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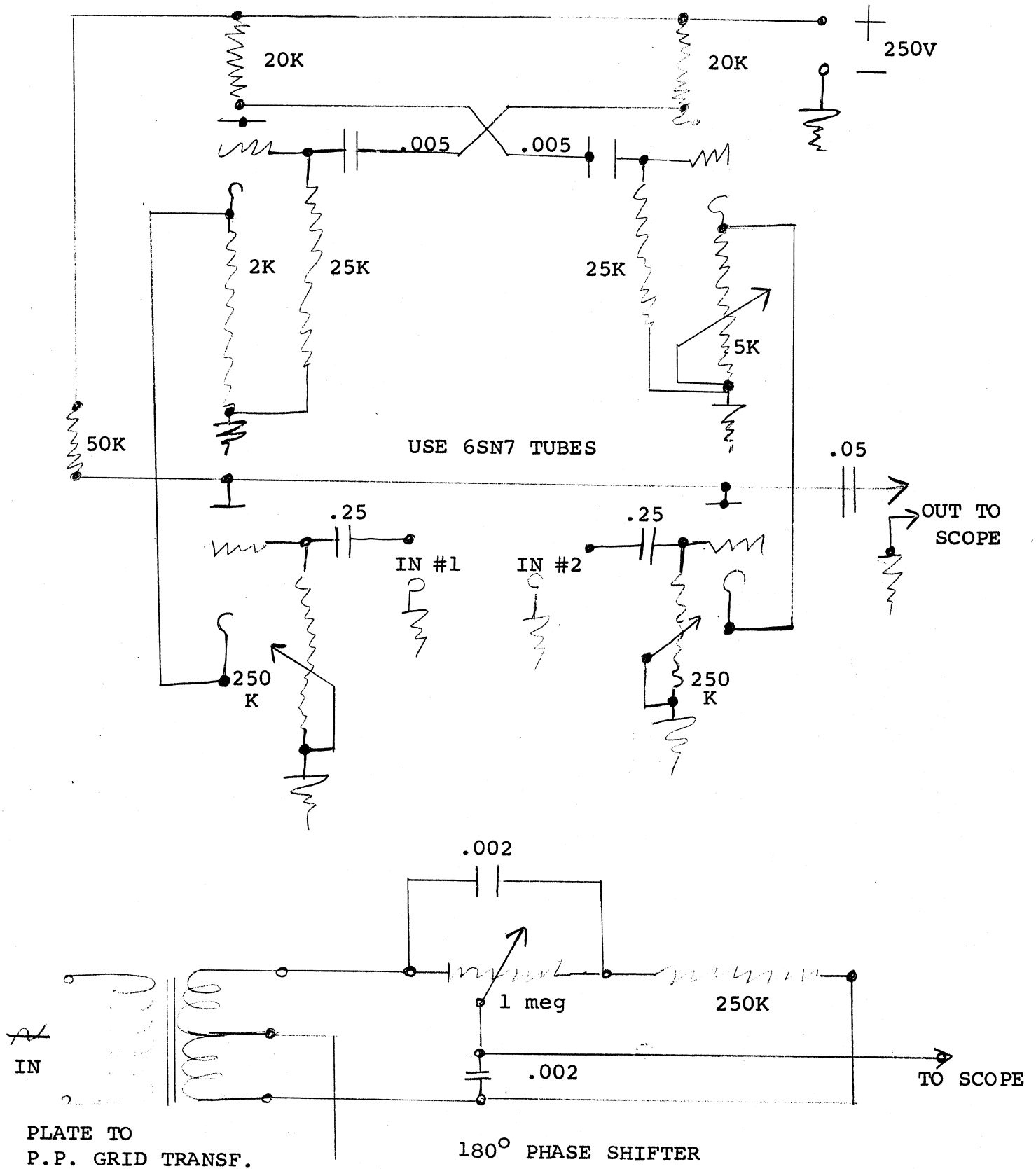
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EITEL-McCULLOUGH, INC.

Electronic Switch That Allows Displays Over Each Other
Or Separated As Desired



R
K6BJ

TETRODE NEUTRALIZATION CONSIDERATIONS

The use of a tetrode results naturally from a consideration of its low driving power and high power-gain as well as from the point of view that it is almost free of the necessity for applying neutralization at the lower frequencies.

It is interesting to note that the very element in the tetrode which makes the tetrode so valuable from the power-gain point of view, the screen, is capable of producing at but one frequency a feedback preventing out-of-phase voltage at the control grid. When the magnitude of this out-of-phase voltage is smaller or greater than necessary, it is capable of causing feedback, hence the need for neutralization even though all necessary precautions have been observed to prevent direct coupling between the output and input circuits external to the tube.

While the fundamental concept of the need for neutralizing triodes is clearly understood it is not so easy to understand that the very nature of the tetrode can cause it to create slight feedback conditions below, but greater feedback conditions above a critical frequency. This critical frequency is usually indicated in the tube data as the frequency at which semi-neutralization occurs. At any higher than the critical frequency and more so when the higher plate voltages are applied, a greater current will flow through the plate-screen capacitance than through the plate-grid capacitance, consequently a greater voltage will develop across the screen-lead inductance. This voltage is 180° out of phase with the voltage from the plate to the control grid and only when the currents flowing to the control grid from these two sources, i. e. plate and screen are equal and opposite will the control grid be at a neutral potential and the tube then will be perfectly neutralized. This state of affairs occurs at the self-neutralizing frequency.

At frequencies below the self-neutralizing frequency of the tetrode external neutralization is generally not necessary except when the higher plate voltages are applied, since the out-of-phase current appearing at the control grid from the screen may then not be large enough to balance out the current that appears at the control grid from the plate. Though usually unimportant, when neutralization is required, this may be accomplished by the insertion of additional inductance between the screen-grid terminal and the cathode, or by conventional cross neutralization.

At frequencies above the self-neutralizing frequency of the tetrode, the out-of-phase voltage developed in the screen-to-ground inductance has become too large and can be compensated for by extending a few inches of wire or tab from the control grid terminal up along the side of the tube, thus adding to the plate-grid capacitance and consequently again balancing the current appearing at the grid from the plate and screen.

John L. Reinartz, K6BJ

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San Bruno, California

A HIGH CAPACITY ANTENNA FOR VERTICAL
HEIGHTS LESS THAN 1/4 WAVELENGTH

By - John L. Reinartz, K6BJ

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It is known that the field strength generated by a vertical antenna is proportional to the current indicated at the base of the antenna. It is desirable, therefore, to increase this current indication by any means available. One way is to increase the power output from the transmitter. When the power of the transmitter is a finite value and limited, only an increase in the effectiveness of the radiating system will then increase the field strength.

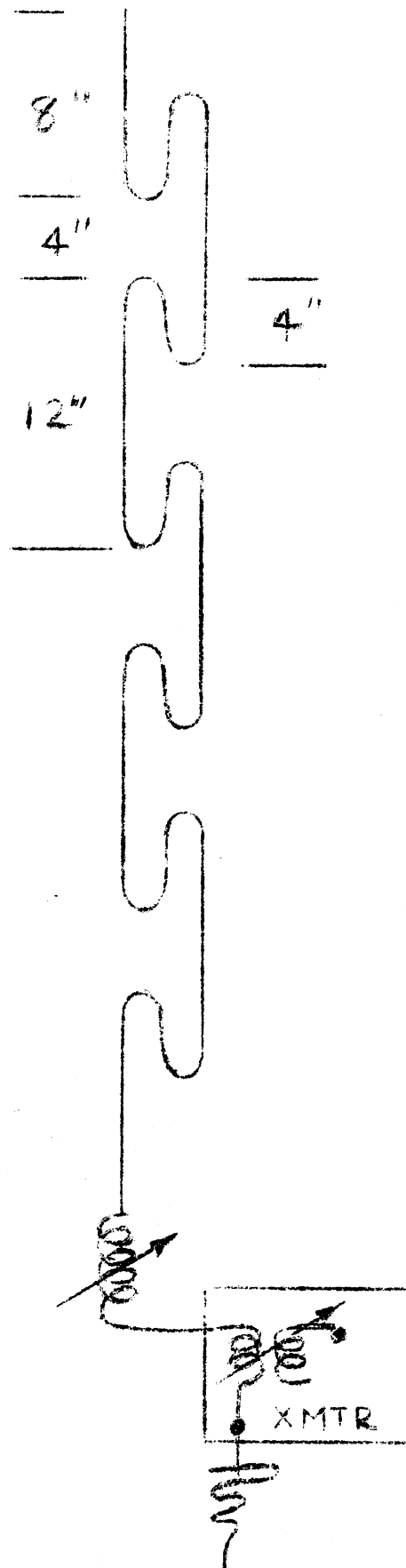
Top loading with a capacity ring or plate as well as center loading with a high Q inductance, are both being used to raise the current indication at the base of the antenna with a resultant increase in the field strength. For mobile operation, a capacity ring or plate on top of a vertical antenna is not always mechanically feasible, while center loading with a high Q coil limits the frequency range over which such a system is effective.

A new and novel method for overcoming the shortcomings of the systems described above has been developed by the writer and it has indicated a capability to increase the field strength generated over a simple vertical wire of comparable height. This new antenna is made up as follows:

Starting from the top of any vertical support, such as a bamboo fishing pole or glass rod of any desirable height, an enamel covered copper wire of #16, 14 or 12 B and S gauge is started at the top after fastening with rubber tape. Then at the 8" point the wire is turned back up for 4" then down for 12", again turned up for 4", again turned down for 12", etc. to the bottom of the supporting pole. As the wire is turned first up and then down, it is placed 120° from the other wire so that the three wires are spaced equidistant around the supporting pole.

The effect of turning the antenna wire up and then down is to decrease the total inductance of this radiating system and to increase its effective capacity resulting in an increased current indication at the base of this antenna for the same power input previously applied to a conventional simple wire antenna of the same height. In addition to the increased field strength generated by this antenna it was also found to be capable of operating over a greatly increased frequency range with comparable signal strength reports for the 2638 Kc boat frequency and the 3995 amateur frequency used by mobiles, the transmitter power remaining the same. At 3995 Kc the antenna was 1/16 wave length high.

1/3 is added to the usual formula for finding the length of wire in feet required for 1/2 wavelength. For a grounded 1/4 wavelength vertical antenna but one-half of this length is required. In each case the resulting physical length of the antenna after folding would shrink to two-thirds of the conventional antenna length. For very high frequencies the folded antenna would allow 3/4 wave operation at the usual length of the conventional 1/2 wavelength antenna and with improved results



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 (Hi Capacity Antenna)

SIMPLE TUBE CALCULATIONS

It is interesting to use the Constant Current Characteristics given in the data sheets of EIMAC tubes and to determine from them the operating conditions a tube will be faced with.

When you have determined that a certain value of plate voltage and grid bias will be used, you place the operating point in a definite location on the constant current chart and thus determine the start of the dynamic characteristic line **for** either Class A, B or C operation.

If you now draw a straight line from this operating point upward and toward the left side of the chart, stopping at the diode line in the case of triodes and at the bottom of the deep curve in the case of tetrodes you can, with the aid of a simple device obtain the value of the instantaneous plate, screen or grid currents as well as their average value. By plotting the instantaneous values on graph paper, a picture of these several current components as a function of time can be formed.

Elsewhere in this paper are plotted the sines of angles from **zero** to 90° in steps of 15° and are shown as lines across the page. If you will take a piece of clear plastic film $3\frac{1}{2}" \times 8"$ and lay it across the lines so that you can, with a large darning needle, scratch these same lines on the clear film, you will have the tool with which to obtain the instantaneous and average current values in the plate, screen and grid pulses.

It will be well to draw your pen along the scratched lines to fill them in with ink for better visibility. You also prick a hole through the zero line in the right hand corner with the same needle so you may hereafter set that point on the starting point of the dynamic characteristic line, moving the top of the film until the 90° line A meets the top end of the dynamic characteristic line E_{min} .

You now read the plate current at each crossing of the dynamic line with the successive lines on the clear plastic, starting with the 90° A line. The average current, that which the dc meter reads, will be given in the formula. If you plot the instantaneous values obtained on cross-section paper, you will form a picture of the current pulses. The X ordinate will then go from zero to 180° for the + Y ordinates and from 180° to 360° for the - Y ordinates.

$$I \text{ average} = \frac{\frac{1}{2}A + B + C + D + E + F}{12}$$

$$\text{Power output} = \frac{\frac{1}{2}A + B + C + D + E + F}{12} \times 0.9 \times (E_b - E_{\min})$$

$$\text{Load Impedance} = \frac{(E_b - E_{\min})^2}{2 \times \text{power output}}$$

The average grid current times the crest excitation voltage (- bias) + (E_{c1} Max) = exciter power required.

Plate efficiency = Power output divided by power input x 100.

These essentials are presented so you may determine how your tube will operate under conditions not given under typical operating conditions in the tube data sheet.

90° A
75° B
60° C

45° D

30° E

15°

0°

Pin hole

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A CARRIER LEVEL AND MODULATION
INDICATING METER

By - John L. Reinartz, K6BJ.

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A most useful device in the radio shack is a carrier level and modulation meter.

The circuit diagram and parts list show how simply such a meter can be constructed. The only item not readily found in the Ham Shack is the 0-50 microammeter. An 0-100 or even an 0-200 microampere meter can however, be substituted if found necessary, although this article assumes the use of a 50 microampere meter for the circuit constants shown.

If similar diodes are used, i.e. 1N34's, 1N38's or 1N52's as pairs, no trouble should be experienced in the calibration of the device.

If a small power supply using full wave rectification and provided with a switch that can switch an 8 mfd filter condenser across the rectified output is used to drive a small crystal controlled or other type oscillator, both carrier shift and 100% modulation can be demonstrated and used to calibrate the device.

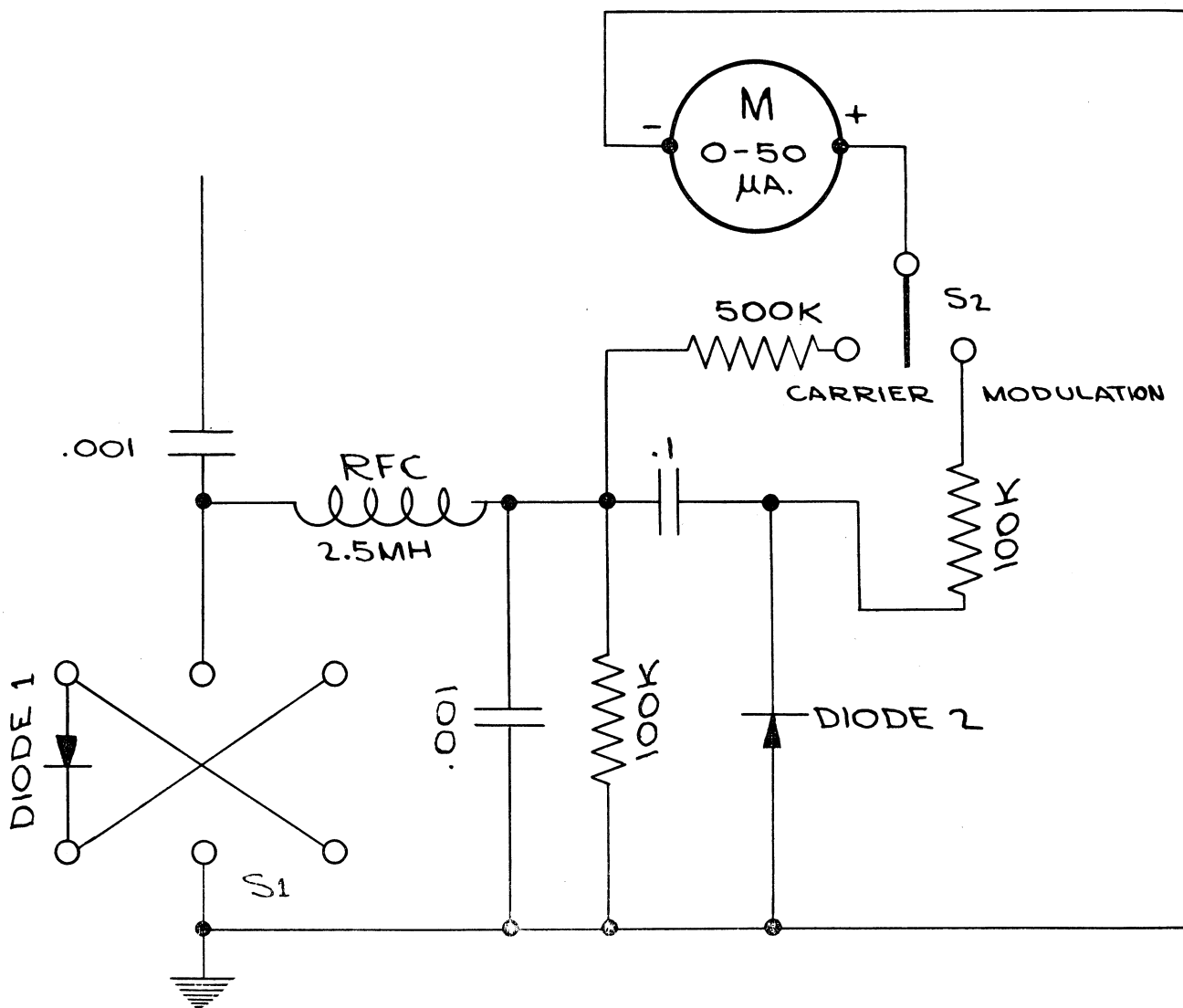
Since with full wave rectification but without a filter across the rectified output the direct current axis is $.9 \times .707$ of the voltage peak, and goes nearly to the peak when an 8 mfd capacitor is connected across the rectified output, the effect of raising the dc axis causes an increase in the input to the oscillator connected to this power supply, with a consequent rise in carrier output. Our carrier level meter indicates this power output change.

To calibrate our meter for 100% modulation, it is but necessary to note its reading as a carrier level meter when the rectified output of the power supply connected to the rf oscillator is not filtered and then to switch our meter to read modulation level, these two readings should be alike. If these two readings differ, resistor 500K should be adjusted until no marked change occurs in the meter reading when its switch is flipped from carrier level indication to modulation level indication.

Our meter is now calibrated at the 100% modulation level. When coupled to a voice modulated transmitter hereafter, it is necessary only to first read the carrier level and then to switch the meter to modulation level indication and whistle into the microphone, if the modulation indication goes to the same reading as the carrier did, 100% modulation can be obtained from your transmitter.

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As an additional refinement, a double pole double throw switch can be inserted into the circuit at diode #1. Either positive or negative modulation peaks can then be measured by reversing the polarity of diode #1.





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STL CIRCUIT DESIGN

THE EIMAC 4X150A POWER TETRODE FOR CIRCUITS
OPERATING ON 940 TO 952 MC. — *By* BYRON O. BALLOU

SINCE it is highly desirable to locate FM transmitters as high above surrounding terrain as possible, the better transmitter sites are frequently beyond the range of line facilities suitable for carrying 15,000-cycle, low-noise FM programs. Recognizing this problem, the FCC allocated the 940- to 952-mc. band for studio-transmitter link service and, more recently, for FM relay broadcasting. This band has proved to be a good choice because suitable antennas with reasonably high gain are not cumbersome in size, permitting the use of relatively low transmitter power for reliable circuit levels.

Mindful of this and other similar applications, the Eimac organization started development of a power tetrode suitable for use at frequencies above 500 mc. In the design of such a tube, there are essential demands which must be satisfied. The more important of these are:

1. Close spacings between internal elements to reduce transit time loss
2. Low-inductance leads
3. Short elements
4. Adequate input-output shielding
5. Adaptability to UHF circuitry

These requirements dictated that the tube should be of external anode construction, using low-current-density seals, oxide-coated cylindrical cathode, a low-inductance screen grid lead to provide isolation between input and output cir-

cuits, and the location of the control grid lead so as to be easily adapted to coaxial circuit design.

The outcome of this project was the new 4X150A. Carefully conducted tests showed stable amplifier operation at frequencies well in excess of 500 mc., and operation as a frequency multiplier to frequencies above 1,000 mc. proved excellent, showing that the tube is admirably suited to UHF circuit design.

During the test program, it was found that outputs of 10 to 15 watts in the new STL band were obtained without difficulty by using a single 4X150A as a frequency doubler, easily driven by another 4X150A as a tripler from approximately 157.5 mc. with less than 2 watts of driving power.

For use in conjunction with the 50-kw. FM transmitter¹ for Radio Diablo, Inc., the Eimac laboratory built a prototype STL transmitter. It was put in daily service on March 12, 1948, transmitting programs from the studios in San Bruno (San Francisco) to the KSBR transmitter site atop 3,850-ft. Mt. Diablo, 32 miles distant. This STL project was undertaken in collaboration with Radio Engineering Laboratories, who designed and built the serrasoid modulator unit. The REL modulator-exciter is capable of some 2 watts output at one-sixth of the antenna frequency, with a deviation of ± 16.6 kc. for 100% modulation. This gives a swing of ± 100 kc. at the antenna frequency.

The circuits were adapted from a wide-range, coaxial-line resonator arrangement used for much of the 4X150A testing in our laboratory. This circuit is unusual in the way the arrangement differs from the run-of-the-mill tetrode VHF circuits. In the usual circuits, the screen is by-passed to a ground plate with the grid circuit on one side and the anode circuit on the other side. In single-tube coaxial circuits this arrangement becomes a bit awkward in that the tube is rather inaccessible. In the 4X150A circuit, the screen grid remains by-passed to the ground (cathode) plate with the coaxial grid circuit extended downward from it. Then the anode circuit is pulled down around the grid circuit. In so doing, the anode end of the tube is made easily accessible from the top of the anode cavity. This circuit operates in a thoroughly normal fashion, with no signs of instability even when working as a straight-through amplifier above 500 mc.

The portion of the STL transmitter described here has to do with the two stages, which multiply the output of the modulator-exciter unit up to the final frequency. The first stage operates as a frequency tripler, the grid drive being supplied in the 156.66- to 158.66-mc. range by the output stage of the REL modulator-exciter at a level of 1.5 to 2.0 watts. The anode circuit is tuned to the third harmonic in the 470- to 476-mc. range. In this stage, both input and output circuits operate on the fundamental

or $\lambda/4$ mode. The second stage is a frequency doubler, with the grid circuit tuned in the 470- to 476-mc. range, and the anode circuit tuned in the 940- to 952-mc. band. In the doubler stage, both the input and output circuits are tuned to the $3\lambda/4$ mode, hence this unit is physically larger than is the tripler stage.

Fig. 1 is a cross-section view of the doubler stage, illustrating the assembly of the socket and resonators. It will be seen that the grid line is a three-quarter wavelength section of coaxial transmission line, shorted on the lower end, and tuned to resonance by the input capacitance of the tube on the open end. There is a small trimmer capacitor located at approximately one-quarter wavelength from the shorted end. At the tube end of the inner conductor is a cylindrical blocking capaci-

tor, and the screen-grid connector collet. This grid line and socket assembly is then mounted inside the inner conductor of the anode cavity, also a coaxial line three quarter-wavelengths long. The line length is such that with the tube output capacitance across the open end, a very small amount of variable capacitance at about

anode cavity. The coupling is varied by means of a small variable capacitance in series with the ground side of the loop.

Cooling air for the anode is supplied through a rectangular opening at the anode end of the assembly, and is exhausted through the anode cooling fins. Fig. 2 shows the blower arrangement.

RESONATOR DIMENSIONS:

Grid line, inner 0.437 in. O.D.
Grid line, outer 1.312 in. I.D.
Anode line, inner 1.75 in. O.D.

Anode line, outer 3.875 in. I.D.

DOUBLER STAGE

11.25 ins. long
6.0 ins. to ground side of screen grid by-pass
7.063 ins. inside length

TRIPLER STAGE

6.62 ins. long
2.813 ins. long
3.875 ins. long

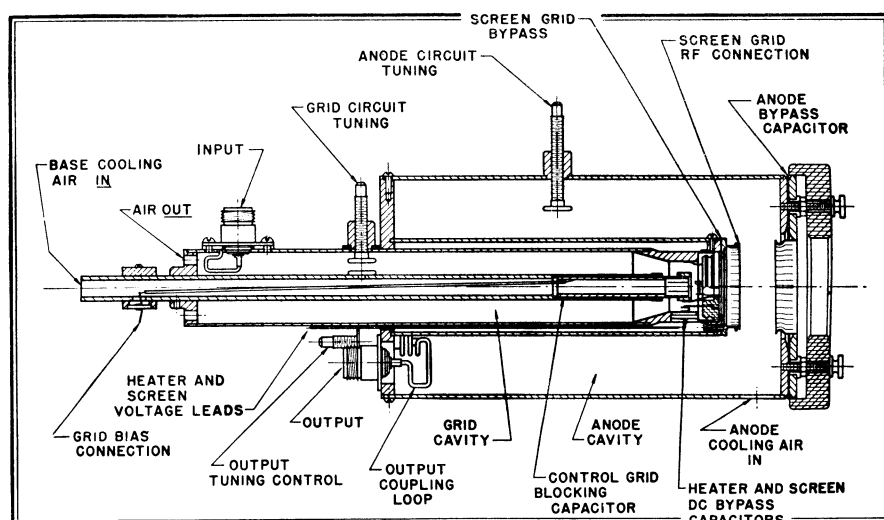


FIG. 1. A cross-section view of the doubler stage, illustrating the assembly of the socket and resonators, and the two air inlets for cooling the base and anode

TYPICAL

OPERATION	TRIPLER	DOUBLER
Anode	700 volts	700 volts
Anode	125 milliamps.	150 milliamps.
Screen	200 volts	200 volts
Screen	5 milliamps.	5 milliamps.
Grid	8 milliamps. ²	8 milliamps. ²
Useful output	15 watts	15 watts

² Through 15,000-ohm grid leak.

TRANSMITTER CHARACTERISTICS

Input Level: plus 10 dbm for 100% modulation (± 100 kc.)

FM noise: better than 70 db below 100% modulation.

Distortion: less than 0.5% between 50 and 15,000 cycles at 100% modulation.

Deviation: ± 100 kc. for 100% modulation, 50 to 15,000 cycles.

Cool air passes over the seals of the tube, a feature most desirable from the standpoint of vacuum tube operation. A static air pressure of .25 in. of water is adequate for proper cooling of the tube anode. A small amount of air is introduced into the inner conductor of the grid line to maintain the base seals of the tube at a reasonably low operating temperature. This air exhausts through holes in the grid-line shorting plug. Static pressure should be .5 in. in this case.

The grid line of the tripler stage is tuned to resonance by means of a 15-mmf. midget variable air capacitor, mounted as far from the shorting plug in the line as is possible. All parts are silver plated to a depth of .001 in. to reduce circuit losses to a minimum.

Specifications for the use of development and design engineers are presented in the tables above.

tor of approximately 150 mmf., using .005-in. Teflon tape as a dielectric. The inner electrode of this capacitor plugs into a ring which is soldered to the grid connection.

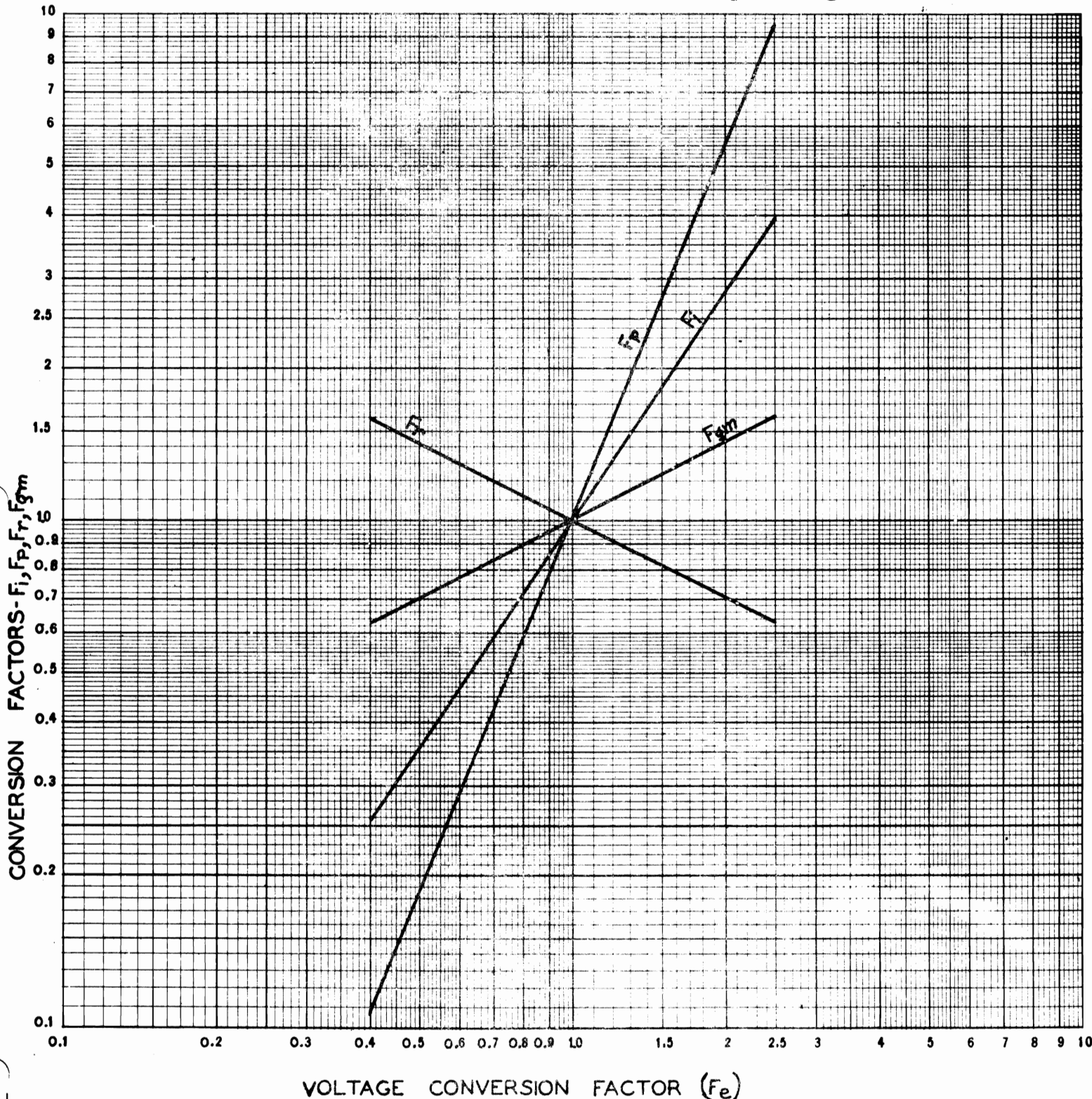
The grid connection is the center pin connector of the standard loktal socket. The control grid bias lead is brought down inside the inner conductor of the line and to the outside through a 500-mmf. ceramic button by-pass capacitor. The drive to the stage is through a standard N chassis connector, mounted on the outer conductor with a small loop attached to it. The socket assembly mounts on the end of the outer conductor of the grid line. This assembly consists of the socket, two ceramic button by-pass capacitors, the annular silvered-mica screen by-pass capaci-

tor, and the screen-grid connector collet. This grid line and socket assembly is then mounted inside the inner conductor of the anode cavity, also a coaxial line three quarter-wavelengths long. The line length is such that with the tube output capacitance across the open end, a very small amount of variable capacitance at about

the center of the inner conductor is quite adequate to tune over the required range. The anode by-pass capacitor is made up of two silvered brass rings, annular in shape. One is soldered to the outer conductor of the anode cavity, and the other carries the connector spring for the anode connection. These rings, separated by clear mica .003 in. thick, provide a capacitance of approximately 1,500 mmf. on DC for the anode is connected to the outer disc by means of a spring connector mounted inside the bakelite protecting cover. In this way the anode potential is positively removed when the cover is lifted for changing tubes. The output coupling loop is mounted on the coaxial cable connector. This, in turn, is mounted on the shorting plug at the bottom of the

CONVERSION FACTORS FOR POWER AMPLIFIER TRIODES & PENTODES

To use these curves: First, determine the ratio of the new plate voltage to the published plate voltage nearest the desired new conditions. This ratio, the Voltage Conversion Factor (F_e), is then used to determine the new screen and the new control-grid voltage. It is also used to determine from the curves, factors for the other new operating conditions.



F_e Applies to Screen Voltage, Control-Grid Voltage, and Plate Voltage.

F_i Applies to Plate Current and to Screen Current.

F_p Applies to Power Output.

F_r Applies to Load Resistance and to Plate Resistance.

F_{gm} Applies to Mutual Conductance.

CLASS B LINEAR SHORT-CUTS

John L. Reinartz, K6BJ

For those who wish to get a Class B Linear on the air with the least amount of research work of the published literature, the following is presented as a method for determining the minimum fundamental operating parameters for a 180° current flow and a minimum distortion factor.

The efficiency of an ideal linear amplifier having a linear characteristic and operating so that the plate current flows exactly 180° during each cycle is:

$$\text{Efficiency} = \frac{\pi}{4} \left(1 - \frac{E_{\min}}{E_b}\right) 100 \text{ percent}$$

where E_b is the plate supply voltage, and $E_{\min}$ is the minimum instantaneous plate voltage during the cycle. Also, power input is defined by:

$$\text{Power In} = \frac{E_b I_{\text{peak}}}{\pi} \text{ and}$$

$$\text{Power Out} = \frac{(E_b - E_{\min}) I_{\text{peak}}}{4}$$

where I_{peak} is the instantaneous peak current at the zero grid voltage point on the constant current characteristic curves at $E_{\min}$.

In the case of the tetrode or pentode, $E_{\min}$ does not drop below the screen voltage, in a triode it can be assumed that $E_{\min}$ is 1/3 of the plate supply voltage E_b .

With this start, we next find the cutoff bias from $\frac{E_{c2}}{u}$ i.e. the screen voltage divided by the screen amplification factor in the case of a tetrode or pentode and in the case of a triode we divide the plate voltage E_b by the amplification factor of that tube.

The next step is to so adjust the bias, that without a signal input to the grid, the plate current reading is I_{peak} , and of course, for this condition the output from the plate is zero.

Now let's see what we can expect as output when a signal is applied to the grid. Let's assume a tetrode, a 4-250A where:

$$E_b = 2000 \text{ volts and } 500 = \text{the screen voltage.}$$

We note that at 500 volts I_{peak} is 0.5 amps on the 4-250A constant current characteristic curve at zero grid voltage and so the power input;

$$= \frac{0.5}{\pi} E_b = \frac{0.5}{\pi} 2000 = 318 \text{ watts}$$

and the power output;

$$\frac{E_b - E_{\min}}{4} I_{\text{peak}} = \left(\frac{1500}{4}\right) 0.5 = 188 \text{ watts}$$

this then proves the first formula of:

$$\text{Efficiency} = \frac{\pi}{4} \left(1 - \frac{E_{\min}}{E_b}\right) 100 \text{ percent.}$$

In the case of a triode, it must be understood that more output can be obtained from a low amplification factor tube than from a high amplification tube. Therefore, let's see what we can get from a 250TL where:

$$E_b = 3000 \text{ volts and } E_{\min} = 1000 \text{ volts.}$$

On the constant current characteristic curves for the 250TL we note that I_{peak} at $E_{\min}$ of 1000 volts = 0.12 amperes at zero grid voltage. So,

$$\text{Power input} = \frac{0.12}{\pi} 3000 = 115 \text{ watts}$$

and the

$$\text{Power Output} = \left(\frac{3000-1000}{4} \right) 0.12 = 60 \text{ watts.}$$

No grid current flow is allowable consequently, no grid driving power is required when the Class B Linear is operated as outlined above. In the case of a tetrode or pentode, the maximum recommended screen voltage should be applied.

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San Bruno, California
JLR:vb

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Eitel McCullough, Inc.

FORMULA FOR GROUNDED GRID OPERATION:

$$\text{DRIVING POWER} = \frac{\text{POWER OUTPUT X GRID VOLTAGE}}{\text{PLATE VOLTAGE}}$$

$$\text{INPUT IMPEDANCE AT FILAMENT TRANSFORMER} = \frac{(\text{PEAK R F DRIVING VOLTAGE})^2}{2 \text{ X DRIVING POWER}}$$

EITEL-McCULLOUGH, INC.
San Bruno, California

Haywire

See QST

See May 1962

QST

INCREASED AUDIO WITHOUT SPLATTER

By - John L. Reinartz.

It is not generally realized, especially in cases where the Class B modulating transformer has poor regulation that a higher ratio of audio output in the positive direction can be obtained if the secondary of the transformer has an additional assymetric load placed on it through the use of a diode tube and appropriate resistor. See attached diagram.

If the negative voltage excursion of the Class B transformer secondary just equals the plate supply voltage, then the positive voltage excursion cannot be as far because in the first instance there is no load across this secondary at the end of the negative voltage excursion and consequently no voltage drop, while in the second instance there is a maximum load across the secondary as the voltage swings in the positive direction with its consequent IR voltage drop and this subtracts from the positive voltage swing.

Placing an extra load on the transformer secondary through the use of a diode and series resistor so that this combination allows loading only on the negative voltage swing, provides a means for preventing the negative voltage excursion from exceeding the positive voltage excursion to any degree desired.

Adjustment of the resistor value can be such that the voltage and modulating power output of the Class B transformer secondary is even greater in the positive direction than in the negative direction with consequent greater modulation effects without, however, exceeding 100% modulation.

A summation of $(E_1 I_1)$ and $(E_2 I_2)$ will show that the requirement of a 50% modulating power for 100% modulation factor is not changed. It is interesting to note that there is a 16% load on the Class B transformer secondary during the negative voltage excursion and an 84% load on it during the positive voltage excursion with respect to the polarity of E_1 . Thus there already exists an assymetric loading of the Class B transformer secondary during an audio cycle.

The addition of the extra load proposed in B during the negative voltage excursion, if of proper value, results in nearly perfect symmetry loading of the Class B transformer secondary during the negative as well as the positive voltage excursions.

1.

(Summary)

Let us consider a Class C rf stage that is to be amplitude modulated. This Class C rf stage is capable of a linear plate current increase with a linear plate voltage increase from a value that may be considered to be the dc plate voltage E_b to twice this voltage $2E_b$. Thus the plate input will go from X watts to 4X watts. Consequently the stage being linear, it can be considered to represent a constant load R, determined by E_b/I_b , I_b being the average plate current and E_b being the plate voltage E_1 .

If we now add the modulating transformer to the Class C rf stage in the usual manner, shown in A of the attached graph, we can consider the secondary of the modulating transformer E_2 to be the equivalent of E_1 in voltage output, since 100% modulation is to be achieved. By definition, 100% modulation is achieved when $E_1 - E_2 = 0$, $I_b = 0$ and the output from the Class C rf stage is zero. Anytime that $-E_2$ is greater than E_1 , overmodulation occurs, no current flows through the Class C rf stage since it is a unidirectional valve and the carrier power is interrupted.

Interruption of the carrier power can result in spurious side bands that interfere with other uses of those frequencies and are to be avoided. These spurious frequencies are also generated when negative clipping is resorted to unless a filter is inserted in the circuit to round off the sharply clipped edges. However, it usually is impossible to reestablish the original wave form, thus some distortion is invariably present in the modulated output. The only reason for negative clipping and subsequent filtering being the desire on the part of the radio amateur for an increased modulated carrier output during the positive voltage swing of the modulation transformer secondary, yet prevent the modulation transformer secondary voltage swing in the negative direction from exceeding the normally applied dc plate voltage, consequently preventing the generation and emission of spurious frequencies.

It occurred to the writer that it should be possible to increase the emission in the positive direction of modulation without the need for clipping in the negative direction of modulation and thereby prevent both overmodulation and audio distortion. As an aid in visualizing the action of the modulation transformer secondary and its effect on the Class C rf stage, an analysis of the several instantaneous values was made and the results shown as curves on cross-section paper.

Several interesting items not previously detailed in textbooks became clear and obvious. The separation of the total power output due to the instantaneous sum of the dc plate and the ac modulation transformer secondary voltages into their component factors E_1 and E_2 is clearly shown in the graph. It is also shown quite clearly that while the full $-E_2$ voltage is developed during the negative voltage generation of the modulation transformer secondary equal to the dc plate voltage E_1 , there actually is no load on either of these two generators. Only when the full $+E_2$ voltage is developed during the positive voltage generation of the modulation transformer secondary equal to the dc plate voltage E_1 are these separate

generators fully and equally loaded.

The load curves for generators E_1 and E_2 are clearly a function of their instantaneous voltages and the instantaneous current I in the circuit system. Obviously E_1 is constant, I varies from zero to $2I$ and E_2 varies from zero to the value of E_1 additive, or subtractive. The load curve for generator E_1 is therefore a straight line between zero and 200% load, while the load curve for generator E_2 is a curve in the positive load direction from zero to 200% and a curve in the negative load direction from zero to a maximum of 25% load and continuous toward zero load at its maximum negative voltage swing. The sum of these two generator loads is of course $I(E_1 \pm E_2) = I^2 R_1 \pm I^2 R_2 = \frac{(E_1 \pm E_2)^2}{R}$.

We now have brought R , R_1 and R_2 into our analysis and it is interesting to note that the resistances across generators E_1 and E_2 vary as the sum of $E_1 \pm E_2$ varies. When the modulating voltage is zero, the system R obviously is the value E_1/I and represents the Class C stage load at the average carrier point shown in the graph. However, when the generator voltage E_2 equals generator voltage E_1 and is additive, the resistance R divides into R_1 and R_2 and each generator sees $\frac{1}{2}R$. This can be more readily understood if we analyze Fig. 2A through 2C. When E_2 equals E_1 but is subtractive, I equals zero, generator E_1 sees a positive infinity resistance and generator E_2 sees a negative infinity resistance and the sum therefore equals zero. However, the sum of the separate resistances that generators E_1 and E_2 see is always equal to the value R at all other points on the varying $E_1 \pm E_2$ voltage curve.

Having noted that the generator E_2 sees no load when its voltage equals the voltage of E_1 but is subtractive and sees $\frac{1}{2}R$ when its voltage equals E_1 but is additive, it follows that no voltage drop occurs in the first instance and maximum voltage drop occurs in the second instance due to the losses that occur in the modulating transformer secondary winding, constituting its own I_r drop and since this I_r drop subtracts from the positive voltage swing, if we can still equal E_1 at this point, we must somehow prevent the negative excursion from exceeding E_1 by the amount of the I_r drop, otherwise we exceed 100% modulation.¹

We have now come to the crux of the whole analysis. We require a means to produce an I_r drop in the modulation transformer secondary when its voltage swing is in the negative direction to equal the I_r drop when its voltage swing is in the positive direction. A diode tube with a resistor in series, connected across the modulating transformer secondary so that current flows in this auxiliary circuit only when the voltage swing is in the negative direction is therefore indicated. The value of the resistance we have already determined to be $\frac{1}{2}R$, while its watt rating should be $\frac{I^2 R}{4}$.

The loading on the modulation transformer secondary when its voltage swing is in the negative direction will now be as indicated by the dotted line in the graph. It will be noted that the modulator system now has a symmetrical loading during the negative as well as positive voltage swing, resulting in reduced second harmonic distortion, full modulation on the positive voltage swing and prevention of overmodulation on the negative voltage swing.

TABLE I.

$\%E_1$	$\%E_2$	$\%E_1 \div E_2$	$\%I$	$\%R_1$	$\%R_2$	$\frac{R}{R_1} \div R_2$
100	0	100	100	100	0	100
100	20	120	120	83.3	16.7	100
100	40	140	140	71.4	28.6	100
100	60	160	160	62.5	37.5	100
100	80	180	180	52.4	47.6	100
100	100	200	200	50.0	50.0	100
100	20	80	80	125.	-25	100
100	-40	60	60	166	-66	100
100	-60	40	40	250	-150	100
100	-80	20	20	500	-400	100
100	-100	0	0	00	-00	00

TABLE 2.

<u>GEN.1</u>			<u>GEN.2</u>		
$\%E_1$	$\%I$	$\%Po$	$\%E_2$	$\%I$	$\%Po$
100	0	0	0	100	0
100	20	20	20	120	24
100	40	40	40	140	56
100	60	60	60	160	96
100	80	80	80	180	144
100	100	100	100	200	200
100	120	120	-20	80	-16
100	140	140	-40	60	-24
100	160	160	-60	40	-24
100	180	180	-80	20	-16
100	200	200	-100	00	0

TABLE 3.

$\% E_1 * E_2$	$\% I$	$\% P_o$
100	100	100
120	120	144
140	140	196
160	160	256
180	180	324
200	200	400
80	80	64
60	60	36
40	40	16
20	20	4
0	0	0

Fig. A.

$$E_1 = E_2$$

$$E = E_1 + E_2$$

$$R = E_1 + E_2$$

$$I$$

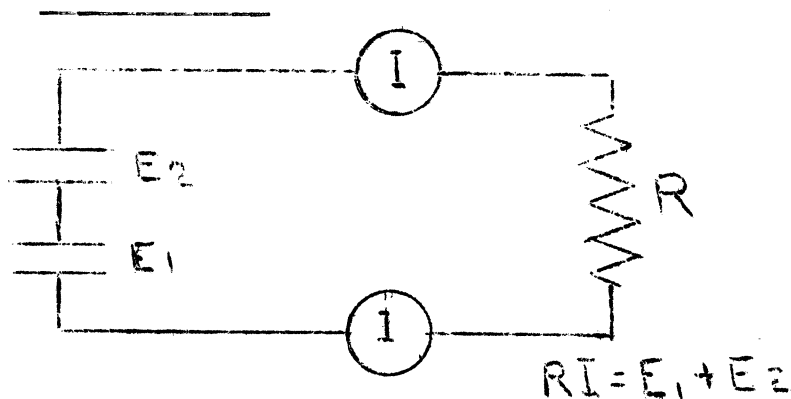


Fig. B.

At $\frac{1}{2}R$ exists a zero voltage point with respect to the connection between E_1 and E_2

I has not changed.

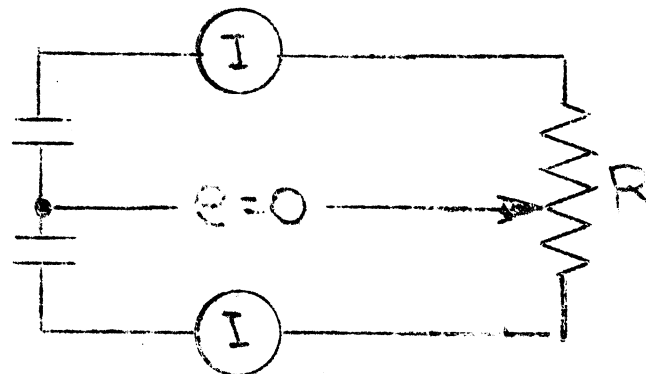
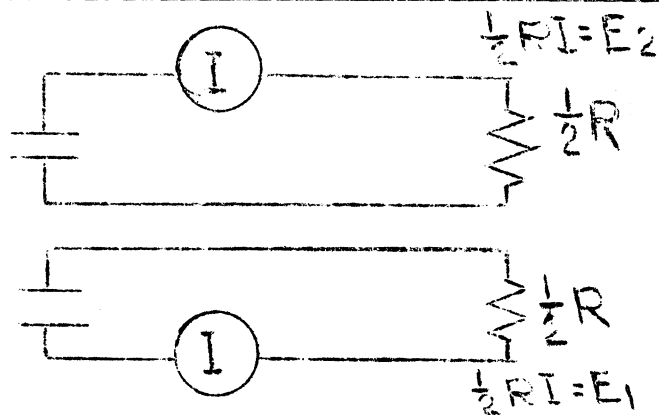
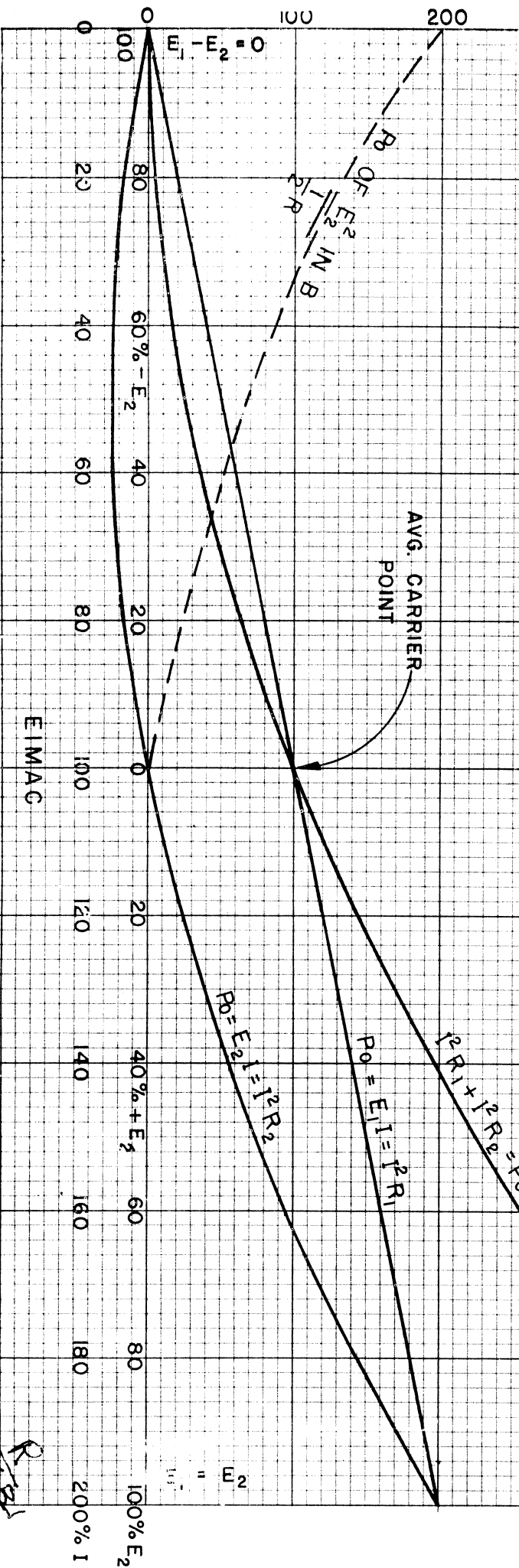
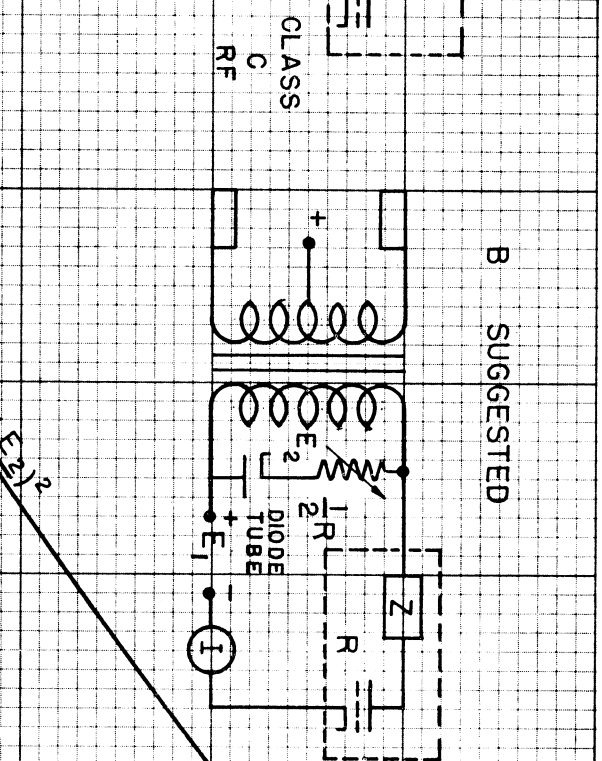
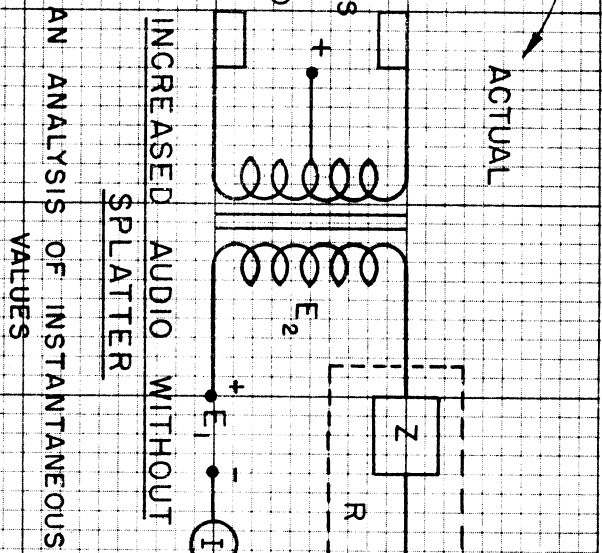
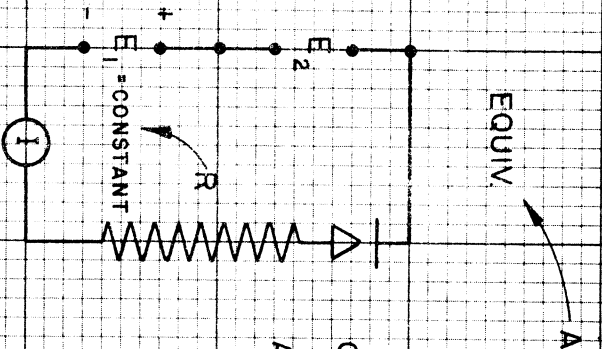


Fig. C.

The two circuits can be separated without changing I , and each generator sees but $\frac{1}{2}R$.





$\% R E_1$	$\% R E_2$	$\frac{+400}{100}$	$\frac{-150}{100}$	$\frac{.6}{100}$	$\frac{25}{100}$	$\frac{+0\% R_2}{100\% R_1}$	$\frac{16.7}{100}$	$\frac{28.6}{100}$	$\frac{37.5}{100}$	$\frac{47.6}{100}$	$\frac{E}{10}$
0	0	500	250	156	125	100	83.3	71.4	62.5	52.4	50

$\frac{R}{K \cdot E_1}$

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A FURTHER EXPLANATION OF THE PAPER
"INCREASED AUDIO WITHOUT SPLATTER"

By - John L. Reinartz

- - - - -

It appears that I should have added a separate analysis to the article on "Increased Audio Without Splatter", in the form of an application to a regular transmitter.

Let us take a normal transmitter where:-

$$E_b = 2000 \text{ volts}$$

$$I_b = 0.2 \text{ amps or a 400 watt transmitter}$$

therefore the Class C load is equal to $2000/0.2 = 10,000$ ohms.

We assume that the Class C stage is capable of a linear current increase with a linear increase in voltage applied to the plate of the tube or tubes, i.e., the electron emission from the filament is more than sufficient for 100% amplitude modulation.

If we now tabulate the instantaneous values of plate voltage increase and decrease, we find that the current likewise increases and decreases, but of course the Class C stage continues to indicate a load of 10,000 ohms, i.e., 2000 volts dc plus 1000 volts of modulating voltage equals 3000 volts, the current will then be

$$3000 \text{ volts}/10,000 \text{ ohms equals } 0.3 \text{ ampere}$$

However, each voltage source sees $2000/0.3$ equals 6667 ohms and $1000/0.3$ equals 3333 ohms, the total of which is 10,000 ohms, the value we started with.

When the modulating voltage is 2000 volts and equals the 2000 volts dc, we get:

$2000 + 2000$ or 4000 volts to the plate of the tube and consequently 0.4 ampere. So each source now sees $2000 \text{ volts dc}/0.4 = 5000$ ohms plus $2000 \text{ volts mod}/0.4 = 5000$ ohms, the total again being the 10,000 ohm plate load we started with.

At this point each voltage source sees 2000 volts dc x 0.4 equals 800 watts and 2000 volts mod. x 0.4 equals 800 watts, the total being 1600 watts which is what we expect for peak power input in a 400 watt transmitter.

That was the positive modulation swing of the modulation transformer. Now let us look at the negative modulation swing. We will take 2000 volts dc and -1000 volts modulation:

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Here the sum is $2000 - 1000$ or 1000 volts to the tube plate. Since the current is also a linear function, it now is 0.1 ampere. However, the plate load is still $1000/0.1 = 10,000$ ohms. But each voltage source now sees 2000 volts dc/ $0.1 = 20,000$ ohms and -1000 volts mod/ $0.1 = -10,000$ ohms, the total being $20,000 - 10,000$ or the same old $10,000$ ohm plate load. Also, the power each voltage source sees is $2000 \times 0.1 = 200$ watts and $-1000 \times 0.1 = -100$ watts, the Class C stage power input being $200 - 100 = 100$ watts which is the same as 2000 volts dc - 1000 volts mod. = 1000 volts $\times 0.1$ amperes or 100 watts.

Now we come to the point where the negative voltage excursion from the modulating transformer is equal to but opposite in sign to the 2000 volts dc. As the negative modulating voltage nears -2000 volts, the sum of the 2000 volts dc and the -2000 volts modulation nears zero. Let us suppose the sum of the two voltages is ± 1 volt, then $1/10,000 = 0.1$ milliamperes, then 2000 volts dc/ $0.0001 = 20$ megohms and -1999 volts mod/ 0.0001 equals -19.99 megohms, the sum again being our old Class C stage of $10,000$ ohm value. But, the load each sees is 2000 volts dc $\times 0.0001 = 0.2$ watts and $-1999 \times 0.0001 = .1999$ watts or a total of $0.2 - .1999 = 0.0001$ watts. This is also the same as 1 volt $\times 0.0001$ amperes or 0.0001 watt. However, when the two voltages, i.e., the 2000 volts dc and the -2000 volts mod. cancel to give zero, the Class C stage has no voltage applied to the plate, no current flows, and there is no load on the 2000 volts dc nor the -2000 volts mod. since these two voltages then are effectively in parallel with $10,000$ ohms as one connecting link. For all practical purposes the $10,000$ ohm Class C stage could be a short circuit for all the influence this would have on the two voltage sources in parallel.

To recapitulate:-

$E_{dc} = 2000$ volts $I_{dc} = 0.2$ ampere
Class C stage load = $2000/0.2 = 10,000$ ohms

$E_{dc} \div E_{mod.} = 2000 \div 1000 = 3000$ volts

I_{dc} goes to 0.3 ampere $3000/0.3 = 10,000$ ohms

$2000/0.3 = 6667$ ohms for dc source
 $1000/0.3 = 3333$ ohms for mod. source
Sum = 10000 ohms

$2000 \times 0.3 = 600$ watts
 $1000 \times 0.3 = 300$ watts
 $3000 \times 0.3 = 900$ watts

In order that the action of the instantaneous summation of the dc and ac voltage values in a high level modulated Class C stage may be observed at desirable incremental values for any length of time, a demonstration unit was assembled comprising a 110 volt variac, two identical 110 volt primary to 110 volt secondary 100 watt transformers as well as the necessary wattmeter, voltmeter and resistances.

Since the instantaneous summation of two ac voltages is readily accomplished and is valid when the load R is essentially non-reactive, the use of ac and transformers is allowable and preferred.

As may be noted from the wiring diagram, the two transformer secondaries are connected in series across the load R, and only the primary of T_1 is constantly connected to one half of the variac voltage between connection 1 and the center tap. It should be noted that the primary of T_2 is connected between the center tap and the variable arm of the variac. It follows therefore that the transformer T_2 secondary voltage will either add to or subtract from the transformer T_1 secondary voltage as the variable arm of the variac is rotated. Only when the variable arm is set on the center tap turn of the variac is the secondary voltage of transformer T_2 zero. At this point only the secondary voltage of T_1 acts on the load R. The analogy of high level modulation of a Class C stage can now be observed at will, the carrier condition without modulation is created when the rotating arm of the variac is at the center turn where no voltage is generated in the transformer secondary T_2 . The full positive modulation condition is created when the movable arm on the variac is rotated to allow the secondary voltage of T_2 to be equal to the secondary voltage of T_1 and additive. Twice the voltage of T_1 then acts on R and since the current through R increased linearly with voltage, this now is $2I$, therefore, we have the 400% peak power condition for 100% positive modulation.

At this point, if the rheostat in series with T_2 is operated, the effect of poor regulation of a modulation transformer can be observed through the reduction of the voltage output of T_2 and the reduction in the power output at load R. When the secondary voltage output of T_2 is subtractive with respect to the constant secondary voltage output of T_1 , the manipulation of the rheostat has no effect since neither T_1 nor T_2 have any load since their combined voltage is zero across R.

The zero carrier condition is created when the rotating arm on the variac is turned so that the secondary voltage of T_2 is equal to but subtractive from T_1 , consequently the voltage acting on R is zero and we have the 100% negative modulation condition at the point where the carrier power is zero.

(1-a)

(Increased Audio
Demonstration)

It will be noted that when the voltage output of T_2 is one-half of the voltage output of T_1 and subtractive, the wattmeter reads 1/8 of the 100% positive modulation condition described above and reads in the negative direction, necessitating a reversal of the polarity of the voltage connection to the wattmeter. The double pole double throw switch is for this purpose.

The entire power curve of the Class B modulation transformer as shown in the graph can be duplicated with complete fidelity. Since it now is quite obvious that T_2 is loaded only when its voltage is additive to the voltage of T_1 and has zero load when its voltage is equal to but subtractive from T_1 , the connecting of $\frac{1}{2}R$ across the secondary of T_2 at this point will load T_2 symmetrically in each direction of its voltage excursions. Of course $\frac{1}{2}R$ must be taken out of the circuit whenever the variable arm on the variac is moved to create the condition where the output voltage from the secondary of T_2 crossed zero and moves to the additive value with respect to T_1 .² In a high level modulated Class C stage, this is automatically accomplished through the use of a diode and series resistor as shown in the graph.

This paper is written in the hope that it may clarify through demonstration the complex action of a modulating voltage in series with a constant voltage acting on a Class C power stage.

- John L. Reinartz -

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The Home of Quality Tubes

JLR/fe

(2-a)

(Increased Audio
Demonstration)

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$$E_{dc} \div E_{mod.} = 2000 \div 2000 = 4000 \text{ volts}$$

$$I_{dc} \text{ goes to } 0.4 \text{ ampere, } 4000/0.4 = 10,000 \text{ ohms}$$

$$2000/0.4 = 5000 \text{ ohms for dc voltage source}$$

$$2000/0.4 = \underline{5000} \text{ ohms for mod. voltage source}$$

$$\text{Sum} = \underline{10000} \text{ ohms}$$

$$2000 \times 0.4 = 800 \text{ watts}$$

$$\underline{5000} \times 0.4 = \underline{800} \text{ watts}$$

$$4000 \times 0.4 = \underline{1600} \text{ watts}$$

$$E_{dc} - E_{Mod.} = 2000 - 1000 = 1000 \text{ volts}$$

$$I_{dc} \text{ goes to } 0.1 \text{ ampere } 1000/0.1 = 10,000 \text{ ohms}$$

$$2000/0.1 = 20,000 \text{ ohms for dc source}$$

$$-1000/0.1 = \underline{-10,000} \text{ ohms for mod. source}$$

$$\text{Sum} = \underline{10,000} \text{ ohms}$$

$$2000 \times 0.1 = 200 \text{ watts}$$

$$\underline{-1000} \times 0.1 = \underline{-100} \text{ watts -- see curve}$$

$$1000 \times 0.1 = 100 \text{ watts}$$

$$E_{dc} - E_{Mod.} = 2000 - 1999 = 1 \text{ volt}$$

$$I_{dc} \text{ goes to } 0.0001 \text{ ampere } 1/0.0001 = 10,000 \text{ ohms}$$

$$2000/0.0001 = 20 \text{ megohms for dc source}$$

$$-1999/0.0001 = \underline{-19.99} \text{ megohms for mod. source}$$

$$\text{Sum} = \underline{10000} \text{ ohms}$$

$$2000 \times 0.0001 = 0.2 \text{ watts}$$

$$\underline{-1999} \times 0.0001 = \underline{-0.1999} \text{ watt}$$

$$1 \times 0.0001 = 0.0001 \text{ watt}$$

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At this point, since we have no load on the modulation transformer, it is subject to negative voltage fluctuations or excursions due to audio frequency resonance which are also a cause of over modulation. The system described in my paper effectively limits such excursions so they cannot reach the dc voltage to produce a zero voltage total or even a negative value.

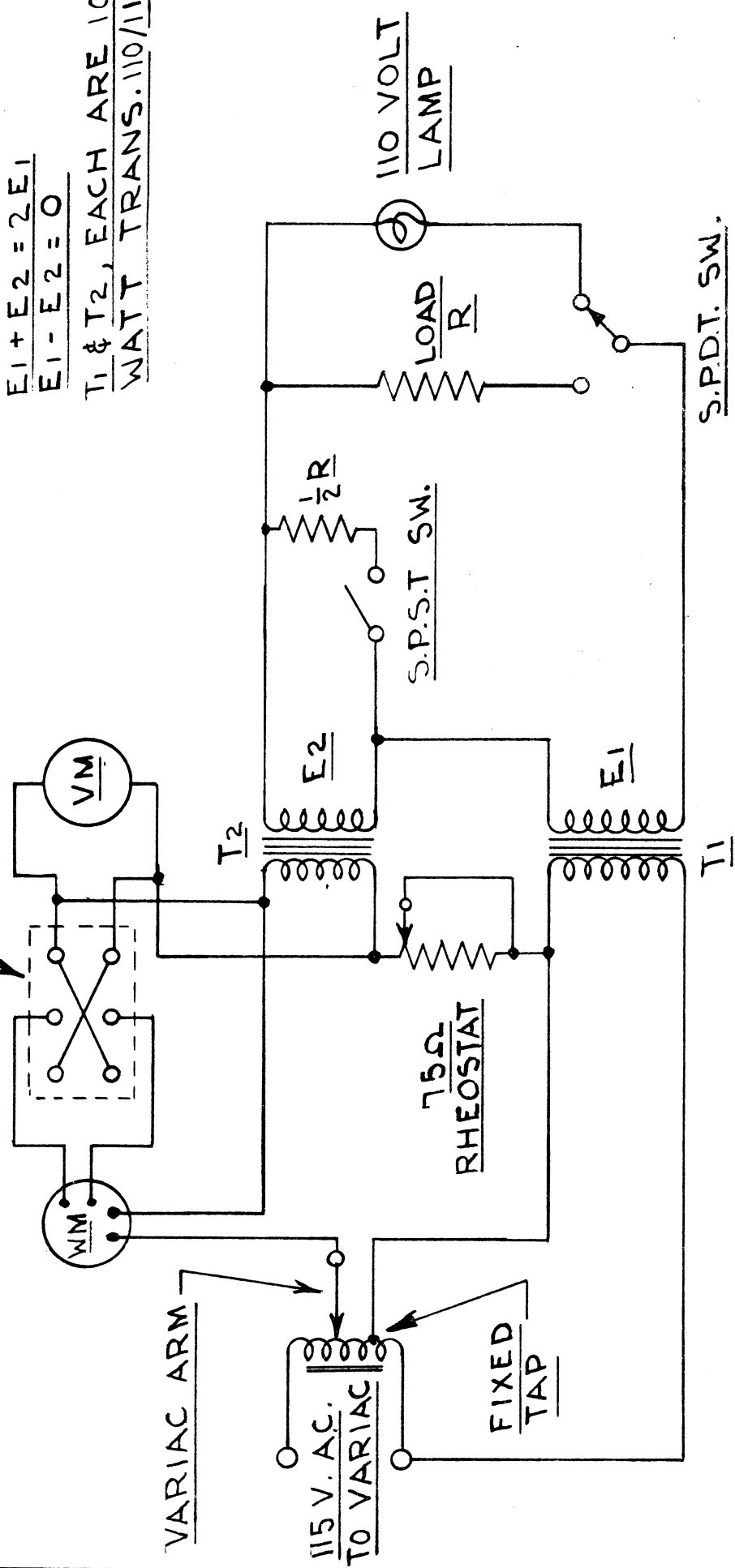
- John L. Reinartz, K6BJ

COMMERCIAL PARTS	
No. REQ'D	DESCRIPTION

DRWG. No.	
SUPERSEDES	
SUPERSEDED BY	
ASSEM. DRWG. No.	

POLARITY REVERSING SWITCH

$E1 + E2 = 2E1$
 $E1 - E2 = 0$
 $T1 \& T2, \text{ EACH ARE } 100$
 $\text{WATT TRANS. } 110/110V.$



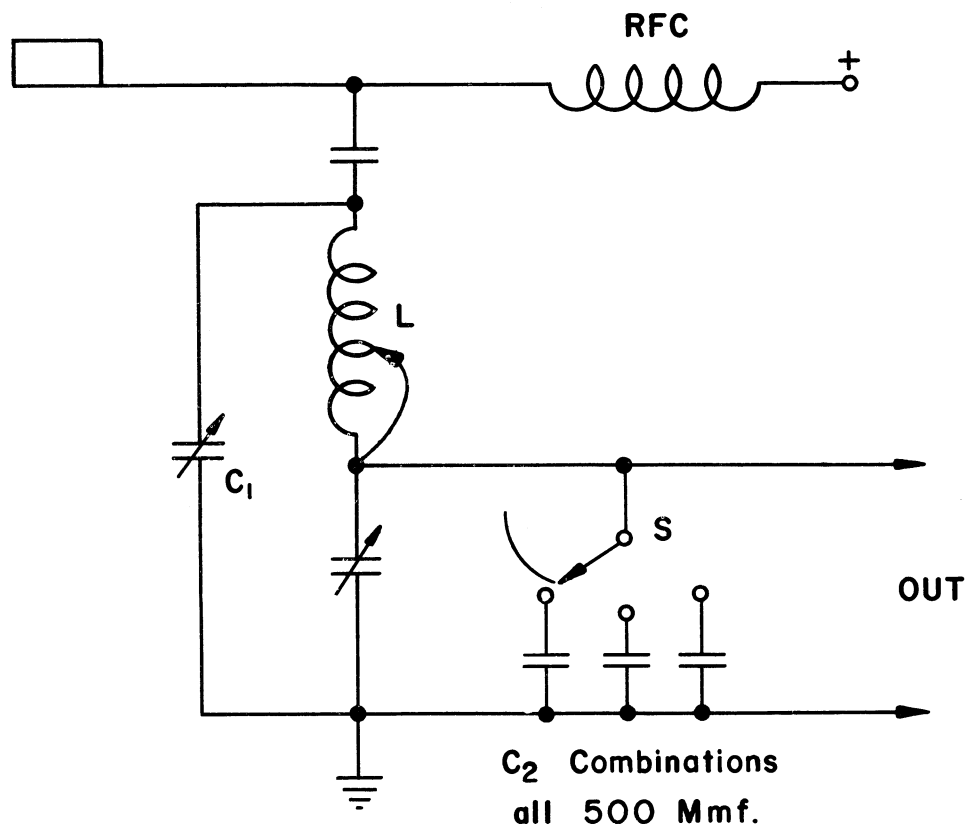
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 SAN BRUNO CALIFORNIA
Eitel

MODULATION ACTION
 DEMONSTRATION UNIT

DRN. AEM	DATE 6-25-52	CHKD.	DATE	CHANGE No.	DESCRIPTION	BY	DATE
TRCD.		APPVD.					
SCALE		NO. REQ'D					
MATERIAL							
FINISH		TOLERANCE					
HEAT TREAT							
REMARKS							

PI NETWORK SHORTCUTS

The pi network looks like the diagram below and by rule of thumb allows deductions that serve as a good start for maximum values for the lowest frequency in any amateur band that is to be covered.



Since the capacitance values of C_1 and C_2 are intimately tied into the reactance ohm value of the inductance L that will be used and because the value of XL is determined by the plate voltage, power output and Q of the circuit, we can use the values given in the attached table without further work.

If the reactance ohm value of X_L is taken as 300 for a typical case then the sum of the values of X_{C_1} and X_{C_2} must also equal - 300 for resonance to take place in such a tuning system.

The next step is to determine the value of the capacitance at the lowest frequency in the band to be covered and for the lowest output ohms the system is to match. If the system is to match 50 ohms we merely place a straight edge from say, 3.5 mc through 50 ohms and read 1000 mmfd's as the proper value of C_2 . The X_{C_1} value obviously is 300-50 or 250 ohms and its capacitance value from the same table is 200 mmfd's. In order that we may juggle values to control output, C_2 should be made up of three fixed sections of 500 mmfd's each plus a 500 mmfd variable while C_1 would be a 200 mmfd variable.

In the same manner and using the value of $X_L=300$ we can also obtain the microhenry value of L and in this case we find it to be 15 microhenry. Because this problem was based on operation from 3.5 to 4 mc, the inductance should be variable to allow changing to the higher amateur bands.

$$\text{Plate resistance} = \frac{E_b}{2xI_b}$$

Where I_b = Plate Current Amperes and E_b = Plate Voltage

$$X_c = \frac{E_b}{2QI_b}$$

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Pi Network Shortcuts
John L. Reinartz, K6BJ
Eitel-McCullough, Inc.
9/3/54

$$\text{TETRODES}$$

$$X = \frac{(.57 E_b)^2}{Q P_o}$$

<u>E_b</u>	<u>P. In.</u>	<u>P. Out.</u>	<u>Efficiency</u>	<u>Q</u>	<u>Ohms Reactance</u>
1000	133	100	75	12	270
1500	267	200	75	12	305
2000	536	400	75	12	270
2500	800	600	75	12	282
3000	1000	800	80	12	305

An average value of 300 would serve.

1000	133	100	75	15	217
1500	267	200	75	15	243
2000	536	400	75	15	217
2500	800	600	75	15	225
3000	1000	800	80	15	243

An average value of 225 would serve.

$$\text{TRIODES}$$

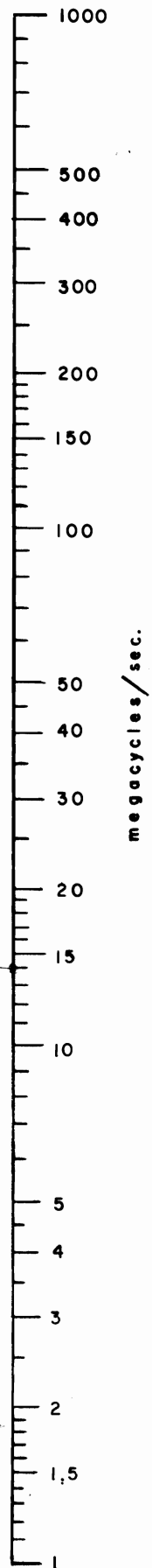
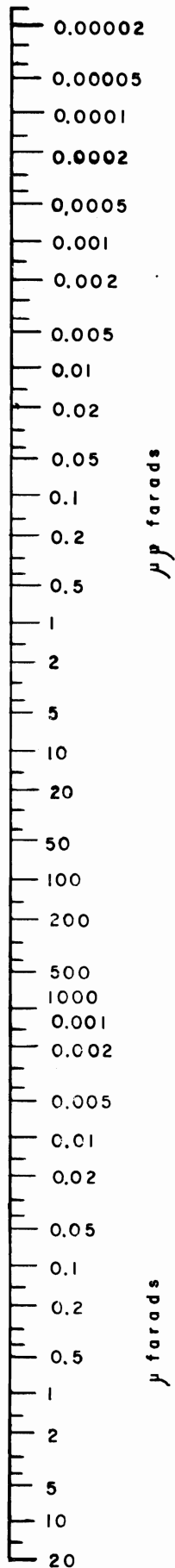
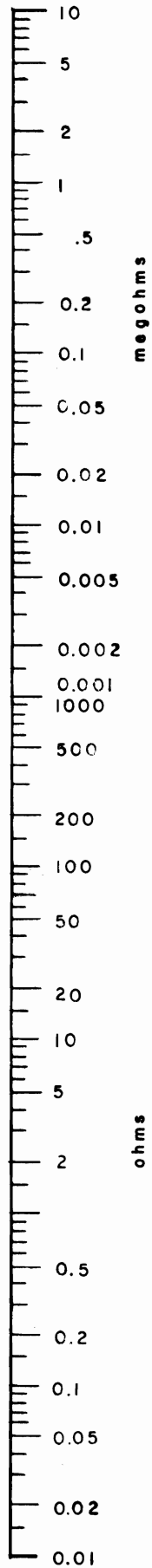
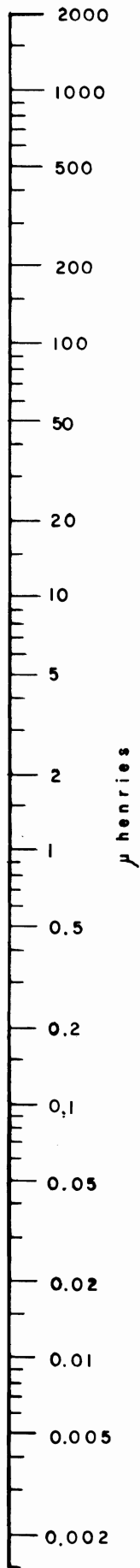
$$X = \frac{(.63 E_b)^2}{Q P_o}$$

<u>E_b</u>	<u>P. In.</u>	<u>P. Out.</u>	<u>Efficiency</u>	<u>Q</u>	<u>Ohms Reactane</u>
1000	133	100	75	12	330
1500	267	200	75	12	372
2000	536	400	75	12	330
2500	800	600	75	12	345
3000	1000	800	80	12	372

An average value of 350 would serve.

1000	133	100	75	15	263
1500	267	200	75	15	297
2000	536	400	75	15	263
2500	800	600	75	15	276
3000	1000	800	80	15	297

An average value of 275 would serve.



THAT ELUSIVE SWR

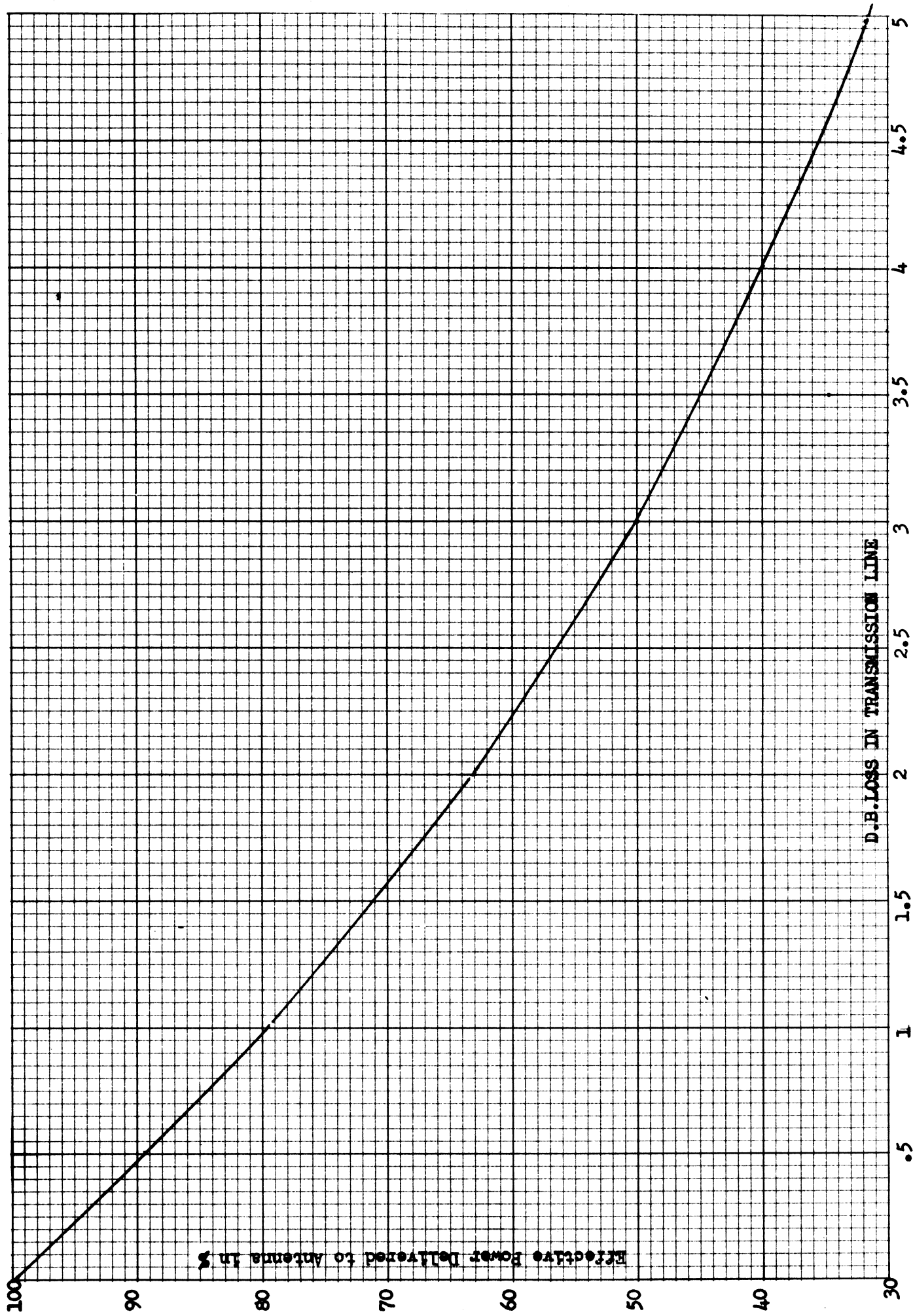
The economics of operating with an SWR as near to 1/1 as practicable may not always be apparent, and may mean little when the coax cable is short and the operating frequency is low. Yet there are many of us who, when operating on 10 meters with low power want to get as many of our few watts to the antenna connection as possible. Nor is the high power man immune; he paid lots of money for his KW rig in order to have many watts available at the antenna connection. Even he doesn't want to lose any! Most of us even put up nice three element beams, home-made or expensive, to further put these watts to work in the most desired direction. Thus the main idea is, lose as few watts as possible between the transmitter and the antenna.

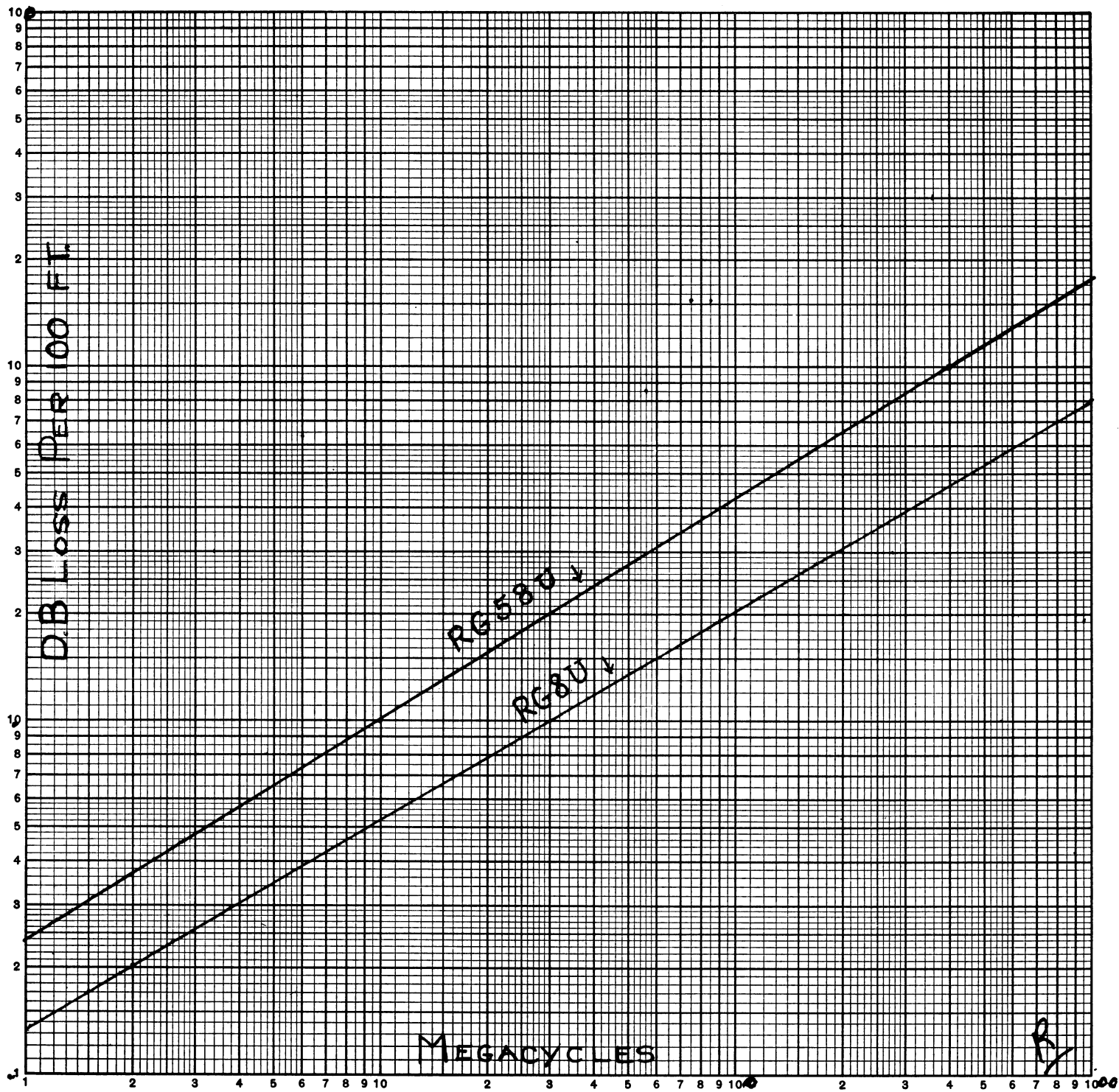
Many of us use RG8U coax cable; it is good and sometimes is available inexpensively. If we take a look at the loss graph for this cable, we note that it is not lossless! Even at 4 m.c. it has approximately 0.3 db loss per 100 feet, and this goes up to nearly 1 db when we operate in the 10 meter band. A look at the percent-power-delivered-to-the-antenna-graph, tells us that this 100 feet of RG8U lets us deliver only 94% of the transmitter output to the antenna connection when we operate in the 75 meter band, but only 79% when operating in the 10 meter band. However, there is another joker, the SWR must be 1/1! For any other SWR value we have another line loss as shown in the SWR-vs-line-loss-multiplier-graph. This graph tells you that if you are operating with an SWR of 3.7/1, your coax line loss is multiplied by two.

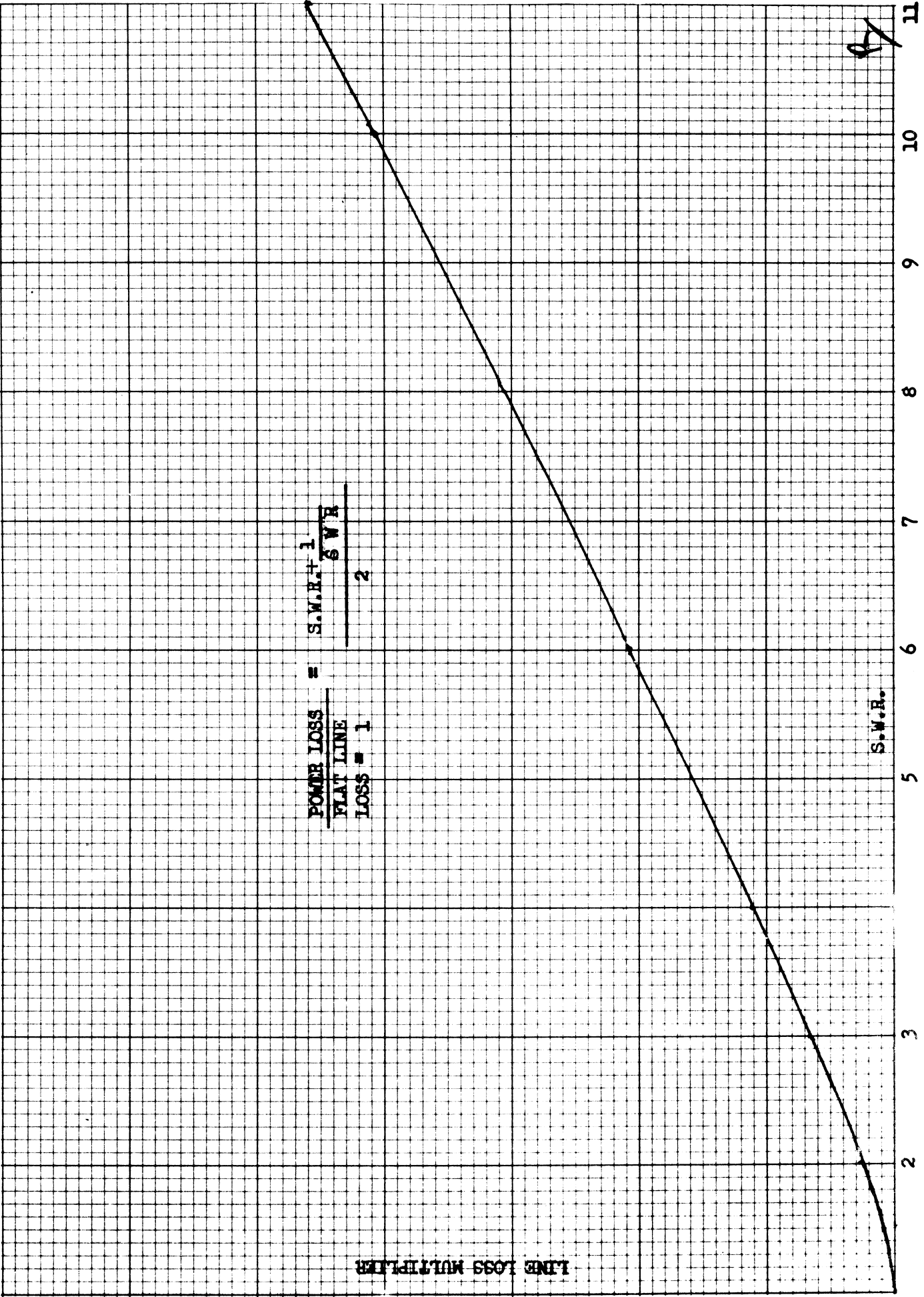
Let's keep the SWR down and let's not lose any more watts between the transmitter and the antenna than we can prevent with a little extra care in the antenna design at the operating frequency; also let's keep the coax cable as short as practicable.

JOHN L. REINARTZ

K 6 B J







$$\frac{\text{POWER LOSS}}{\text{FLAT LINE LOSS} = 1} = \frac{\text{S.W.R.} + 1}{\text{S.W.R.} - 1}$$

LINE LOSS MULTIPLIER

S.W.R.

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EITEL-McCULLOUGH, INC.
San Bruno, California.

THE REDUCTION OF HARMONIC POWER
OUTPUT IN AMATEUR TRANSMITTERS.

By-John L. Reinartz, K6BJ.

It is not generally realized that most amateur transmitters using but one tuned circuit in the final output stage cannot meet the FCC rule stated in Art. 17, Act of 1947 in regard to the reduction of the radiation of harmonic frequencies to not less than 40 db below the fundamental frequency output. This comes about in the following manner:

A tube generates harmonic components as a matter of course when using Class C conditions. Each time the grid of the tube is driven positive, a pulse of current flows in the plate circuit of the tube. The amplitude of these harmonics depends on the angle of plate current flow. The magnitude of these harmonics is given by Terman in "Radio Engineer's Handbook" Fig.66. For example, for a plate current flow angle of 140° the harmonic amplitudes are as follows:

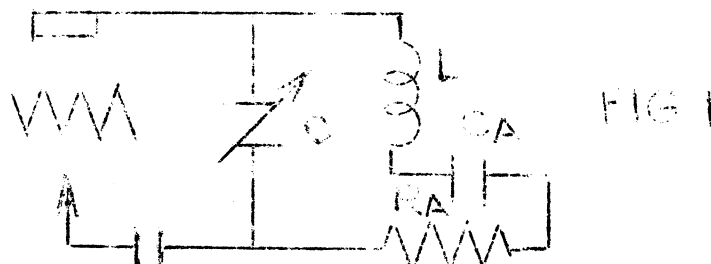
$db = -20 \left(\log \frac{I}{I_n} \right)$ where I = the fundamental and I_n = the harmonic current and gives us Table I.

<u>Component</u>	<u>TABLE I.</u>	
	<u>Amplitude in % of Fundamental</u>	<u>Amplitude in db</u>
Fundamental (N = 1)	100%	0
Second Harmonic (N = 2)	69.4%	- 3.2
Third Harmonic (N = 3)	30.8%	-10.3
Fourth Harmonic (N = 4)	4.6%	-25.8

The voltages produced across the output circuit by these harmonic components are dependent on the magnitude of the impedances presented to each harmonic component by the tuned circuit and are dependent to a large degree on the Q of the tuned circuit.

The performance of any amplifier in a transmitter is determined by both the characteristics of the tuned circuit used and the tube. The choice of the tube has been made easy for us by the manufacturer who has supplied us with the necessary tube characteristics and operating values. It is therefore necessary only for us to give consideration to the rf circuit constants that should be used. C, L and R can be of variable values within rather large limits and if frequency were the only consideration the capacity could be made small and the inductance large or vice versa, while the action of the reflected load resistance on the tuned tank circuit is to decrease the sharpness of tuning as its shunt value is made smaller. In actual practice there is a compromise value for the three components comprising a tuned tank circuit at resonance which results in high efficiency and good harmonic suppression.

Now the larger the tuning capacitance the smaller the impedance presented to the harmonic components present in the plate current pulse consequently the harmonic voltage produced across this capacitance is smaller. Also we have a larger circulating current in a larger capacitance for a given watt output. It is this ratio, also called Q , of the circulating volt-amperes (rf voltage times circulating current) to the power output that determines the harmonic power that can be passed on, being higher for low Q and lower for high Q tuned tank circuits at resonance. Harmonics are suppressed to a considerable extent even by a simple tuned circuit. Labus and Roder (Proceedings of IRE, Vol. 19, pp 949-962, 1931) derived expressions for the suppression of harmonics. For example, if the tuned tank circuit is as shown in Fig. 1



where R_A and C_A are the antenna resistance and capacitance, then the db reduction of harmonics in the antenna due to the tank circuit Q is

$$db = -20 \left(\log N^2 \left(1 - \frac{1}{N^2} \right) Q \right)$$

where Q is the operating Q of the tank circuit at the fundamental frequency. This gives us Table II.

TABLE II.

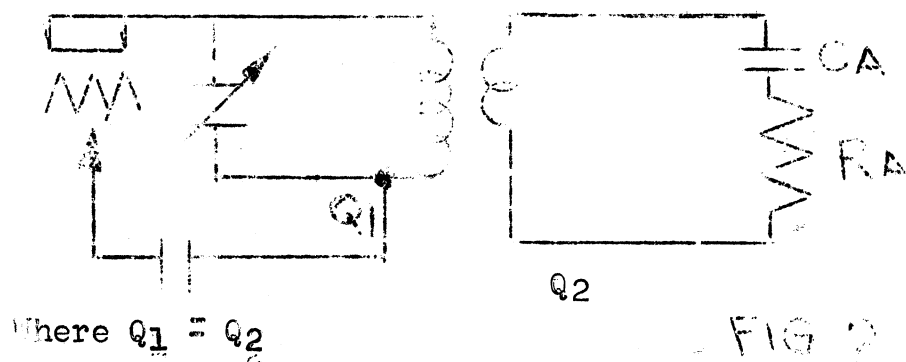
<u>Q</u>	<u>N = 2</u>	<u>N = 3</u>	<u>N = 4</u>
5	-23.5	-32.0	-37.5
10	-29.6	-38.1	-43.5
15	-33.0	-41.6	-47.0
20	-35.6	-44.1	-49.6

adding these db values to those obtained in Table I we get

<u>Q</u>	<u>N = 2</u>	<u>N = 3</u>	<u>N = 4</u>
5	-26.7	-42.3	-63.3
10	-32.8	-48.4	-69.3
15	-36.2	-51.9	-72.8
20	-38.8	-54.4	-75.4

We see from these tabulations that every time we double the Q of the tuned circuit the harmonic level goes down by 6 db, but that the reduction of the second harmonic still is insufficient to comply with the FCC rule of -40 db even when the Q of the tuned circuit is -20.

A usual Q for amateur operation is 15. If the circuit is doubly tuned as in Fig. 2, there is a greater reduction of harmonics as shown in Table III.



$$\text{db} = -20 \left(\log \frac{N}{I_n} \right) + -20 \left(\log N^2 \left(1 - \frac{1}{N^2} \right)^2 Q_1 Q_2 \right)$$

TABLE III

<u>Q</u>	<u>N = 2</u>	<u>N = 3</u>	<u>N = 4</u>
5	-38.2	-54.4	-76.8
10	-50.2	-67.4	-88.8
15	-57.3	-75.1	-96.2
20	-62.3	-79.4	-100.8

It can be seen from these expressions that whenever a Q is doubled the harmonics are all reduced by 12 db. Another important fact that can be seen from these expressions is that it is better to have a Q of say 10 in each circuit of Fig. 2 than to have a Q of 20 in the circuit of Fig. 1 and we now can meet the FCC rule of -40 db for harmonic radiation when we use a Q of 10 or better in each of the tuned circuits. This -40 db represents .01 watt for an amateur station radiating 100 watts on the fundamental frequency.

Now let us consider the field strength produced by an antenna. This problem has been treated by W. T. Wintringham in Publication R4-2860-A (July 1948) of the Radio Manufacturers Association. He shows that the field strength produced by a horizontal half-wave dipole is

$$E = 23 \sqrt{\frac{P}{d}} \text{ volts per meter}$$

where P is the radiated power in watts and d is the distance in feet from the radiator to the point where E is measured. This can be considered an average value. Actually the field strength varies with the

distance between a lower and higher value due to subtraction and addition of the wave reflected from the ground and the direct wave as well as the configuration of the lobes due to the length of the transmitting antenna for a particular harmonic of the fundamental frequency for which the antenna may be cut. The formula should not be used for distances greater than about 650 feet. Since the amateur would be concerned with distances down to say 50 feet the above formula for field strength applies. Inversion of the formula will give the power required to produce a given field strength.

$$P = 1820 (Ed)^2 \text{ microwatts}$$

where E is in volts/meter and d is in feet.

The FCC defines the limit of the television service area as a field strength of 500 microvolts/meter. It has been determined that an interfering signal of 1/100 of this value is not objectionable. We therefore must not produce an interfering field greater than 5 microvolts/meter. Let us now find the power required to produce such a field at 500 feet.

$$P_{uw} = 1880 (5 \times 10^{-6} \times 500)^2 = .012 \text{ uwatts.}$$

Compare this .012 microwatts with the .01 watt the present FCC ruling allows for harmonic level radiation.. This .01 microwatt represents a power ratio of fundamental to harmonic of 10^{10} when the power radiated at the fundamental is 100 watts, this increases to 10^{11} when the fundamental power is 1000 watts and is -100 db and -110 db respectively. These values are far more severe than the -40 db level currently required, but are what the amateur must attain if he wants to stay on good terms with the general public.

Since even two tuned tank circuits may fail to reduce an interfering signal to the -100 or -110 db level, other means must be found to do this and several good articles have appeared in amateur magazines. Max Seybold has shown in the August 1947 issue of QST that the addition of trap circuits in the plate lead of the final Class C stage will reduce the harmonic level some 40 to 50 db and if considered along with two tuned tank circuits may reach the desired -100 or -110 db level.

In cases where even the processes outlined above fail to reduce the interference to television reception for shorter distances than 500 feet, it will be found advantageous to resort to additional grounded trap circuits tuned to the interfering harmonic with such trap circuits closely coupled to the hot end of each plate tank circuit of every stage in the transmitter. Such a system devised by the writer was found capable of apparently complete cancellation of an harmonic. Because every rf stage in a transmitter amplifies the harmonic components present in its grid excitation voltage, the first place to get rid of the harmonic is at the crystal oscillator plate tank circuit. What may be left can be taken care of in the subsequent

stages at their respective plate tank circuits.

In order that we may first locate the offending harmonic, we must utilize some method of measuring its relative amplitude. The old standby and universally used absorption-type wavemeter is ideal for this purpose when modified as shown in Fig. 3. Devised by the writer, it has proved to be extremely sensitive yet simple to construct.

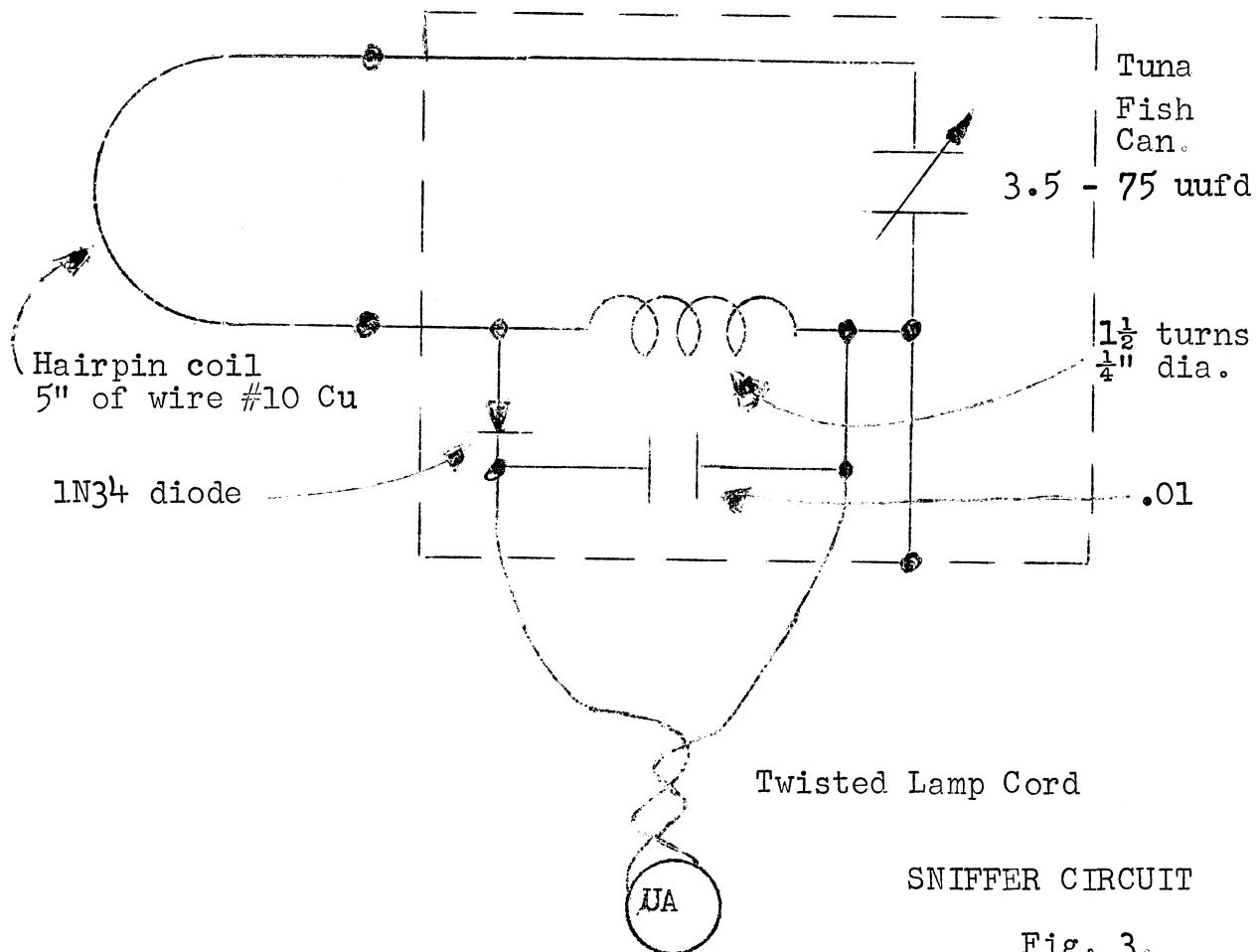


Fig. 3.

As the check for harmonics progresses, make a note of the location and relative value of the harmonics for future reference. Harmonic indications may be found in almost any lead in the transmitter but may be expected to give greatest indication near the plates of a tube. Care should be used to prevent actual contact with any high voltage point.

To reduce the harmonic radiation from leads, these should be

by-passed with 0.001 mmf at as many points as may be required to suppress the harmonic. For the cancellation of the harmonics found in the plate lead of the tubes, a tunable circuit is coupled to the hot end of the plate tank, one side is grounded and the other side connected through a 10 mmf capacitor to the cathode near the socket as shown in Fig. 4. The circuit should tune from about 33 mc to as high a value as possible, with a good variable air tuning capacitor. The method of use is to tune the circuit at the offending harmonic frequency in such a manner that a counter emf is produced which effectively cancels the harmonic in the cathode stream of the tube as indicated by a complete cancellation of the harmonic radiation reading on the sniffer micro-ammeter.

Each stage in the transmitter should be so treated if there is indication of harmonic output at any stage.

The use of metal chassis even though not completely shielded is better than breadboard construction. However, the operation of this system of reducing harmonic interference does not specifically require shielding.

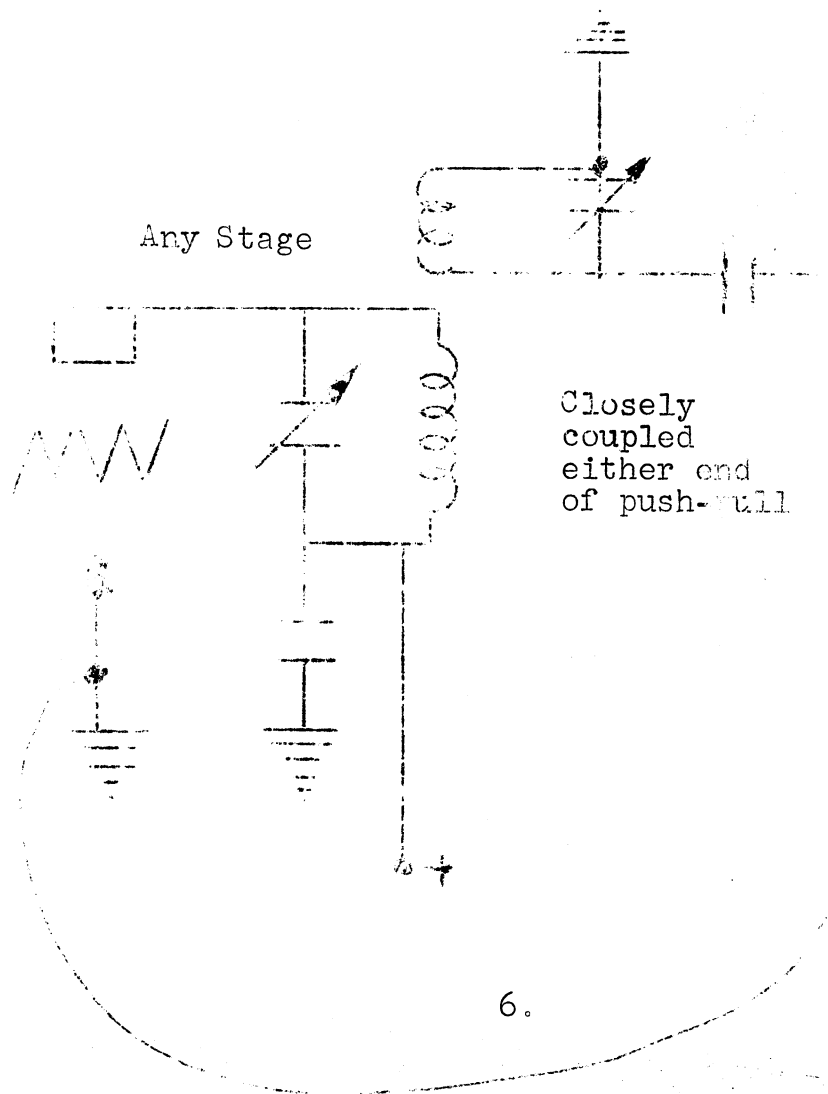


FIG. 4

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COUPLING CIRCUITS BETWEEN
STAGES AND TO ANTENNA.

By - John L. Reinartz -
K 6 B J

The choice of coupling methods between the several stages in a transmitter is essentially one of expediency and is determined mostly by the physical location and arrangement of the circuit components. However, the power transfer from the final stage to the antenna system should be at the operating frequency and with maximum harmonic suppression in order that we may comply with the regulations of the FCC that harmonic radiation be at least 40 db below the fundamental power radiation. Here the coupling methods are intimately tied in with the choice of the constants of the tuned tank circuits. These constants deserve careful consideration first.

A tuned circuit contains capacity inductance and resistance of variable values. The resistance is made up of the ohmic resistance of the inductance and the reflected resistance of the load and if frequency were the only consideration, the capacity could be small and the inductance large or vice versa, while the action of the reflected load resistance on the tuned tank circuit is to decrease the sharpness of tuning as its shunt value is made smaller. In actual practice there is a compromise value for the three components comprising a tuned tank circuit at resonance which results in high efficiency and good harmonic suppression. The values of these circuit elements are important and are not hard to determine.

The performance of any amplifier in a transmitter is determined by both the characteristics of the tuned circuit used and the tube. The choice of the tube has been made easy for us by the manufacturer who has supplied us with the necessary tube characteristics and operating values. It is therefore only necessary for us to give consideration to the rf circuit constants that should be used. Values of C, L and R can be chosen so as to reduce the amplitude of harmonic frequencies and also will contribute in no small way to the successful operation of a well-engineered transmitter.

Let us analyze the tuned circuit for the final stage in our transmitter after we decide upon the output in watts each intermediate stage shall be capable of supplying to the next stage as driving power. Let us then decide on a 100 watt final stage. Everything will now center around this value. The next requirement is to choose the operating dc voltage. This is usually given by the manufacturer for a stated tube with some allowable variation. For a 100 watt stage a reasonable value is 1000 volts. We now have the

operating dc voltage and the power output requirements, two of the essentials of what is to follow.

When the tube is operating, there is an ac voltage component (at radio frequency) superimposed on the dc plate supply voltage. This alternating voltage is produced across the tuned-circuit thru the action of the grid driving voltage applied to the tube from the exciter stage. In a Class C stage each time the grid of the tube is driven positive, a pulse of current flows in the plate circuit of the tube. This current is not sinusoidal and contains many harmonic components. The voltages produced across the load circuit by these harmonic components are dependent on the magnitude of the impedance presented to each harmonic component by the tuned circuit.

From Ohms Law for ac, we know that E^2/R watts is equal to the equivalent load resistance and it is also known that the rms ac voltage across the tuned circuit is approximately equal to $0.9 \times 0.707 E_p$. In our case this is $0.9 \times 0.707 \times 1000$ which equals 630 rms volts. Now R equals $630^2/100$ and is 4000 ohms (approx.) shunt and is the equivalent ac resistance of the tuned circuit. For push-pull R equals $(2 \times 630)^2/100$ and is 16000 ohms (approx.) shunt. Later we need this information in the determination of the capacity value that is to be used in the tuned circuit.

Now the larger the tuning capacitance the smaller the impedance presented to the harmonic components present in the plate current pulse, consequently the harmonic voltage produced across this capacitance is smaller. Also we have a larger circulating current in a larger capacitance for a given watt output. It is this ratio, also called Q , of the circulating volt-amperes (rf voltage times circulating current) in the tank circuit to the watts output expected from the tank circuit that determines the harmonic power that can be passed on, being higher for low Q and lower for high Q . Table A shows the relation of the harmonic to the fundamental in db as a function of the circuit Q .

From Table A it can be seen that in order to keep the harmonic power that can be generated in a tuned circuit low, the ratio of volt-amperes to watts output should be high, but if we are to keep the I^2R losses due to the volt-amperes down then a practical upper limit for Q is 20. Let us use a ratio of 15. This is the loaded Q of the circuit. Since $C = \frac{Q}{WR}$, if we now put $C = 15 \times 10^6 / 2\pi FR$ and calcu-

late for 14.25 mc., we obtain 42 micro micro farads, this includes the tube capacitance across the tuned circuit. Therefore this 42 micro micro farads is the proper total value of capacitance for the volt-ampere component of 15 times the watts output in order that the harmonic power level in the tuned circuit may be initially reduced as shown in Table A.

Continuing the analysis of the tuned circuit we find that the value of the inductance that must go with the 42 micro micro farads for resonance at 14.25 Mc. is found by first finding the reactance ohms of the capacitance, $10^6 / 2 \pi F C = 266$ ohms and since at resonance $X_L = X_C$ and as $L = X_C / 2 \pi F$ we have $266 / 2 \pi F = 3.0$ micro henries.

From the values we have just obtained we learn more about the tuned circuit. For instance, $630/266$ equals the current flow thru the condenser or coil and is equal to 2.36 amperes. The coil wire size should be such as to carry this current without heating. The voltage across the condenser under the loaded condition is equal to $I_a X_C \sqrt{2}$ and is equal to $2.36 \times 266 \times 1.414$ and is equal to 900 volts peak. With no load across the tuned circuit the condenser peak voltage will be nearly twice the plate supply voltage if no dc voltage appears across the condenser. In actual practice the condenser plate spacing should be capable of withstanding three times the dc voltage without breakdown. We now have engineered one tuned circuit of our prospective transmitter. This procedure should be followed for each tube before we are ready to consider coupling methods from one stage to the next or from the final stage to the antenna.

The method of coupling between stages is determined somewhat by the constructional features of the transmitter. If the low power frequency multiplying stages are adjacent to each other, capacitive coupling can readily be used and thereby we can obviate tuned grid systems. In such cases the coupling problem is considerably simplified. If the intermediate stages are located at different levels, link couplings using a coaxial cable suggests itself since we are also interested in the reduction of energy radiation by the coupling device. Therefore a coaxial cable should be used because of its complete shielding and the easy construction of an insulated link, while the outer shield can be bared and grounded at all necessary places. The link should be coupled to the tank circuit at the point nearest rf ground.

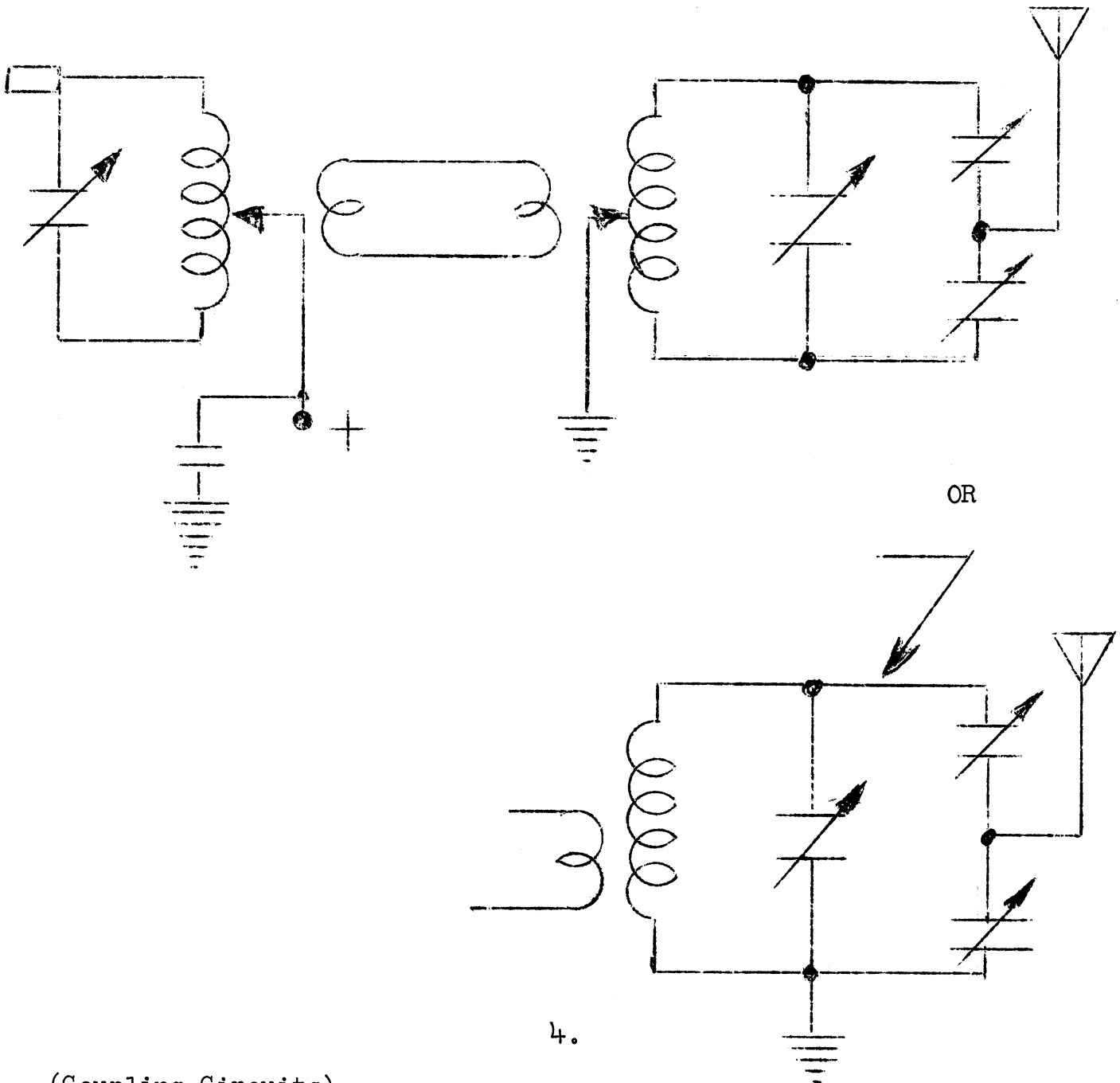
We now come to the consideration of the method to use for transferring power from the final tank circuit to the antenna system and to do this at the fundamental frequency with as much suppression of harmonic components as will meet the minimum FCC requirement of 40 db down from the fundamental or as much down below 40 db as may be required in a given case. Table A shows that the suppression of the second harmonic cannot be sufficient in a single tank circuit even with a Q of 20. Therefore we must follow the final tank circuit with another of comparable Q. If necessary three circuits can be used. In addition, Labus and Roder* point out that harmonic transfer can be reduced to nearly zero when a tank circuit similar to the familiar push-pull circuit is used for a single-ended final and that instead of by-passing the capacitor to ground, a point near the center turn of the inductance is located

*Reduction Harmonics, IRE June '31³.

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for by-passing to ground at which point nearly complete cancellation of the harmonic components is possible.

Such an arrangement has been used in actual practice. The harmonic component transfer to the next stage and from that stage to the antenna was thereby reduced to a negligible quantity. Links using coaxial cable were found to be most practicable and satisfactory.



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TABLE A

Harmonic level in db due to 140° tube current pulse for $3/2$ power* law and harmonic order:

$$N = \frac{2}{3.2} \quad \text{db} \quad \frac{3}{10.3} \quad \text{db} \quad \frac{4}{25.8} \quad \text{db}$$

Harmonic reduction in db due to one tank circuit of

<u>Q</u>	<u>N=2</u>		<u>N=3</u>		<u>N=4</u>	
5	23.5	db	32.0	db	37.5	db
10	29.6	db	38.1	db	43.5	db
15	33.0	db	41.6	db	47.0	db
20	35.6	db	44.1	db	49.6	db

Harmonic reduction in db due to two tank circuits of equal Q

<u>Q</u>	<u>N=2</u>		<u>N=3</u>		<u>N=4</u>	
5	35.0	db	44.1	db	51.0	db
10	47.0	db	57.1	db	63.0	db
15	54.1	db	64.8	db	70.4	db
20	59.1	db	69.1	db	75.0	db

TABLE B

Harmonic reduction in db due to 140° tube current pulse and one tank circuit of

<u>Q</u>	<u>N=2</u>		<u>N=3</u>		<u>N=4</u>	
5	26.7	db	42.3	db	63.3	db
10	32.8	db	48.4	db	69.3	db
15	36.2	db	51.9	db	72.8	db
20	38.8	db	54.4	db	75.4	db

Harmonic reduction in db due to 140° tube current pulse and 2 tank circuits of equal Q.

<u>Q</u>	<u>N=2</u>		<u>N=3</u>		<u>N=4</u>	
5	38.2	db	54.4	db	76.8	db
10	50.2	db	67.4	db	88.8	db
15	57.3	db	75.1	db	96.2	db
20	62.3	db	79.4	db	100.8	db

* Terman

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SCREEN AND GRID MODULATION OF EIMAC TETRODES

Because of the popularity of simple methods of modulation, especially in mobile applications where power consumption may present a problem, the applicability of a combination of controlled carrier, screen and grid modulation to Eimac Tetrodes was investigated. The results are gratifying, 100% modulation with negligible distortion being possible with the circuit arrangement shown in the circuit diagrams.

Adjustment is simple. The variable resistor in the cathode of the 6L6 is adjusted so that the screen voltage of the tetrode is at the normal Class C typical operating voltage. During modulation this voltage will rise to about 120% of the normal Class C typical operating voltage, thus giving rise to controlled carrier conditions. The only other adjustment for 100% modulation is to reduce the grid drive to approximately 60% of the normal grid current requirement.

Some screen voltage reduction may be necessary in order that it may swing from nearly zero to twice the indicated dc voltage during modulation along the linear portion of the screen voltage characteristic curve. Adjust the tap on the tank coil to obtain desired results. This series condenser aids the modulation effect.

In order that the screen voltage may swing down toward zero during modulation, it is necessary that the grid voltage also swing downward at the same instant. The 1/10 microfarad capacitor connected between the screen and grid as shown in the diagram accomplishes this, and the result is practically distortionless 100% modulation.

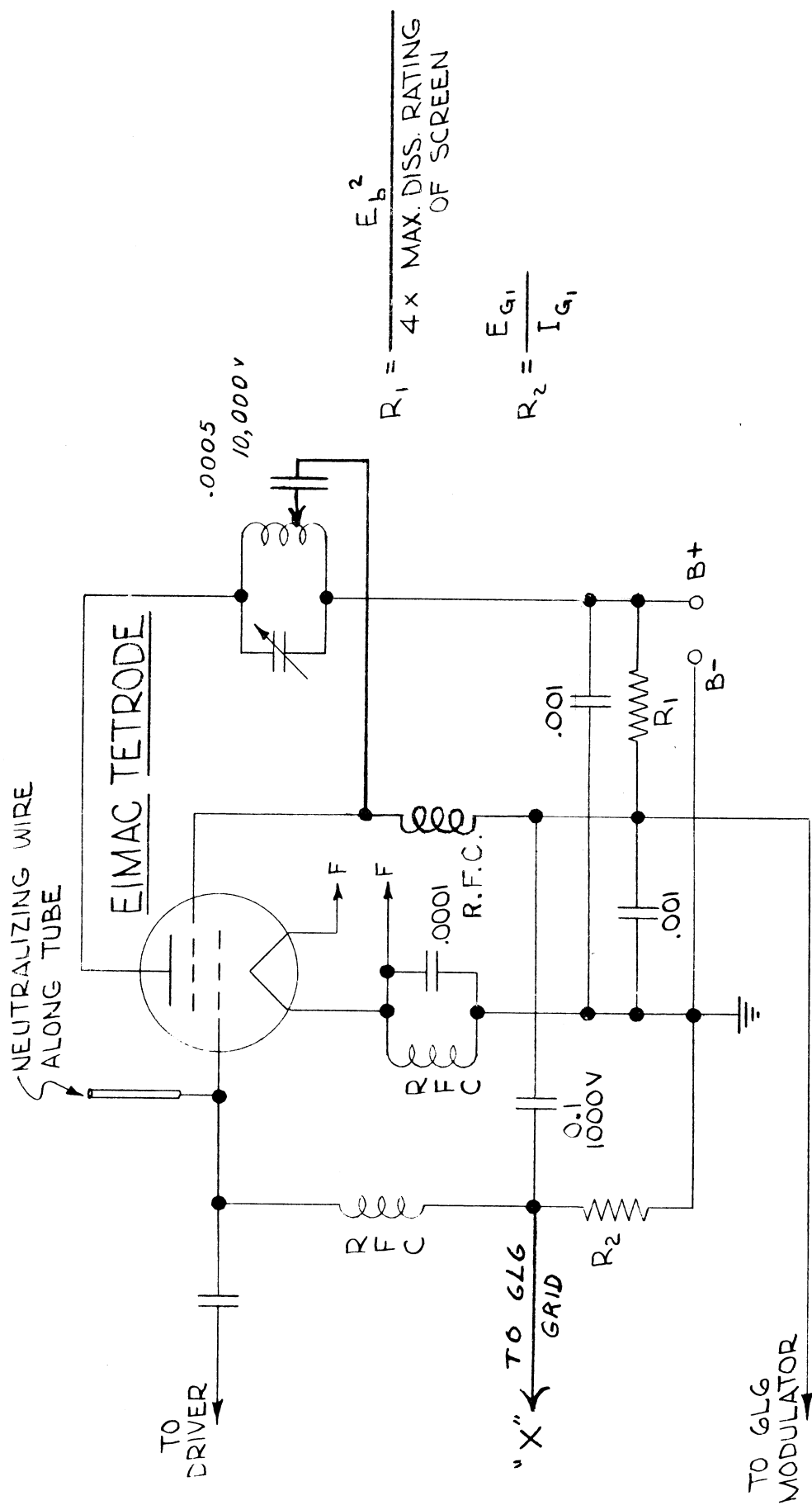
Plate voltage values up to the maximum Class C telegraphy conditions allowable for the tetrode in question may be used. Loading should be heavy but should be adjusted so that without modulation applied, the plate dissipation rating is not exceeded.

We are extremely well satisfied with the results of this investigation to determine if the modulation system devised would be applicable to our tetrodes. You also will be well satisfied should you use it.

John L. Reinartz (K6BJ)
Oct. 14, 1952
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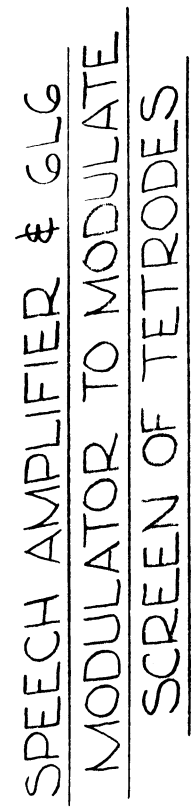
Attach Voice picture
Combination
Speech





COMBINATION SCREEN &
GRID MODULATION

$\frac{R}{K6B-1}$


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- - - - -

THE TUNA FISH CAN RF BRIDGE VOLTMETER

Or

OHMS - REACTIVE AND OTHERWISE

By - John L. Reinartz, K6BJ
- - - - -

Would you like to add a simple instrument to your collection with which you can ascertain resonance, the impedance of a circuit containing resistance and reactance: the reactance of a condenser or coil as well as resistance values within its range? Then build yourself a tuna fish can rf bridge voltmeter.

For instance, in tank circuit calculations we can use a sufficiently accurate yet simple formula to obtain the values of the capacitative or inductive reactance for resonance and need to know only the plate voltage to be used, the power output desired and the desired Q of the circuit. The form of this ~~ex~~pression is:

$$\text{For triodes: } \frac{(.63 E_b)^2}{Q P_o} = X_L \text{ or } X_c$$

$$\text{For tetrodes: } \frac{(.57 E_b)^2}{Q P_o} = X_L \text{ or } X_c$$

Where (.63 E_b) equals the RMS plate voltage swing, E_b equals the dc plate voltage, Q equals the desired circuit Q and P_o equals the desired watt output.

From the attached reactance table we can very readily determine both the capacitance value and the inductance value at the desired operating frequency, needing merely the reactance value obtained from the values used in the formula.

To measure the reactance of a condenser or coil, we can use the simple voltmeter bridge circuit shown in Fig. 1.

Because this bridge measures ohms or impedance and since the resistance is negligible in a good condenser or a small inductance of good design, the value read off the dial at a specific driving frequency is the reactance ohm value. When a series resonant circuit is measured at its resonant frequency, the value read off the dial is the resistive component. Only when a resistor is being measured that is non-reactive, will the reading of the dial be the same for any driver frequency up to 30 mc., the bridge accuracy limit.

The characteristic impedance of a line less than one-half wave length long at the highest measuring frequency may readily be determined if previously unknown. It is but necessary to connect one end to the bridge and a variable carbon potentiometer to the other end. When the voltmeter no longer changes its zero reading with a 2 to 1 change in driver frequency, the bridge will be in balance and the dial reading will be the Z of the line, as will be the setting of the carbon potentiometer at the end of the line.

A quarter wave mobile antenna may be measured for resonance, also its resistive component at resonance. We simply connect our bridge between the base of the antenna and ground, or the point at which the coax, when used, is connected into the antenna system. We can check the driver frequency at the car receiver if the rf driver frequency is not accurately calibrated.

Calibration of the bridge can be by the use of an ohmmeter - rotation of the dial will indicate the resistance on the ohmmeter. If a zero to 100 ohm potentiometer is used in the bridge, calibration can be for every 5 ohms. When a 500 or 1000 ohm potentiometer is used, calibration may be at 25 ohm points. If values under 100 ohms are expected to be measured, a 100 ohm potentiometer is suggested otherwise, a 500 ohm potentiometer will serve most purposes.

Initial balance with a resistor of known value connected to the bridge after calibration can be accomplished through the addition of a 1000 to 5000 ohm resistor placed across any one of the points marked with an X, since it cannot be determined beforehand how the diodes and resistors will balance.

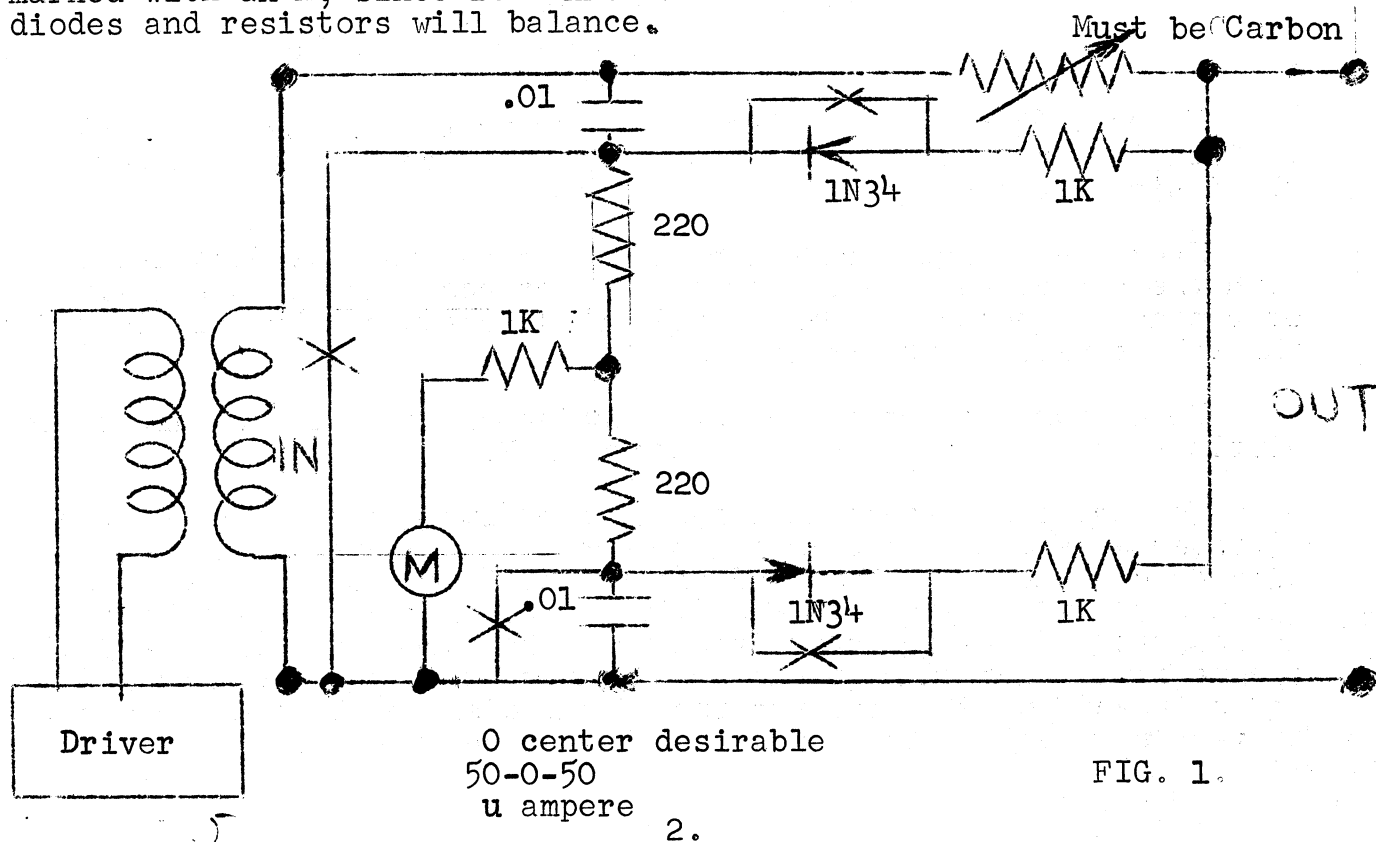


FIG. 1.

(rf bridge)

TETRODES

$$X = \frac{(.57 E_b)^2}{Q P_o}$$

<u>E_b</u>	<u>P. In.</u>	<u>P. Out.</u>	<u>Efficiency</u>	<u>Q</u>	<u>Ohms Reactance</u>
1000	133	100	75	12	270
1500	267	200	75	12	305
2000	536	400	75	12	270
2500	800	600	75	12	282
3000	1000	800	80	12	305

An average value of 300 would serve.

1000	133	100	75	15	217
1500	267	200	75	15	243
2000	536	400	75	15	217
2500	800	600	75	15	225
3000	1000	800	80	15	243

An average value of 225 would serve.

TRIODES

$$X = \frac{(.63 E_b)^2}{Q P_o}$$

<u>E_b</u>	<u>P. In.</u>	<u>P. Out.</u>	<u>Efficiency</u>	<u>Q</u>	<u>Ohms Reactane</u>
1000	133	100	75	12	330
1500	267	200	75	12	372
2000	536	400	75	12	330
2500	800	600	75	12	345
3000	1000	800	80	12	372

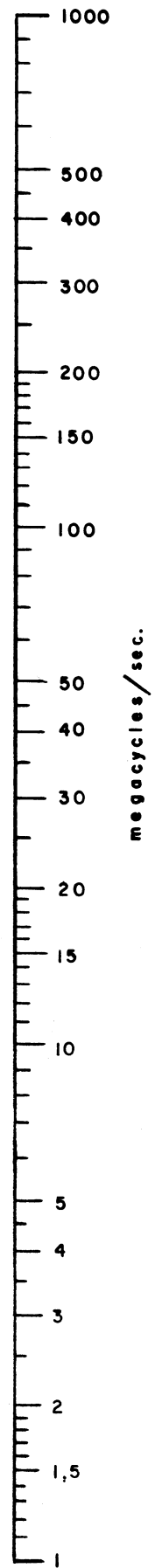
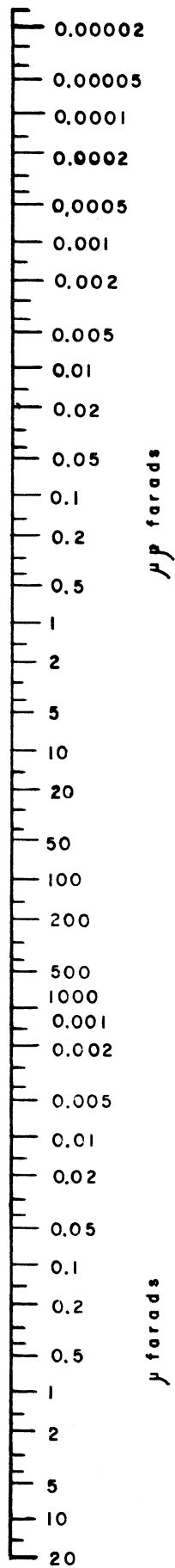
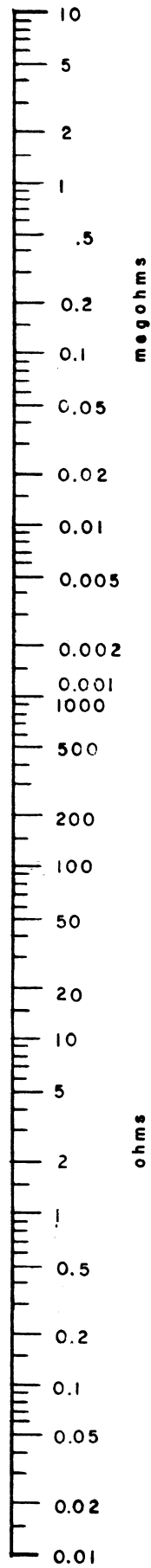
An average value of 350 would serve.

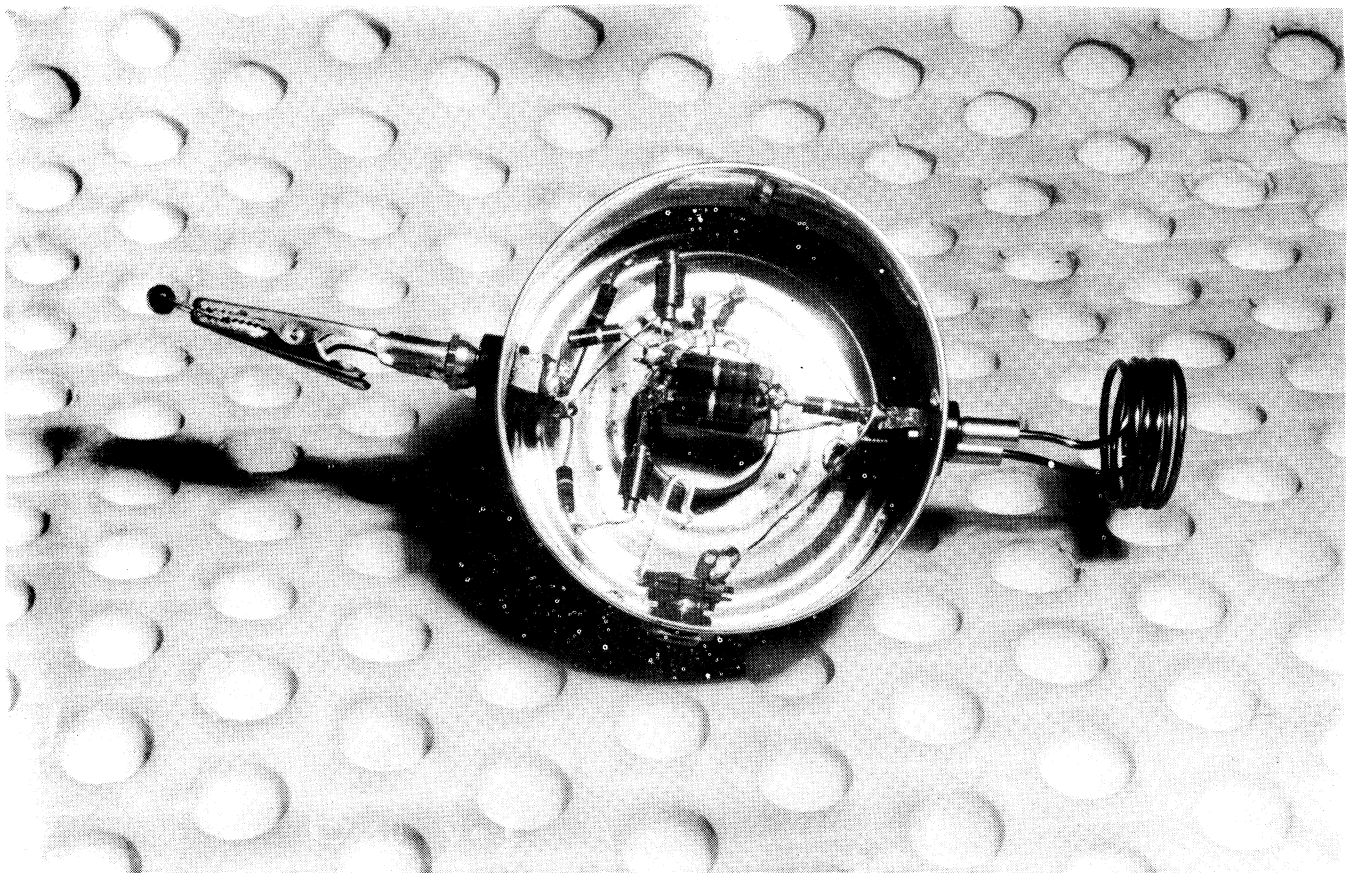
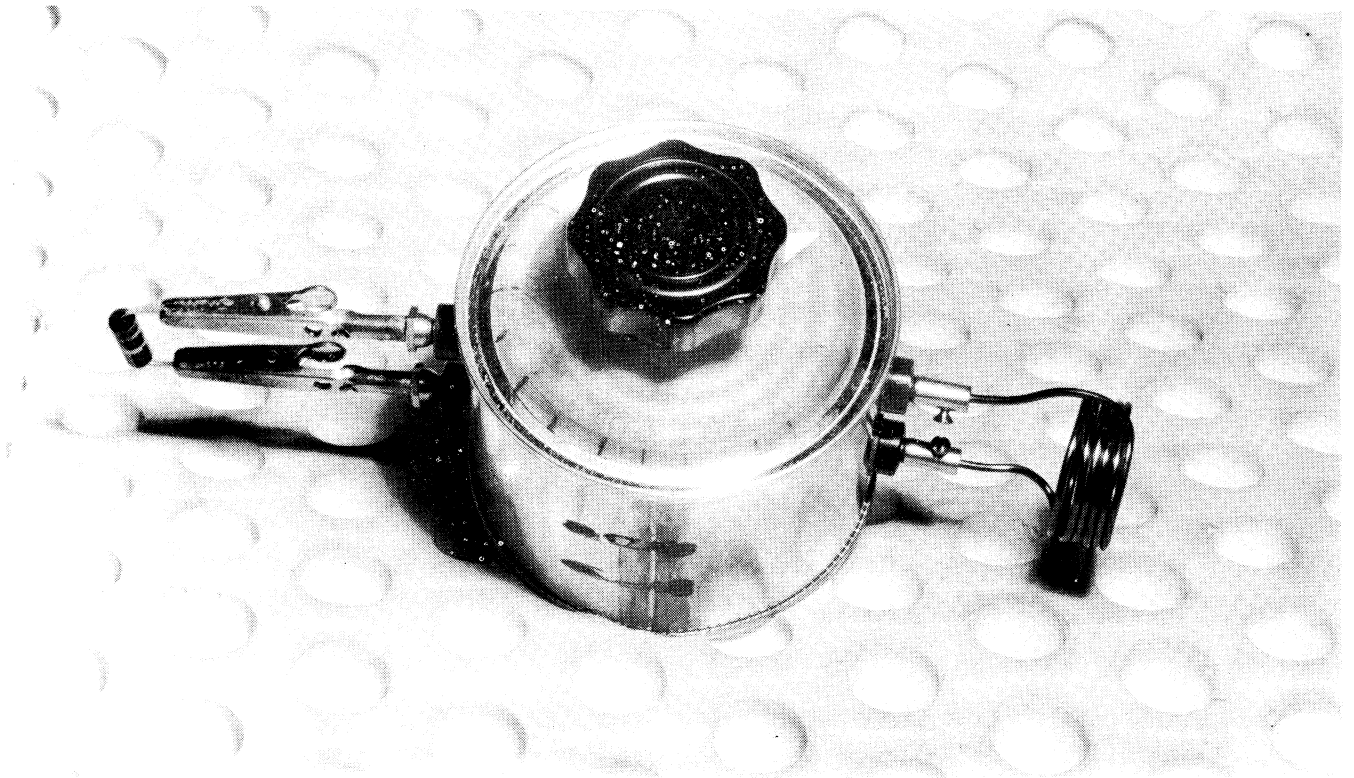
1000	133	100	75	15	263
1500	267	200	75	15	297
2000	536	400	75	15	263
2500	800	600	75	15	276
3000	1000	800	80	15	297

An average value of 275 would serve.

3.

Eitel-McCullough, Inc
Home of Quality Tubes
San Bruno, California
(Rf Bridge)





EITEL-McCULLOUGH, INC.
San Bruno California

In audio calculation in its simplest form, the method to use is to ascertain the turns or voltage ratio, secondary to one-half primary of the modulation transformer, and to understand that the voltage output at the secondary of the transformer must equal the dc plate voltage applied to the rf stage for 100% modulation.

Because the audio tubes work one at a time, only one-half of the primary winding of the audio transformer is working at that time. If there were no losses in the transformer, and if the ac swing generated by each modulator tube were equal to the applied dc voltage, the voltage output at the modulation transformer secondary would understandably then be a function of the applied dc voltage to the plate and of the turns ratio of one-half primary to secondary.

Because the tubes and transformer are not perfect, the losses must be made up and we accomplish this by an increase in the number of turns in the secondary with respect to the turns in each half of the primary. Therefore, if we assume a modulation transformer efficiency at full load of 85%, we need to increase the turns ratio, secondary to one-half primary. Also, since the ac swing with respect to the dc voltage applied to the plate in Class AB₁ and AB₂ can be taken as 60% and 85% respectively, we again must increase the turns ratio, secondary to one-half primary.

The required turns or voltage ratio, secondary to one-half primary for any values of dc voltage applied to the rf, or audio stages can be ascertained from:

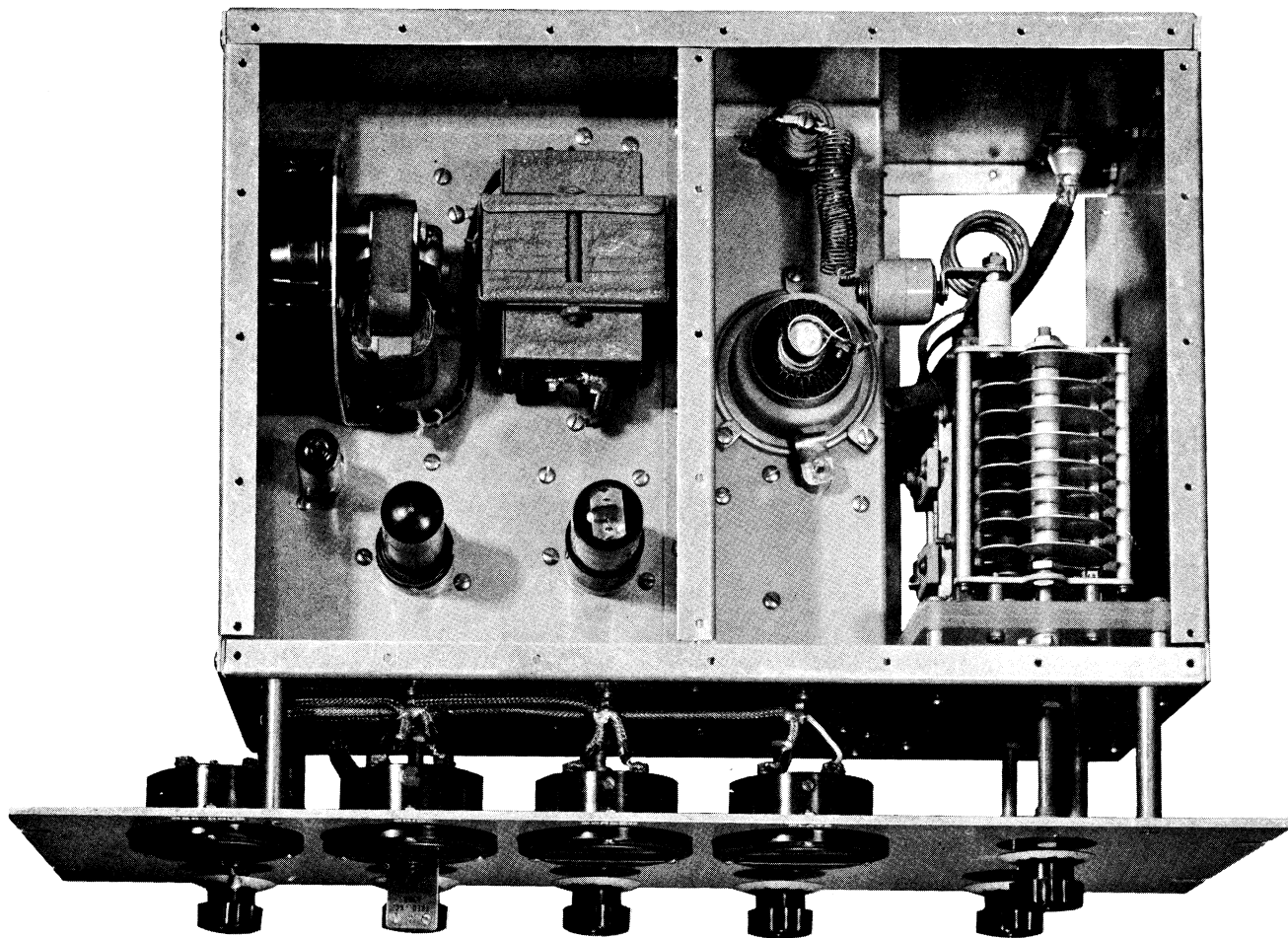
$$\begin{aligned}\text{Turns ratio} &= \frac{E_b \text{ rf stage}}{E_b \text{ mod.} \times .85 \times .6} && \text{for AB}_1 \quad \text{and} \\ &= \frac{E_b \text{ rf stage}}{E_b \text{ mod.} \times .85 \times .85} && \text{for AB}_2\end{aligned}$$

Where the rating plate on the transformer gives the impedance ratio, then the turns ratio is the square root of the impedance ratio. When neither this value nor the turns or voltage ratio are given, the voltage ratio at no load can be determined by the application of any reasonable ac voltage to one-half of the primary, and then measuring the output voltage of the secondary. The turns ratio is the same as the voltage ratio.

The output watt rating of the transformer should be equal to one-half of the power input to the rf stage for 100% modulation.

John L. Reinartz, K6BJ

Eitel-McCullough, Inc.
San Bruno, California
Home of Quality Tubes



Top View: In this illustration the blower and filament transformer locations are seen to be in the left hand shielded compartment at the rear of the chassis. In the same compartment, from left to right, are the 0A2 regulator, 6AG7 oscillator and 6V6 doubler. The 4CX300A amplifier is in the right-hand compartment, along with the plate circuit components.

Compact, Powerful 6-METER RIG

Uses Eimac Ceramic 4CX300A



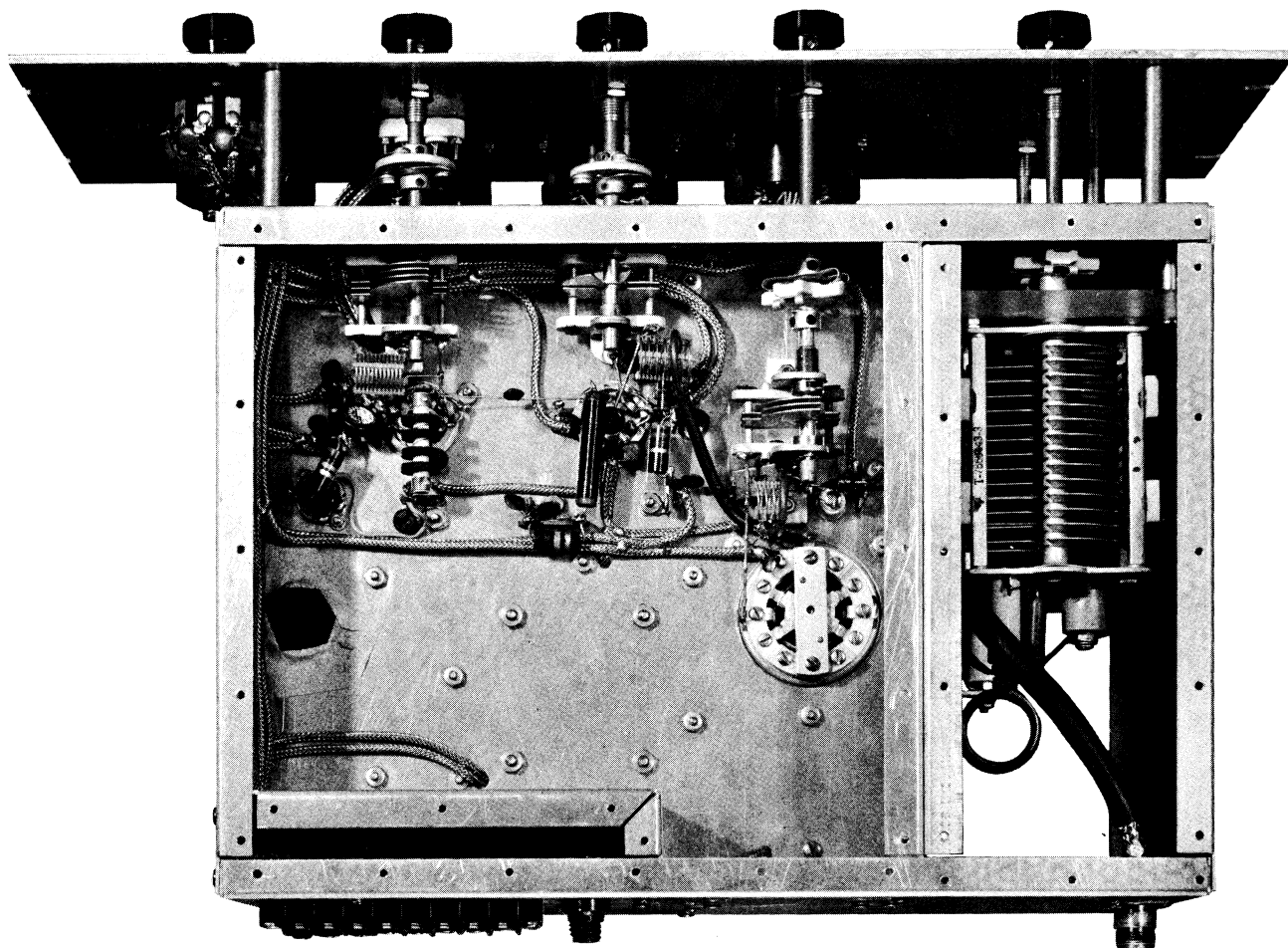
By RAY RINAUDO, W6KEV

The transmitter shown in the photographs is a three stage crystal controlled RF unit capable of running 625 watts input CW or 300 watts AM phone on 50 mc. The tube line-up is a 6AG7 oscillator-multiplier, 6V6 doubler, and a 4CX300A amplifier.

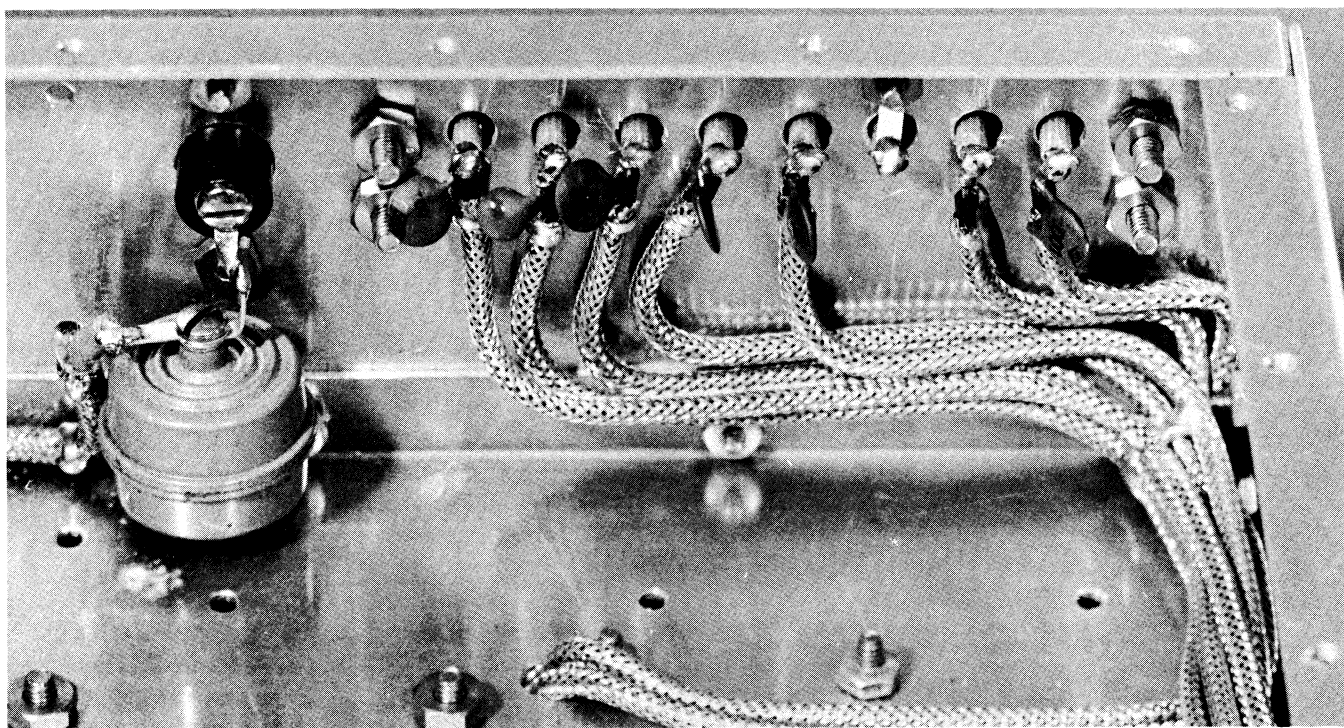
The 6AG7 crystal oscillator is used in a modified Pierce circuit with the plate circuit tuned to 25 mc. Crystals in either the 6 or 8 mc range work satisfactorily without circuit changes. The screen voltage is regulated at 150 volts by means of a 0A2. The 25 mc output is capacity coupled to the following doubler.

The doubler stage uses a 6V6GT in a conventional circuit. The screen voltage is adjustable from

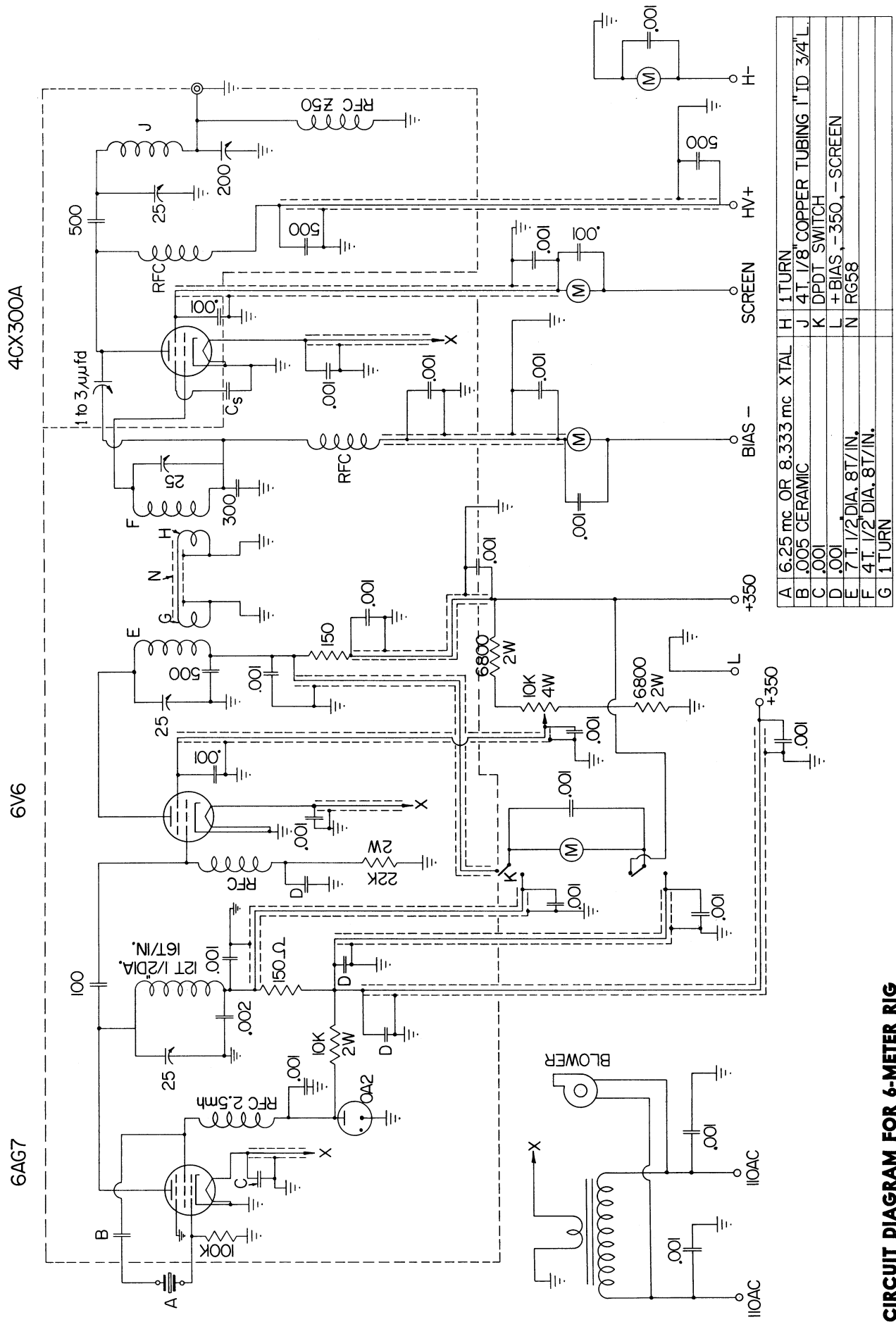
(Continued on page 4)



Bottom View: The tuning condensers visible in this view, from left to right, are oscillator plate circuit, doubler plate circuit, final amplifier grid circuit and the final amplifier plate loading condenser. The tube sockets are for the 0A2, 6AG7, 6V6 and the 4CX300A. The blower exhaust into the chassis is at the extreme left. The small shield in the lower left hand corner of the chassis encloses the terminal strip connections.



Bottom View—Terminal Strip: By-passing details of power input terminal strip. The high voltage terminal and associated by-pass condenser are shown at left.



CIRCUIT DIAGRAM FOR 6-METER RIG

the front of the panel by a potentiometer; by this means the drive to the final amplifier is controlled. The output from the plate circuit is transferred to the grid circuit of the final amplifier by link coupling through a short length of RG-58U. Inductive coupling through two tuned circuits is necessary in order to minimize the coupling of 25 mc, 75 mc and higher harmonics to the final amplifier grid.

The 4CX300A final amplifier stage uses a pi network plate circuit. The tube is neutralized by means of a bridge circuit which is completed to ground by the 200 mmfd tuned grid circuit "by-pass" condenser. The neutralizing capacity is obtained by presenting a $\frac{1}{2}$ inch by $\frac{1}{2}$ inch surface adjacent to the anode of the tube, the separation being about $\frac{3}{8}$ inch. This condenser can be seen in the top view of the transmitter, appearing as a small bracket next to the 4CX300A and toward the front panel. The 4CX300A is mounted in an Eimac SK-710 socket, a socket which has a built-in screen by-pass condenser and is designed for the tube.

The entire RF unit is enclosed in an aluminum box measuring 14 inches wide by 10 inches deep by $6\frac{1}{2}$ inches high. It is supported from a standard 7 by 19 inch panel with four 2-inch long pillars, thus allowing the mounting of meters outside of the RF fields. The top of the box and the small portion of the bottom directly below the final amplifier plate tuning condensers are covered by perforated aluminum sheet. The chassis portion has a solid aluminum sheet bottom cover making possible pressurization by the blower, the air escaping through the SK-710 socket and the Eimac SK-606 chimney and cooling the 4CX300A anode. A small shield has been placed in one corner of the enclosure in order to prevent the radiation of harmonics via the terminal strip connections; a close-up view of the under side of the chassis with the corner shield removed shows the details of the by-passing of the terminal strip leads.

The blower used for cooling is a home brew device. An L-R No. 2 squirrel cage blower is driven by a phonograph motor which turns at about 3300 RPM. The unit moves enough air to adequately cool the tube up to about 150 watts plate dissipation, but would be inadequate for the full 300 watt rating. However in neither the keyed CW mode (625 watts input) or the AM mode (300 watts input) does the average plate dissipation exceed 100 watts.

For AM phone the final amplifier is plate modulated. As with other tetrodes it is necessary to provide modulating voltage to the screen. A choke of 10

henrys or more in series with the screen supply circuit does give adequate modulation but the maximum is somewhat less than 100 percent. Correct operation can be achieved only by applying audio voltage directly to the screen from a tap on the modulation transformer or from a separate low power modulator capable of up to 5 watts output. A single 6V6 will supply more than enough power. It is important that the screen and plate modulation be correctly phased! A simple reversal of the leads from the screen or plate modulator (but not both) will indicate the proper phasing.

Four meters are used to indicate the currents drawn by the various stages. One measures the plate current of either the oscillator or the doubler: a double pole double throw switch connects the meter to the desired circuit. Individual meters are used in the grid, screen and plate circuits of the final amplifier. It will be noticed that plate current is measured in the negative lead from the high voltage supply. This requires that the negative lead of the high voltage supply not be connected to ground at any point in the power supply.

Although the oscillator and the doubler stages operate at the same plate voltage, separate terminal strip power connections are provided. This is done so that a suitable keying point is available for CW operation. Keying is done by breaking the plate and screen voltage to the doubler, allowing the oscillator to run continuously; breaking the circuit must be done using a suitable filter in order to suppress key clicks.

Three power supplies are needed. A 350V supply capable of 125 ma will power the oscillator, doubler and the amplifier screen—the correct screen voltage being obtained through a series dropping resistor. The amplifier plate supply can be at any voltage up to 1500 for phone and up to 2500 for CW at a current of 250 ma. The amplifier grid bias supply should apply -90 volts to the grid, and should be capable of withstanding 30 ma of grid current without serious loss of voltage regulation.

It is important that screen voltage never be applied to the amplifier unless plate and grid bias voltages are also on the tube. It is permissible to apply or remove all voltages simultaneously. Similarly the amplifier should never be operated lightly loaded unless the drive power is also reduced. The above applies to all tetrodes and pentodes and failure to observe these precautions will cause excessive screen dissipation, resulting in very short tube life.



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