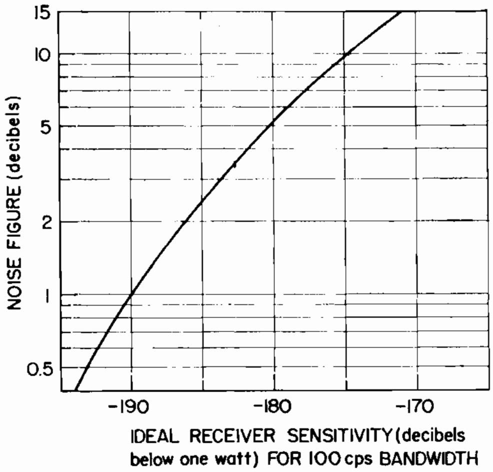


Fig. 3—Ideal receiver sensitivity *vs.* noise figure.

k is Boltzmann's constant, 1.38×10^{-23} joule/°K

T is temperature in degrees K

B is bandwidth in c.p.s.

If one uses room temperature of 290 degrees K, then it can be shown that:

Ideal receiver sensitivity (−dbw.) =

$$204 \text{ db.} - 10 \log B - \text{db. noise figure.}$$

This relation is correct for systems with noise figures greater than 3 db. (system temperature greater than 290 degrees K), but needs revision to be correct for present-day low-noise amplifiers. By using an equivalent system temperature for T instead of 290 degrees K, we can still satisfy the IRE definition for noise figure and be consistent with present practice. All of this is simply saying that it is possible for a directive antenna and receiver at u.h.f. to look at a portion of the sky that is colder than 290 degrees K.

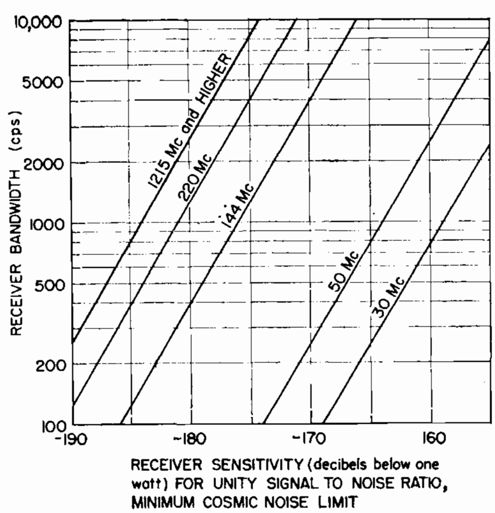


Fig. 4—Receiver sensitivity *vs.* bandwidth for the various amateur bands.

A plot of the above interpretation for a receiver with a 100-cycle bandwidth is shown in Fig. 3. There are some surprises in this graph that arise from proper use of noise figure. For instance, an improvement in receiver noise figure from 3 db. to 2 db. improves receiver sensitivity not just 1 db., but nearly 3. Going from a 10-db. crystal mixer to a 2-db. paramp yields a sensitivity improvement of 12 db. This makes it pretty obvious that the best possible noise figure and the lowest possible line losses are all important, at frequencies where system noise is greater than cosmic noise.

Fig. 4 uses the information of Fig. 2, and assumes that the system performs to the lower cosmic-noise limit. It shows what receiver sensitivity to expect in each case, for unity signal-to-noise ratio with various bandwidths. If the system is not cosmic noise-limited, the number obtained from Fig. 4 should be decreased by the number of db. difference between the ideal case of Fig. 4 and the actual system. Again, both noise figure and transmission-line loss enter here. The number from Fig. 4, as modified by reality, is the receiver sensitivity in decibels below 1 watt, and can be added algebraically to the path loss of Fig. 1 to obtain the two-way antenna gain necessary.

For example, select 1296 Mc. and assume a parametric-amplifier front end with a 2-db. noise figure [5] and 2 db. of feed-line losses. From Fig. 1 the total path loss is 244 db. and from Fig. 2 the system is definitely not cosmic noise-limited.

Example:

Fig. 1: Total path loss for 500 watts power output **244**
Feed-line loss **2**

246 db.

Fig. 4: Cosmic-noise-limited receiver sensitivity (500 c.p.s. bandwidth) **−187 dbw.**

Fig. 2: Receiver n.f. = 2 db. = 170° K
Line loss = 2 db. = 170° K

340° K = 3.4 db.

Cosmic noise = **0.5 db.**

Fig. 3: Difference between 0.5 db. cosmic noise and 3.4 db. actual receiver system **+ 10 db.**

− 177 dbw.

69 db.

$$\frac{69}{2} = 34.5 \text{ db.}$$

This is the antenna gain required at each station for unity signal-to-noise ratio in a 500-c.p.s. bandwidth, but as W1FZJ has pointed out,[6] the ear can be a narrower filter if properly trained.

[5] Troetschel and Heuer, "A Parametric Amplifier for 1296 Mc.," *QST*, January, 1961.

[6] Harris, "The World Above 50 Mc.," *QST*, June, 1961.

(Continued on page 6)

These figures show, among other things, that the initial 1296-Mc. moonbounce, with an 18-foot dish (35 db.) and 300 watts of transmitter power, was both a technical triumph and an operating feat. It is also clear that the higher frequencies are the logical choice for both commercial and amateur work of this type.

As Soifer [7] recently noted, the basic problem is to obtain adequate signal-to-noise ratio, and the graphs presented here should help the equipment- and antenna-oriented amateur get a feel for the moonbounce problem. It should be remembered, however, that marginal systems give marginal results (if any at all), and that these numbers should be used conservatively if reliable communication is the goal. QST

[7] Soifer, "Space Communication and the Amateur" *QST*, November, 1961.

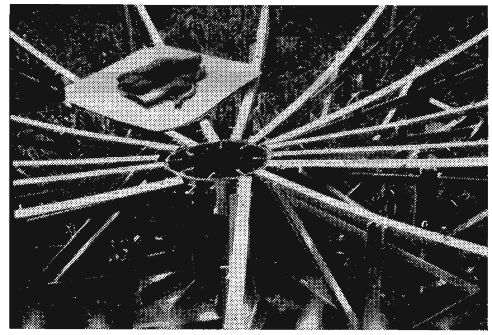
The World Above 50 Mc.

CONDUCTED BY BILL SMITH,* WB4HIP

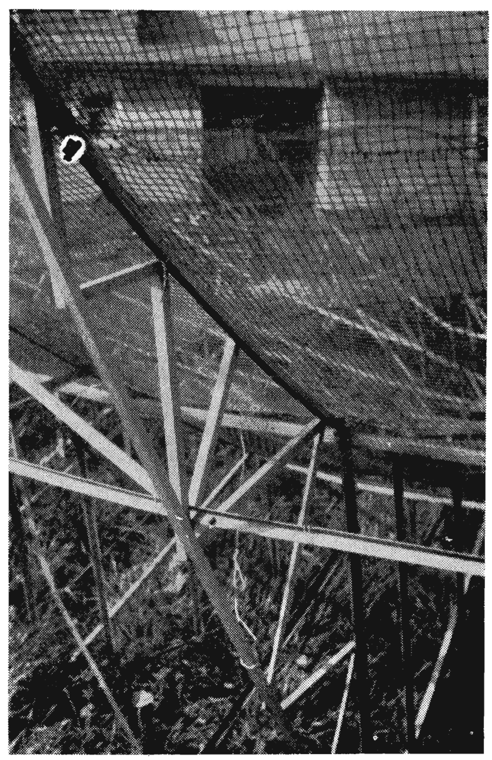
Plain Language E. M. E.

Our correspondence indicates an ever-increasing interest in amateur space communications. For those turning skyward for new adventure in v.h.f., the possibility of working e.m.e. seems indeed exotic. At my request, Mike Staal, K6MYC, has agreed to offer his guidelines for developing a successful e.m.e. system. These are the result of much work, Mike having traveled both unsuccessful and successful avenues. The discussion is not intended to illustrate a cut-and-dried system that must be used, but rather to point out what equipment is being used, and in some instances, how successfully. We hope this will stir your imagination and interest in e.m.e. communications.

The station at K6MYC is probably as basic and simple as one should consider for e.m.e. To illustrate how little is actually required, the following is *all* that is needed and used at K6MYC. An SBE-34 s.s.b. and c.w. transceiver is used with a receiving converter and transmitting mixer for 144 Mc. The only thing unusual is a common local oscillator tripled to 130 Mc. permitting transceive operation on 144. The transmitting converter is similar to that on page 159 of the ARRL *V.H.F. Handbook*. The receiving converter is an old, much-modified Ameco tube-type with a 6 db. noise figure. A 50-ohm pad is used between the SBE-34 and the transmitting converter to swamp most of the 40 watts of 14 Mc. output. The converter output is 5 watts which drives a *linear* amplifier through a relay. The 5 watts is adequate to drive a pair of 4CX250Rs in the WØMOX configuration (December, 1961, *QST*) to one-kw. input. The amplifier delivers 650 watts which is fed through thirty feet of ½-inch heliax to coaxial switches at the antenna. Two relays are used at the antenna, one for the transmitted signal and the other for double protection of the FET preamp located in the same housing. Belden 8214 carries the preamp output to the receiving converter. A 1-Mc. crystal oscillator running into a tunnel diode provides both calibration at 144.000 and a weak-signal source, which is absolutely necessary for observing receiving-system performance. A noise blanker, 60-cycle audio filter and tape recorder are occasionally used. That is it, aside from the antenna and mount. Compare your station with the aforementioned and you'll probably find



Careful examination of this photo will reveal some construction ideas for the hub of a parabolic reflector. The dish belongs to W3SDZ.



W3SDZ used hardware cloth for the covering on his 432-Mc. dish. Note the construction of the struts and supports.

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very little keeping you from beginning in e.m.e. work except the antenna.

The bare minimum gain required from the 144-Mc. antenna is 20 db. over a dipole. This does not mean that echoes are not possible with slightly less gain, but for any hope of reliability through the moon's cycle, 20 db. is the line when using "normal" receiving systems.

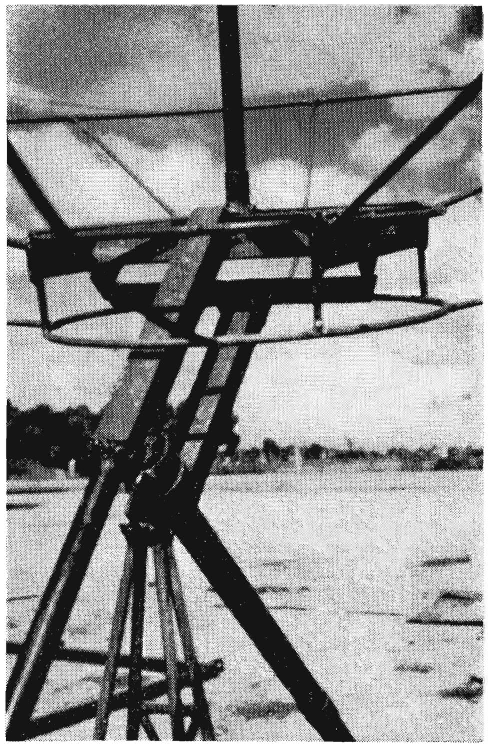
Now about the antenna. To my knowledge, no one has yet been satisfied with the performance of Yagis on an e.m.e. circuit. W6DNG used them but has since changed to an extended expanded collinear which he says is the best of more than 50 e.m.e. antennas he has tried. F8DO has a Yagi array, but doesn't feel it is performing as well as it should. However, short Yagis of 4 or 6 elements may be the answer if you must try them. VE3BZS/VE2 has an array of sixteen, 4-element Yagis and is now doubling that number. He's had some success in hearing his own echoes. K6HCP, using two 26-foot boom Yagis, ran several hours of tests over a period of days with K6MYC with completely negative results. Transmitting at K6HCP and listening at K6MYC produced nothing. The opposite was also tried without success although K6MYC could hear his own echoes.

The antenna at K6MYC is a 160-element collinear which I believe is producing close to the theoretical 24 db. gain. Echoes can be received almost anytime during the moon's 28-day cycle, assuming the Faraday rotation (polarization rotation in the ionosphere) is correct. We will discuss Faraday rotation later as well as the 28-day cycle, which is related to sky temperature (cosmic noise) at various "look" angles. F8DO, VE3BZS/VE2, ZL1TFE, ZL1AZR, WB6DEX, and of course VK3ATN have all heard K6MYC on e.m.e. WB6KAP has an antenna almost identical to K6MYC's and has had equally good results.

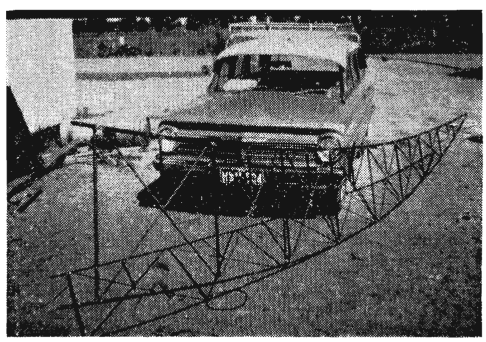
The cubical quad looks good since it fits into the low-Q class with collinear types. ZL1TFE heard signals with four 5-element quads patterned after those by W1CER and modified by W7FS. Don't rule out expanded quads as they should be quite practical. Size and weight seems to be their chief drawback.

Rhombics, the king of the h.f. antennas, seem to have a place in v.h.f. circles as well. This antenna does not allow much moon time each month, but the gains achieved can be extremely high, at much less cost than most other arrays. VK3ATN uses four rhombics stacked one above the other with six-foot spacing between for his 2-meter e.m.e. antenna. The antenna is 342 feet long per leg and has an apex angle of about 10 degrees. The gain is calculated at 33 db., over perfect ground, but actual gain is probably closer to 27 to 30 db. VK3ATN has been very successful using this antenna and 150 watts input. The LaPort rhombic is being tried and seems to have possibilities. ZL1AZR has a single-layer one and has copied K6MYC and possibly VK3ATN. More layers or a side-by-side configuration may be in order. The antenna is only 70 feet long. The disadvantages of rhombics are immobility and low elevation angles.

All of the antennas thus far discussed have been linearly polarized. Now let's consider some sort of circularly-polarized antenna. First a definition of circular polarization is in order; let us use the helix to simplify the explanation. Since a helix has no linear element, it theoretically radiates equally in all planes and the wave is launched in the direction of the spiral. Depending on whether the helix is wound clockwise or counterclockwise, the antenna would be called right- or left-hand



Shown is the hub assembly of the VK3ATN dish. The declination and hour angle drive motors are yet to be mounted as is a 16-inch diameter bearing.



This is one of the more than 20 steel tubing trusses being used in the 50-foot dish at VK3ATN. The 20-foot long trusses weigh 30 pounds each and are within ⅛-inch tolerance of a parabolic curve.

circularly-polarized respectively. For point-to point communication using helices, both antennas should be wound the same direction. When listening for one's own echoes, a right-hand signal radiated at the moon will return left-hand. This means that to hear your own echoes the direction of circularity must be switched. Circular polarization can also be achieved with properly phased crossed dipoles orthagonally mounted.

WB6DEX currently uses nine 20-element crossed Yagis, but runs only the horizontal 90 elements when testing with another station using horizontal polarization. Otherwise he would lose about 3 db. by putting half of his power into the vertical elements. However, if both stations used 180 elements circularly-polarized arrays, 3 db. would be gain on both ends — obviously very worthwhile.

Also the problem of long term fading due to Faraday rotation would be eliminated. This polarization mismatch can cost as much as 20 db. when using linear-polarized antennas. Helix antennas have not yet been successfully used for a two-way amateur e.m.e. contact as far as I know, but maybe W8JK's helix will intrigue some of you who need a new challenge. (See W1CER's article on page 20, November, 1965 *QST*.)

Certainly one antenna that should not be overlooked is the parabolic reflector, be it circular or cylindrical. However, dishes of a useful size at 144 Mc. are impractical for the average amateur, but at 432 and higher the picture brightens. (K2UYH described a homebuilt dish in the August, 1966 *CQ*.)

To summarize on antennas, my personal experience tells me circularly-polarized antennas for 432 and above, if at all possible, and below 432 shoot for maximum gain in a low-*Q*, linear polarized array.

Now let's look at a smaller but still important component in the e.m.e. station, the preamp. It may not be entirely necessary if your converter has a noise figure of 3 db. or less, but if located near the antenna the preamp can reduce feedline losses on receiving and possibly lower the system noise figure a bit. The noise figure to aim for at 2 meters is 2 db. You can try for less, but don't expect a noticeable increase in sensitivity, because the lowest sky temperature encountered at 144 Mc. is about 1.9 db. At 432 and above cosmic noise is less and very low-noise devices become more useful. It is doubtful that your system will be cosmic-noise limited. On 144 and 432 transistors appear to be the way to go, and more specifically, FETs or the steadily improving MOS dual-gate FET. Many types and brands are available for under $2. Many good preamp circuits have been published, but most lack protection for the transistor. A pair of diodes, typically 1N100s, back-to-back at the input to ground will save much grief. If you insist on using regular bipolar transistors, be sure to build a good stripline filter to help eliminate overloading of the transistor by strong local stations in the broadcast band and higher. Normally a filter is not needed ahead of a FET.

Little need be said about the balance of the converter except that crossmodulation (overload) of the mixer stage can sometimes be a problem. The use of FETs as mixers is a current solution. Recently RCA began marketing a dual-gate MOS FET pair that look ideal for converters, a 3N140 front end and a 3N141 mixer. Both are under $2 and may be the best yet for 144 and 220.

Next month we'll look at methods used during e.m.e. tests and pass along some time-saving hints. Also a thorough examination of the problems encountered is in order, as is a discussion of antenna mounts and drive mechanisms. In the meantime, you should read W6UGL's article, "The Moonbounce Problem, 28 Mc. and up," on page 20, September, 1963, *QST*.

A Layman's Look at E.m.e.—Part II

K6MYC continues his discussion this month of propagation problems effecting e.m.e. communications and what the amateur can do to alleviate some of them.

Although there are electrons everywhere in our atmosphere and beyond, those in the ionosphere have the greatest effect on v.h.f. and u.h.f. signals leaving this planet. This cloud of electrons is in a constant state of flux, their number either increasing or decreasing, or moving about to form clouds or blobs, much the same as vapor clouds. For our discussion, however, think of the ionosphere as a homogeneous layer with no irregularities. A plane-polarized 2-meter signal entering this layer is gradually rotated and may go through several rotations before passing through the ionosphere and into space. If electron content is high, as it normally would be during daylight hours, the signal may rotate many more times than it would during early morning hours. This phenomenon is known as *Faraday rotation*.

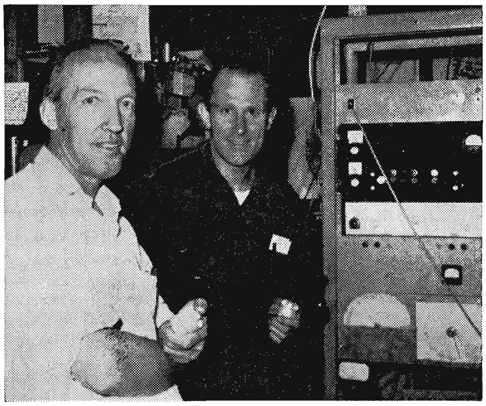
Regardless of the plane of polarization originally, the wave may come through the layer in any plane until it strikes the moon. As an example, consider the direction of rotation to be clockwise. When the signal strikes the moon and is reflected, it maintains its plane of polarization until beginning to re-enter the ionosphere, where again it begins to rotate, still in a clockwise direction, until returning to the antenna from which it was transmitted. An originally horizontal signal may have rotated six times plus 45 degrees leaving the ionosphere, and another six times plus 45 degrees upon re-entry, adding up to a net 90-degree rotation change, or vertical polarization. The signal received on a horizontal antenna may suffer a 20- to 30-db. loss from polarization shift alone.

The problem of Faraday rotation is further complicated when contact with another station is attempted. The transmitted signal must pass through two probably-different ionospheric sections before arriving at the other antenna. The polarization of the arriving signal may match the plane of one of the two antennas, but not necessarily both, or either. To put it simply, your own echoes may be coming back well, but the other station may not hear anything. But if transmissions are continued for an hour or so, chances are your own echoes will fade and the other station may start hearing you.

(A demonstration of this occurred on Dec. 20, when K6MYC and VK3ATN had another e.m.e. QSO. During the entire QSO, 1302 to 1310 GMT, neither was able to hear his own echoes. VK3ATN also heard W6YK for 8 minutes following. — EDITOR)

Another interesting fact about Faraday rotation is the relation to the hemispheres involved. A plane-polarized signal leaving the northern hemisphere twisting clockwise will return in the southern hemisphere counter-clockwise. It is possible that the effects of Faraday rotation can be nullified if the electron content of the ionosphere were the same for both paths. The shift related to hemispheres does not occur if both stations are in the same hemisphere. Schedule times should be chosen when both stations can use approximately the same antenna elevation angle (the moon the same distance above the horizon at both station) as there are usually two to three times as many electrons in the horizon path to the moon as in the path at a 45-degree elevation angle. The best time for ionospheric stability is between 2200 and 0600 local time at both stations.

Another factor entering into echo quality is *scintillation*, which cannot be corrected with circular polarization. An uneveness of electron density forms in the ionosphere and acts on a signal much like a lens on light. These "blobs" can have a focusing or defocusing affect on a signal producing unrealistically strong echoes, or no echoes at all. Scintillation, from my observations, is more apparent at frequencies below 144 Mc.

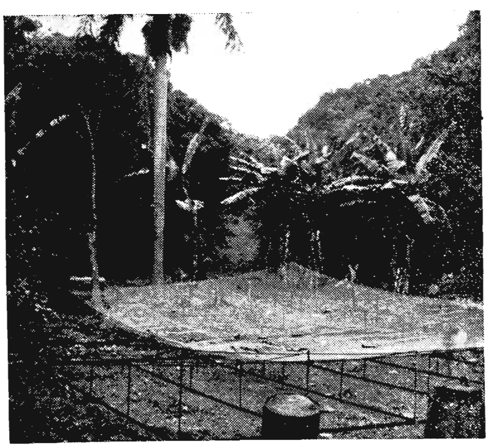


New Zealand, e.m.e. buff Ralph Carter, ZL1TFE, (left) recently visited K6MYC in Saratoga, California. Ralph is actively working towards e.m.e. contacts on 144 and 432 from his home in New Zealand.

Reprinted from February, 1968, QST

Libration fading caused by the rocking motion of the moon also effects echoes. For short periods the path loss can be reduced by as much as 6 to 10 db. The moon is a rough surface and acts like many reflectors. Sometimes they add up in phase, while on the average they give a seven per cent πr^2 reflectivity. Libration spread is more troublesome at frequencies higher than 144 Mc.

Another factor having a large bearing on whether or not contacts can be made with marginal systems is *cosmic noise.* On 432 and above this should not cause much concern, but at 144 it is a different story. The minimum cosmic noise at 144 Mc. is about 1.9 db., which is quite easily heard with modern transistors. Cosmic noise is greatest is the direction of the Milky Way, or the galactic center. From my experience cosmic noise can make a 2-db. receiving system perform like a 6-db. system, or worse, when the moon is near the galactic center. There is usually a period of five to seven days each month when the moon is at its lowest declination angles. These days should be avoided if success depends upon



W1FZJ/KP4 is building this 50 foot square "dish" for 432-Mc. e.m.e. tests. The dish will later be expanded to 150 feet for use on 144. A movable feed will be mounted atop a 60-foot tower in the center of the dish

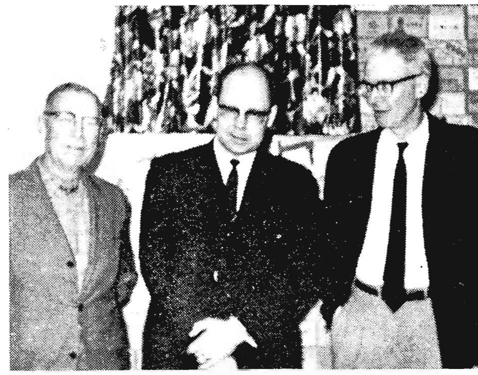
optimum receiving capabilities. Even the period as the moon is increasing its declination to its peak of approximately 27 degrees is not especially quiet. In my opinion the ten-day period after the moon has reached its declination peak is the best.

The following suggestions are offered as possible solutions to the problems just discussed:

1) Faraday rotation can be handled with circular polarization, or in part by carefully planned schedule times.

2) Larger than minimum antennas help overcome scintilation, libration fading and cosmic-noise effects.

3) An effective method of reporting and confirming signal reports helps in completing information exchanges. Avoid using code characters requiring dots, such as the letters I, E, S and H, and the numbers 2 through 7. The following sys-



Willis Brown, W3HB, Bethesda, Maryland, recently hosted Andy Kalt, DL8PK, Wahn, West Germany (center), and Bill Smith, W3GKP, of early moonbounce fame. DL8PK is active on 2 meters in Germany. By the way, Massachusetts meteor jockey W1JSM is the son of W3HB.

tem is currently being used by those scheduling VK3ATN:

T — signals detected
M — letters or portions of calls copied
O — Both calls and report copied
MT — nearly solid copy
5 — solid copy, no need for code

By this system an O plus both calls received at both ends and confirmed with RRR establishes a contact. Had this sytem been in use for my November 22nd test with VK3ATN we probably would have made another contact. However, by the old system VK3ATN was sending 3s represented by the letter E. Es are easily lost to fading and are sometimes not discernible from noise pips ringing in narrow-bandwidth audio filters. Especially after many hours of listening for weak signals, dashes are much easier to detect.

4) Receiving system modifications such as post detection, phase lock, noise blanking and cancellation all can help find signals in the noise. F.s.k. should offer a 3 db. signal-to-noise improvement and is an area for experimentation.

5) Keep transmitting and receiving periods short. I prefer 1 to 2-minute periods, particularly in daylight hours when Faraday rotation is rapid; 90 degrees every 15 to 30 minutes. Echoes can appear, peak and fade in 5 minutes or less. Five-minute periods are used by many, since some detection schemes require 3 to 5-w.p.m. c.w. speed for proper integration time.

6) Use relatively slow-speed c.w., under 10 w.p.m. When testing with VK3ATN my transmission periods are two minutes long. During the first minute each call is sent 2 or 3 times, and the report is sent the second.

7) Be sure of your frequencies, times and calling sequences. Frequencies must be within one kilocycle.

8) Keep your antenna as close as possible on the moon. If your antenna has a 5-degree beamwidth at the 3 db. points, you probably can't afford to be 5 degrees off. It is worthless to build a good antenna system and then waste it with poor

aiming. This has been the principal cause of many e.m.e. failures.

9) Don't start listening for echoes in a narrow bandwidth (under 500 cycles) unless you are experienced or have a receiving system that requires it. I prefer an 800 to 1000-cycle bandwidth but most of my receiving is done in a 2.1-kc. bandwidth, with the ear providing the "selectivity."

10) When searching for weak echoes, continuously sweep the 500 to 1000-cycle portion of the band where the signal should be. I've found I can detect signals this way that might otherwise be lost in the noise. The ear can detect pitch changes easier than a steady note.

11) Doppler shift on two meters is not much of a problem. I've never heard an echo shift more than 500 cycles at 144 Mc. If the moon is rising, the signal will appear high in frequency. As the moon passes due south there will be little or no shift; then as the moon begins to set, the echo will appear lower in frequency. When listening for your echo from a rising moon, set the receiver so the transmitted signal produces a 200 to 300-cycle note. The echo will then produce a 500 to 700-cycle note. The opposite is true of a setting moon. Doppler on 432 and higher is of more concern and will produce a 1-kc. shift or more, except when the moon is due south of your antenna.

Next month's concluding discussion of this series will cover antenna mounts, drives and readout systems.

E.M.E. for the Layman — Conclusion

This month we conclude a three-part discussion of e.m.e. (earth-moon-earth) principles by Mike Staal, K6MYC. The final section covers antenna mounts, drive systems and readout mechanisms.

First the prospective moonbouncer must decide if he is going to use his antenna system for anything other than e.m.e. experiments. This decision governs the selection of an appropriate mount and drive system. A very simple mount can be constructed if the antenna is to be used only for e.m.e. and thus be aimed at a specific point in space. This may be a logical place to begin, but you will probably soon become frustrated at being limited to perhaps 5 or 6 hours each month when the moon passes through the antenna's pattern. I suggest at least a partially-steerable array.

If only e.m.e. is contemplated, a polar (or equatorial) mount would be a wise selection as it requires only one drive mechanism for tracking and some form of manually tilting the array slightly from day to day to set the *declination*[1]. To accomplish this, your antenna mast or tower must be mounted parallel to the axis of the earth. Thus, if your station location is at 35° north, the mast would be fixed at an angle of 35° from the earth's surface at such location, oriented in a north-south direction (see fig. 1.). The declination (manually-tilted axis) changes from day to day. Information may be found in *The American Ephemeris and Nautical Almanac*, 1968, available through the Superintendent of Documents, U.S. Government Printing Office, Washington, D.C. for a nominal price. All that is necessary now is that your drive mechanism rotate the antenna at a rate of 15° per hour to track the moon.

This is all fine and dandy for e.m.e., but if you want to use your array for satellites, meteor scatter, aurora or something similar, a polar mount is not much good. A drive system permitting the array to be fully steerable in both azimuth and elevation (az-el) is the answer.

The array at K6MYC is mounted atop a homemade 12½-foot tower. The four legs of the tower are fastened to a platform which in turn is bolted to the roof of the garage directly above the operating position. A large unmodified prop pitch motor is mounted inside the top of the tower. A husky steel plate is welded to the rotating gear and another plate is attached to the first with ordinary door hinges, see the photographs. These hinges are employed in the elevating mechanism. To this plate a 3-inch aluminum channel is attached and the main boom of the array is clamped in this channel. A jack screw with right-hand left-hand square threads starting from the center out raises and lowers the array. At the lower end of the jack screw is a 20-to-1 gear reduction box giving a zero to 90° elevation time of three minutes. With the plates together the array is pointing straight up. The entire elevation drive rotates with the array.

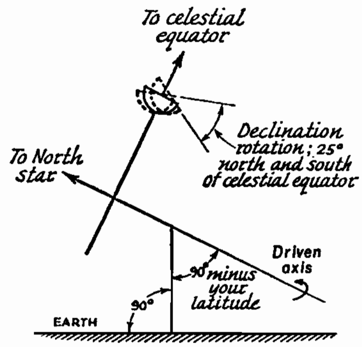
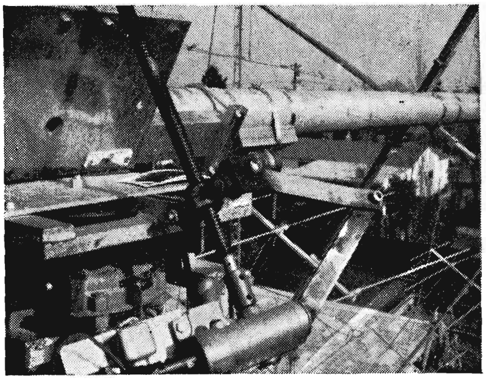
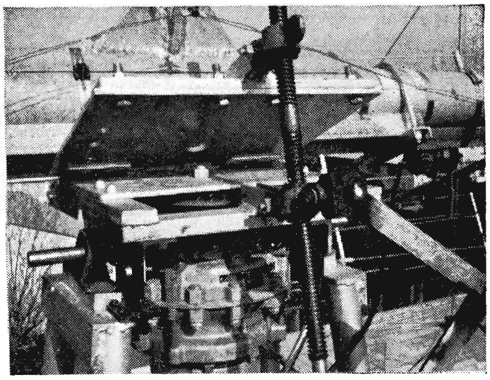


Fig. 1.

Selsyn hookups are used for direction readout and may be varied to suit the particular builder. I'll let you work out your own azimuth system, but my elevation selsyn mount is quite simple. The selsyn is attached to the main array boom and aligned with it. A weight was tightly affixed to the selsyn shaft and, of course, the weight always hangs straight down regardless of the position of the array. The mates to both selsyns are mounted on a panel in the shack. Crude, perhaps, but it gives *one-degree accuracy*, and in e.m.e. you can't afford less!

A handy item for telling if your array is pointing at the moon is the RCA SQ2520 photo-cell costing about $2, or its equivalent. This device is sensitive enough to detect the light of even a small sliver of moon. When placed at the end of a 20-inch long one-inch diameter tube and the leads connected to an ohm meter, it is an


Mounted on the lower end of the jack screw is the 20-to-1 reduction system. Note the collinear elements and main boom.


The elevation selsyn is mounted on the boom to the right of the mount. Note the jack screw, elevation plates and channeling holding the main boom on the mount assembly.

accurate indicator of proper aiming. Obviously it must be mounted so to be aimed along the exact plane of your array. It is useful only at night when the moon is visible.

As can be seen, the problems of mounting, steering and controlling an e.m.e. array are mostly mechanical and must be left to the ingenuity of the builder. Following the basic principles given here on locating the moon the builder may develop his own system.

It has been a pleasure to present these notes on e.m.e. problems, and it is my hope that many of you will become interested in building your own e.m.e. system. — *K6MYC*

The World Above 50 Mc.

CONDUCTED BY BILL SMITH,* K4AYO

Beginning Moonbounce—101

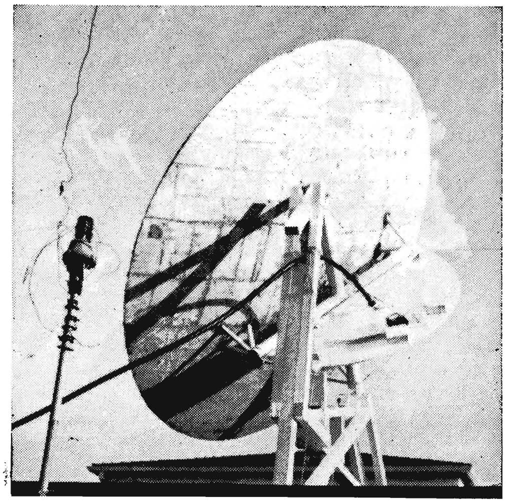
Each month we receive letters from prospective moonbouncers inquiring for reference material, and hints how to begin their e.m.e. project. In this column for January, February and March 1968 appeared a three-part series by K6MYC, designed especially to answer the most common questions. For those who do not have these issues, we'll paraphrase some of the highlights this month, but suggest you obtain the originals from a friend, or ARRL Headquarters at the nominal fee of 75 cents each.

Basically, this is what is required: 500 watts or more of transmitter output, the best possible receiver front end, a bigger antenna than most of us will ever erect, the means of aiming the array at the moon, and much perseverance. All, but the latter, may be store-bought, if you're so inclined.

Lets look at each. The transmitter power is easily acquired at 144 MHz., the most popular e.m.e. band, 220 MHz., where apparently there is no active e.m.e. work, and at 432 MHz. 1296 and up are progressively more difficult. There are numerous transistors capable of achieving the necessary noise figure at 144, many in the one dollar price range. The picture doesn't change too much at 432; at 1296 the device will cost 10 dollars or more.

The antenna, its type, size and aiming, may be considered together. Success has been had with collinears, Yagis, rhombics and dishes, or parabolic reflectors. The most popular, because it is tolerant of less-than-optimum amateur construction techniques, is the collinear. K6MYC designed, and later discussed in the April, 1967 edition of this column, a modification of a commercially available collinear. The modified version of that antenna is now on sale. At 2 meters, it is probably the best available commercial antenna.

Both SM7BAE and ZL1AZR, who together hold the world's 144-MHz. e.m.e. record, use multiple-Yagi arrays. Another promising 2-meter Yagi array was described by Oliver Swan at the recent West Coast V.H.F. Conference. In tests at K6MYC, a four-bay array of these Yagis, spaced 80 inches both horizontal and vertical, recovered the same amount of e.m.e. signal from KØMQS as did a 40-element collinear array. Physically the collinear array is about three times as large as the Yagi array. Details of this antenna will appear soon in *QST*.



Mounted on the roof of his Los Angeles home, this is the homemade dish of WB6IOM. He used this dish to successfully work G3LTF and establish a new 1296-MHz. moonbounce record. The 16-foot diameter dish consumed 450 square feet of sheet aluminum and 70 pounds of epoxy to bond the aluminum sheets. (WB6IOM photo)

Rhombics, used with much success by VK3-ATN and KØMQS, are capable of developing gain in excess of 30 db. over isotropic. Their disadvantages are physical size (several hundred feet in length) and fixed direction, except in the case of VK3ATN who varied the direction a few degrees by a pully and track arrangement. Rhombics are not feasible at the average city amateur location.

The parabolic reflector, more commonly known as a "dish," is essentially a low-efficiency antenna, something in the order of 35 percent. In addition, because of its physical size, especially at 144 and 432, it is not practical for the back-yard e.m.e. enthusiast. However, at 1296 and higher, good gain can be developed from a modest size dish. A picture of WB6IOM's 16-foot dish, used in establishing the world's e.m.e. distance record on 1296, appears elsewhere in this column.

Even more important is how you aim the array. It matters not how much gain the array has if it can not be aimed at the moon. Three systems are available; fixed position, partially steerable (polar mount), and fully steerable. A fixed-position array is the simplest to build. You have only to determine the place in space where the moon will travel through the array's pattern at a given time, and fix the array in that

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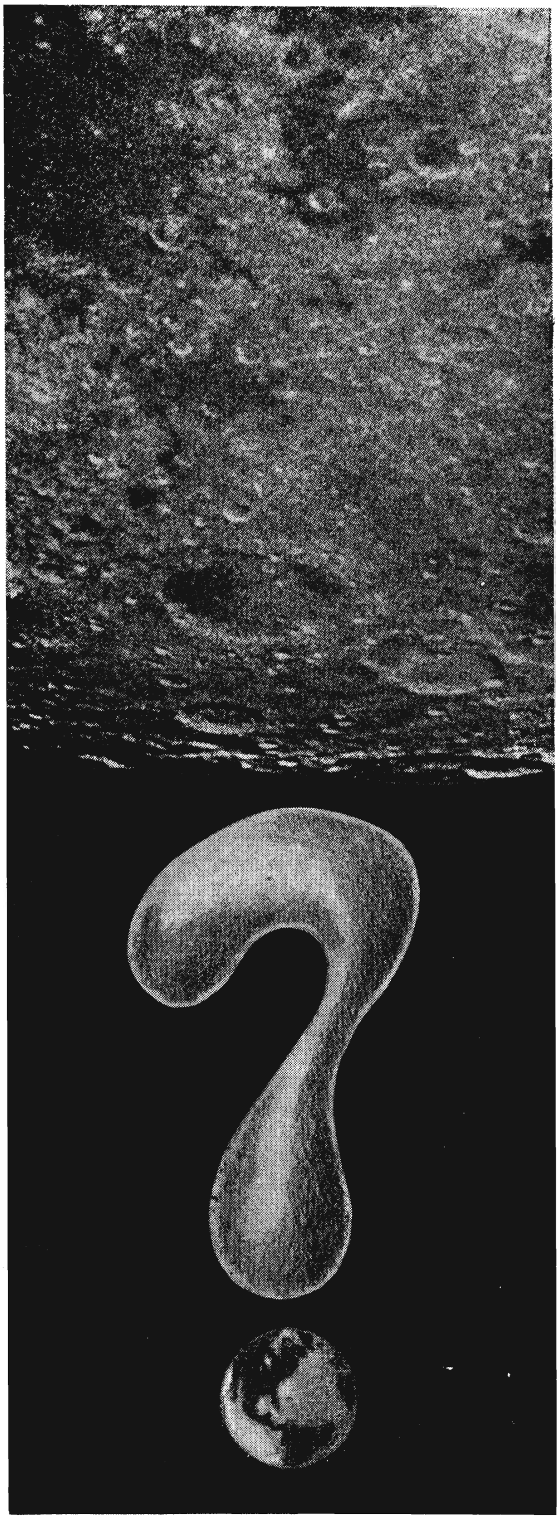
position. This method, however, limits the time each month the moon will pass through the antenna's pattern, and who you can work because of matching the "window." The window is a mutual place in space where antennas at both stations are pointed at the moon simultaneously.

The partially steerable, or polar mount, antenna is especially suitable for e.m.e. work. It needs to be set only once daily for declination (the angle in degrees north or south of the celestial Equator, or elevation angle) and then rotated in azimuth (horizontal plane) to track the moon. The moon travels across the sky at approximately 15 degrees per hour.

A fully steerable array, in both azimuth and elevation (az-el mount), is more flexible for use on other propagation modes, but is difficult mechanically to construct and calibrate for e.m.e. purposes. This is the most desirable type for satellite work.

All right, we've thrown out some facts; what do they boil down to? For the e.m.e. neophyte I'd suggest the following, and you e.m.e. greybeards may sit back and stroke them. Try 144 MHz., there is more activity, and technically 2 meters is more easily achievable. Construct a collinear array of at least 160 elements. That puts you into the 20-db. gain e.m.e. ballpark. Mount the array in a fixed position, taking into consideration who you wish to schedule. The mount may be modified at a later date to a polar configuration, after you become more familiar with e.m.e. techniques.

Much of this discussion may be directly applied to satellite programs, hopefully to soon again grace the amateur horizon through the Amsat and Nastar projects. E.m.e. and satellite work is within the grasp of many of us. As KØMQS recently said, "if I can work e.m.e., anyone can." What Dick said is true — *if* you have the perseverance to put the system together, and stay with it until it works. You still can't buy that!



How High the Moon

BY DON LUND,* WAØIQN

Reprinted from July, 1965 QST

THE YEAR 1964 will long be remembered by v.h.f. moonbounce enthusiasts. First, the patient work of Bill Conkel, W6DNG, and Lenna Suominen, OH1NL, paid off with the first two-meter moonbounce contact, and then KP4BPZ really showed the possibilities of such work. Postmortems on the week end of June 13 and 14 were held wherever v.h.f. men gathered, but one aspect of our participation seems clear. Many groups and individuals depended upon their ability to visually align their antennas on the moon. Cloudy weather meant failure: partly cloudy weather meant disastrous breaks in tracking the moon.

Getting around this trouble is really pretty easy. First, you need to know where your antenna is pointed. If you're using good rotators, the indicators tell you. If you're using an "Armstrong" system, attaching "setting circles," which are circular dials with 360-degree markings, will tell you. Now the only thing you need to know is where in the heck the moon is!

The purpose of this article is to show two ways of calculating where the moon will be in the sky at a given time on a given date. The first way is quick and dirty. With no mathematics and no references other than this article, it will predict the moon's position to an accuracy of 5 degrees or so for observers within the United States. Since an antenna with an honest 20-db. gain will have a half-power beam width of about 13 degrees, 5-degree accuracy should be acceptable for most applications. If this isn't good enough, a second way is described. It is both accurate and tedious. To use it, one needs a table of trigonometric functions and one reference book. Either of these methods will help you aim your antenna at the moon in fair weather or foul.

All of this discussion will be in terms of elevation and azimuth coordinates. Elevation is the height in degrees of the center of the moon above the horizon. Azimuth is the bearing of the moon, measured clockwise from North. For example, the elevation of the horizon is 0 degrees, and the elevation of a point directly overhead is 90 degrees. The azimuth of the eastern horizon is 90 degrees, while the azimuth of the southern horizon is 180, and so on. We are going to stick to "az-el" coordinates because this is the simplest type of mounting for an amateur to build and align. Also, because of the moon's rapid motion in declination (declination is the same, in celestial coordinates, as latitude in geographical coordinates), other types of mountings do not offer the advantage for the moon that they do for heavenly bodies with fixed declinations.

The Moon's Position; Quick Way

If we watch the moon's path across the sky for a month or so, we see that it shows a cyclic variation. The moon might, on the first night, rise quite high in the sky. The next night it would not rise quite as high, and the next night it would be even lower. After about 13 days it would be lowest in the sky, and the next night it would be

* 430 S. 45th St., Boulder, Colorado.

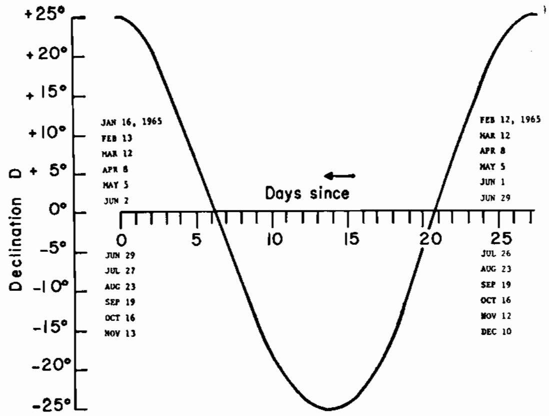


Fig. 1—The average declination of the moon during 1965.

higher again, until after about 27 days, it would be at its highest again. This is because the moon wobbles in declination. The wobble is almost sinusoidal, with a period of about 27 days, as shown in Fig. 1. The dates given are the starting dates for the oscillation. Since the period isn't exactly 27 days, it is necessary to slip a day every so often, as on September 18–19. The maximum amplitude changes about 1½ degrees during 1965, so the curve shows the mean declination, D, during 1965. Thus on July 26, 1965, the moon's declination is + 25 degrees. On August 2, seven days later, the declination is 0, while on August 3, D = − 4 degrees.

Knowing the moon's declination, we may compute its path across the sky (see Fig. 2). We see that when the moon's declination is most positive, it passes highest in the sky; when the declination is most negative, it is lowest in the sky. At some time, call it T, the moon is due South. At $T - 1$, that is one hour before T, the moon is on a solid line corresponding to the declination from Fig. 1, where it crosses the dashed line marked "$T - 1$." An hour and a half later, the moon is still on the same solid line, and is where the dashed line marked "$T + \frac{1}{2}$" crosses it. At $T - 1$ and $T + \frac{1}{2}$, we can read the moon's azimuth and elevation off the bottom and side scales. One word of caution about Fig. 2: It has been computed for an observer whose latitude is 40 degrees North, which is on a line passing through San Francisco, Indianapolis and Philadelphia. For observers north and south of this line, the elevation scale is squeezed or stretched. However, for the kind of accuracy we need, the curves will produce acceptable results over most of the continental United States, except Texas, Florida and Maine.

All that is needed now is to find the time, T, at which the moon is due South. This is shown in Fig. 3. Again, the dates are the starting times of the periods, which are about 29 days long. The time can then be read directly in local standard time. For example, the moon is due South at midnight on July 12, 1965. On August 3, 22 days later, the moon should be due South at about 4:40 P.M. local standard time. As before, Fig. 3 represents an average curve for 1965, computed for an observer at the middle of the United States. East and West Coast times may be off by several minutes.

In summary, the complete procedure is:

a) Given the date, find D from Fig. **1**.

b) Given the date, find T from Fig. 3.

c) Knowing D and T, enter Fig. 2, reading off azimuths and elevations at hourly intervals before and after T. For illustration, let's say we want the azimuth and elevation of the moon on August 3, 1965. From Fig. 1, $D = -4$ degrees, and from Fig. 3, $T = $ 4:40 P.M. In Fig. 2, the $D = -4$ degrees curve must lie a sixth of the way down from $D = 0$ degrees to $D = -25$ degrees. Pencilling a curve like that in, at $T - 3$, that is at 1:40 P.M., the azimuth is 127 degrees and elevation is 29 degrees. At 2:40 P.M., the azimuth is 141 degrees, and elevation 32 degrees. Following along, we can find elevation and azimuth every hour. Sounds a little complicated at first, but with some practice, it becomes quick and easy.

The Moon's Position: Exact Way

For the man who has everything — a 300-foot dish and an IBM computer — the easy way may be neither satisfying nor accurate enough. For the man with such excellent capabilities, we offer a cookbook which shows one way of computing the moon's elevation and azimuth. We won't define things like hour-angle, for these definitions would constitute a full course in astronomy. Rather, we will just tell you how to compute, and let you study the references if you wish.

The first step is to compute the local sidereal time, which we call T_s. Pick a Greenwich Mean Time, T_g, for which we want to compute the moon's position in the sky. At this point we must refer to *The American Ephemeris and Nautical Almanac, 1965* (or whatever year you wish) which is available from the Superintendent of Documents, U. S. Government Printing Office, Washington, D. C. Copies are often available at nearby observatories, and occasionally at nearby universities. In the *Ephemeris*, under the section

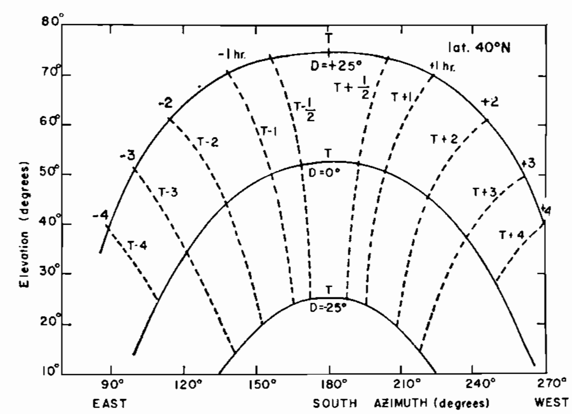


Fig. 2—Azimuth and elevation of the moon.

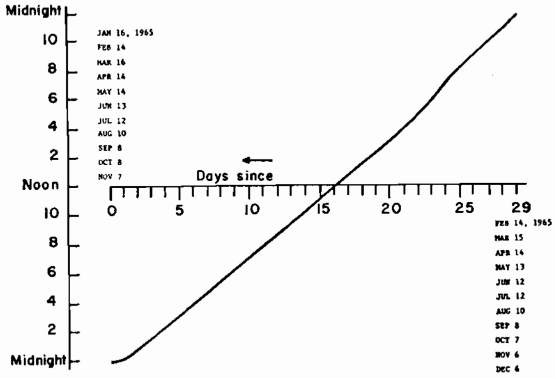


Fig. 3—The average local standard time at which the moon appears due south.

titled "Universal and Sidereal Times, 1965" there is a column called "Sidereal Time, Hour Angle of the First Point of Aries — Apparent." One entry is given for each day of the year. Look up the value for the date you wish, and call the value T_a. Then the local sidereal time may be computed from

$$T_s = 1.002778T_g + T_a - 24\frac{l_1}{360}$$

where l_1 is your longitude in degrees, west of Greenwich. Next, compute the hour angle of the moon; call it *h.a.m.* To do this, in the *Ephemeris*, in a section titled "Moon, 1965, For Each Hour of Ephemeris Time," for each date there is a column showing "Apparent Right Ascension" for each hour of time. Look up the Apparent Right Ascension for the date and time of interest; call it *r.a.m.* Then the hour angle of the moon, in degrees is:

$$h.a.m. = (T_s - r.a.m.)\frac{360}{24}$$

Next, we compute the elevation of the moon; call this angle E. This is computed from

$$\sin E = (\sin D \times \sin l_2) + (\cos D \times \cos l_2 \times \cos h.a.m.)$$

where l_2 is the latitude of the observer and D is obtained from the column "Apparent Declination" which is just to the right of the "Apparent Right Ascension" column in the *Ephemeris*. Having found E from the tables of trigonometric functions, look up $\cos E$. Then the azimuth, A, can be computed from:

$$\cos A = \frac{\sin D - \sin l_2 \times \sin E}{\cos E \times \cos l_2} \text{ and}$$

$$\sin A = \frac{\cos D \times \sin h.a.m.}{\cos E}$$

From the trigonometric tables, we can then look up A.

For the person who needs this accuracy, and has access to an IBM computer, a Fortran program for the above may be obtained by writing the author.

Summary

To permit aiming antennas at the moon through cloudy skies, we have shown two ways of computing the position of the moon in the sky. The first way is as simple as we know how to make it. Its accuracy is poor by astronomical standards, but should be sufficient for most amateur applications. The second way is more accurate, but involves tedious computations. We comment that we have ignored certain fine points in the second method, such as the difference between Ephemeris and Greenwich Mean Time and the fact that the Ephemeris values of right ascension and declination are as seen from the center of the earth. Such refinements can be introduced if the need for ultimate accuracy arises.

References

For a general reference which provides an excellent introduction to the terms and ideas used here, we would recommend *Astronomy*, by R. H. Baker (D. Van Nostrand Co., Inc., 1960). For more detailed information, which includes the derivation of expressions like those which we have used in the Exact Method, we could recommend *Elementary Mathematical Astronomy*, by C. W. C. Barlow and G. H. Bryan, as revised by H. S. Jones (University Tutorial Press, Ltd., 1961). Tables in Figs. 1, 2 and 3 were supplied by the High Altitude Observatory, Boulder, Colorado.

QST

POLAR MOUNTS FOR MOON TRACKING

Technical Editor, *QST*:

"How High the Moon?" in July 1965 *QST* was an excellent article on el-az antenna aiming, but there is one statement that gives me concern: "Az-el is the simplest type of mounting for an amateur to build and align. Also, because of the moon's rapid motion in declination, other types of mountings do not offer the advantage for the moon. . . ."

In my article (January 1965 *QST*), a discussion of designing and building a polar mount for moon tracking was discussed. While certainly condensed, it was complete enough for the serious amateur to get started on design work.

So let's compare mount methods. First of all, two movements are required for either type of mount, and they are made essentially the same way. The only real difference in a polar mount is that its axis is inclined to point at the North Star (northern latitudes). So there is practically no difference in the construction materials or work required to construct a polar mount.

The mention of alignment, I assume, means calibration of the mount. Here is where you can begin to appreciate the polar mount. Large high-gain antennas with sharp patterns don't always point electrically where you think they do by bore-sighting methods. Nature has provided us with the sun to find out where the antenna is pointing electrically. (If you can't hear sun noise, you don't have enough gain to work moonbounce, except possibly KP4BPZ.)

The *Nautical Almanac* (which is cheaper than the *Ephemeris & Nautical Almanac* mentioned in "How High the Moon?") has a simple hourly table which tells you in astronomical coordinates exactly where the antenna is pointing, once you have found the sun with the antenna. You simply calibrate your readouts to these figures. Now any time you want to find the moon, a check of the table will give you the exact settings for your mount.

If automatic tracking is desired, a simple clock-controlled drive on the hour-angle axis will keep your antenna on the moon from horizon to horizon. A change in the declination is required once a day, as the moon only moves about 2 degrees per day in declination.

In contrast, an el-az mount requires that two corrections, for both elevation and azimuth, must be constantly fed to the antenna. The corrections must be made manually, as no simple way is available to make an el-az mount auto-track, short of an IBM computer. — *Victor A. Michael, W3SDZ, Box 345, Milton, Penna.*

Reprinted from September, 1965 QST

Tracking the Moon—In Simple English

Practical Ideas for Designing and Aligning a Polar Mount

BY VICTOR A. MICHAEL,* W3SDZ

A MAJOR pitfall facing the prospective moonbouncer is the antenna mount and tracking system. Even a 50-foot dish is of no value in lunar communication, if it cannot be pointed at the moon and kept there. When we began our moonbounce efforts, many hours were spent pouring over astronomy texts. It was determined rather quickly that a whole new language would have to be learned for a proper understanding of the moon-tracking problem. Gathered together here are some of the essentials involved.

Earth-Space Relationship

Understanding the earth-moon relationship in space is the first step in solving the moon-tracking problem. This relationship is best illustrated with a polar mount, as in Fig. 1. A polar mount is simply an elevation-azimuth mount with its azimuth axis parallel to the axis of the earth. Thus a polar mount at the equator would have its axis parallel to the earth (horizontal, to the viewer on the ground), while at the North Pole the axis would be vertical, or at a 90-degree angle to the plane of the earth. Your latitude determines the position of the polar axis with respect to the earth's surface, as illustrated in Fig. 1. Once this is determined, we can proceed to a few other terms.

Celestial Equator. An extension of the earth's equator; the circle that would be formed at a right angle around the polar axis.

Meridian. The north-south line directly overhead.

Hour Angle. The angle in degrees to the right of the meridian. (Degrees can also be transferred into time: 15 degrees equals 1 hour; 1 degree equals 4 minutes.)

Declination. Angle in degrees north or south of the celestial equator.

Using the Nautical Almanac

This is the most important tool you will use in setting up, calibrating, and using your moonbounce antenna. It is available from the U. S. Government Printing Office for $2.00. Be sure you get the right book; there is a similar publication from the same source titled *The American Ephemeris and Empirical Nautical Almanac.* This is more expensive, and harder to use for amateur applications.

On page 39 is a portion of the tables found in the *Nautical Almanac.* It will be noted that the position of the moon is plotted for each hour of GMT. As an example, at 1200 GMT Jan. 1, 1965, the GHA (Greenwich Hour Angle) is 15 degrees 16.5 minutes. This means that the moon has passed overhead at Greenwich, and is now 15 degrees 16.5 minutes, or just over one hour, to the right of the meridian, as the observer faces south. The declination is given as S 23 degrees 39.5 minutes, which means that the moon is at this position south of the celestial equator.

Once you know where the moon is at Greenwich, a simple formula may be applied to determine its position with respect to your own location. The declination is always the same, no matter where you live. The only factor that changes is the hour angle. The Local Hour Angle (LHA) can be obtained by the formula

$$\mathrm{LHA} = \mathrm{GHA} \begin{array}{c} -\ \text{west} \\ +\ \text{east} \end{array} \text{longitude.}$$

Getting back to our example, suppose you l've at 75 degrees west longitude. We find that the moon would have an LHA of 300 degrees 16.5 minutes, or approximately 4 hours before meridian.

Mount Design Considerations

After you examine your almanac you will discover a few facts about the moon's habits that will help you to design a mount. First of all, the moon spends about two weeks above the celestial equator and about two weeks below. The maximum declination is about 25 degrees

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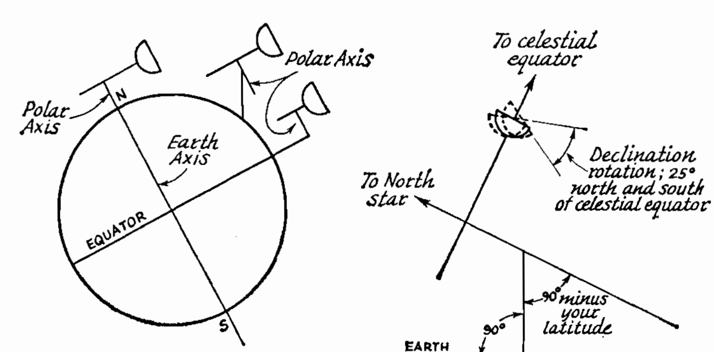

Fig. 1—Principles of the polar mount for moon tracking. At the left it can be seen that the polar axis is always parallel to the axis of the earth. Its position with respect to the earth's surface depends on the latitude of the observer. Two planes of rotation are required; the declination, which may be varied a small amount from day to day as required; and hour-angle, which should be controlled with a clock drive to follow the moon.

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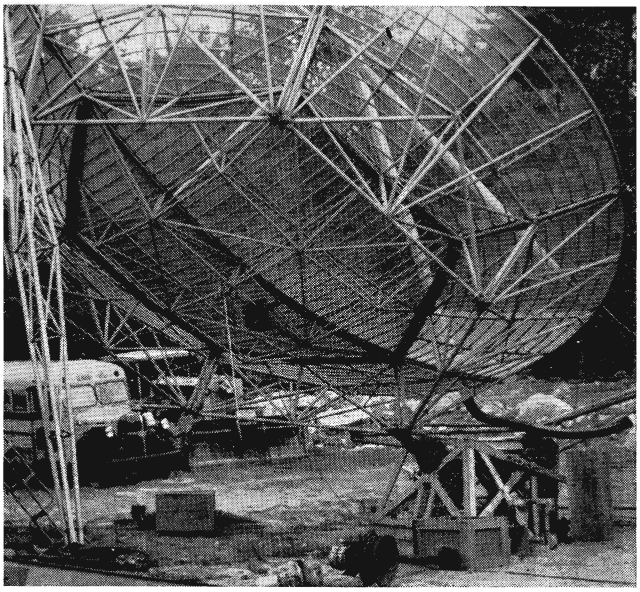

Fig. 2—Simplified polar mount and 28-foot dish at W1BU. Principles of the mount and its clock drive are explained in the text.

north or south of the celestial equator. Tracking ability for about 3 to 5 hours of hour angle each side of local meridian should be satisfactory.

At this point it is possible to make some compromises in order to simplify the mount in favor of a larger antenna. For instance, at W1BU, Sam gave up two weeks out of a lunar month in order to use the 28-foot dish of Fig. 2 in recent moonbounce tests. He can elevate the antenna above, but not below, the celestial equator. The high edge (upper left in the picture) is elevated to the desired position, while the lower edge rests on the pedestal at the lower right. The hour angle is controlled by a clock drive, just visible at the lower center. Though complex enough, this is far simpler than the true polar mounts used on the 18-foot dish at W1BU, or the mount and drive for the 256-element collinear array at W3SDZ, Fig. 3.

Calibration of the Mount

Obviously, if you are going to use the information in the *Nautical Almanac* with your mount, there must be some system of readout. There are

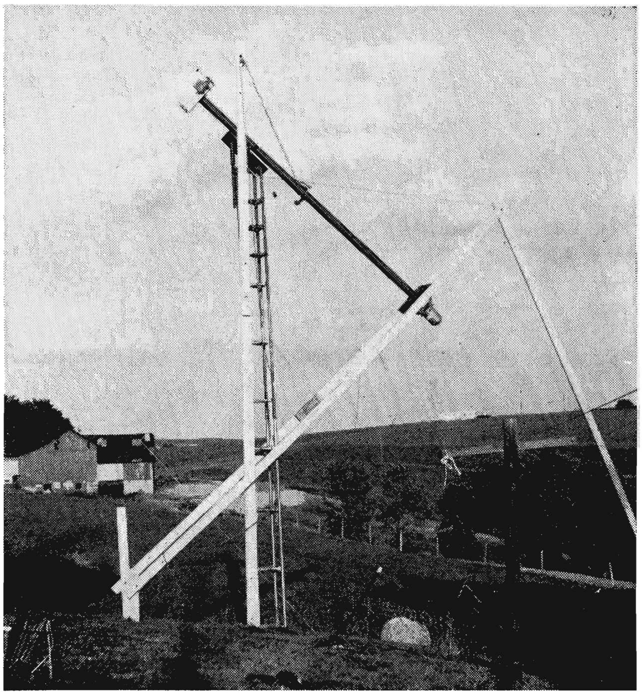

Fig. 3—Polar mount at W3SDZ, before the 256-element 432-Mc. collinear array was in place. The complete array is pictured in November 1964 QST, page 75.

many possibilities, and many different systems will evolve. As a starting point, a few ideas will be discussed here, and then "to each his own."

As a practical matter, the declination need be set no more than once per day, for it changes less than 2 degrees in 24 hours. For a few hours of moonbouncing effort each day, less than 1-degree variation is involved. Unless your antenna pattern is much sharper than the best amateur efforts to date on frequencies below 1300 Mc., this error is no problem. Declination readout can be rather simple: a calibrated scale on the antenna mount, a selsyn readout, or even a good 10-turn pot geared to the declination axis, and connected to a mercury battery and a meter. Anything accurate to plus-or-minus 1 degree should be all right.

Hour-angle readout and automatic tracking are the chief problems in moon tracking. The moon appears to move across the sky at slightly less than 15 degrees per hour. Actually, it is the earth that is moving at 15 degrees per hour. The moon is also moving, but at less than 1 degree per hour. Thus our basic problem is to drive the polar or hour-angle axis at 15 degrees per hour with a clock. The simple procedure of turning off the hour-angle drive for about 3 minutes once each hour, until the moon catches up, keeps things more than accurate enough for antenna tracking.

Now a "clock" doesn't necessarily have to look like a clock. For instance, a large synchronous 60-cycle motor driving a gear train at 1 revolution per day, coupled directly to the hour-angle axis, will work. The W1BU system is shown schematically in Fig. 4. Actual readout can be by any method that will develop plus-or-minus 1-degree accuracy.

When the mount is made, the antenna mounted, and the readout devices reading, the next question will be where is the antenna *really* pointing? This may sound simple, but most would-be moonbouncers have had trouble with this problem. Fortunately, nature has provided us an almost constant radio signal that permits a rather accurate calibration of the antenna system to be made. That signal comes from the sun. It will be seen that the almanac gives identical information on GHA and declination for the sun; thus, by listening to solar noise you can calibrate your polar mount in the same terms of reference as you will use in moon tracking. Some time spent reading K2LMG's "Antenna Patterns from the Sun," *QST* for July, 1960, will be well spent at this point.

What you have just read covers some of the essentials. It is hoped that enough information has been given to enable the prospective moonbounce enthusiast to determine his requirements for mounting and tracking. If any serious experimenter in this field needs help at this point, the author will be glad to try to be of assistance.

QST

G.M.T.		SUN		MOON				
		G.H.A.	Dec.	G.H.A.	v	Dec.	d	H.P.
d	h	° ′	° ′	° ′	′	° ′	′	′
1	00	179 09·2	S23 02·3	201 09·9	11·7	S22 48·3	4·9	54·0
	01	194 08·9	02·1	215 40·6	11·7	22 53·2	4·7	54·0
	02	209 08·6	01·9	230 11·3	11·7	22 57·9	4·7	54·0
	03	224 08·3	· · 01·7	244 42·0	11·6	23 02·6	4·5	54·0
	04	239 08·0	01·5	259 12·6	11·6	23 07·1	4·4	54·0
	05	254 07·7	01·3	273 43·2	11·5	23 11·5	4·4	54·0
	06	269 07·4	S23 01·1	288 13·7	11·6	S23 15·9	4·2	54·0
	07	284 07·1	00·9	302 44·3	11·5	23 20·1	4·1	54·0
	08	299 06·8	00·7	317 14·8	11·4	23 24·2	4·0	54·0
F	09	314 06·5	· · 00·5	331 45·2	11·5	23 28·2	3·8	54·0
R	10	329 06·2	00·3	346 15·7	11·4	23 32·0	3·8	54·0
I	11	344 05·9	23 00·1	0 46·1	11·4	23 35·8	3·7	54·0
D	12	359 05·6	S22 59·9	15 16·5	11·3	S23 39·5	3·5	54·0
A	13	14 05·3	59·7	29 46·8	11·4	23 43·0	3·5	54·0
Y	14	29 05·0	59·5	44 17·2	11·3	23 46·5	3·3	54·0
	15	44 04·8	· · 59·3	58 47·5	11·3	23 49·8	3·2	54·0
	16	59 04·5	59·1	73 17·8	11·2	23 53·0	3·1	54·0
	17	74 04·2	58·8	87 48·0	11·3	23 56·1	3·0	54·0
	18	89 03·9	S22 58·6	102 18·3	11·2	S23 59·1	2·9	54·0
	19	104 03·6	58·4	116 48·5	11·2	24 02·0	2·7	54·0
	20	119 03·3	58·2	131 18·7	11·2	24 04·7	2·7	54·0
	21	134 03·0	· · 58·0	145 48·9	11·1	24 07·4	2·5	54·0
	22	149 02·7	57·8	160 19·0	11·2	24 09·9	2·5	54·0
	23	164 02·4	57·6	174 49·2	11·1	24 12·4	2·3	54·0

Table I—Section of a page from the *Nautical Almanac*, showing solar and lunar data for each hour of January 1, 1965, at Greenwich.

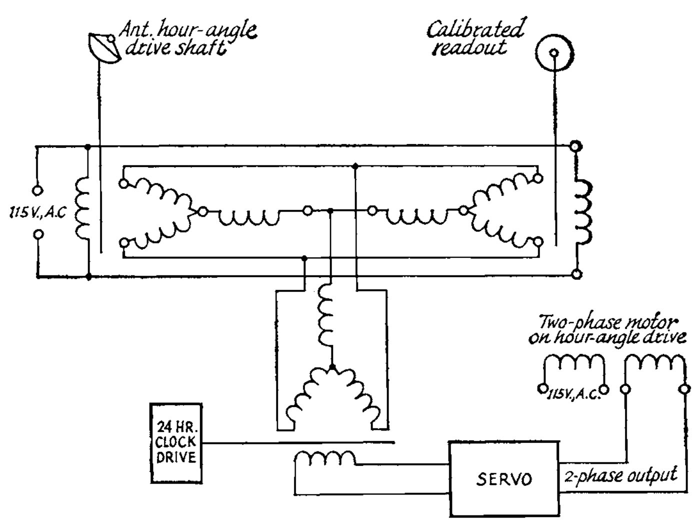

Fig. 4—Schematic diagram of the clock drive and readout system at W1BU.

USING SUN NOISE

BY DON LUND,* WAØIQN

One may hear the question "How many db's of sun noise do you get?" asked among serious v.h.f. men. Checking system performance of an advanced v.h.f. station by measuring the amount of noise received from the sun can be most useful, but there are some pitfalls that must be avoided. Our aim here is to set out the relationship between the power radiated by the sun, the antenna characteristics, and the receiver performance. If we know any two of these sets of parameters, we can measure the third. Finally, we'll explore some of the pitfalls inherent in talking about "db's of sun noise."

Solar Temperature, Antennas and Receivers

Twenty years or so ago, astrophysicists were arguing over whether the outer atmosphere of the sun was hotter than the visible surface. Radio astronomy provided some of the evidence that the outer atmosphere was much hotter than most astrophysicists had previously imagined. The result was that the "apparent temperature" of the sun increased with wavelength, at all wavelengths longer than a centimeter or so. (Apparent temperature comes in because the size of the sun is different at different wavelengths. So the sun is taken to be the same size as the optical sun, and apparent temperature is the temperature it would have, to radiate the measured power, at given wavelength, from this size of disc.)

What happens if we point an antenna at the sun? If the beamwidth of the antenna is just exactly the size of the sun, the antenna temperature will be the same as the temperature of the sun at this wavelength. Antenna temperature doesn't mean that we could burn a finger on the antenna; it means that the antenna is delivering the same amount of power to the transmission line that would be delivered by a resistor heated to the antenna temperature. This means that if we took a 50-ohm resistor, and heated it to 400,000°K, it would generate the same amount of noise at 432 Mc. as would be delivered to a 50-ohm resistor by an antenna with a ½-degree beamwidth pointed at the sun.

Antennas used by hams are not that sharp. If an antenna with a 10-degree beamwidth were pointed at the sun, its gain would be less, and it would deliver less power to the transmission line than a ½-degree antenna. Said differently, the antenna temperature would be lower for the broader antenna. In equation form

$$T_a = T_s \frac{\Omega_s}{\Omega_a}$$

where T_a is the antenna temperature, T_s the apparent temperature of the sun, Ω_s is the solid angle subtended by the visual sun (7×10^{-5} steradians), and Ω_a is the solid angle corresponding to the half-power beamwidths of the antenna[1]. If θ_H and θ_V are the half-angles to half-power beamwidths in the horizontal and vertical planes in degrees, then

$$\Omega_a = \frac{\pi}{4}\,\frac{\theta_H}{57.3}\,\frac{\theta_V}{57.3}, \text{ approximately.}$$

For illustration, an antenna which was 15° to the −3 db. points in the horizontal plane and 10° in the vertical plane would "see" a solid angle

$$\Omega_a = \frac{\pi}{4}\left(\frac{7.5}{57.3}\right)\left(\frac{5.0}{57.3}\right) = 8.99 \times 10^{-3} \text{ steradians}$$

The antenna is connected to a feed line which has some loss. If we call the feed line loss, when expressed as a ratio, A, we have

$$T_b = A\,T_s\frac{\Omega_s}{\Omega_a} + (1 - A)\,T_o$$

for T_b, the temperature at the receiver terminals due to the power received from the sun. T_o is the earth's temperature, usually taken as 290°K.

With no signal input, the receiver temperature is

$$T_R = (N - 1)\,T_o$$

where N is the noise factor of the receiver (noise factor is related to noise figure in the following way: if we express noise figure as a ratio, and add 1, we have the noise factor. A 6-db. noise figure corresponds to a noise factor of 5.).

If the sun noise at the output of the receiver is d decibels above the receiver noise, and if we converted to a ratio, call it D, then $d = 10 \log_{10} D$, and combining all the above, we have:

$$A\,T_s\frac{\Omega_s}{\Omega_a} + (1 - A)\,T_o = D\,(N - 1)\,T_o$$

The answer to "how many db.'s of sun noise" then is

$$D = \frac{1}{N-1}\left[A\,\frac{T_s\,\Omega_s}{T_o\,\Omega_a} + (1 - A)\right]$$

An equation much like this has appeared here before[2]; perhaps this presentation, which shows where such an equation comes from, will help in understanding what will be said later.

Let's work an example, showing how practical results may be predicted. If the receiver has a noise figure of 5 db., then $N = 4.16$. If the feed-line loss is 2 db., then $A = 0.631$, and if we are interested in 432 Mc. the apparent solar temperature is about 500,000°K for a condition when the sunspot number is 50 (see below). T_o is 290°K

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[1] For further discussion, see Pawsey and Bracewell, *Radio Astronomy*, Oxford University Press, Oxford, England, 1955, p. 21. Steradian: The solid angle subtended at the center of a sphere by a portion of the surface whose area is equal to the square of the radius of the sphere.

[2] See Bray and Kirchner, "Antenna Patterns from the Sun," *QST*, July 1960.

(about room temperature) and $\Omega_s = 7 \times 10^{-5}$. If the antenna beamwidth is 10° by 10° to the half-power points, its half-beamwidth to half-power points is 5° by 5°, and $\Omega_a = 6.0 \times 10^{-3}$. Then

$$D = \frac{1}{4.16 - 1}\left[0.631 \frac{5 \times 10^5}{2.9 \times 10^2} \frac{7 \times 10^{-5}}{6.0 \times 10^{-3}} + (1 - 0.631)\right] = 4.14$$

Converting this back to decibels, the sun noise should be almost 6.2 db. above the receiver noise for this system. There is one problem with this calculation: The sun radiates noise of both vertical and horizontal polarization (usually equal amounts) while most antennas accept only one polarization. If this is the case, the antenna only accepts half the incident radiation, and we must subtract 3 db. for polarization loss. In such a case, the sun noise would be 3.2 db. above receiver noise.

Making Measurements

The radio astronomer would measure N, A, and the antenna parameters, and then knowing these would measure T_s daily by measuring daily values of D. As hams, we are probably more interested in measuring the antenna parameters, or in monitoring our receiving system to make sure everything is working the way it should. This way leads to some trouble, simply because we don't know enough about T_s. At frequencies below about 1000 Mc., the apparent solar temperature isn't very well known for several reasons. The first is that not too many solar observatories have measured solar temperatures daily over a long period of time in this frequency range. While Potsdam, Ottawa, and Toyokawa, among others, measure daily solar temperatures between 1,000 and 10,000 Mc., and have over most of a sunspot cycle by now, not very many protracted measurements are available for the frequencies we are talking about. The second reason is that the solar temperature varies from day to day. Radiation at these frequencies comes from high in the solar atmosphere, and there is still much to be learned about this region of the sun. Therefore, solar temperatures often show little correlation with sunspot number, which is really a measure of activity in the lower part of the sun's atmosphere. The best guess that can be made as to solar temperature as a function of frequency, and the amount it increases for a Wolf Sunspot Number of 100, is shown in Table 1.

TABLE 1

Frequency	Temperature (Sunspot No. = 0)	Percentage Increase (Sunspot No. = 100)
144 Mc.	1,100,000°K	10%
220	1,100,000	12%
432	400,000	50
1296	150,000	100

These values have been obtained by comparing the reported results of Allen[3] with the daily values reported by the Toyokawa Observatory of the Research Institute for Atmospherics of Nagoya University. The accuracy of these values is not very good.

With this caution in mind, some good information can be obtained from monitoring solar temperature. One thing that can be done is to find, experimentally, what the beamwidth of an antenna is. If D turns out to be more than 2 (that is, 3 db. above receiver noise), we can find the half-power beamwidth ($2\theta_H$ and $2\theta_V$) by pointing the antenna at a point in the sky that the sun will cross, and letting the sun slowly drift through the antenna pattern. When the sun is in the center of the antenna pattern, put a 3-db. attenuator *between the antenna and transmission line* (not between the converter and i.f. strip). Such an attenuator is easily made from coaxial cable (about 29 feet of RG-58/U for 432 Mc.). Clock the times at which the receiver output from sun alone is the same as with the sun at the center of beam and the 3-db. attenuator in line. Since the sun drifts one degree every four minutes, dividing the minutes (between calibrated −3 db. points) by 4 gives the half-power beamwidth in degrees (2θ). Turning the antenna on its side and repeating will measure the beamwidth in the other plane. If the antenna does not give more than 3 db. of sun noise, you will have to use a signal generator, and rotate the antenna to measure these beamwidths.

Knowing the beamwidth and the feed-line loss, one can measure the receiver noise figure (assuming a value for T_s). This can be compared with the noise figure measured by using a noise generator. If by measuring solar temperature, using the values you think are correct for your system, you come close to the values shown in Table 1, then you can be sure that your system is performing properly. By measuring these things daily, you can check the performance of your total system.

Summary

In the preceding sections, we have discussed how to measure receiving system parameters, and how to monitor system performance to guard against deterioration. Should you suddenly get 1 db. of solar noise, when you have been getting 3 db., you know that your system needs some checking. Finally, we discussed some of the reasons why this is not an exact measurement, but rather should be taken as an indicator of system performance. QST

[3] Allen, "The Variation of Decimetre-Wave Radiation with Solar Activity," *Monthly Notices of the Royal Astronomical Society*, p. 174 (1957).

Antenna Patterns from the Sun

Using Solar Noise for Plotting Vertical Patterns of V.H.F. Arrays

BY D. W. BRAY,* K2LMG AND P. H. KIRCHNER,* W2YBP

Your v.h.f. antenna is probably aimed at the horizon, but that's not where your signal is going. The radiation in the vertical plane is actually pointed up in the air. This is true whether your antenna is horizontal, vertical or circularly polarized. It occurs because half of the antenna is looking at the ground when it is directed at the horizon. Since the beam strikes the ground, energy bounces off the earth and combines with the energy that is headed skyward.

For a horizontally polarized antenna, which most v.h.f. men use, the reflection is such that the signal is cancelled at the horizon and a sharp beam is formed a few degrees above the horizon. Above this lobe many other weaker and higher-angle lobes are formed. A typical vertical antenna pattern for an antenna a few wavelengths above the ground is illustrated in Fig. 1. A vertically polarized antenna would, if the ground were a perfect conductor, add the reflected and direct signal to produce a lobe with its maximum right on the horizon. But the ground isn't a perfect reflector and it causes the same effect as in the horizontal antenna: a lobe structure that is pointing upward.

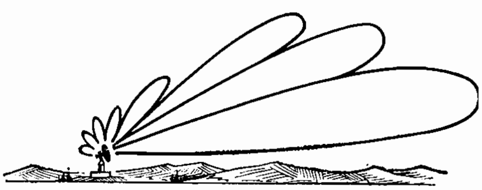

Fig. 1—No antenna ever has a single lobe, aimed precisely at the horizon. As shown in this artist's conception, the main lobe is always at least a few degrees above the horizontal, and other lobes appear above it at higher angles.

The only way to beat this is to raise your antenna up as high from the ground as practical to make the first vertical lobe as low as possible. Even though your antenna is as high as you could put it, it still poses the question: is it really high enough; what would another ten feet in elevation buy? Or another question can come to mind as it did at K2LMG. Is that h.f. array, which is below the v.h.f. beam, acting as a ground plane for the latter, causing a high angle of radiation?

The way to answer such questions is to make a vertical antenna pattern plot. That sounds easy but it can't be done as simply as taking a horizontal radiation pattern. To plot a horizontal pattern all you have to do is to have a nearby friend turn on his transmitter, rotate your antenna, and read his signal strength on your S meter.

Although not so easy as the horizontal pattern, the vertical pattern can be plotted by taking a clue from the radio astronomers. Mother Nature has provided a strong and fairly constant source of radio energy a long way off: the sun. In the course of an afternoon the sun sweeps through a range of elevation angles as the earth turns. If you track the sun in azimuth, you can measure the noise level received by your v.h.f. converter as the sun moves down the vertical plane of your antenna. Using the level of the received signal and the calculated position of the sun, the antenna pattern can be plotted.

By this method the question of the h.f. beam at K2LMG was answered. It showed that there was no interaction. If we have excited your curiosity to the point where you would like to run through this experiment for your own antenna, the method is outlined below.

Solar Noise

Since the signal to be received is wide-band noise, the receiver should be opened up to the widest possible bandwidth, so that the greatest amount of noise energy will be collected. But, since this energy is noise a special detector-integrator voltmeter circuit should be used on the output of the receiver in order to smooth out the variations in the noise for more accurate voltage readings. The receiver should be operated with the b.f.o. set for normal c.w. reception. Connect one of the detector-integrators of Fig. 2 to the audio output. One is for use with a volt-ohmmeter, where R = (lowest full scale voltage) × (voltmeter ohms per volt) and $C = \frac{10^6}{R}$ μf., with at least a 6-volt rating. The other is for use with a vacuum-tube voltmeter. If the v.o.m. is used the detector must be connected to the high-impedance tap (the higher impedance the better) of the receiver audio output transformer. It doesn't matter which audio output is used for the v.t.v.m. circuit. Even a high-impedance headphone circuit can be used. The received signals will be read on the voltmeter.

Since the gain of even the very best of receivers will change with time, the effects of

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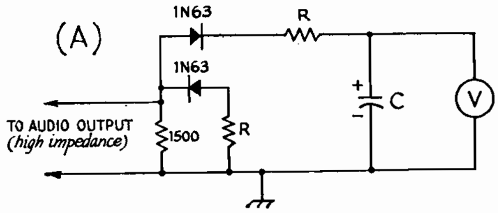


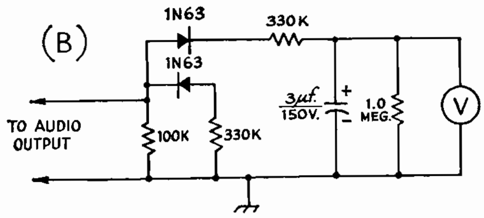


Fig. 2—Detector-integrator circuits for use in taking noise readings. Circuit A is for volt-ohmmeters, B for vacuum-tube voltmeters. See text for information on R and C values in circuit A.

receiver gain variations can be removed by putting a resistor equal to the impedance of the transmission line on a coax connector and substituting it for the antenna and transmission line at the converter input just before each reading is taken. Noise from the resistor provides a "standard signal."

Taking Data

Data can be taken at either sunset or sunrise. By the method outlined in the following section a graph of the angle to the sun as a function of time should be plotted for the selected day previous to taking data. This graph will then give the beginning and ending times of the test. For a sunset measurement, start at the time of maximum desired elevation angle and continue taking readings until about ten minutes after sunset. For a sunrise measurement you should start about ten minutes before sunrise and continue until the time for the maximum desired elevation is reached.

During the run the horizontal position of the sun will change, so it must be tracked in azimuth. Since your beam is relatively broad in azimuth, probably only a few changes in the horizontal position will have to be made.

The readings should be taken every few minutes, and about once a minute at low elevation angles. Each reading should include the time, the voltage output of the receiver with the resistor substituted for the antenna, and the voltage output with the antenna connected. Set the receiver gain so that the voltmeter is about two-thirds full scale on the most sensitive d.c. scale. At times the sun noise will flare up, or a car will go by the house, so caution should be observed to be sure that the reading of the antenna signal is really a good average value. The reading of the sun won't be much larger than that of the resistor and in fact at times will be lower than the resistor, so don't be alarmed by only small changes of voltage. Since the changes are small, care should be taken to achieve accurate results.

Calculating the Sun's Elevation Angle

The apparent motion of the sun is fairly complicated, and you will have to be prepared to do some work here. If you are interested only in the elevation angles below 15 degrees, and are willing to settle for about 1-degree accuracy, the job isn't too bad. The first step is to find your latitude and longitude from a map, and select a date for making the measurements. From Table I, look up the sun's latitude on the selected date, interpolating between tabulated dates.

Second, find out what time the sun will set (or rise) on the chosen day. If you live in a large city you can simply consult a local newspaper or the TV weatherman. The authors have found that in smaller cities these sources sometimes quote times which actually apply to a larger city nearby, and are not accurate enough for our purposes. The same applies if you live more than 15 miles out in the suburbs. In this case, find the correct time from one of the references listed at the end of this article, following the instructions given with the tables.

Next, calculate the number A from the following formula:

$$A = \sqrt{\cos^2 L - \sin^2 D}$$

where L is your latitude and D (declination) is the sun's latitude.

Now, for any time which is M minutes before sunset (or after sunrise), the sun's elevation angle in degrees is equal to A times M divided by 4, or

$$\phi^\circ = \frac{A}{4} M$$

To extend your pattern to elevation angles higher than 15 degrees you will have to work a little harder. In addition to finding the number A, find another number B from this formula:

$$B = \sin D \sin L$$

where D and L are as before. Remember that

TABLE I
Sun's Latitude Variations

Date	Sun's Latitude	Date	Sun's Latitude
Jan. 1	−23.0	July 4	23.0
9	−22.0	12	22.0
21	−20.0	24	20.0
29	−18.0	Aug. 1	18.0
Feb. 8	−15.0	12	15.0
22	−10.0	27	10.0
Mar. 8	− 5.0	Sept. 10	5.0
20	0.0	23	0.0
Apr. 3	5.0	Oct. 6	− 5.0
16	10.0	20	−10.0
May 1	15.0	Nov. 3	−15.0
12	18.0	14	−18.0
21	20.0	21	−20.0
June 1	22.0	Dec. 2	−22.0
10	23.0	11	−23.0
22	23.4	21	−23.4

when D is negative, sin D is also negative.

For a time M minutes before sunset (or after sunrise) find an angle X degrees by dividing M by 4. That is,

$$X^\circ = \frac{M}{4}$$

Now find the elevation angle ø from the equation

$$\sin \phi = A \sin X + B\,(1 - \cos X)$$

The procedure described above gives the elevation angle of the sun at any time it is above your horizon, to about 1 degree accuracy. To get a better picture of the fine structure of your antenna pattern, especially at the low angles which are most important, better accuracy is needed. About 0.2 degree can be achieved by careful calculation and by applying certain corrections.

Read your latitude and longitude to 0.1 degree or better, and find the sunset (or sunrise) time to the nearest minute. Now adjust this time slightly by $\frac{3.3}{A}$ minutes. Add this to the sunrise time, or subtract it from the sunset time. Use this adjusted time to calculate the elevation angles as described above, and then apply a final correction by adding the amount shown in Fig. 3 to the calculated values. These corrections take into account the refraction of the signal (and the light) by the atmosphere, and the difference in size between the radio sun and the visible sun.

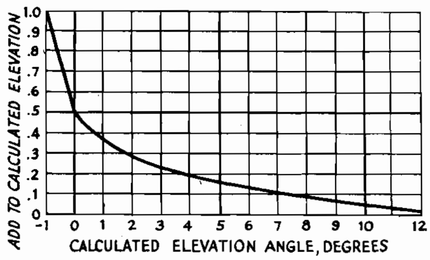


Fig. 3—Chart showing elevation angle corrections to be applied for results of high accuracy.

Finding the Sun's Azimuth

If you do the job on a sunny day, the simplest way is to have a friend keep the beam pointed toward the sun in azimuth, lining it up by eye. Alternately, calculate the sun's azimuth in advance and rotate the antenna from time to time as required. When readings are taken, the beam should be within about one-fifth of a beamwidth of the sun's azimuth.

Azimuth is found from the formula

$$\cos \Theta = \frac{\sin D - \sin L \sin \phi}{\cos L \cos \phi}$$

ø is the elevation angle already calculated. Again, remember that when D is negative, sin D is also negative. The azimuth, Θ, is measured eastward from north in the morning, and westward from north in the afternoon. When cos Θ comes out negative, Θ is larger than 90 degrees and the sun is more south than north.

Plotting the Results

Now that the angle of the sun and the signal-strength readings have been obtained, the antenna pattern can be plotted. Fig. 4 is a curved-earth grid with elevation angles plotted on it. Taking the readings that were made as the sun ran its course, divide the signal voltage from the sun (E_S) by the signal voltage from the resistor (E_R). Do this for each reading taken. Now square each of these values of (E_S/E_R) to obtain the value of $(E_S/E_R)^2$. The next step is to compute the value Y, using the equation

$$Y = \sqrt{\frac{\left(\frac{E_S}{E_R}\right)^2}{\left(\frac{E_S}{E_R}\right)^2_{\min}}} - 1$$

where $(E_S/E_R)^2$ is each of the readings that were taken as the sun crossed your antenna and $(E_S/E_R)^2_{\min}$ is the value of the reading after the sun is below the zero-degree elevation angle by 5 minutes or more. Then find the greatest value of Y. At this reading, calling it $Y_{\max}$, assign an arbitrary value of slant range — 500 miles. This is then one point on the plot: 500 miles and the angle to the sun at that time. Now take 500 miles and divide it by $Y_{\max}$ and multiply all of the other Y values by this amount. Plot on Fig. 4 the angle for each signal-strength reading and distance just found. Drawing the curve, you now have your antenna pattern in the vertical plane.

There is one caution. The sun is not really a point source of radio waves. It can be represented as a ring of about 1-degree angular diameter on the outside and about one-half degree on the inside. Because of this, the nulls in the antenna pattern will not appear to be sharp. For this reason, a sample antenna pattern is shown in order to guide you in your plot. When the curve shows a dip, it probably is a very deep null as indicated by the dotted lines on the same curve, Fig. 5. Because the depth of the nulls cannot be determined, the antenna pattern taken by this method would probably not satisfy an exacting scientist, but in practice the signals that are received on such an antenna, amateur or otherwise, are not from point sources either. Thus the antenna pattern taken by this method is truly an operational pattern.

For those interested in meteor scatter an estimate of optimum range can be made. The memeteor trails will be most prevalent at a height of 50 miles. From your antenna pattern note the range at which the elevation angle line through the peak of your lowest lobe intercepts the 50-mile height. Multiplying this number by 2 will yield your approximate optimum meteor scatter range. In the example shown in Fig. 5, this would be about 1000 miles.

Noise Figure and Antenna Gain Check

There is another interesting sidelight to this

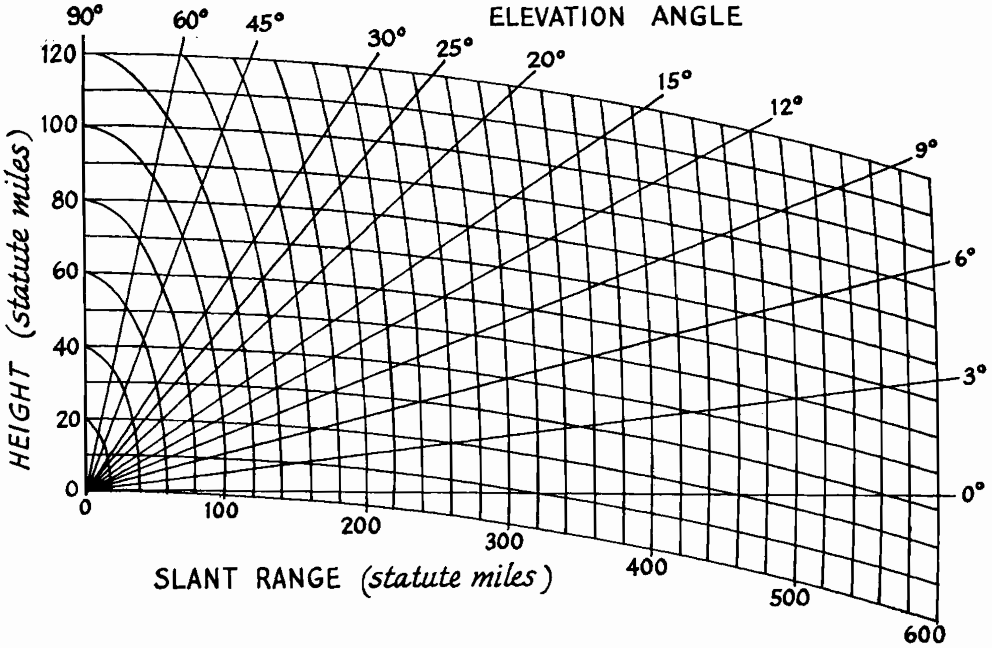


Fig. 4—Curved-earth grid for plotting results obtained from solar noise readings.

subject. The "quiet" sun is a more or less calibrated source of radio energy. Thus by a few simple calculations you can get an idea of your antenna gain or noise figure for actual received signals. Because the amount of energy that is received is a function of both the noise figure and antenna gain, you can start with one of the known values and find the other. The equation which applies is

$$\frac{G_P}{F_P} = 290\,\frac{L_P}{K}\left[\left(\frac{E_S}{E_R}\right)^2_{max} - \left(\frac{E_S}{E_R}\right)^2_{min}\right]$$

where G_P = the power gain on your antenna
F_P = the noise figure expressed as a power ratio
L_P = the transmission line loss for your cable and your length
K = a constant dependent upon the frequency band

and

$$\left(\frac{E_S}{E_R}\right)^2_{max}$$

is the maximum signal ratio from the antenna pattern data taken above. This value will occur at the peak of the first vertical lobe. $(E_S/E_R)^2$ min is the signal ratio at the time the sun was a few minutes below the horizon.

This formula will only apply when the sun is quiet. If the answers are out of line the test should be repeated until a quiet day is found. A quiet sun radiates the lowest amount of energy; all other conditions produce greater received power.

Your antenna gain is probably the least well-known number of your radio system.

TABLE II

Frequency Band	*Value of K*
144 Mc.	2.9
220 Mc.	2.8
432 Mc	2.7
1296 Mc	0.65

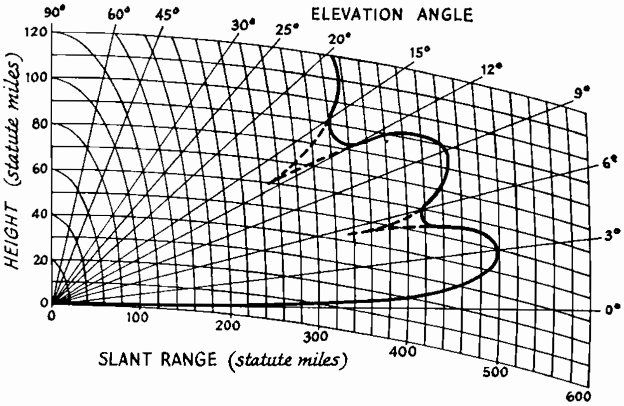
To calculate the antenna gain:

1) Estimate the noise figure of your converter by taking the manufacturer's noise figure, or from tube data if it is a home-brew model. This value will probably be expressed in db. Convert the db. noise figure to a power ratio by the common db. formula, F_P = antilog $F/10$ where F is the noise figure in db. from above. This conversion can also be made using the decibel chart in the *ARRL Handbook*.

2) The factor K is listed in Table II for the various amateur bands above 50 Mc. The 6-meter band has been omitted because of the strong background of radio energy in this frequency range in large areas of the sky, which could adversely influence the results. For the higher bands the background radiation is much less. It is possible that one of the bright radio stars could be near the sun when the measurement is being taken, and would therefore influence the readings on the higher frequencies, also, but the chances of this are remote.

3) L is the line loss. This figure is easily estimated by looking up the transmission-line manufacturer's data for your frequency. It is usually expressed in db. loss per hundred feet. Thus, calculate the db. value for your length and convert the db. loss to a power ratio as you did above for the noise figure.

Fig. 5—Representative vertical antenna pattern. Dips in the heavy line represent nulls in the pattern which are actually much deeper than data will indicate. This is due in part to the fact that the sun is not a true point source for radio noise. Antenna pattern may be more like that shown in dotted lines.



Substitution of the values in the formula will yield the power gain G_P of the antenna. This can be converted to db. gain by the common formula

$$G = 10 \log G_P,$$

or by using the *Handbook* table.

What we have really been talking about here is a practical use of radio astronomy. The methods used here also apply to the detection of radio stars. Many interesting experiments can be performed. For those who are interested, take a radio look at Cygnus A or the center of our galaxy in Sagittarius, when they are rising or setting. Both are good strong noise sources, and real DX!

References

American Ephemeris and Nautical Almanac, issued yearly by U. S. Naval Observatory. Consult it at library.

Astronomical Phenomena for the Year, published annually. A reprint of selected pages from the above, 25 cents from Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C.

The Telephone Almanac, issued annually. Free from Bell System Telephone Company business offices.

Information Please Almanac, published by Macmillan Company, New York City; sold at newsstands and bookstores

MORE 50-MC. MOONBOUNCE EXPERIMENTS

Aylmer
Quebec
Canada

Technical Editor, *QST*:

The purpose of this letter is to give corrections and additions to a previous one,[1] and to describe further 50-Mc. moonbounce experiments at VE3BZS/2.

In the formula for the Doppler shift, the transmitter, frequency, f, should have been expressed in cycles, not megacycles. Also, it was mentioned that the antenna was usually aimed optically. However, the following formulae were referred to when the moon was obscured by clouds:

$$\sin E_T = \sin L_T \sin D + \cos L_T \cos D \sin H_T$$

$$\sin A_T = \frac{\cos D \cos H_T}{\cos E_T}$$

where E_T is the elevation of the transmitting antenna
A_T is the azimuth from true north
L_T is the latitude of the transmitter
H_T is the hour angle of the moon, and is approximately $\frac{360}{(transit)} \times t$, where t is the time in hours after local mean time of moonrise at the equator, and *(transit)* is the time in hours between ephemeris transits of the moon (approx. 25 hours).

Similarly for the receiver.

The formulae are approximations, since E_T is the elevation of the moon at the earth's center, not at the station. However, the difference is less than one degree, at most. Also, the local hour angle definition may not be standard.

The moonbounce experiments were continued with different antenna polarizations to see if improvement could be obtained. It was pointed out by Soifer[2] that crossed Yagi antennas transmit and receive the same sense of elliptically-polarized radiation. Thus, theoretically, assuming specular reflection of the radio wave at the moon's surface and therefore reversal of the sense of the polarization, the antenna used in the experiments previously described should not have received the transmitted echoes. Neglecting effects of the ionosphere and lunar surface, the fact that some echoes were received is probably attributable to ground reflection effects and/or transmission of elliptically-polarized waves due to mismatch in the antenna system.

[1] "50-Mc. Moonbounce Experiments," Technical Correspondence, *QST*, May, 1962, p. 49.

[2] Soifer, "Research, Tracking and Reporting, Project Echo A-12," *QST*, June, 1962.

A trial was made recently using two of the crossed Yagis transmitting radiation polarized in one sense and the other two receiving in the opposite sense. Stronger and more frequent echoes were recorded, even though the system gain (neglecting the problem of reverse polarization) was only one-fourth that previously used.

Another trial was then made with the four antenna units transmitting vertically. This was chosen over horizontal to have the major lobes of the upper and lower bays as nearly coincident as possible, in the event of ground reflection. During this trial the moon was obscured, and some power-line interference was present, but results were the best so far.

The 50-Mc. trials were brief and incomplete, but results seem to indicate that the echo amplitude varies widely; that the average signal-to-noise ratio of the echoes is less than that given by the formula in the previous letter; that, when using circular polarization, improved performance results if the transmitting and receiving antennas have opposite polarizations; and that, with the system parameters used, no distinct advantage between circular and linear polarization was noticed.

— *Alan Goodacre, VE3BZS/2*

SUN NOISE

Technical Editor, *QST:*

Some questions have arisen over the definitions used in the article on sun noise (April 1968 *QST*, page 42), and perhaps a note clarifying these is in order. To conform to common usage, as stated in the *IEEE Standards*, the relation between noise figure and noise factor should be:

Noise Figure = 10 log (Noise Factor).

Thus, ". . . add 1 . . ." is incorrect, and a noise figure of 6 db. corresponds to a noise factor of 4. Also, as used in the equations, A is the fractional transmission of the feed line, rather than its fractional loss. Fractional transmission equals one minus fractional loss, resulting in A being correctly given in the example. However, the redefinition of N, as above, changes the results of the example so that the sun noise, taking one linearly-polarized component, should be 4.8 db. above the receiver noise, for the other constants assumed in the example.

One last word of caution: the whole presentation was based on the ratio signal/noise rather than signal + noise/noise. For systems where the sun is only slightly above the noise, reduction to signal/noise ratio, as described in January 1968 *QST*, page 34, may be required. — *Don Lund, WAØIQN, P.O. Box 1664, Boulder, Colorado 80301.*

50-MC. MOONBOUNCE EXPERIMENTS

119 Fourth Ave.
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Ontario, Canada

Technical Editor, *QST:*

Communication by moon-reflected radio waves offers amateurs the opportunity for making v.h.f. contacts anywhere in the world, for periods of from a few minutes to several hours each day, if the considerable technical problems can be solved. The following describes some attempts to obtain moon echoes on 50 Mc.

Previous work with the reception of 50-Mc. transmissions by W7RDY at VE7AIZ gave evidence of perhaps two or three consecutive weak echoes during each of half a dozen trials.[1] Lack of more consistent results was assumed to have been due to Faraday rotation of the plane of polarization in the ionosphere, causing loss of signal. Antennas at both ends were horizontally-polarized Yagis. It seemed worth while to make another attempt at VE3BZS, using circular polarization, in a manner similar to that used by K1HMU [2] to overcome the Faraday-effect problem.

Four Yagis, each with 5 elements in a horizontal plane and 5 in a vertical plane, were arranged in box configuration approximately 20 feet square. The vertical driven elements were fed 90 degrees out of phase with the horizontal ones. The antenna could be rotated only in azimuth, and was usually optically aimed. The transmitter used a heterodyne exciter for maximum stability. An external tunable oscillator at 1 Mc., with 50 times frequency multiplication, gave receiver injection at 50 Mc., plus the Doppler shift, plus or minus the audio filter frequency of 940 cycles. This beating signal and the returned signal, if any, are fed into the regular 50-Mc. converter, and then into the station receiver, set for 609-cycle bandwidth. Then follows the 940-cycle audio filter, with a bandwidth of 20 cycles, and a tape recorder.

An important receiver point is that the gain of the receiver should be set so that the noise at the output of the audio filter disappears when the external injection is turned off. Under this condition the effective predetection (i.f.) bandwidth of the receiving system is determined by the audio filter. The heterodyne system for the transmitter allows the oscillators to run continuously, permitting better frequency stability than when turning the oscillator on and off. Heterodyning also reduces the drift at the signal frequency, for a given amount of oscillator drift, compared to a conventional oscillator-multiplier system. Absolute frequency stability was not extremely good, due to lack of temperature control of crystals and transistors, but relative drift to the rece ver was from one to two cycles per minute. This is good enough to permit audio filter selectivity of 10 to 20 cycles to be used.

Because of this narrow bandwidth the Doppler shift had to be calculated. The approximate formula used was:

$$\Delta f = f\left[\frac{37.04}{(\text{transit})}(\cos L_T \cos H_T + \cos L_R \cos H_R)\cos D \times 10^{-6} + 5.54\left(\frac{1}{(\text{semi})_1} - \frac{1}{(\text{semi})_2}\right) \times 10^{-2}\right]$$

where Δf = Doppler shift in cycles ;

f is = transmitter frequency in megacycles

$(\text{semi})_1$ = semidiameter of the moon expressed in seconds of arc

$(\text{semi})_2$ = semidiameter of the moon expressed in seconds of arc 12 hours later for the day concerned

(transit) = the time in hours between ephemeris transits of the moon (approx. 25 hours)

L_T = latitude of the transmitter

H_T = hour angle of the moon and is approximately $\frac{360}{(\text{transit})} \times t$ where t = time in hours after local mean time of moonrise at the equator

L_R, H_R similarly for the receiver

D = apparent declination of the moon

The necessary information for the calculation may be obtained from a current *American Ephemeris and Nautical Almanac.* The first term in the square brackets is usually dominant and at moonset at 45 degrees latitude amounts to about − 110 cycles at 50 Mc.

Three trials were made and only one or two weak but identifiable echoes were received. Signal-to-noise power ratios were of the order of 1:1, or less. This means that little or nothing can be heard of the return signal by ear, but a visual presentation shows evidence of a return. The advantage of visual methods in detection of very weak signals increases with very narrow receiver bandwidth, since signal and noise tend to sound the same under these conditions.

The average signal-to-noise ratio at the output of the audio filter was calculated using the following formula, which neglects fading effects produced by the motion of the moon's surface, Faraday rotation and ground reflection:

$$\left(\frac{S}{N}\right)_{\text{POWER}} = \frac{1.6 \times 10^{-26}\, G_R \lambda^2\, G_T P_T\, 10^{\frac{-2KL}{10}}}{4.1 \times 10^{-21}\left(.22\lambda^{2.4}\; 10^{\frac{-KL}{10}} + F - 1\right) B}$$

where P_T = transmitter power output in watts

K = attenuation of transmission line in db./100 ft.

L = transmission line length in units of 100 ft.

λ = wavelength in meters

G_R = gain of receiver antenna over isotropic radiator

G_T = similarly for transmitter

F = noise figure of receiver at wavelength λ

B = effective noise bandwidth of receiver in c.p.s.

It should be noted that P_T, K, and F vary with λ for given components. Frequency stability problems make minimum B vary with λ also. For given conditions there is an optimum λ to produce maximum average signal-to-noise ratio. The words "average signal-to-noise ratio" are used, since the instantaneous noise power may deviate from the average value, but the actual signal-to-noise ratio should be within a factor of two of the average about 50 per cent of the time.

The calculated signal-to-noise power ratio using:

P_T — 400 watts
K = 3 db./100 ft.
l = 100 ft.
λ = 6 meters
G_R = 64
G_T = 64
F = 2
B = 20

gives $\left(\frac{S}{N}\right)$ = 1:1.

A possible explanation for lack of consistent (although weak) echoes may be that the image antenna produced by ground reflection (assumed perfect for sake of argument) is causing cancellation and reinforcement[3] of the circularly polarized radiation in such a manner that alternate zones of radiation are produced where the polarization changes from being completely vertical to being completely horizontal. When the moon is in a zone where the radiation is predominantly linearly polarized, Faraday rotation may cause loss of signal, when transmitting with circular polarization.

A comparative test between VE3BZS/2 and another local station, with distant stations using horizontally polarized antennas, gave signals several db. lower than expected. Probably this was due, in part at least, to ground reflection producing predominantly vertically polarized radiation at low angles, when using circular polarization at VE3BZS/2.

The results of these 50-Mc. tests seem to show that 50 Mc. is not too practical under present conditions for amateur moonbounce work. Also circularly polarized antennas may suffer a loss in efficiency under conditions of good ground reflection in combating Faraday-rotation effects.

The narrow-band methods used in the receiver and transmitter should be adaptable for use on higher frequency amateur bands to allow use of existing equipment with little modification and cost. — *Alan Goodacre, VE3BZS*

[1] *The VHF Amateur*, February, 1961, pp. 13–16.

[2] See photos in *QST*, November, 1961, p. 89.

[3] *The A.R.R.L. Antenna Book*, pp. 46–48.

Reprinted from May, 1962 QST

MOONBOUNCE OPERATING AIDS

DIVISION OF VARIAN
301 Industrial Way
San Carlos, California

MOONBOUNCE OPERATING AIDS

BY: Robert I. Sutherland
W6PO

Four "moonbounce" operating aids are included in this collection of EME notes. They are: 1)- An EME Path Loss Nomogram, 2)- An EME Operating Chart, 3)- Universal EME Window Chart for USA and Europe, and 4)- Reprint of a QST article on Noise Behind the Moon.

1- An EME Path Loss Nomogram (figure 1)

The nomogram was calculated by Joe Reisert, W6FZJ. It shows the change in path loss for the six "moonbounce" bands as the moon travels from perigee to apogee. "Sky and Telescope" magazine gives the distance to the moon and the dates for perigee and apogee. The semi-diameter (SD) of the moon is given in the "Nautical Almanac" on the right-hand pages at the bottom left.

The semi-diameter is half the moon diameter and is given in minutes of arc. The larger the SD number, the closer the moon is to the earth and therefore the lower the path loss. Scheduling for an EME contact at apogee could end in failure for systems which are marginal, as the 2 decibel difference in signal strength at apogee could "use up" all the signal-to-noise margin of the circuit.

2- EME Operating Chart (figure 2)

This chart has proven useful in keeping track of what has been sent and copied during an EME schedule. The operator can quickly determine whether he should be transmitting or listening at a given time within the schedule. After a series of schedules early in the morning, the operator usually needs help in keeping things straight! The chart will do the job.

The chart shown is made up for a California station scheduling Europe, Australia or the eastern USA. East coast operators will need two charts, one for working west and the other for working Europe.

3- Universal EME Window Chart for USA and Europe (figure 3)

Many experimenters have suggested that a Universal EME Window Chart be established for each of the bands used for moonbounce contacts, the argument being that a fixed array could be erected, pointing towards the window. Such an array could be built close to the ground, eliminating high towers and complicated aiming systems. In addition, it would not be hindered by trees and buildings if properly placed, would be easy to maintain and (hopefully) would not offend the neighbors.

Stations using a fixed array and the Universal EME Window could work each other via the Window, as well as other stations using steerable antennas.

In all probability, the well-equipped tropo station could not use the Window as he cannot raise his array to the proper angle of elevation. But these stations could, as they are now doing, work the steerable EME stations and others near the same Longitude who are using the rising or setting moon. The ultimate answer, of course, is for all experimenters to have antennas steerable in both azimuth and elevation to some degree.

While it is impossible to have a Universal EME Window to satisfy all continents, I propose a Universal Window for 144 MHz work between USA and Europe be chosen. Such a Window is illustrated in the graph, calculated for a setting moon in Germany. Various locations in the USA could point to this Window using the typical azimuth and elevation data in the table of figure 4 to aim a fixed array.

The table indicates that if an array with a sharp pattern is used, some steering will be required to hit the entire Window. If the array is not steerable, then the last one-half hour of the Window should be favored in order to work Europe. The table suggests the antenna should have a broad azimuth pattern and a sharp vertical pattern because of the small change in elevation required during a one hour schedule.

4- Noise Behind the Moon

The following reprint of the article "Sky Temperature Behind the Moon" (QST, October, 1964) is of interest to the serious moonbouncer. There are times when the moon is in front of a noisy part of the sky and during these periods the echos returning to earth from the moon can be completely masked by the noise. This would prevent a contact from being made, without an apparent reason for the failure.

The magazine "Sky and Telescope" has a map showing the moon's path across the sky for each month. By knowing the location of the strong celestial radio frequency sources, an operator can avoid scheduling on days when chances of success are low. Generally speaking, if the moonbounce schedule is held to periods when the moon has a north declination, the chances of success are enhanced, as the moon is in a "quiet" area of the sky. Southern declinations have a high probability of being noisy.

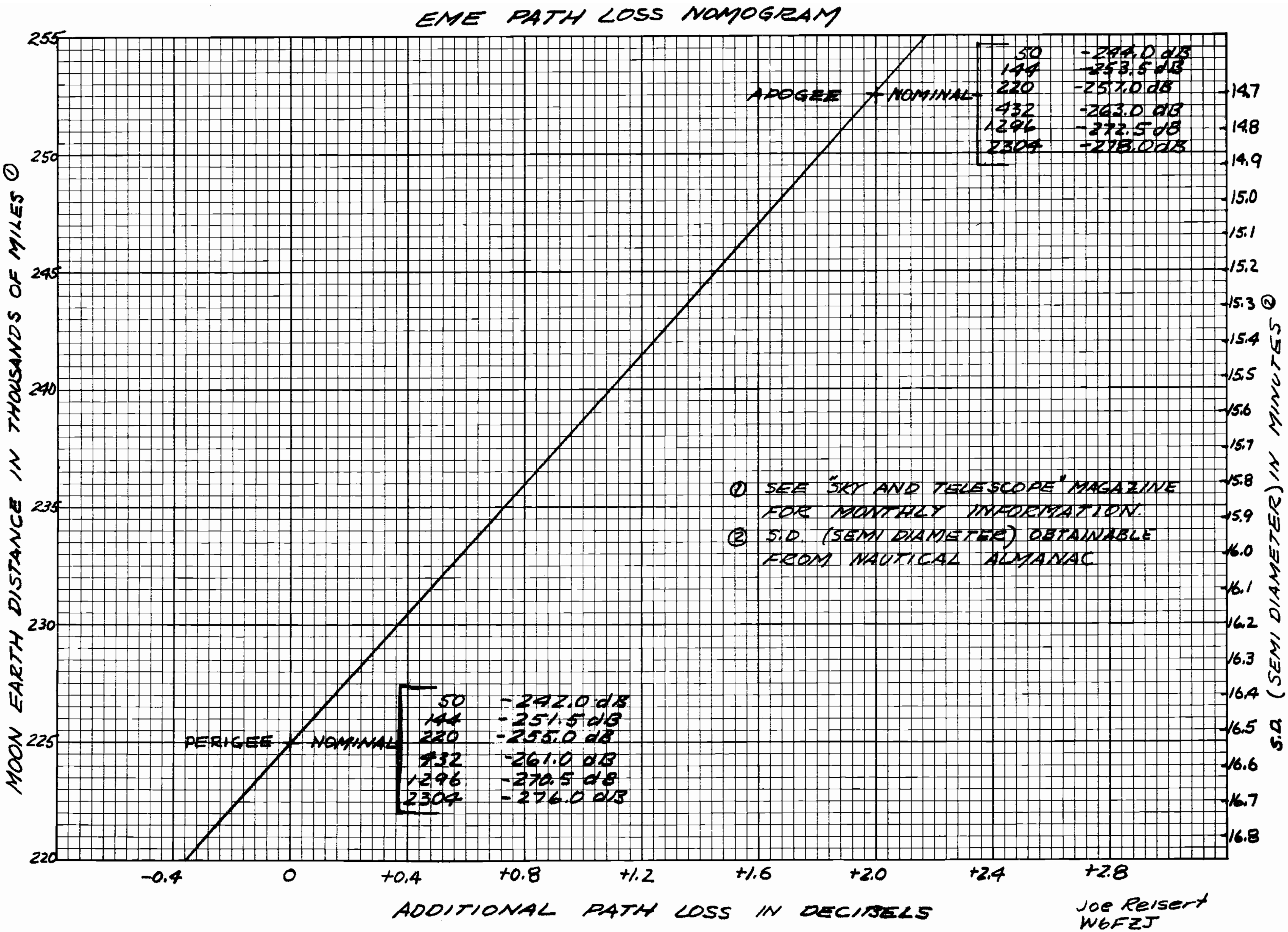
EME PATH LOSS NOMOGRAM
APOGEE
NOMINAL
50 -244.0 dB
144 -253.5 dB
220 -257.0 dB
432 -263.0 dB
1296 -272.5 dB
2304 -278.0 dB
PERIGEE
NOMINAL
50 -242.0 dB
144 -251.5 dB
220 -255.0 dB
432 -261.0 dB
1296 -270.5 dB
2304 -276.0 dB
① SEE "SKY AND TELESCOPE" MAGAZINE FOR MONTHLY INFORMATION.
② S.D. (SEMI DIAMETER) OBTAINABLE FROM NAUTICAL ALMANAC
MOON EARTH DISTANCE IN THOUSANDS OF MILES ①
255
250
245
240
235
230
225
220
S.D. (SEMI DIAMETER) IN MINUTES ②
14.7
14.8
14.9
15.0
15.1
15.2
15.3
15.4
15.5
15.6
15.7
15.8
15.9
16.0
16.1
16.2
16.3
16.4
16.5
16.6
16.7
16.8
-0.4
0
+0.4
+0.8
+1.2
+1.6
+2.0
+2.4
+2.8
ADDITIONAL PATH LOSS IN DECIBELS
Joe Reisert
W6FZJ

FIG. 1

DATE ____________ STATION ____________

TIME ________ GMT ________ NOISE ____________

0 -- 2 RECV	30 -- 32 XMIT
2 -- 4 XMIT	32 -- 34 RECV
4 -- 6 RECV	34 -- 36 XMIT
6 -- 8 XMIT	36 -- 38 RECV
8 -- 10 RECV	38 -- 40 XMIT
10 -- 12 XMIT	40 -- 42 RECV
12 -- 14 RECV	42 -- 44 XMIT
14 -- 16 XMIT	44 -- 46 RECV
16 -- 18 RECV	46 -- 48 XMIT
18 -- 20 XMIT	48 -- 50 RECV
20 -- 22 RECV	50 -- 52 XMIT
22 -- 24 XMIT	52 -- 54 RECV
24 -- 26 RECV	54 -- 56 XMIT
26 -- 28 XMIT	56 -- 58 RECV
28 -- 30 RECV	58 -- 60 XMIT

FIG. 2

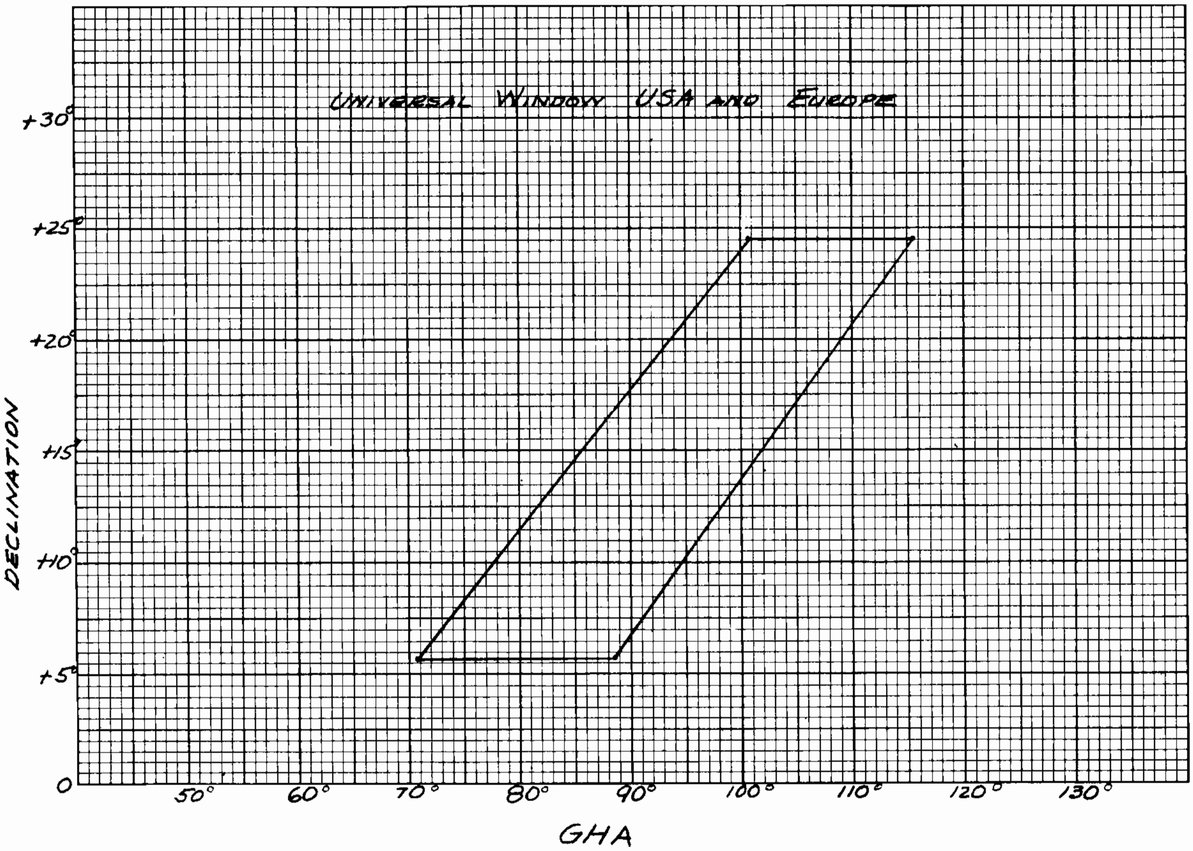

FIG. 3

TYPICAL ELEVATION AND AZIMUTH DATA
TO HIT PROPOSED UNIVERSAL WINDOW
AT VARIOUS GEOGRAPHIC LOCATIONS

Date	GMT	Seattle, Washington		Nashua, New Hampshire		San Mateo, California		Minneapolis, Minnesota		Frankfurt, Germany	
		El (°)	Az (°)	El (°)	Az (°)	El (°)	Az (°)	El (°)	Az (°)	El (°)	Az (°)
7/17/73	0700-0812	28-37	117-131	52-51	179-202	32-43	111-124	45-50	147-167	11-0	265-276
7/18/73	0830-0930	39-46	123-145	57-50	191-226	44-56	115-136	54-55	159-191	8-0	277-286
7/19/73	0945-1045	47-54	126-145	60-52	212-234	52-62	116-134	60-60	168-197	8-0	284-294
7/21/73	1220-1332	60-65	137-164	58-46	240-259	66-75	122-167	67-59	198-232	7-0	297-305
7/22/73	1330-1432	61-66	136-164	59-47	242-260	67-76	119-166	68-61	199-234	7-0	300-308
7/25/73	1550-1650	52-57	130-147	59-53	224-240	58-65	115-135	63-61	172-202	7-0	288-298
7/26/73	1615-1715	43-50	120-141	59-48	203-235	47-61	114-143	57-56	162-201	9-0	280-290
7/27/73	1635-1735	35-41	122-133	55-50	191-214	40-49	115-130	50-53	155-179	9-0	270-282

FIG. 4

Sky Temperature Behind the Moon

BY C. R. SOMERLOCK,* W3WCP

It is common knowledge that if an antenna is pointed at the sky, it will receive radio noise. This noise is generated chiefly by our galaxy, the milky way. The intensity of the emissions will, of course, depend on the particular part of the sky at which the antenna is pointed: the "hottest" direction being that of the galactic center ($\alpha = 17^h 43^m$, $\delta = -28.8°$.).

If, now, one desires to communicate with some object that is silhouetted against the sky, and whose position in the sky changes, it is important to know the background sky temperature in the vicinity of that object. (The "temperature" is a measure of the radiated noise power.) In a situation such as an amateur v.h.f. moonbounce attempt, this information can make the difference between success and failure.

Method

To obtain the data, the position of the moon was determined (in equatorial coordinates) from *The American Ephemeris and Nautical Almanac* for 12 noon each day of the month of December 1965. These positions were then converted into galactic coordinates with the aid of appropriate coordinate conversion table.[1] Knowing the position of the moon in galactic coordinates allowed the background sky "temperature" to be read from a radio map of the galaxy. The particular one used was plotted by Baldwin[2] at 81 Mc. using a beamwidth of 2 by 15 degrees.

Since the moon's precession rate is small, the gross features of the data are valid for a year or so. Because the data vary in a periodic manner, similar plots can be made for any desired month in this valid period simply by shifting the abscissa of the curve by a multiple of one lunar orbital period (about 27¼ days). The values can also be extrapolated to other frequencies by the approximate formula:

$$T_x = T_o \cdot \left(\frac{f_o}{f_x}\right)^{2.7}$$

While the details of the radio sky may vary with frequency somewhat, the main features will not change.

Results

The end result of all this work is the curve shown in Fig. 1, plotting background sky temperature behind the moon for 144 Mc. Note that for about 3 days out of the month the moon is passing directly across the center of the Milky Way, resulting in very high noise temperatures in the antenna beam. As an example of what effect this can have on 2-meter moonbounce communications, consider the following case:

Bandwidth = 1 kc.
Preamplifier noise figure = 2 db.
Sky temperature = 400° K.
Received signal = −140 dbm.

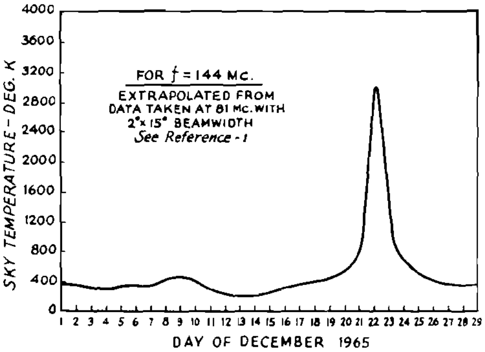

Fig. 1—Background sky temperature behind the moon, for December, 1965, for the 144-Mc. band.

The noise figure can be converted to an equivalent noise "temperature" by the relation:

$$T = 290\ (N\text{-}1) \text{ degrees Kelvin}$$

where N is the noise *factor*. The sky radiation can then be added directly to this preamp noise temperature, since they have the same effect and are in the same units. The total noise can then be converted back to more familiar units of power by:

$$\text{Noise power} = KTB$$

where: $K = 1.6 \times 10^{-23}$ = Boltzman's Constant
T = sky temp. + pre-amp temp.
B = system bandwidth

Under these conditions, the noise power is computed as being −142 dbm, and the received signal-to-noise ratio will be:

$$S/N = +2 \text{ db.}$$

Table I

Frequency Conversion Factors

For a Freq. of:	Divide Values of Curve by:
50 Mc.	0.058
144	1.00
220	3.15
430	19.3
1296	380

Now consider what happens if the same system is used when the moon is in front of the galactic center. In this case:

Sky temperature = 3000° K,
so Noise power = KTB = −133 dbm.

The signal, however, is still only −140 dbm, so the received signal-to-noise ratio under these conditions has dropped to −7 db. This kind of variation can easily make the difference between success and failure.

Values for other frequencies can be determined by dividing readings from Fig. 1 by the appropriate factor from Table I, showing the relative noise strengths for various amateur bands. Clearly the problem lessens with increased frequency, and becomes negligible at 1296Mc.

Table II

Dates of Noise Maxima for a Moon-Tracking Antenna.

1964	1965
July 27	Jan. 1
August 23	Jan. 28
September 20	Feb. 25
October 17	March 24
November 13	April 21
December 10	May 18
	June 14
	July 11
	August 8
	September 4
	October 1
	October 29
	November 25
	December 22

Table II gives a list of approximate dates of future noise maxima which should be avoided for v.h.f. moonbounce attempts, unless the system is designed to handle these conditions. QST

* NASA, Goddard Space Flight Center, Greenbelt, Maryland 20771.

[1] Annals of the Lund Observatory, No. 15, 16, 17

[2] Baldwin, J. E. — "A survey of the Integrated Radio Emission at a Wavelength of 3.7M.", Monthly Notices of the Royal Astronomical Society *115*, Pages 684–689.

MORE ON THE MOONBOUNCE UNIVERSAL WINDOW FOR 144 MHz

DIVISION OF VARIAN
301 Industrial Way
San Carlos, California

AS-49-3

4142

PROPOSAL FOR A UNIVERSAL EME MOONBOUNCE WINDOW

By: Ken Tentarelli, W1FZA

In much the same way that "calling frequencies" are useful in promoting activity on some VHF propagation modes, a "universal window" could be of great value in promoting EME (earth-moon-earth) activity. The phrase "universal window" is used here to mean a selected area of the sky which the community of EME enthusiasts agree to use as a primary target for aiming their antenna arrays. First let us explore what could result from such an agreement by EME operators and then let us consider which particular area of the sky should be selected.

For EME to be enjoyed by the greatest number of interested operators, those stations with the facilities for erecting vast, high-gain, arrays should be encouraged to build the largest possible antennas. These stations could then furnish EME contacts to stations with much smaller arrays. Unfortunately, at the present time anyone who builds a large fixed, or partially steerable, EME array might find that his window is not shared by many other EME stations. With such a limited prospect of success, only the most courageous among us have invested their time and money in massive EME antenna construction projects. On the other hand, the existence of a universal window would give confidence to stations building large fixed arrays that their efforts would be rewarded by access to all other EME stations. Another advantage of a universal window is the simplification of moon tracking problems; in fact, any practical antenna could be oriented in a fixed position within the window and yield several moon-days per month.

Now let us consider which point in the sky should be chosen for the universal window. The window which is now used for European EME schedules (see AS-49-2) has been suggested by Bob Sutherland, W6PO, as a window which offers several advantages over other choices:

1. It is now being used for regular schedules by several stations which implies that anyone who constructs an antenna aimed at this window can expect some measure of success almost immediately.

2. The window corresponds to a meridian crossing near the middle of the North American continent. This means that all stations in the continental United States would have high antenna elevations when aiming at the window. High elevation angles ease the problem caused by ground-level obstructions (trees, hills, etc.), reduce interference from local stations, minimize TVI, and enable antennas to be mounted near ground level to keep feedline losses low.

3. The window includes the region of maximum northerly lunar declination. Since the moon changes declination in sinusoidal fashion it remains near its maximum declination for several days; thus this window optimizes the number of moon-days per month available to fixed position antennas.

4. The window is designed such that even very high gain antennas need only azimuth steering to further extend the number of available moon-days.

5. Stations on all continents except Asia and Australia can access this window (no single window is accessible from all continents).

Obviously, no window could be truly universal and meet everyone's needs. Therefore, it is proposed only that this window be a primary target; that is, every EME station should insure that his array can be aimed at this window in addition to any other window he might desire to use. If the concept of a universal window is accepted by a majority of EME stations it should not be too long before those fortunate stations with massive arrays have the WAS award within their grasp, while VHF city dwellers with small antennas will be able to enjoy the thrill of communication via the moon.

Explanation of Figures

Figure 1 is a diagram of the proposed universal window in declination and Greenwich hour angle coordinates. The moon is within this window approximately one hour per day for about five days per month. Also shown in this figure are AZ-EL coordinates which bound the window for three geographic locations across the United States.

Figure 2 is diagramed similar to Figure 1. Shown in this figure are the 3 dB beamwidths of two antennas which are fixed in position and aimed at the midpoint of the window. One antenna is a 320-element collinear, and the other is a single 24-element Yagi.

Figure 3 provides AZ-EL coordinates of the midpoint of the window as a function of station location. This data may be interpolated with good accuracy.

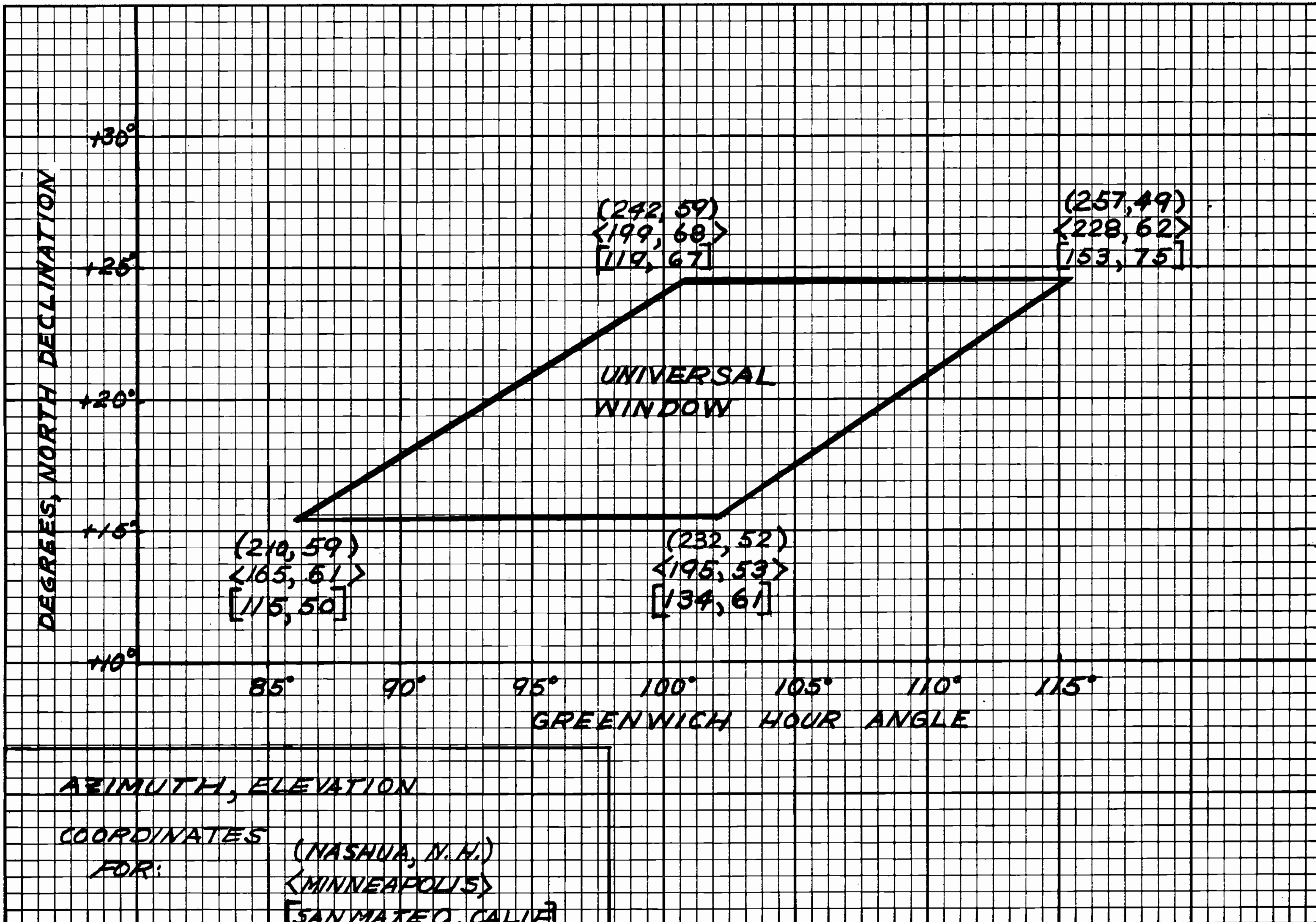

FIGURE 1

W1FZA

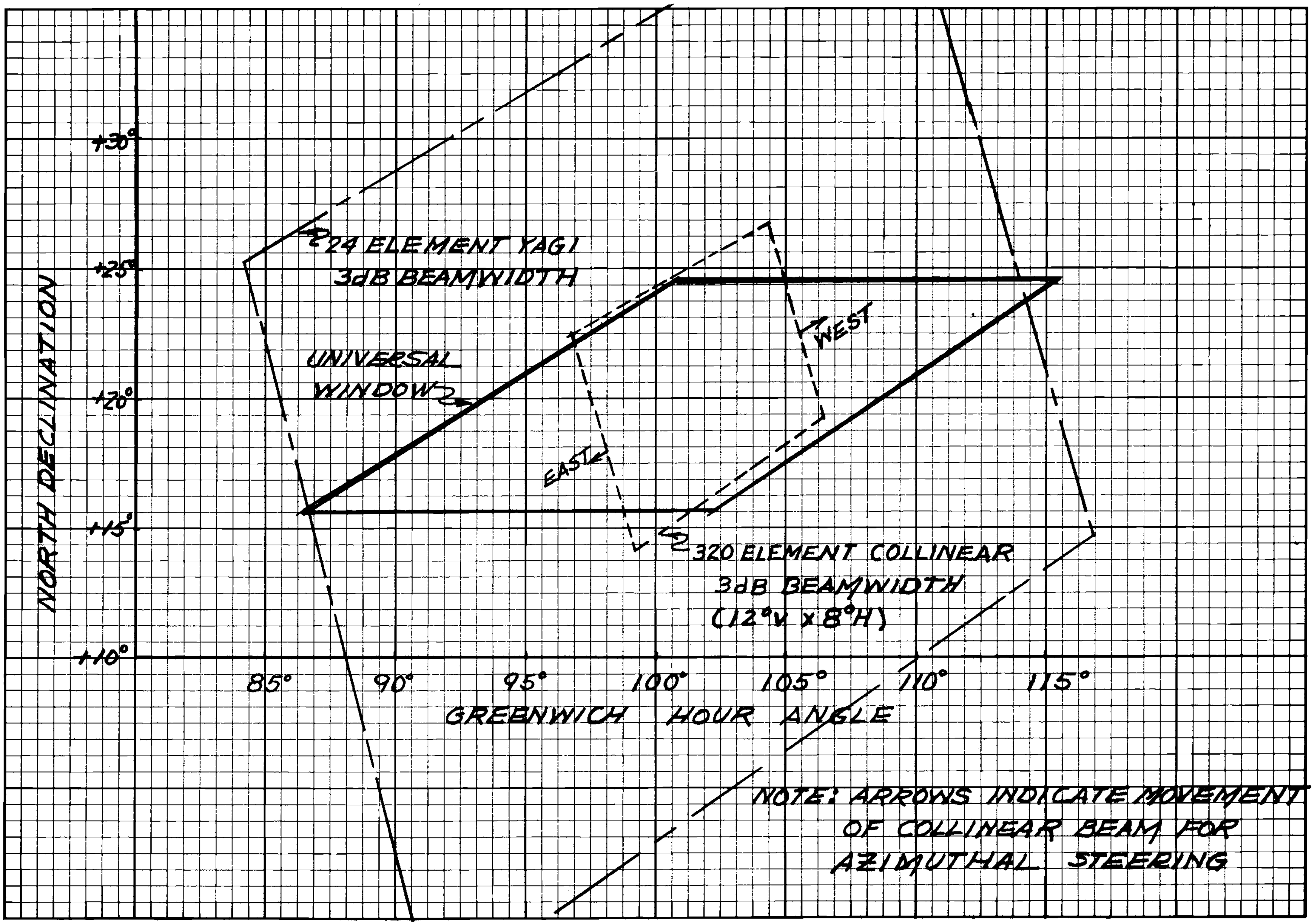

FIGURE 2

W1FZA

ELEVATION-AZIMUTH COORDINATES VS LOCATION
FOR MIDPOINT OF UNIVERSAL WINDOW
(GHA = 102°, DEC = 20°N)

NORTH LATITUDE / WEST LONGITUDE	25°		30°		35°		40°		45°	
	El	Az	El	Az	El	Az	El	Az	El	Az
70°	60.1	266.9	59.4	258.4	58.1	250.4	56.1	243.3	53.6	237.1
75°	64.6	264.1	63.7	254.0	61.9	244.9	59.5	237.1	56.5	230.6
80°	69.1	260.6	67.8	248.4	65.5	238.0	62.5	229.7	59.1	223.2
85°	73.5	255.7	71.7	240.9	68.8	229.4	65.3	221.0	61.3	214.9
90°	77.8	248.1	75.2	230.1	71.6	218.3	67.5	210.7	63.1	205.6
95°	81.8	233.7	78.2	213.9	73.8	204.2	69.1	198.7	64.3	195.3
100°	84.7	200.7	79.8	190.7	74.9	187.2	69.9	185.5	64.9	184.4
105°	84.3	150.4	79.6	164.1	74.8	169.2	69.8	171.8	64.9	173.3
110°	81.1	122.5	77.7	142.3	73.4	152.7	68.9	158.7	64.1	162.6
115°	77.0	110.0	74.6	127.4	71.1	139.2	67.1	147.1	62.8	152.5
120°	72.7	103.1	70.9	117.4	68.2	128.7	64.7	137.1	60.9	143.3
125°	68.2	98.6	66.9	110.4	64.8	120.5	61.9	128.7	58.6	135.2

FIG. 3

SUCCESSFUL 144 MHz

EME ANTENNAS

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AS-49-5

4144

SUCCESSFUL 144 MHż EME ANTENNAS

One of the most important decisions that an amateur planning a moon-bounce system must make is the choice of antenna. There are many variables that enter into the decision as there are both practical and technical topics to consider. In the practical category are:

1. What will the system cost?
2. What is the availability of the materials?
3. What is the available real estate?
4. What are the esthetics as judged by the neighbors and the wife?
5. What are the electrical and mechanical abilities of the amateur and his helpers?

In the technical category some of the considerations are listed below. Quite often the technical desires must give-in to some of the previously listed practical considerations.

1. What system gain should be sought after?
2. Should the array be fully steerable, partially steerable, or fixed?
3. Should the array be on a high tower, or on the ground?
4. Is the array for EME only, or is it to be used for other propagation modes?
5. If the array is to be steerable, should an AZ-EL mount or a Polar mount be used?
6. What kind of feedline should be used?
7. Should the preamplifier be mounted at the antenna?
8. What kind of transmission line should be used in the phasing lines?
9. Should power dividers be used?
10. What type of antenna should be used? (Probably the hardest decision is the choice of antenna type.)

Amateurs tend to become quite emotional when discussing the relative merits of collinears, yagis, log-periodic yagis, rhombics, and dishes. All of the antennas mentioned have been used successfully in EME systems. The relative merits of each antenna probably change as the discussion moves from band to band. For example, the dish is quite acceptable on 432 MHz, but is too big to be practical on 144 MHz.

Included in this EME Note are pictures of successful 144 MHz antennas. Pictures of antennas for other bands are not available at this time but as pictures are collected they will appear in subsequent issues of the EME Notes.

When deciding whether to use a collinear, a yagi or a log-periodic yagi, the capability of the amateur must be considered. The lower the "Q" of the antenna, the better the chances of operational success. It is difficult to fail with a low-Q collinear array. The log-periodic yagi is a band-pass type of antenna of medium-Q and is therefore easy to assemble into an array. Short yagi antennas are a little more critical. The long and very long high-Q yagis are even more critical to assemble into an array. Also, their performance is optimum only over a narrow part of the two meter band.

It is very important to be certain that your chosen yagi design has been carefully checked out before building eight or sixteen identical antennas. If this is not done, a lot of time and money can be spent on an array with inadequate gain for EME work. With proper antenna-to-antenna spacing, the yagi and log-periodic yagi tend to have a cleaner pattern than other types. The cleaner the pattern, the better the antenna will perform for receiving.

The main idea to keep in mind when thinking about the antenna array to put up is the minimum system requirement as pointed out in earlier issues of these Notes. For two stations each running 500 watts output with a system noise figure below 2 dB, the minimum antenna gain each station should strive for is 20 dB. Actually what is required is a "round trip" gain of 40 dB. That is to say, if one station has 17 dB antenna gain and the other station has 23 dB gain, their total gain is 40 dB round trip. Therefore, if the two stations schedule each other often enough, they will be successful in completing a contact. Two stations with 17 dB antenna gain would probably not be able to make contact very often, if at all. If the constraints on your antenna project relegate you to something under 20 dB, you still can work those stations having an extra amount of antenna gain. There are several 144 MHz stations with antenna gain in the 23 to 24 dB range. There are even some having close to 30 dB gain.

Remember that the previous discussion on system requirements is for 144 MHz. As EME activity and interest increase, there will be bigger and better arrays built.

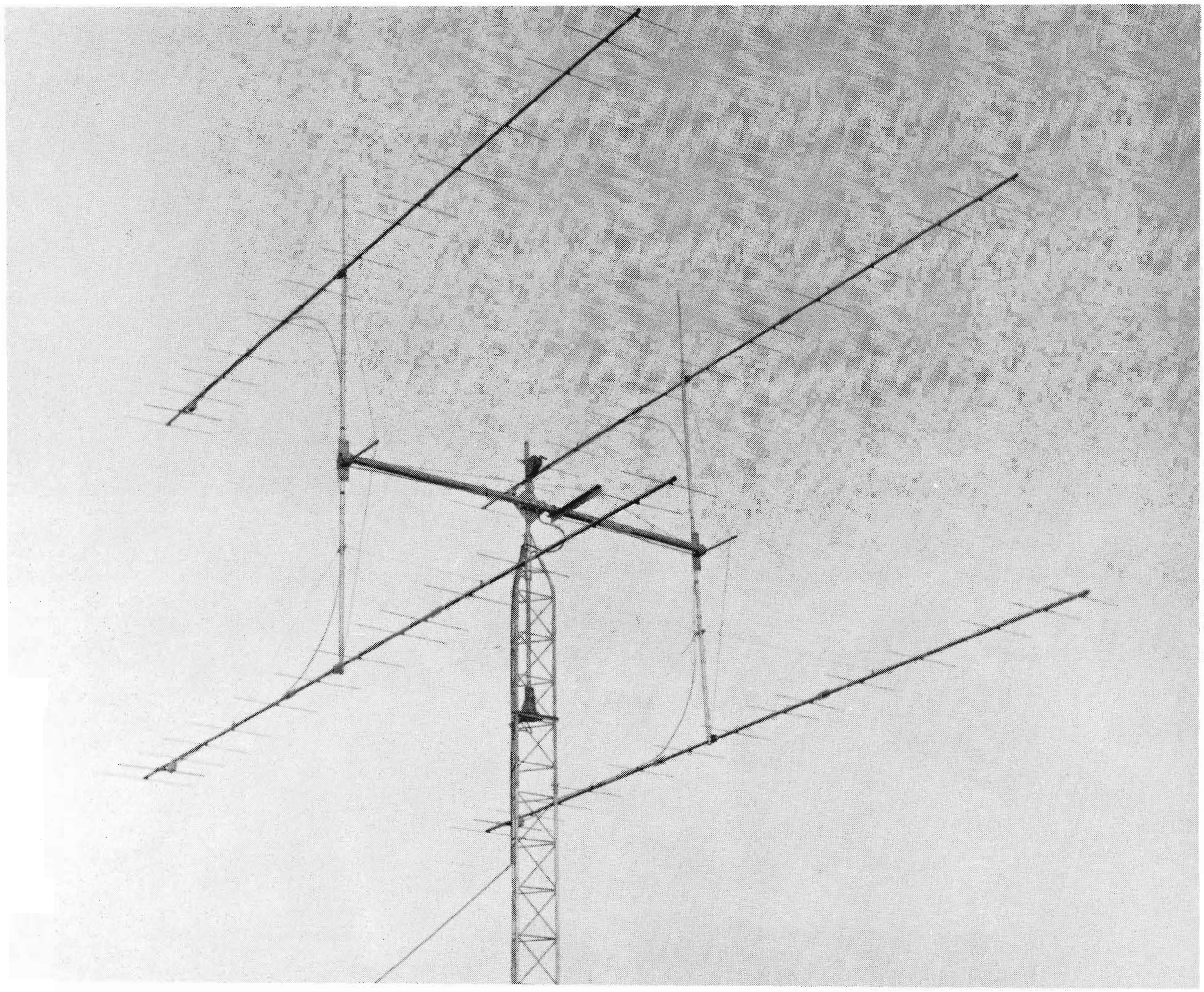


Figure 1

The array of John Allen, WAØCHK consisting of four Hy-Gain 15 element Yagis. The antennas have been modified as per the instruction from K6MYC. As the antennas come all of the parasitic elements appear to be one inch too long. The "H" frame is made from a 14 foot length of three inch diameter 6061-T6 aluminum pipe with an eighth inch wall thickness. The vertical members are made from two pieces of 1.59 inch diameter 6061-T6 aluminum pipe 13' long with a .140 inch wall thickness. The gusset plates are quarter inch 6061-T6 aluminum. The clamps are muffler clamps and electrical conduit "U" bolts. Small antenna tuners are mounted 6 1/2 inches away from each feed point and are tuned for 50 ohms. The phasing lines are 75 ohm .412 O.D. surplus CATV aluminum coax. The antenna is mounted on a 25' Rohn #45 tower. The boat winch mounted above the center of the "H" frame is for lifting the antenna up and down the tower for maintenance. The antenna is elevated manually by loosening the four muffler clamps on the "H" frame center gusset plate and twisting the three inch pipe and antennas to the desired elevation angle.

The rotor shown in the photos is a CDE TR-44. Struts are made from quarter inch polypropolyene rope to keep the antennas aligned when elevated. The gain of this antenna array is probably around 19 to 20 dB.

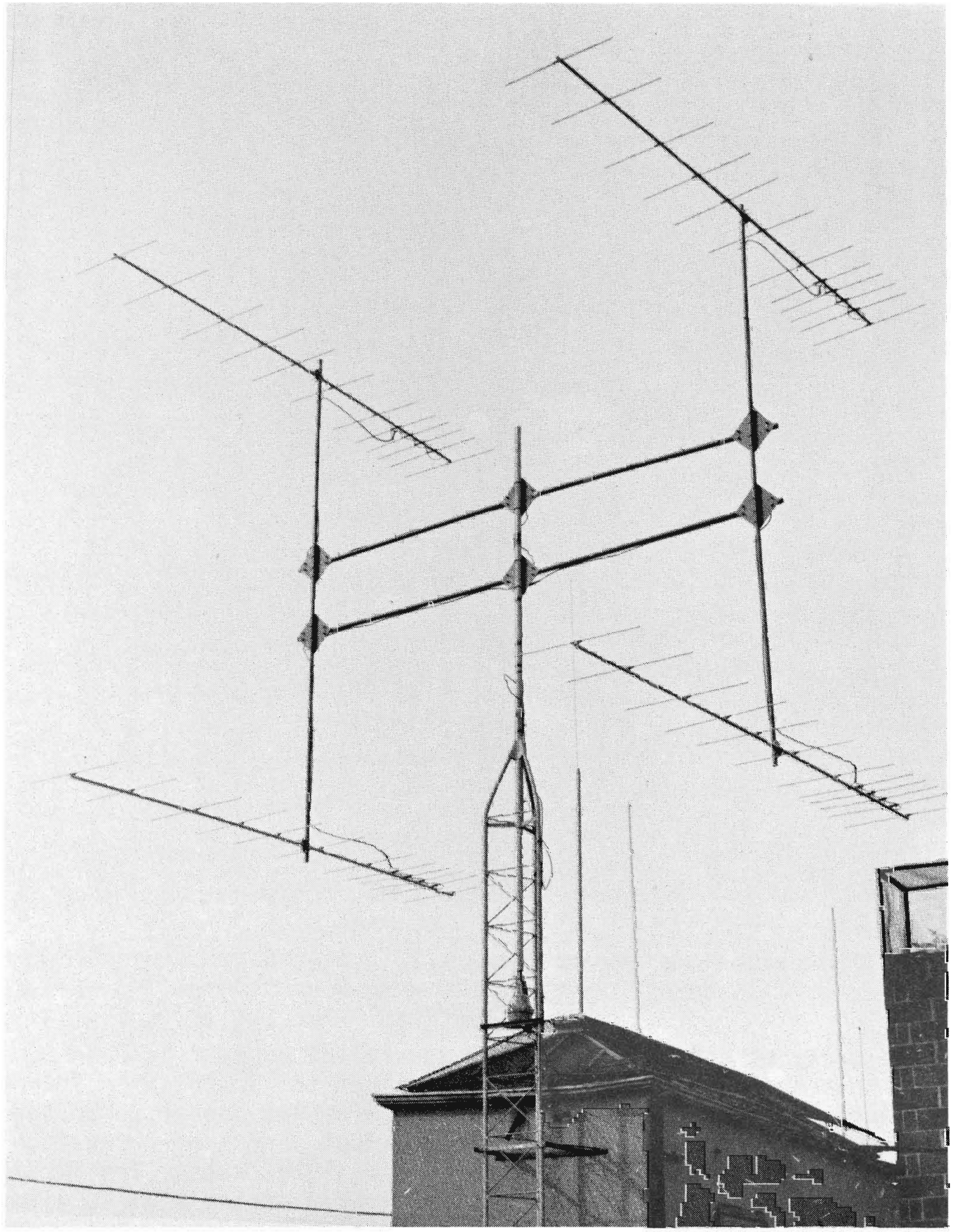


Figure 2

The antenna array of Herb Power, WA2WOM consisting of four KLM log-periodic Yagis. The array is located on top of an apartment house near Prospect Park in Brooklyn. The location is probably not the most desirable for an EME station. The tower is 19' 9" high. The booms are 14' long with an antenna to antenna spacing of 13' 6". Extensive tests between Herb and Carl, W2AZL seem to indicate that the array has a gain of 20.5 dB over a dipole.

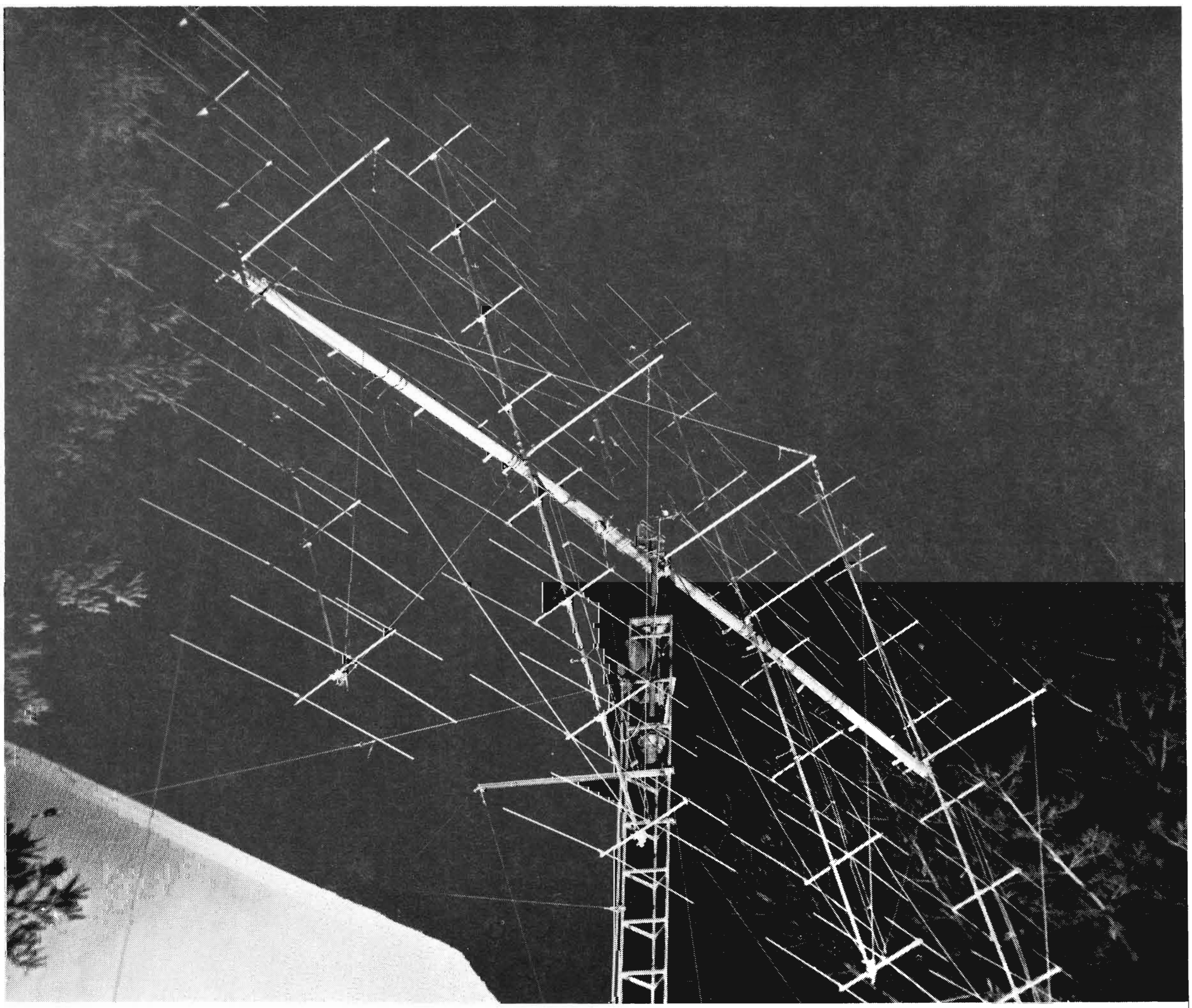


Figure 3

The array of Bob Sutherland, W6PO consisting of eight Cushcraft 20 element collinear antennas. The main boom is 30' of four inch diameter aluminum irrigation pipe. The four secondary booms are each made from two pieces of 10' and one piece of 5' long steel TV mast. The boom guys are all made from quarter inch polypropolyene rope. The phasing lines are made from Belden 8275 tubular twin lead. Each bay of 40 elements is matched with a universal matching stub. All four bays of 40 elements are then fed from another universal matching stub. The antenna array is Az-El mounted with motor drive and selsyn readout. The elevation drive consists of a hinge, a jack screw and a direct current gear motor. The azimuth drive is a converted prop pitch motor. The gain of the array should be between 23 and 24 dB.

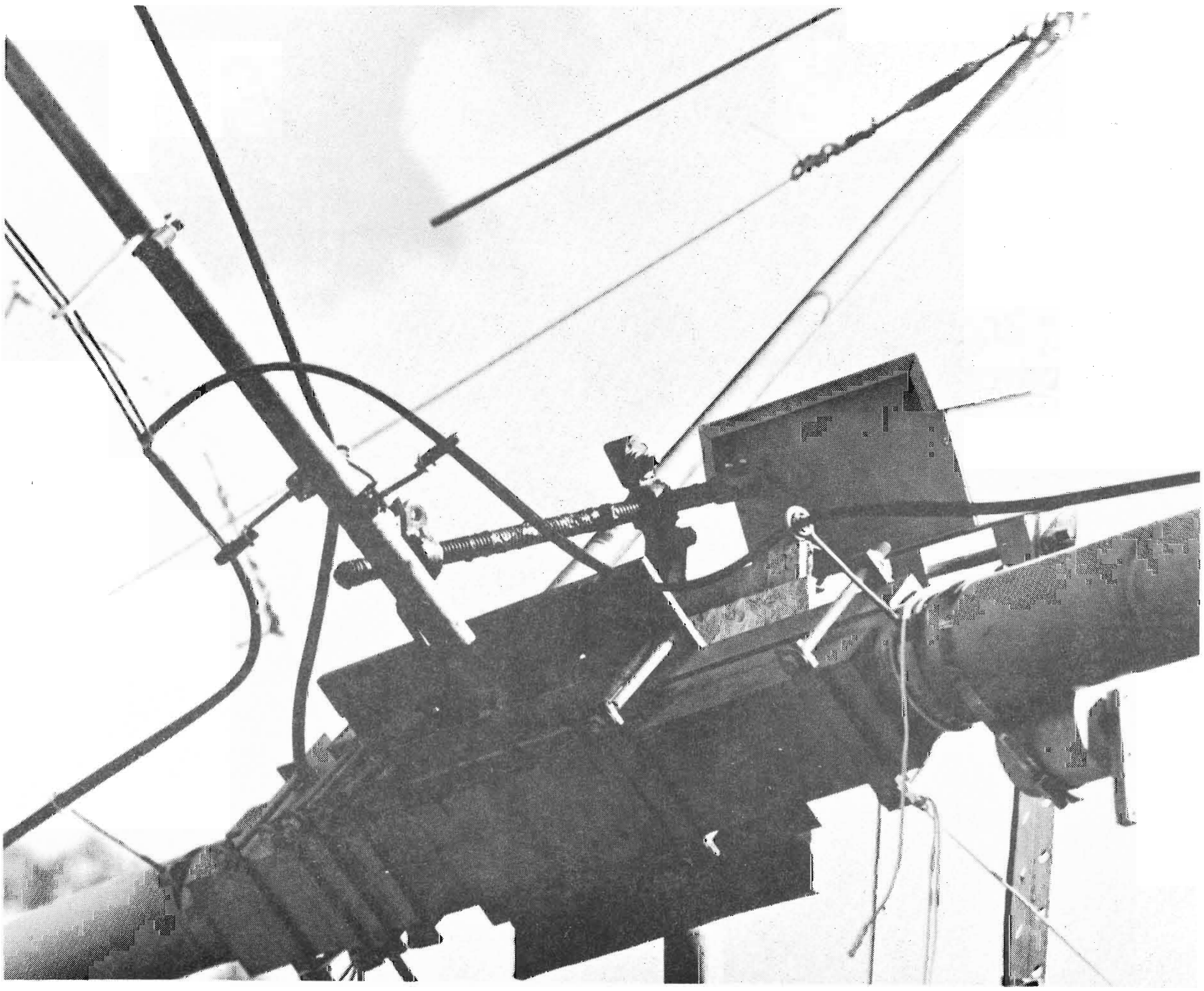
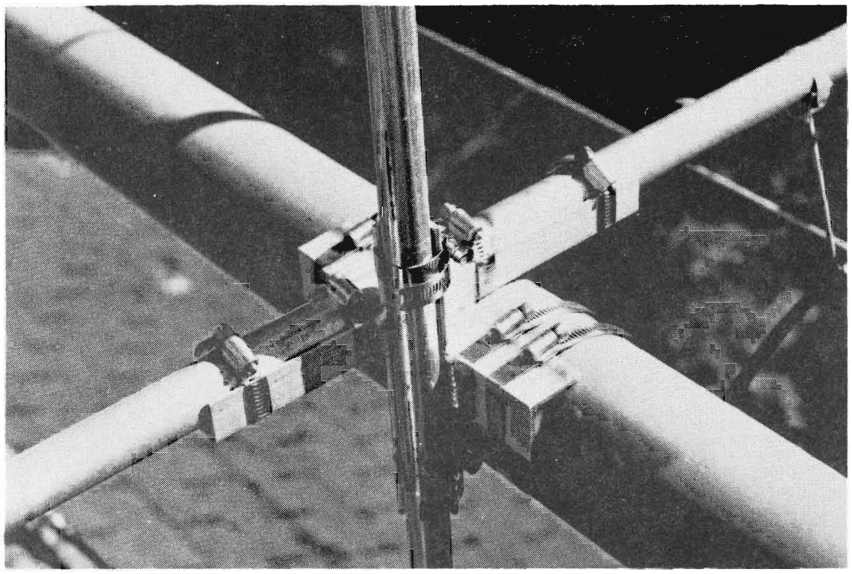


Figure 4

This picture shows the elevation hinge and drive mechanism on the W6PO array. The box at the right end of the jack screw houses the gear motor. Just under the boom to the right is the housing for the elevation selsyn. The housing is held to the boom by means of a stainless steel hose clamp. The half inch square aluminum rod hanging straight down at the end of the selsyn housing, and connected to the shaft of the selsyn, always points to the center of the earth regardless of the elevation angle. Thus the selsyn measures the elevation angle and transmits it to the readout selsyn at the operating position.



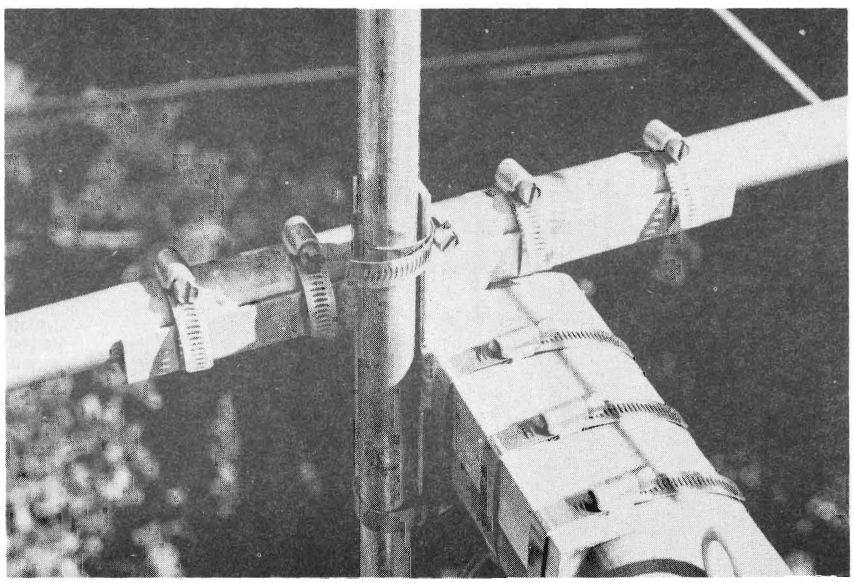


Figure 5

These pictures are of the technique used in the W6PO array for fastening the secondary booms to the main boom and to support the aluminum guy post. The guy post is about four feet long and is used to anchor the polypropolyene rope which supports the secondary booms. Three pieces of aluminum angle are arranged mutually perpendicular to each other and fastened by means of machine screws and nuts. The assembly is then heliarc welded along each common edge. The machine screws can then be taken out if they interfere with the booms, or post, that are positioned in the angle. Stainless steel hose clamps are then used to assemble the array. The result is a strong and light weight assembly. This idea was gleaned from K6MYC's bag of tricks.

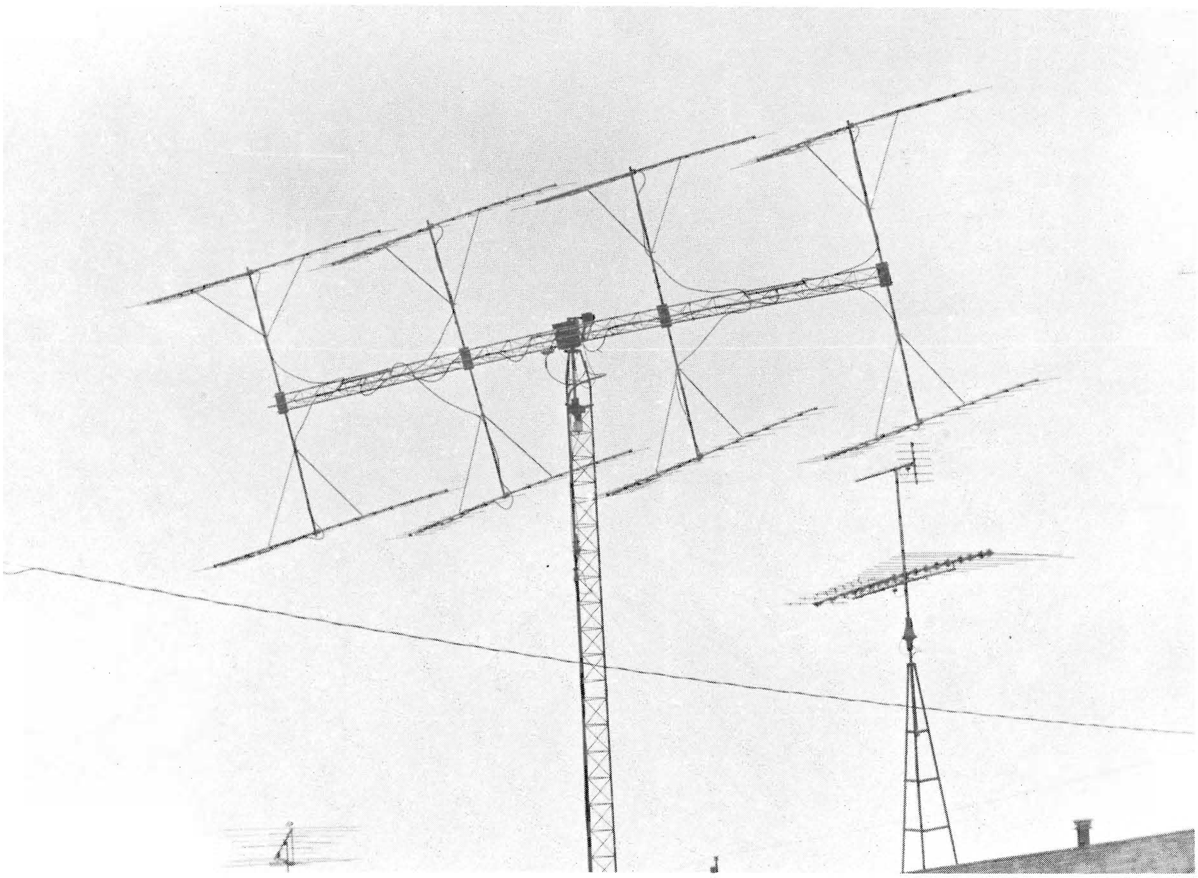


Figure 6

The array of Kelly Scheimberg, W8KPY consisting of eight 16 element KLM log-periodic antennas. The boom is made from ten foot section of 11" Universal aluminum tower. Each section weighs 12 pounds. The vertical spars are 15 feet long and are made from 1 1/2" 6061-T6 aluminum with a .058" wall thickness. The vertical spars are spliced at the 12' point with 1 1/4" tubing. Two four port and one two port power dividers of the WØEYE design are used to feed the antennas. RG-14A/u coaxial cable is used in the phasing lines. The whole array is fed with 7/8" heliax. The antenna is on Az-El mount. The elevation system is a hinge, jack screw and a gear motor drive. The elevation readout system has a selsyn with a plumb mounted on the boom. The readout is at the operating position. The azimuth drive is a HyGain Rotobrake 400 modified for selsyn readout as per the system devised by Vince Vargas, W7JRF/8.

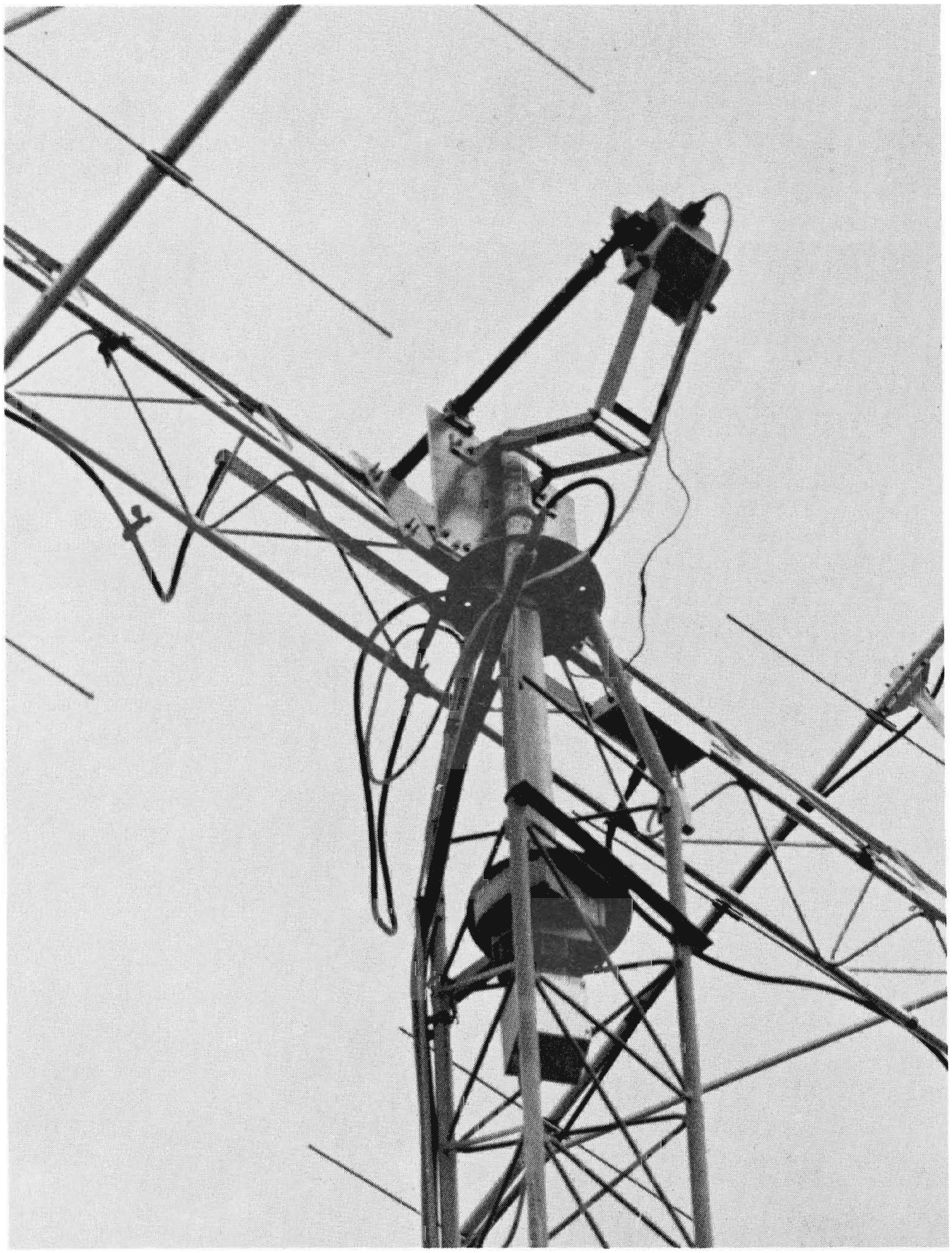


Figure 7

This is the elevation drive mechanism from the W8KPY array. The hinge is driven by a jack screw, or lead screw, from an engine lathe. Lead screws of this type can be obtained from companies that rebuild lathes and milling machines. The gear motor is inside a protective enclosure located at one end of the lead screw. The reduction gear can be seen just outside the enclosure. You will notice that the support for the gear motor is allowed to change positions. This is necessary to allow the hinge to close. The nut traveling along the thread of the lead screw must also be allowed to change attitude to prevent binding. A piece of 1/2" superflex coax cable is used between the 7/8" heliax and the antenna to allow unimpeded rotation and elevation change.

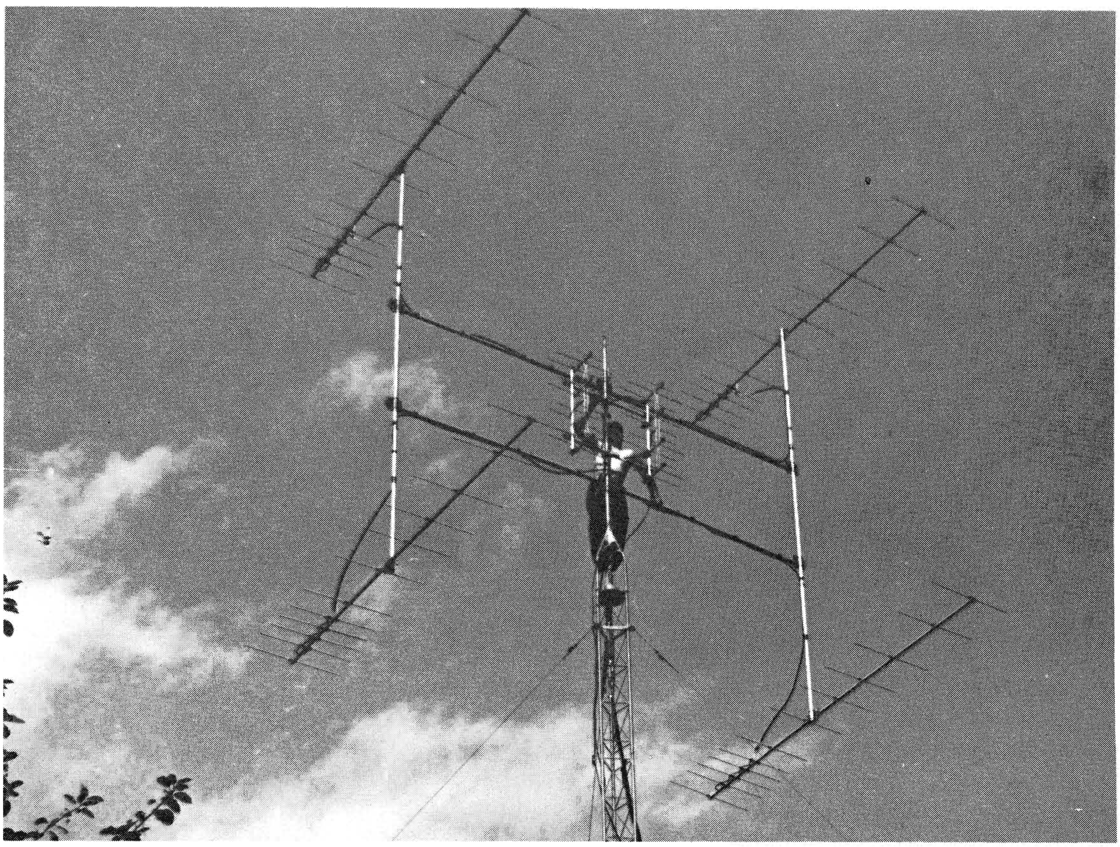


Figure 8

The antenna array of Steve Powlishen, WA1FFO consisting of four 12 element Swan Yagi antennas. The antenna-to-antenna spacing is 16' 8". The center of the antenna is at 55 feet. The antenna is variable in azimuth only. Successful moonbounce contacts have been made using the setting moon. The antenna in the center of the two meter array is a 40 element collinear for 432 MHz.

Figure 9

The array of John Perchalski, K4IXC, is easy to reach from the ground. The array center is nine feet above the ground. The array is made up of eight, seven element Yagis spaced ten feet apart, both vertically and horizontally. The main boom is approximately 30 feet long. The individual Yagis are patterned after the seven element units in the ARRL VHF Handbook. The booms are ten feet long and are made from 1"x1" fir strips. The wooden booms are protected from the elements by several coatings of polyurethane varnish. The parasitic elements are one-eighth inch aluminum wire. The driven element is a folded dipole made from 3/8" tubing in parallel with 1/8" diameter tubing. The phasing lines are cut to multiples of one-half wavelengths of Columbia wire number 05790 Dura-Foam 300 UHF-TV Twin Lead. The mount is made of two pieces of aluminum tubing; one rotating inside the other. The elevation mechanism simplicity can be seen in the picture. Counterweights on the handles balance the array.

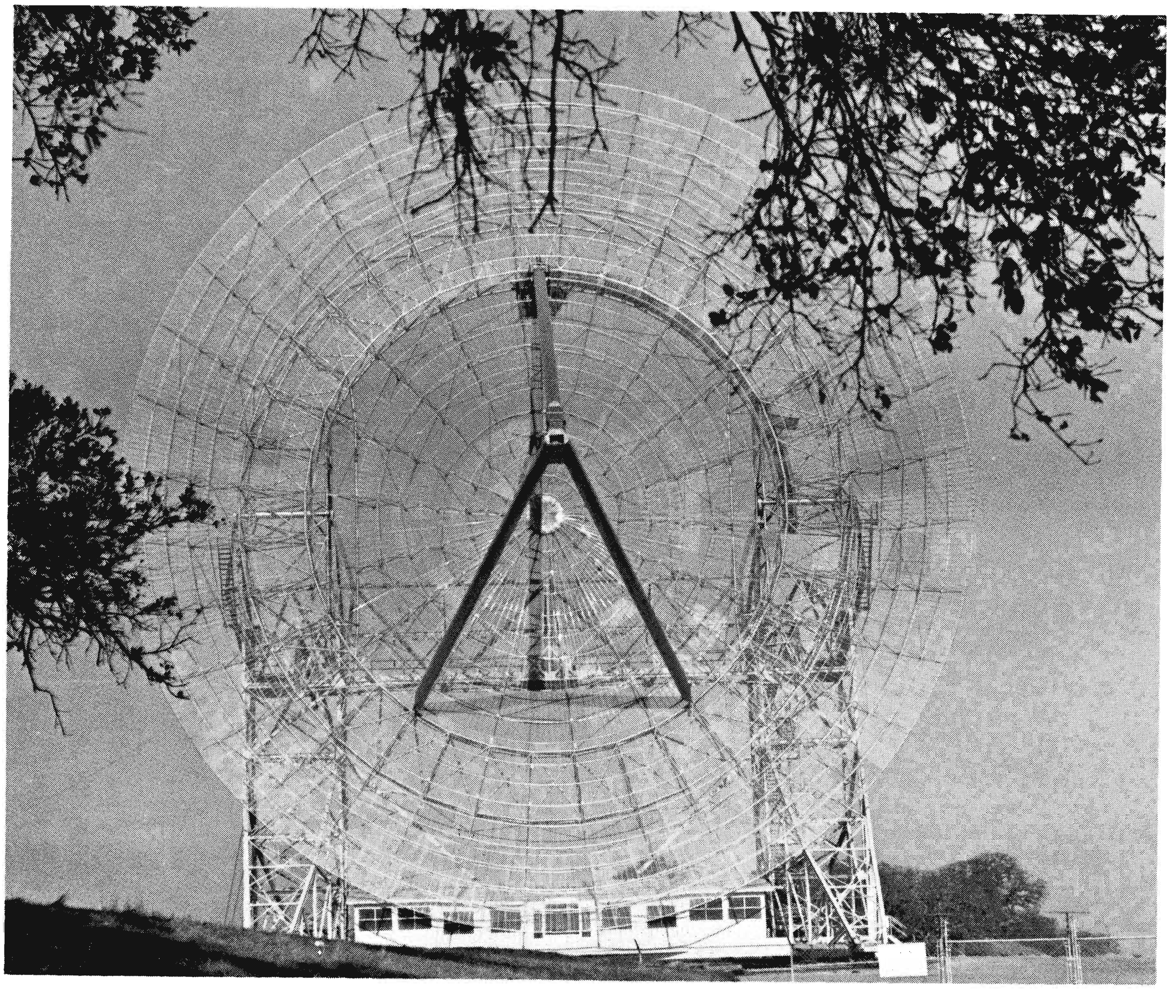


Figure 10

This is a picture of a typical California two meter EME antenna. It was last used by WA6LET to run a series of moonbounce tests from Stanford, California. The antenna is 150 feet in diameter and has a gain of 35 dBI at 144 MHz. The beam width is 2.7 at the 3 dB points. The antenna is fully steerable in azimuth and elevation using two U.S. Navy 5-inch gun mounts. The antenna and the house below all rotate on a track. The aiming of the antenna is accomplished with a PDP-8 computer which up-dates every two minutes. The feed antenna is a NBS Standard 144 MHz antenna and is rotatable from zero to ninety degrees polarity. The feed-line is 1-5/8" coax.

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The idea of DXing via the moon is at once enticing and frightening to many amateurs. Few amateur radio pursuits can match the challenge and the thrill of a completed moonbounce QSO, but the complexity and cost of an e.m.e. system frightens some hams.

Many amateurs think of moonbounce systems in terms of high power, costly receiving preamps, and antennas big enough to fill football fields.

Some e.m.e. systems are that elaborate, to be sure, but K6YNB/KL7 enjoyed considerable success during his 1976 moonbounce DXpedition with a very simple and inexpensive system, including an antenna small enough to be set up by one person in half a day.

The transmitter used a popular commercial transceiver driving a push-pull 4CX250B amplifier. The receiving front end employed low-cost pre-amps similar to the U310 design in this series. (1)

What was unique about the Alaskan moonbounce system was the antenna, a 19 foot by 19 foot array of 16 three-element cubical quads made from wood and #12 copper wire. The antenna mount used nothing more sophisticated than one Ham-M rotator and four small TV antenna rotors, but it was steerable in polarization as well as elevation and azimuth. It could easily be adapted to rotate in all three ways using ropes rather than rotors for steering, reducing the total cost of the antenna system, including phasing cables, well below $100.

CONQUERING FARADAY

The unique feature of the array, of course, is its steerability in polarization. As everyone who has read this far into the Moonbounce series knows, the biggest obstacle to reliable earth-moon-earth communication between properly equipped stations is "Faraday Rotation", with its ability to shift the polarization of signals leaving and re-entering the earth's atmosphere.

Faraday Rotation often renders two well equipped stations unable to hear each other for extended periods of time, or may leave an operator hearing a station that cannot hear him.

On 432 MHz it is not unusual to solve this problem by using a dish antenna and rotating the feed to match the polarization of the incoming signal. On two meters, however, a dish big enough for e.m.e. work is beyond most amateurs' means. Circular polarization will also solve the problem, but the added gain required for e.m.e. work when the signal is omnipolarized again makes amateur-size arrays too small for moonbounce work on two meters.

There are two solutions to the Faraday problem at 144 MHz:

1) to wait until rotation changes, which may mean waiting hours for a completed two-way exchange; or 2) build an array that can be shifted in polarization to match that required for e.m.e. communication at the moment.

The K6YNB Alaskan array takes the latter approach, and does it simply. Obviously, polarization rotation is only possible when the individual antennas being rotated will clear the tower. Of course, even very long yagis can be supported from the rear so they may be rotated in polarization without hitting the supporting frame, but that presents another tough set of mechanical problems. K6YNB chose to use small lightweight antennas with very short (30 inch) booms to simplify the mechanical problems of shifting the polarization.

With these short-boom quads, the entire frame will rotate 360^{o} on its axis to match any incoming polarization with the moon as low as 5^{o} above the horizon.

THE THREE-WAY ROTATOR

As the accompanying photos illustrate, the array uses a CDE Ham-M rotor on a small tower to control azimuth. Atop the Ham-M are two Alliance T-45 TV rotors wired in parallel. These little rotors elevate a mast on which two more T-45s are mounted. These T-45 rotors turn the main axle on which the e.m.e. frame is mounted. As long as the booms of the 16 quads are mounted parallel to this central axle the whole thing will steer in azimuth, elevation and polarization simultaneously.

The method of paralleling Alliance rotors is described in Ham Radio magazine by Forrest Gehrke.[2] Suffice it to say here that the job involves making sure the rotors to be paralleled are in the same position (e.g. both fully clockwise) and then wiring the three motor terminals of rotor #2 to the same terminals of rotor #1. The indicator leads need not be paralleled.

To provide adequate voltage and current capability for two motors, remove the transformer and AC capacitor from one control box, mount them on the rear of the other, and carefully wire everything in parallel. Take care not to reverse the leads on the two transformers.

For those who do not wish to acquire five rotors, even when four of them sell for under $35 apiece, there is an even less expensive way to go. Use ropes instead of the T-45s to steer both elevation and polarization. All you need to do this is some kind of homemade bearings for the rotating pipes.

<u>ABOUT THE CUBICAL QUADS</u>

Many amateurs feel that cubical quads do not work well, especially at VHF. Wherever this myth began, we would like to dispose of it now. As all recent literature on quads indicates, a three-element quad has perhaps 2 dB more gain than a three-element yagi. It takes a yagi with nearly a 5-foot boom to equal the gain of these quads with their 30-inch booms, as J. E. Lindsay pointed out in his definitive study of Quads vs. Yagis.[3] The quads used here have been measured at 9.1 dBd per bay, a little less than the theoretical figure of 9.3 dBd cited by William Orr in the second edition of his cubical quad handbook.[4]

Not only does the choice of quads make it possible to shift the polarization more easily, but it also reduces the cost and simplifies the impedance matching and phasing problems.

The quads are made of clear pine booms (3/4 x 3/4 molding is good for this) with even lighter molding for spreaders. #12 TW covered wire, normally used for AC house wiring, forms the elements. The director and reflector are cut to the dimensions shown, shaped in a square and soldered at the bottom to form a closed loop. The driven element has an SO-239 coaxial connector soldered directly to the two sides of the loop at the bottom. There is no matching or balancing of any kind! For a permanent non-portable installation, the connectors may be eliminated. Attach the ends of each driven element loop to a plexiglass insulator so the phasing lines may be soldered in place.

If the dimensions are followed, the only trick to achieving success with these antennas is to remember that all 16 must be fed with the center conductor in the same position relative to the others. You cannot feed some with the center conductor going to the right side of the driven element while others are fed to the left.

To avoid this mishap, build one quad and then set it aside as a reference antenna and build all the others exactly the same way.

When you mount the quads on the supporting frame, don't spoil everything by mounting some of them upside down or sideways in relation to the others!

THE SUPPORTING FRAME

The supporting frame consists of two 25 foot lengths of aluminum, tapering down from 1½" at the center, and mounted on the main axle (a length of standard TV mast material).

To prevent sag in any plane, the spreaders are connected with a square of rope, so that the assembly resembles one element of a twenty-meter cubical quad with its wire loop. Then the four spreaders are supported with additional outriggers up to the main axle, which extends four feet beyond the main hub for this purpose.

At a point 4'2" from the outer end of each spreader, there is an 8'4" cross spreader of 1" aluminum tubing on which two quads will be mounted. The other two quads on each of the four arms are mounted directly on the spreader itself.

There are many ways to attach these little quads to the spreaders. Any sort of lightweight bracket will do. CUSHCRAFT or KLM element-to-boom brackets work well.

THE PHASING HARNESS

Each quad has a characteristic impedance of about 60 ohms. A full electrical wavelength of lightweight RG-59 type coax reproduces that 60 ohms at a four-way junction, producing 15 ohms at one point on each spreader arm. Four 1.75 wavelength sections of RG-11 transform the 15 ohm up to about 375 ohms. The four RG-11 lines join at the center of the array, producing an impedance of about 93 ohms, which goes through a single quarter wavelength of RG-11 to produce an impedance around 54 ohms and a virtually flat SWR at resonance.

Here is a step-by-step procedure for making the phasing lines.

1) Buy 100 feet of Belden 8221 coax, a lightweight foam line similar to RG-59. If you use any other brand or type, the dimensions given here will probably have to be recalculated because velocity factors vary widely from the nominal published parameters.

2) Cut the 8221 into 16 six-foot lengths and mount a PL-259 connector with a UG-176 adaptor on one end of each. In a permanent installation, the cables may be soldered directly on the quads, avoiding the cost and losses in these connectors.

3) Now measure each 8221 line from the tip of the connector and cut each down to 5"7" (67"), including the length of the connector.

4) Strip back 5/8" of insulation on the open end and solder four 8221 cables to each of four SO-239s, attaching one braid to each mounting hole and bringing all four center conductors into the SO-239s center pin.

5) Now cut four 96" (8') lengths of Belden 8238 RG-11 coax. Mount a PL-259 on one end of each, then strip 5/8" from the other ends and mount all four on another SO-239 (or Type N) at the center.

6) Cut a 13½" length of RG-11, put connectors on both ends, and attach it from the center junction to the antenna relay.

7) Weatherproof all phasing lines thoroughly for permanent installations.

An antenna mounted relay and preamp are strongly recommended, although a short run of low-loss cable such as RG-17 may be used. This will keep the relay and preamp inside, away from the weather, sacrificing a little performance for reliability.

PUTTING IT UP

Assembling the array is pretty simple. Put the rotating mount on your supporting mast or tower. Make sure your array will steer without hitting trees, buildings, or other obstructions. Now assemble the spreaders and connect the supporting ropes.

Finally, mount the quads one at a time, rotating the array after mounting each so the thing doesn't become too imbalanced. A hint: this is easier if the main axle is free to spin. Connect the phasing lines and attach your feedline(s).

HOW IT PERFORMS

After using a moonbounce array that is steerable only in elevation and azimuth, using this array is a revelation. Even though the system as described here is an exceptionally small and lightweight e.m.e. array with only about 20.2 dBd gain, it behaves as if it has much more, because you can always be sure the polarization is optimized. With an ordinary array, you often struggle to hear weak signals (or to be heard yourself) because the polarization is far from correct.

In two weeks of e.m.e. schedules at K6YNB/KL7 it was found that incoming signals were rarely exactly horizontal in polarization.

It was almost always necessary to move to a polarization other than "normal" for best signals. About half the time, the other station seemed to hear K6YNB/KL7 at the same polarization setting that produced the best received signal. On other occasions, various polarizations had to be tried until the other station began responding with "O"s.

Regardless of the Faraday situation at a given moment, however, it was always possible to acquire the signal of well-equipped e.m.e. stations within a few minutes, and most QSOs were completed within 15 or 20 minutes. If no signal was heard in that period of time, it was a pretty good indication that the two stations did not have enough system gain between them to overcome the path loss and make a contact for that day!

With this system, Faraday Rotation ceases to be a serious obstacle to e.m.e. communication.

THE CONCLUSION

This is not the biggest moonbounce antenna around, but it may be one of the simplest and easiest to put together. And it clearly performs well enough to produce lots of good signals off the moon with a pair of 4CX250Bs at 1000 watts input.

At this writing, there are probably 30 stations around the world on two meters with enough system gain to work a station using this antenna consistently.

REFERENCES

1. A Pre-Amplifier for 144 MHz EME - AS49-9

2. Gehrke, Forrest, "Antenna Rotator for Medium Sized Beams," Ham Radio, May 1976, page 48.

3. Lindsay, J.E., "Quads vs. Yagis", QST, May 1968, page 11.

4. Orr, William, All About Cubical Quad Antennas, 2nd ed., Wilton, Conn.: Radio Publications Inc., 1970.

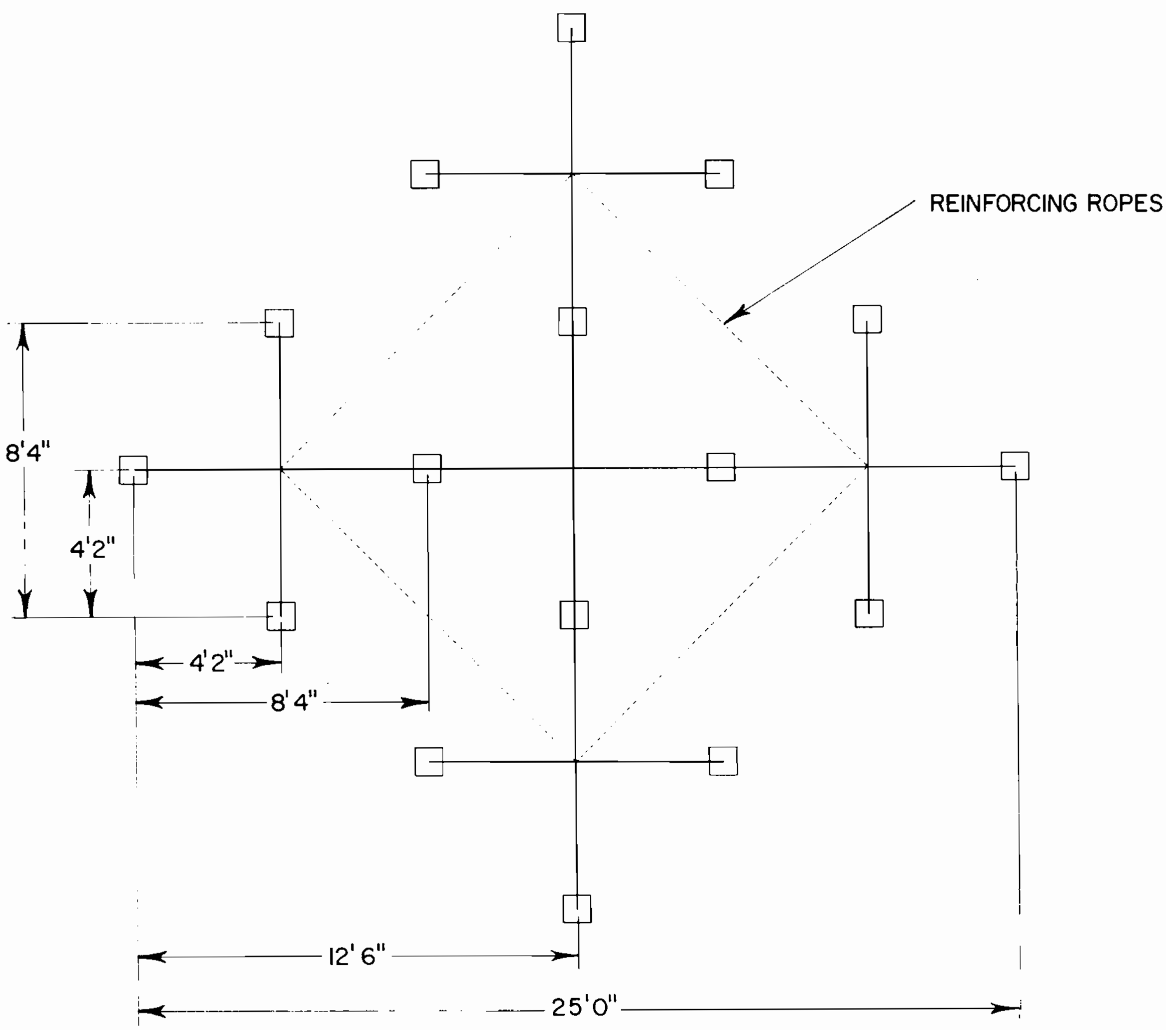
Reflector - 86½" loop of #12 TW copper wire, covered, soldered closed at bottom.

Driven element - 82" loop of #12 wire, with coaxial connector mounted at bottom center.

Director - 77" loop of #12 wire, soldered closed at bottom.

Spacing: 18" Refl. to DE, 12" DE to Director.

Phasing Harness: 16 lengths of Belden 8221 cable, each 67" long, to four four-way junctions; 4 lengths of Belden 8238 (RG-11), each 96" long, from four-way junctions to center of array; single 13½" length of RG-11 to feedpoint/relay.



PHYSICAL DESCRIPTION OF THE ARRAY

The array is made up of sixteen three element Quads.
The Quads are spaced four feet two inches apart.
The total size of the array is 19 feet by 19 feet square.
The spreaders are 25 feet long.

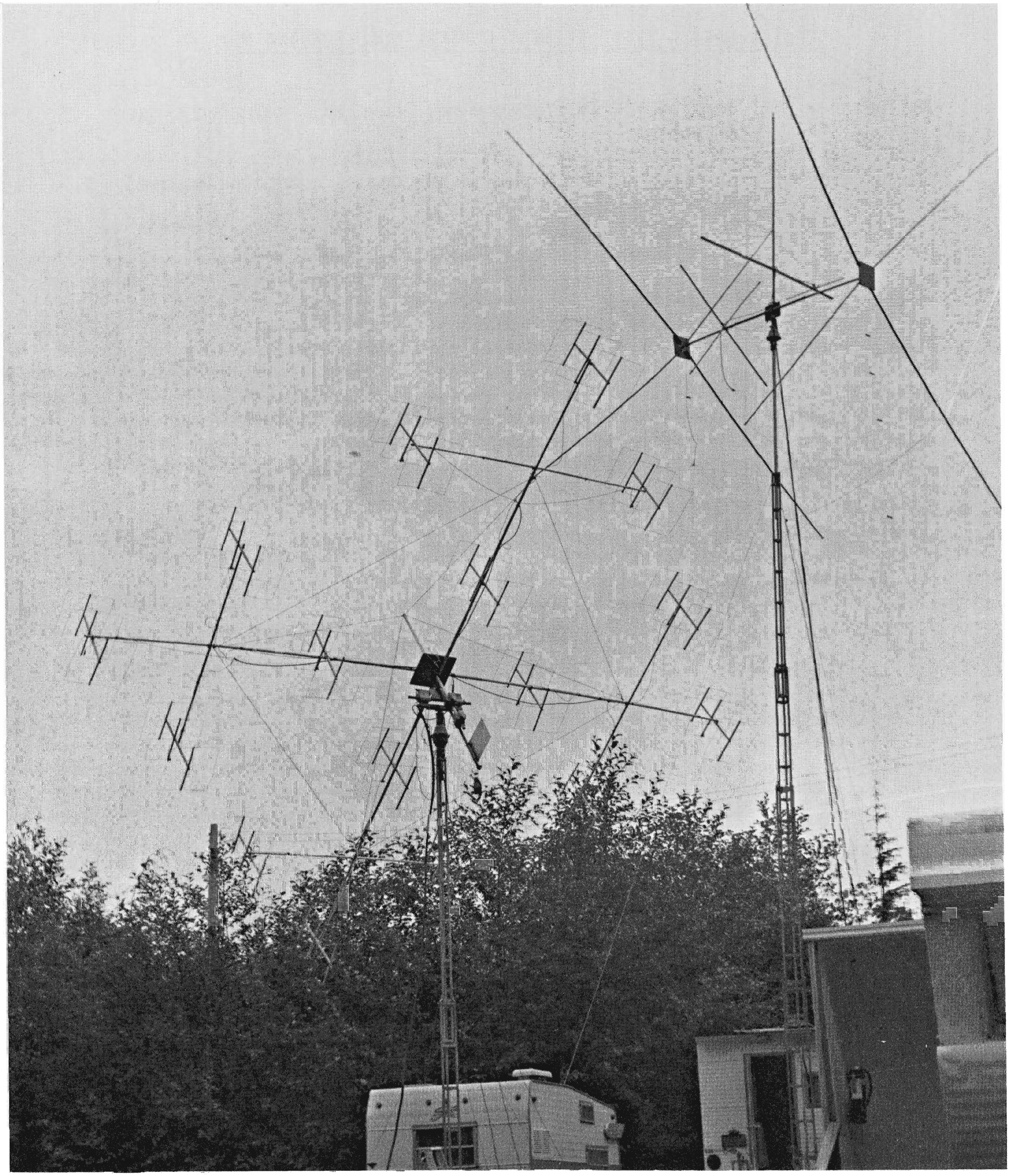


FIGURE 1

This is K6YNB's moonbounce station as it appeared in Ketchikan, Alaska. At left is the array of 16 three-element quads for 144 MHz with its three-way steerable mount. At right, on the tower mounted on the camper truck, is a two-element quad used for liaison communication on 20 meters. With this system, K6YNB/KL7 worked 15 states on two meters in two weeks, 13 of them via the moon!

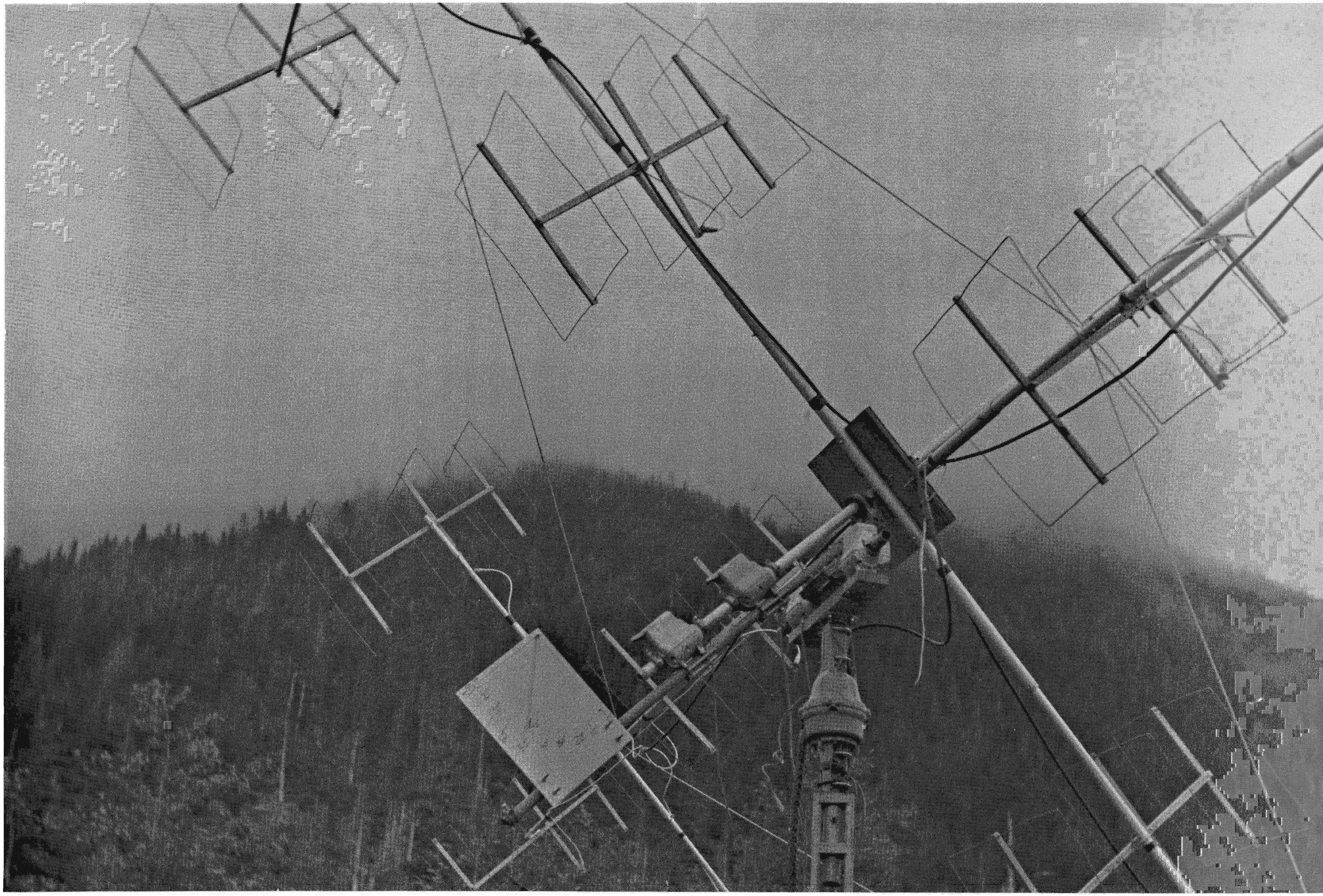


FIGURE 2

This shows the detail of the three-way mount, capable of steering the moonbounce array in azimuth (the Ham-M rotor on the tower), elevation (the two TV antenna rotors on the Ham-M), and polarization (the two TV rotors on the elevated pipe). A homemade protractor accurately reads the elevation, and a filter choke provides counterweight (recommended value: 10 Henry at 200 mAdc).

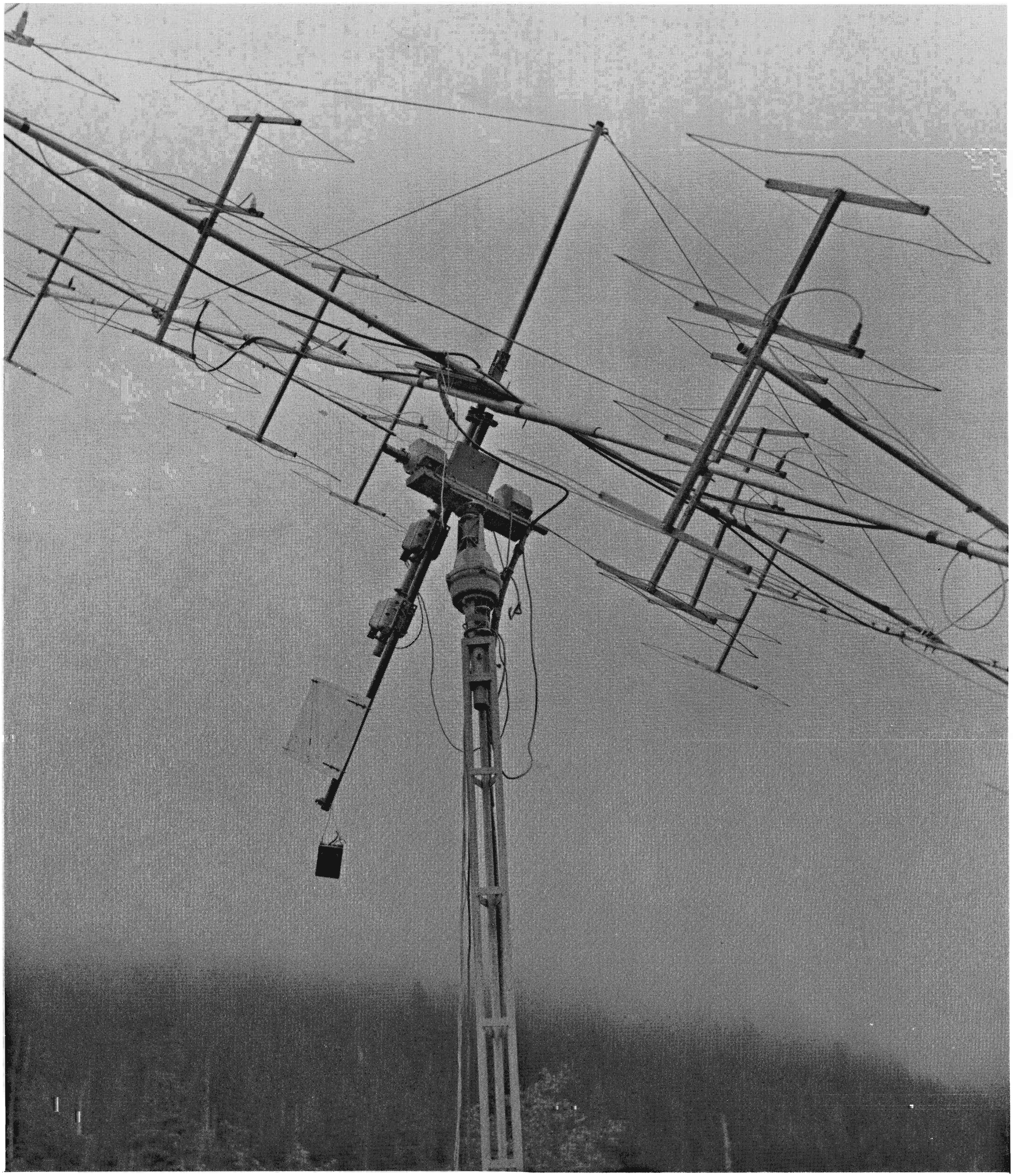


FIGURE 3

Here's another view of the e.m.e. array, against a backdrop of the perennial fog of Southeastern Alaska.

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