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Research and Development Technical Report
ECOM-2836

HIGH-EFFICIENCY TRANSISTOR CW RF POWER AMPLIFIER

by

D. R. Lohrmann

May 1967

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TECHNICAL REPORT ECOM-2836

HIGH-EFFICIENCY TRANSISTOR CW RF POWER AMPLIFIER

By

D. R. Lohrmann

DA Task No. 1E6 34301 D246 01

MAY 1967

COMMUNICATIONS/ADP LABORATORY

ABSTRACT

Research work on VHF transistor power amplifiers of the switched type demonstrates semi broadband designs yielding collector efficiencies as high as 80% at 40W output in the frequency range from 30 to 76 MHz. A laboratory model of such an amplifier was built. It was designed to fit the AN/VRC-12.

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*Cy of Article written by Mr. D.R. Lohrmann, entitled: Amplifier has 85% Efficiency While Providing up to 10W Power over a Wide Frequency Band, March 1966 Issue of the Electronic Design Magazine.

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HIGH-EFFICIENCY TRANSISTOR CW RF POWER AMPLIFIER

I. INTRODUCTION

Recent progress in semiconductor development made available transistors which enabled us to develop high-efficiency power transmitter circuits in the VHF range.

II. PURPOSE

The purpose of the work was to find ways to build RF CW transistor power amplifiers with efficiencies as high as possible. This is especially important for portable sets, where one tends to save weight of batteries and heatsinks.

III. DISCUSSION

1. General Considerations

The approach taken is a semi-broadband mode, in which the transistor is used as a switch. Therefore, the principle is confined to use in CW amplifiers, like in FM sets.

Collector efficiencies up to 80% at 40W output at 10 db gain have been observed, see fig. 1 and fig. 2.

The transistor used was #5067 by TRW Semiconductors Inc. (TRWS). It had been developed under contract DA 28-043-AMC-02162(E).

The main problem out of the many that had to be solved, was parametric oscillations on half signal-frequency or lower. They are caused by the nonlinear collector to base and base to emitter capacities, which are swept from maximum to minimum and back at each cycle of the signal. One way the effect can easily be eliminated is to insert attenuating resistors between collector and emitter of the transistor, however, this results in reduced efficiency and hence other ways were sought and found to overcome the problem. Unfortunately, there is no reasonably useful theory on the instability of circuits containing nonlinear reactances at present, however, we succeeded in making progress towards a preliminary solution to the problem.

2. Advantages found in using this amplifier in VRC-12

2.1 Reduction of battery power

The battery power drawn was reduced from 47W down to 17.8W in receive and from 225W down to 110W in the transmit mode. This reduction is particularly valuable if the set is to be used as a relay station. The reason why the power in receive is cut down so much also

is that the current set uses two indirectly heated tubes in the power amplifier of the transmitter, which have to be kept heated in standby during receive in order to be ready instantaneously for transmit.

2.2 Increased reliability

No tubes would have to be changed any more in the power amplifier. In the old VRC-12, one trouble spot had been the high voltage supply. This would be eliminated, adding greatly to the reliability of the set.

2.3 Continuous transmit duty possible, even at 67°C

2.4 The fan noise has been eliminated with use under normal environment conditions

2.5 Modernizing present sets

VRC-12 sets currently being used could be modernized into models with transistorized transmitter power amplifiers by exchanging subassemblies. This could even be done in the field. The only tool required would be a screwdriver. If the transmit PA module of VRC-12 should be transistorized the procedure in changing would be as follows:

a. Take out old transmit PA module and screw in new one.

b. Take out old power supply module and screw in new one.

c. Take out relay K406 and screw in dummy.

d. The low pass filter in the antenna line is removed.

e. Screw in additional isolation amplifier, containing a transistor 2N918 and some components. (Schematic of isolation amplifier, see figure 4.)

The reason for the latter is that the injection of the transmit RF into the PA module carries a spurious oscillator signal, which is 11.5 MHz off the signal frequency. Formerly, the PA contained additional tuned circuits, which eliminated this spur. However, in our broadband approach we therefore had to solve the problem in a different way by inserting a broadband isolation amplifier stage between transmit driver and HF tuner in the VRC-12. (Schematic of amplifier stage, see figure 5.) See also changes in overall circuits, figure 6.

The necessary connections can be made easily without using a solder gun.

3. Description of the amplifier

3.1 General

The amplifier amplifies 100 W of rf input power up to 32W output power ± 1db.

The frequency range from 30 to 76 MHz is covered in three bands, from 30 to 40, 40 to 53 and 53 to 76 MHz. There are no tuning elements like tuning capacitors or tuning coils in the setup. The nominal battery voltage used is 25.5V.

3.2 The predriver stage, see fig. 5

It is equipped with transistor 2N3866 by RCA. Switch S_1 reduces the input power in the two lower bands, since the final power amplifier stage needs 4W of drive power only in the highest band, and less in the lower ones.

In the collector line there is a broadband transformer. It matches the collector of the 2N3866 to the input impedance of the following stage which is approximately 8 Ω . This next stage requires about 1W of input power.

For a 17V peak voltage swing on the collector and for 1W of output power, the transformer has to transform the load impedance of 8 Ω up to 150 Ω .

The transformer was built with a Q 3 Ferrite Cup Core. There were five turns of #18 magnet wire on the primary and one turn of copper foil on the secondary. The copper foil was 1/4" wide; it was wrapped directly on the cylindrical primary coil. The coupling factor achieved that way was .85.

The resistor of 1.5K ohms in parallel to the collector of the 2N3866, as well as the 5 and 10 Ω resistors at the input of the next stage, was necessary to prevent parametric oscillations appearing on 1/2 signal frequency and lower.

3.3 The driver stage

The driver stage uses a transistor 2N3632 by RCA. It delivers between 3 and 4.5W driving power to the power stage. The transistor is self-biased.

First, we had tried to design this stage broadbanded also using 2N3375. However, we could not safely eliminate oscillations on subharmonics, which occurred on several points in the frequency range. The amplifier worked very well into a broadband 50 Ω real load, however,

when it was connected to the final stage these oscillations appeared. At a given frequency the effect could be made to disappear if the cable length between driver and final stage was varied. It seems possible that the nonlinear reactance in the base emitter circuit of the PA stage was involved in addition.

Since the theory indicated that this effect of subharmonics can be eliminated by showing a very low impedance on one half the signal frequency or lower frequencies to the pumped collector or base capacity respectively, we used a bandfilter for each of the three output ranges in the collector circuit of the driver.

The filters were designed as one section π types, the input parallel coil of the filter serving as a feeding line for the DC to the collector. Thereby the coupling capacitor could be eliminated also.

3.4 The power amplifier stage

The power amplifier is equipped with a transistor 5067 by TRWS. As mentioned, this transistor was developed by TRWS, with a design goal of B_{VCES} = 100V, 50W output power at 76 MHz with 80% collector efficiency at 10 db gain.

3.4.1 The input matching circuit

The input matching circuit is broadbanded. It matches the 50Ω impedance of a driving source to the approximately .9Ω input impedance of the transistor 5067. The input impedance of the base lead of the 5067, at a distance 1/8" from the case was measured under operational conditions. This was done by measuring voltage at the base lead, the current going to the base, and the phase angle between them. For achieving this, a Tektronix sampling scope was used, which was synchronized externally. From a signal source, capable of delivering a maximum of 40W the current was impressed to the base thru a 50 nH coil. From measurements of absolute value and phase angle of the voltage at the input side of the coil and at the base side of the coil the input impedance was calculated. One has to see to it that the ground connection of the probe at the base is very short, and very close to the point of measurement at the base. Since the wave form is distorted, low pass filters were used before entering the scope.

The results are plotted in figure 7. The Smith chart is referenced to 1Ω. One can see that the input impedance is in the order of 1Ω, and in the frequency range between 30 and 76 MHz it behaves approximately like .9Ω real in series with a 4000 pF capacitor and a 3.2 nH inductivity.

The transformer uses a 20K 12 ferrite cup core, airgapped to result in 20nH/turn square. The primary of this transformer consists of 3 turns of mylar coated copper foil, the foil being 1mm wide. On top of these three turns, which were supported by a bobbin, one turn of copper tape, 5 mm wide was wrapped around. This way, a stray factor of 20% can be achieved, or, in other words a coupling factor of .9 is obtained.

The transformer transforms down to about 3 ohms; the broadband transforming network which follows transforms these 3 ohms down to .9 ohm.

For a summary on transformers see appendix of this report. Figure 8 shows the input matching circuit for the #5067 used, and its equivalent circuit. Mechanically, it is very simple and inexpensive. Functionwise electrically, it is not quite so simple. Parallel and stray inductance of the transformer are combined into parts of a general broadband transformation circuit, see figure 9, [1] .

The network transforms and has a bandpass characteristic, having f_u, f_l, and f_m, as upper, lower and center frequencies.

For an impedance transformation of $R_1/R_2 = n = 4$, we read from figure 9:

$$\frac{f_u - f_l}{f_m} = .6$$

resulting in about 20% mismatch at the frequency limits. If $f_u = 76$ MHz, and $f_m = \sqrt{f_l \cdot f_u}$ we get $f_m = 56.6$ MHz, and $f_l = 42$ MHz.

Since our band goes down to 30 MHz, this will give us some increased mismatch on the lower frequency end. We can tolerate this however, because on the lower frequency end we need much less driving power (-3db) anyhow.

R_2, which constitutes the series input resistance of the transistor, was measured to be about .9 ohm; in the actual circuit it seems to be closer to 1.2 ohms. From [1] , figure 8 and figure 9 we see that

$$X_1 = 2\pi f_m L_1 = R_2 \sqrt{n - 1} = 1.2\ \Omega$$

$$\underline{L_1 = 3.5\ \text{nH}}\ .$$

In the actual circuit, this inductivity is in the base lead of the transistor. Therefore it does not appear in the circuit diagram. The series capacity of 4000pF in the equivalent circuit of the transistor input impedance is neglected. Then we get the other values to be

$$X_2 = 1/2\,\pi\, f_m\, C_2 = R_2 \sqrt{\frac{n}{\sqrt{n}-1}} = R_2 \cdot 2 = 2.4\,\Omega$$

$$C_2 = 1170 \text{ pF}$$

$$X_3 = 1/2\,\pi\, f_m\, C_3 = R_2 \sqrt{n(\sqrt{n}-1)}$$

$$C_3 = 1170 \text{ pF}$$

$$X_4 = R_2 \frac{n}{\sqrt{\sqrt{n}-1}} = 2\,\pi\, f_m \cdot L_4$$

$$L_4 = 14 \text{ nH}.$$

The 1100pF capacitors were made out of a 2200pF ceramics capacitor by Mucom, which contains two 1100pF capacitors sandwiched together and paralleled. The wire paralleling the top electrodes of the two chips was cut, this way rendering the two 1100pF capacitors in series. Then the coating of the capacitor was removed on one side, and subsequently soldered with this surface to the ground plain.

3.4.2 Output Circuit

The output circuit is designed as described in [2] see appendix. The DC current is supplied to the collector via three chokes, one for each band. Care must be taken to insure that the collector at no frequency within the band sees a short circuit on the second harmonic. Otherwise at that frequency the power output and the efficiency will be greatly diminished. The same is valid for the base circuit. Seen from the base into the matching circuit there should not be a short circuit on the second harmonic.

For load impedance measured at the collector see figure 11. The ouput power specified is P_{out}=40W, the battery voltage is V_B = 24.4V, and the saturation voltage is assumed to be V_{CESat}=3V.

Then from [2] we have $R_\omega = \frac{.85\,(V_B - V_{CESat})^2}{P_{out}} \approx 10\,\Omega$.

This is the impedance the collector has to see on the fundamental. Therefore, the transformer behind the collector - low pass filter section has to transform from $10\,\Omega$ up to $50\,\Omega$. An example for calculating the low pass filter section following the collector is given for band 3, 53....76 MHz, see [2].

$$Z_\pi = 1.1\,R_\omega = 11\,\Omega$$

$$f_g = 1.5\,f_u \quad . \quad \text{With } f_u = 76 \text{ MHz},$$

$$f_g = 114 \text{ MHz}$$

$$C = \frac{1}{2\pi f_g Z_\pi} = \frac{1}{2\pi \cdot 114 \cdot 10^6 \cdot 11} = \underline{127 \text{ pF}}\,.$$

The dynamic collector capacity of the transistor #5067 is about 110 pF. Therefore, only a small trimmer is required on the collector side of the π - filter section.

$$L = \frac{Z_\pi}{\pi \cdot f_g} = \frac{11\,\Omega}{\pi \cdot 114 \cdot 10^6\ 1/s} = \underline{30 \text{ nH}}$$

is the inductivity of this low pass filter section.

The capacitor on the output side of the filter was replaced by a series circuit, which provides a short circuit on double, mid band frequency, however exhibits a capacitive impedance of an equivalent of 127 pF at the mid band frequency. This has been found to improve the efficiency.

The transformer is coupled to this low pass section by a capacitor, which is designed to compensate the stray inductivity of this transformer in the center of the band.

The transformers in all three bands are cup cores of ferrite 20K 12, 14 millimeter diameter, airgapped to 20nH/turn^2. In the lowest band there are 6 turns of #22 wire, bifilar wound and tapped at 3 turns; in the center band 4 turns with tap at 2, and 5 turns with tap at 2 of #20 wire for the highest band.

These transformers are followed by low pass filters, which provide a total harmonic suppression of better than 75 db, see figure 10. The filters have a pass band attenuation of less than .7 db.

3.5 Protecting circuits

Figure 5 shows the protecting circuit. Basically, it acts by reducing the DC supply voltage for driver and power amplifier stages. The three power transistors (Silicon PNP 2N3791 by Motorola) in parallel act as the regulating unit. The whole circuit performs the following functions:

3.5.1 High current keying switch. (activated by the low current keying relays K401 in the frame.) When the transmitter is keyed, it supplies DC to the power amplifier.

3.5.2 Voltage regulator. Up to 26V battery voltage the transistors 2N3791 are switched in completely. Above that voltage, the output voltage is kept constant at 26.5V, see figure 12.

3.5.3 Transient suppressor. Any transient voltage spikes coming from the DC supply (up to 80V) are suppressed by the circuit.

3.5.4 Current limiter. If the output is loaded with more than 3.5 A, the output voltage degrades, the current rising to approximately 4 A in the short circuit condition. This is a further measure to protect the power amplifier, see figure 13.

3.5.5 Mismatch Protector - Between the low pass filters and the antenna terminal there is a directional coupler inserted, see figure 5. The DC output voltage of this sensor is proportional to the amount of RF power being reflected back from the antenna in case of mismatch, see figure 14. If the output voltage of the VSWR sensor exceeds a certain threshold, being set by potentiometer P2 in figure 5, it starts reducing the DC supply voltage of the driver and power amplifier stage. This way the amount of reflected RF power is kept at a constant value above a VSWR of 3.

3.5.6 Peak Voltage Protector - If the peak voltage on the collector of the power amplifier exceeds a given value, (85V), the DC driving voltage is reduced as well. The peak detector sensor is shown in figure 5. The level at which reduction of driving DC voltage starts can be set by potentiometer P3 of figure 5.

3.6 Influence of mismatch of antenna

As a necessary consequence of this high collector efficiency, rf wise the amplifier somehow acts like a stiff source; its internal impedance is by necessity not matched to the load. Therefore, variations of the antenna load cause much larger variations of the power

delivered to the load than in the "matched" case. In a sense, the amplifier behaves like a power outlet on the wall. If one reduces the internal impedance of, say a solder gun by a factor of three, it will draw three times more power. Similarly, if the antenna impedance is varied along the circle of VSWR = 3, the power delivered to the load will vary approximately between a minimum of 1/3 and a maximum of three times of that rated power, which it would deliver at a load having a VSWR of one. In other words, having an output power of 40W at a real antenna load impedance of 50 ohms, the amplifier will send a minimum of 13W and a maximum of 120W into the load, if we vary this load along a VSWR = 3 circle. Of course, the 120W point is not reached in practice, since the overload protector circuits start acting long before. Otherwise, the power transistor would be destroyed. See Figure 15 for a measurement of output power delivered to the load, where the load was moved on a VSWR=3 circle. Figure 15a shows calculated values.

Here a few remarks about the equivalent circuit of the transmitter output are added. The transmitter stage as described can always be represented as a source of EMF E, a series resistor R_o and a piece of transmission line, having characteristic impedance $Z_o = 50\,\Omega$ and length l, see figure 16. These parameters are vital for understanding the various mismatch conditions. They can be measured in the following way:

The amplifier is loaded with a real load R_2 which differs from the nominal load Z_o by a factor of 1.5...3, in our case, say with $17\,\Omega$. Between the amplifier and this load there is inserted a directional type power meter and a transmission line, having $Z_o = 50\,\Omega$ and being variable in length l. In varying l, the transmitter sees a load which varies along the circle VSWR = constant = 3. The power meter will measure a maximum forward going power $P_{f\,max}$ and a minimum forward going power $P_{f\,min}$. Then, the resistor R_o in equivalent circuit figure 16 is found

$$R_o = Z_o \cdot \frac{1 - \frac{R_2}{Z_o}\sqrt{\frac{P_{f\,max}}{P_{f\,min}}}}{\sqrt{\frac{P_{f\,max}}{P_{f\,min}}} - \frac{R_2}{Z_o}} .$$

$P_{f\,max}$ and $P_{f\,min}$ we find from figure 15, however $P_{f\,max}$ must be referred to the nominal battery voltage of 25.5V that is, without the action of the DC protecting circuit.

$$P_{f\,min} = 21W; \qquad P_{f\,max} = 63W \cdot \frac{25.5V^2}{19V^2} = 113W.$$

R_2 was $17\,\Omega$, $Z_o = 50\,\Omega$, thus with $\frac{Z_o}{R_2}$ = VSWR = 3 we find

$$R_o = 5.5 \cdot$$

Further,

$$E = 2 \frac{R_o + R}{Z_o + R} \cdot \sqrt{P_{fmax} \cdot Z_o} \cdot$$

With $R_o = 5.5\Omega$, $R_2 = 17\Omega$, $Z_o = 50\Omega$ and $P_{fmax} = 113W$ we find $E = 50V$.

It is interesting to note that in going with the load along a VSWR = constant circle, the maximum loss in the transistor does not occur at maximum power output (see figure 15) but at a complex load with 45° capacitive phase angle.

3.6.1 Closer antenna matching tolerance required

One can see from these considerations that closer tolerances of antenna matching are required, using, e.g., centerfed whips and better matching networks, in order to get the whole benefit of the design possibility. The benefits are that power consumption is less, hence smaller batteries can be used, heatsinks become smaller, and blowers for cooling can be avoided.

4. Miscellaneous

4.1 The fan. The power dissipated in the set itself has been cut from 200W down to 70W. Therefore, the fan need not be switched on during transmit keying. However, it has been left in the set in order to meet marginal environmental conditions, like 65^oC ambient temperature, or if the set is heated up by direct sun radiation. In this laboratory model, the switch for the fan is operated by a thermal switch located on the heat sink. This is set to turn on the fan if the temperature on the heat sink exceeds 78^oC.

Currently, the fan motor has been driven by a 400 Hz two phase voltage, being supplied by a converter in the power supply. This converter, which contains two heavy coils and a transformer, can be replaced by a transformerless induction motor, being supplied by Rotating Components Inc., Bayshore, N.Y. (RCI No. 15SD-1A). Preliminary tests indicate that this fan works well. At 25.5V battery voltage it draws .48A of DC current. At 22V battery voltage there is still ample starting torque. The fan has not been built into the feasibility model, since its diameter is slightly greater than that of the old fan (36 mm diam).

It would have been too costly as far as time is concerned to change the mechanics of our feasibility model to fit this fan in; however, if additional models of the amplifiers should be

built, it is recommended to use this transformerless induction motor fan. Probably, however, the 400 Hz source could not be eliminated entirely, since it is used for driving servo motors in the push-button type VRC-12. Otherwise, we would get two different types of power supplies, which might be bad for logistics.

4.2 The power supply module

The power supply module has been stripped of its high voltage converter, see figure 17. The choke L9001 could be reduced to a 1"-diameter ferrite cup core, since the DC current to be handled is only 1/2 of what it was before.

4.3 Remarks on mechanical design. The amplifier in its present form represents a feasibility study. Therefore, a number of mechanical changes would be necessary before entering fabrication. Changes recommended:

1. The three positions of the band switch are furnished as follows: 30...53 and 53...76 MHz by ordinary mechanical bandswitch contained in present VRC-12. The additional switch position for 30...40 MHz is achieved by Ledex, acting as a motor. The information is derived by a cam which is positioned on the shaft that formerly drove the tuning capacitor and the variometer. The cam activates a microswitch. which switches the Ledex. The Ledex seems to be reasonably good, on a life test we obtained in excess of 500,000 operations till failure. After about 650,000 operations the main shaft froze. At that time, the contacts were worn down also. In case a high reliability design would be desired, some changes in the Ledex might be required. Also, the current should be switched by a high reliability transistor, so sparking on the contacts could be eliminated. Eventually, the design should be changed such that all three positions are served by the Ledex, the Ledex having three positions.

2. Along with this, a regrouping of mechanical parts should take place. The three regulator power transistors should go on a common heat sink, mechanically assembled together with the regulator printed circuit board. The power amplifier and the driver amplifier would be built onto a single printed board, eventually replacing the rotary switch on the driver assembly by four miniature shift switches, being located where needed electrically.

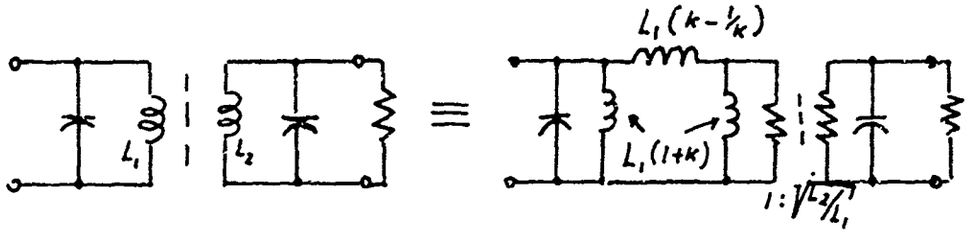
The three output filters and the VSWR detector and switch circuit should be unified in a single block.

5. Appendix A

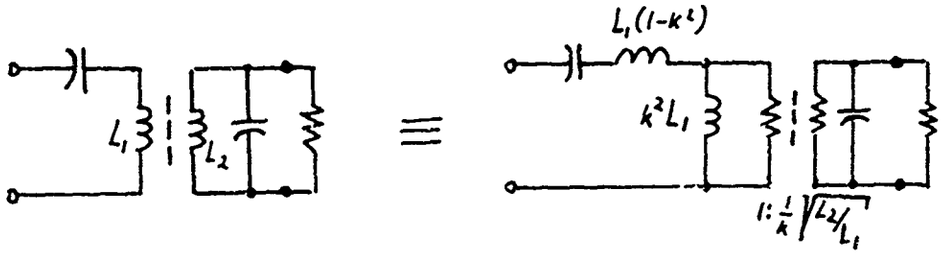
5.1 Some remarks about transformers

To improve matching and to extend bandwidth the following methods can be taken.

5.1.1 The transformer, if taken as an inductive π - network, receives a proper parallel capacitor at the input and output. Thus, the configuration of a wideband - bandfilter is obtained. This way, given the stray factor or coupling coefficient of the transformer, its π - ersatz circuit is calculated, and then the capacities are designed according to well-known broadband-bandfilter design.

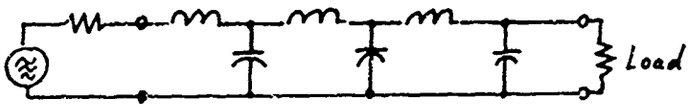


5.1.2 The transformer, if taken as an inductive type L ersatz network, can be complemented into a half-bandpass-filter Tsection.



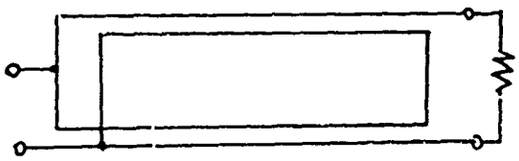
The same thing can be done complementing it to a half π-section. The design goes along well-known principles of bandpass filter design.

5.1.3 **Transformation** by use of "tapered" low pass filter design, *see [3]*



An analogue of this is the exponential line transformer in transmission line technique.

5.1.4 Guanella type transmission line transformer, see [4], [5], for very broadbanded applications.



LITERATURE INDEX

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[2] D.R. Lohrmann, Amplifier has 85% Efficiency While Providing Up to 10W Power Over A Wide Frequency Band, Electronic Design, March 1966, page 38. (See App. B, pp 35-40 of this technical report.)

[3] G.L. Matthaei, Table of Chebyshev Impedance Transforming Networks of Low Pass Filter Form, Proc. IEEE, Aug 1964, page 939.

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ACKNOWLEDGEMENT

The author wishes to express his gratitude to Messrs. A.C. Colaguori and J. Anderl for their encouragement, advice and discussion. His gratitude is also expressed to Messrs R. Chadwick, G. Hessel and H. Spence for their help in design and construction.

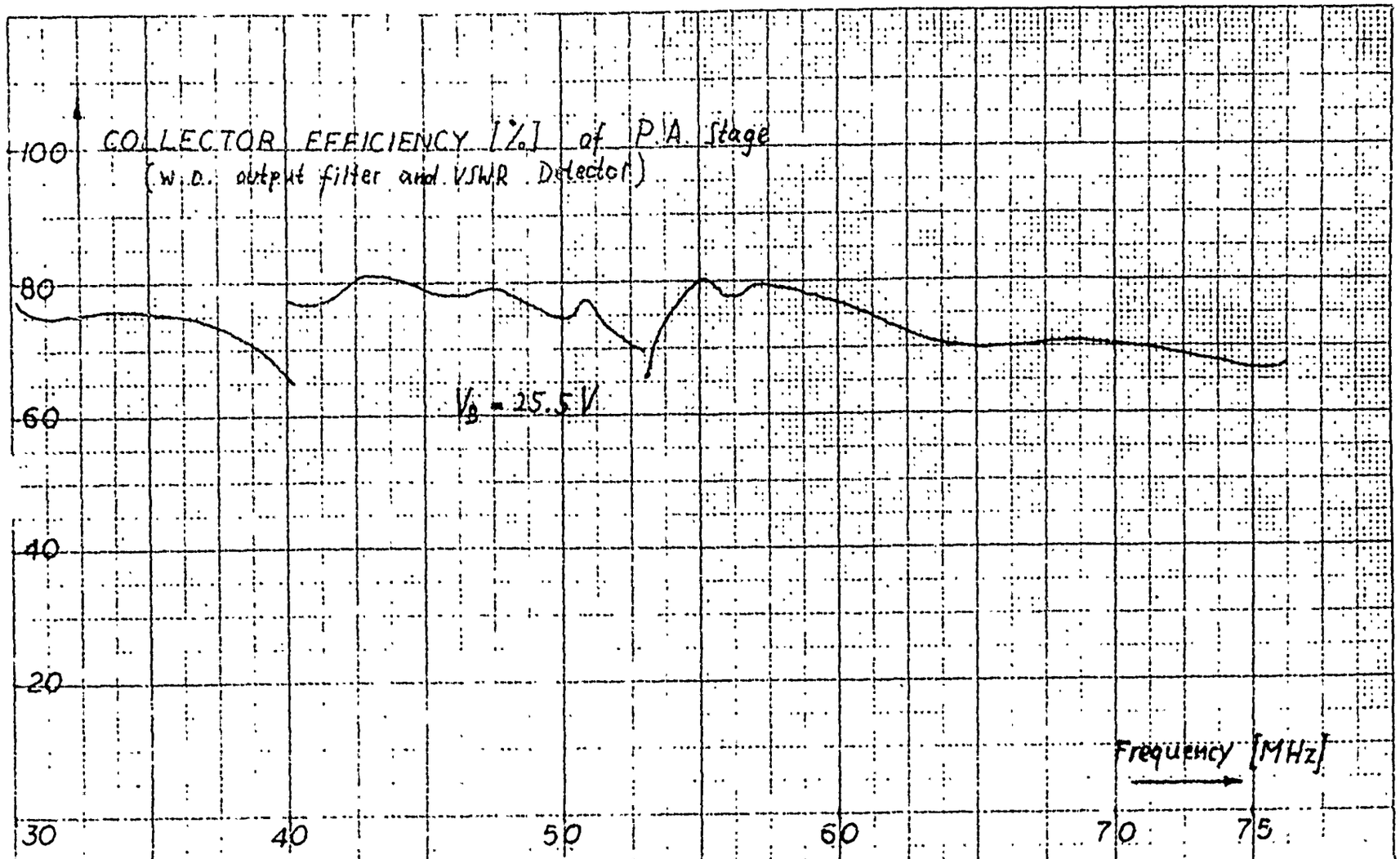


FIG:1 COLLECTOR EFFICIENCY OF PA STAGE

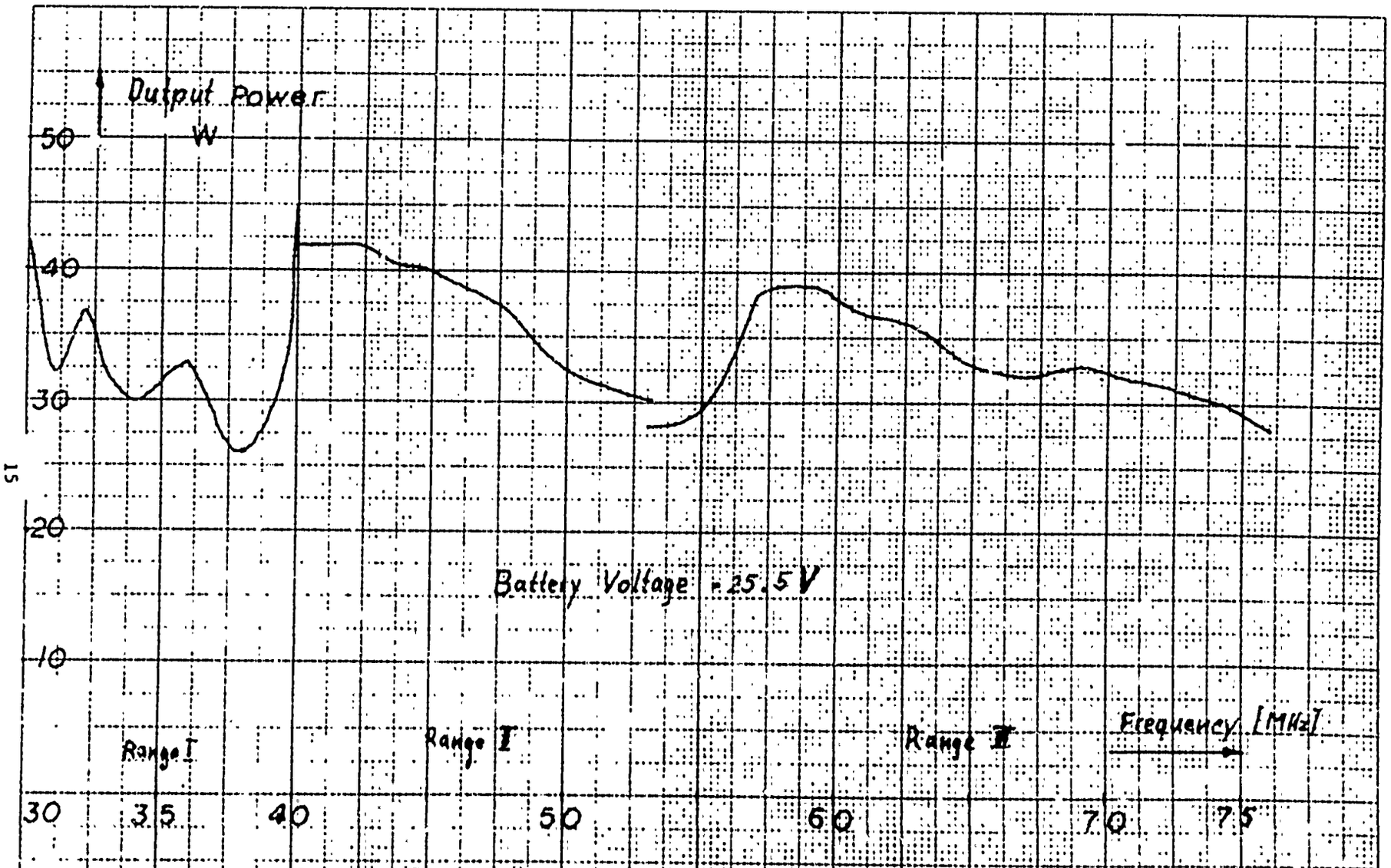


FIG.2: OVERALL OUTPUT POWER

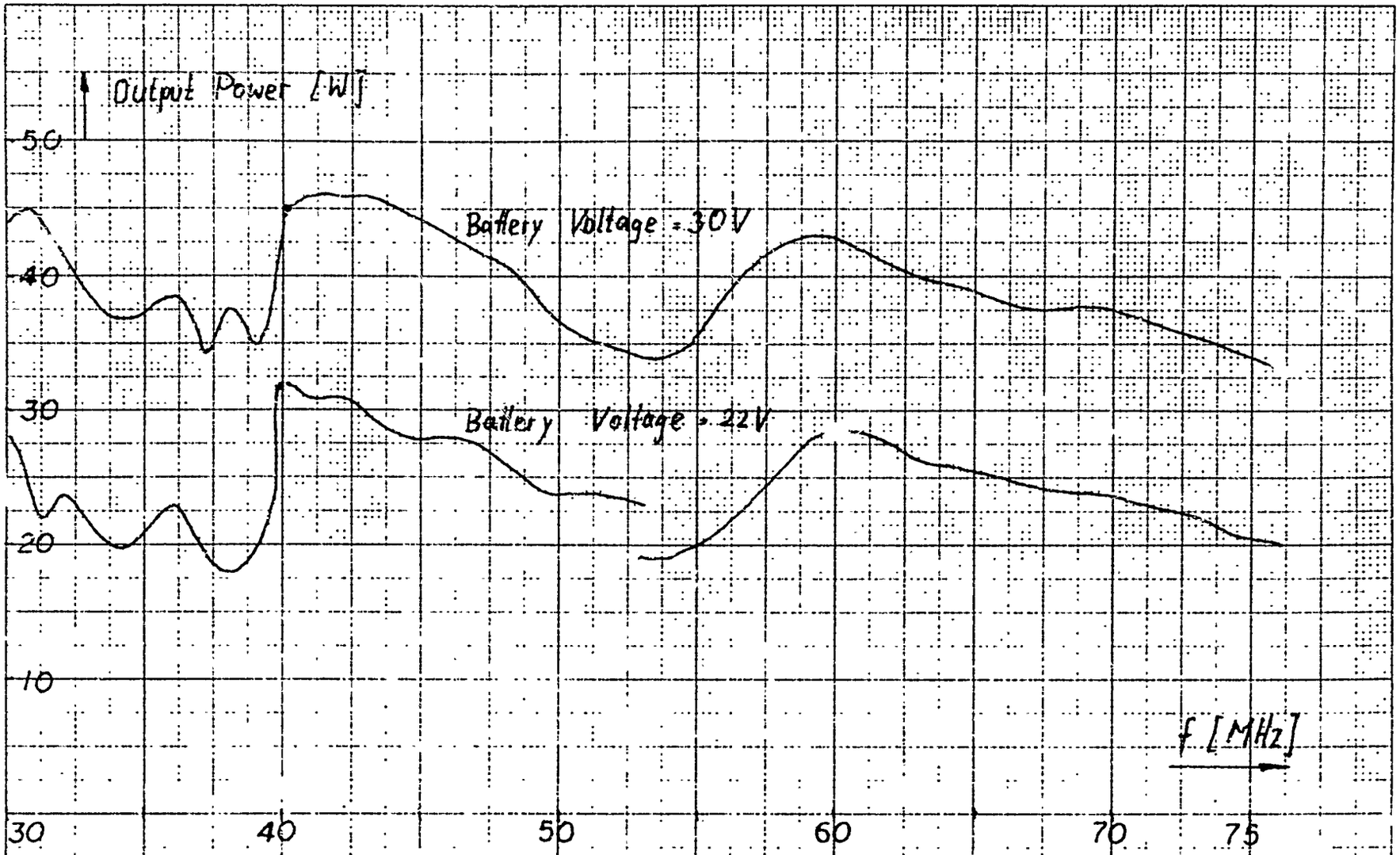


FIG. 2A: Overall Output Power at 22 V and 30 V Battery Voltage

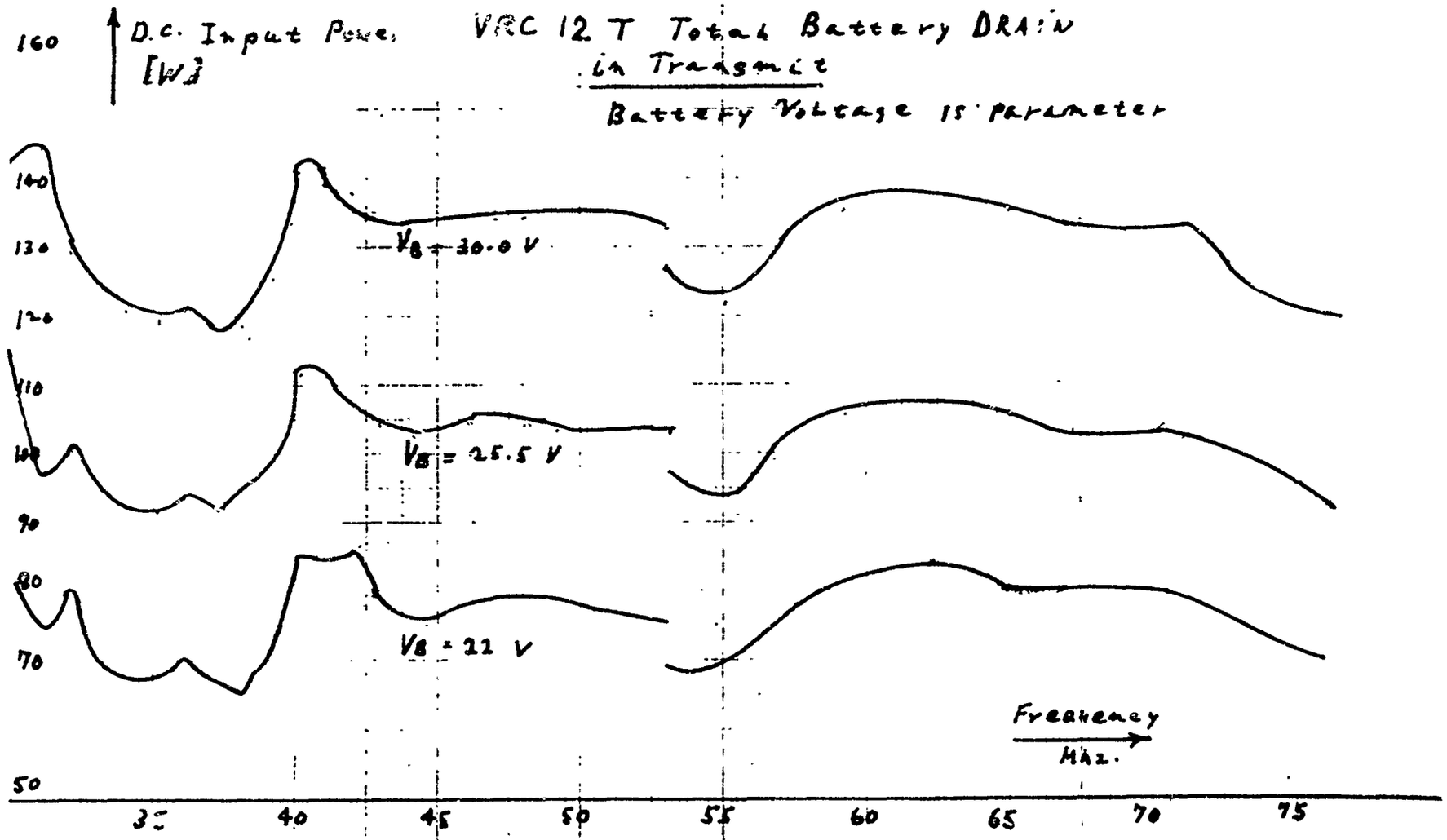

D.C. Input Power [W]
VRC 12 T Total Battery DRAIN in Transmit
Battery Voltage is parameter
160
140
130
120
110
90
80
70
50
V_B = 30.0 V
V_B = 25.5 V
V_B = 22 V
Frequency MHz.
35
40
45
50
55
60
65
70
75

FIG. 3

fig. 3

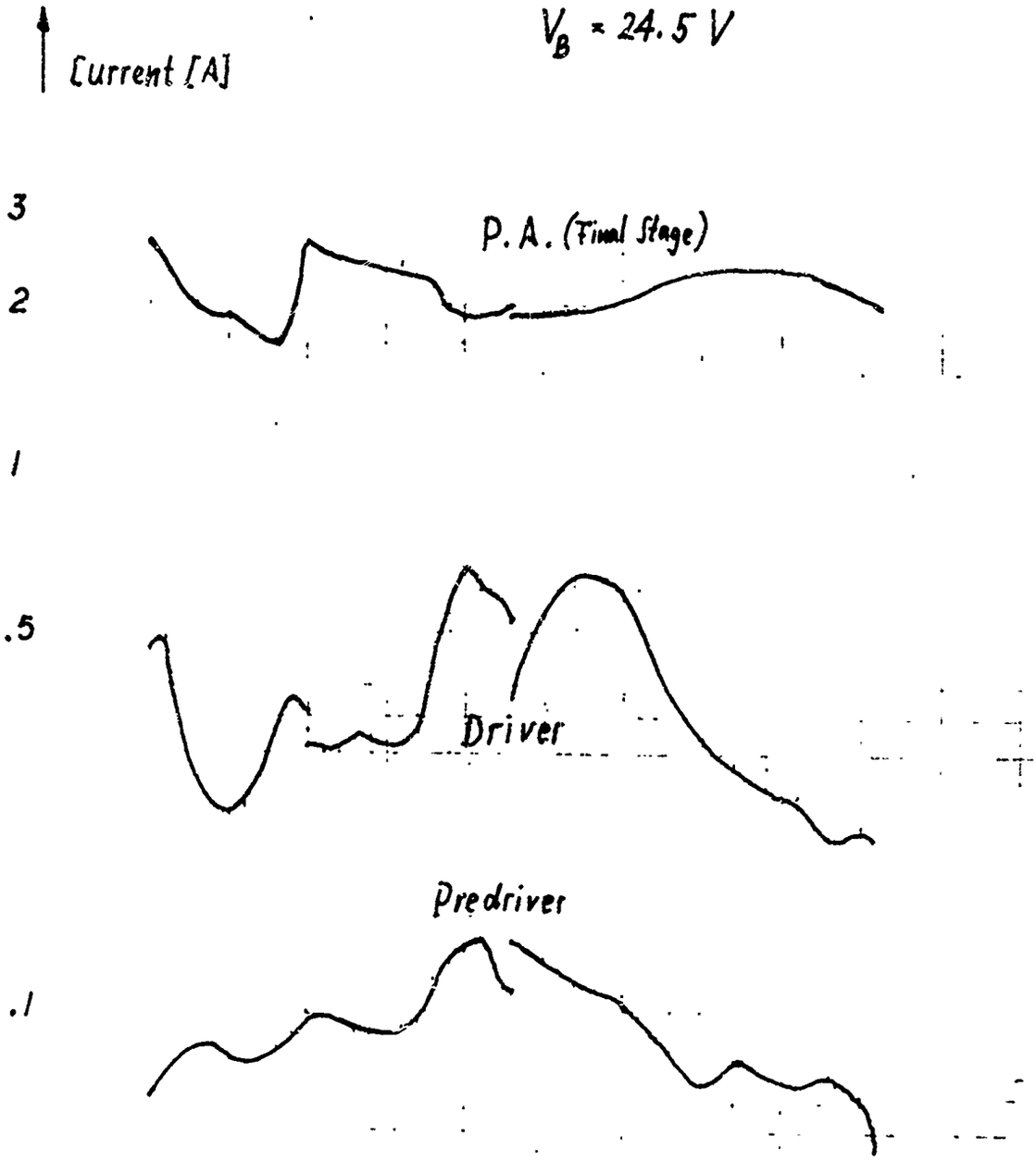


FIG. 3A :

P.A. - DRIVER - and PREDRIVER - CURRENTS

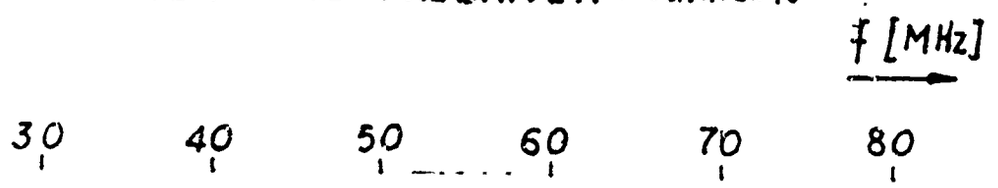
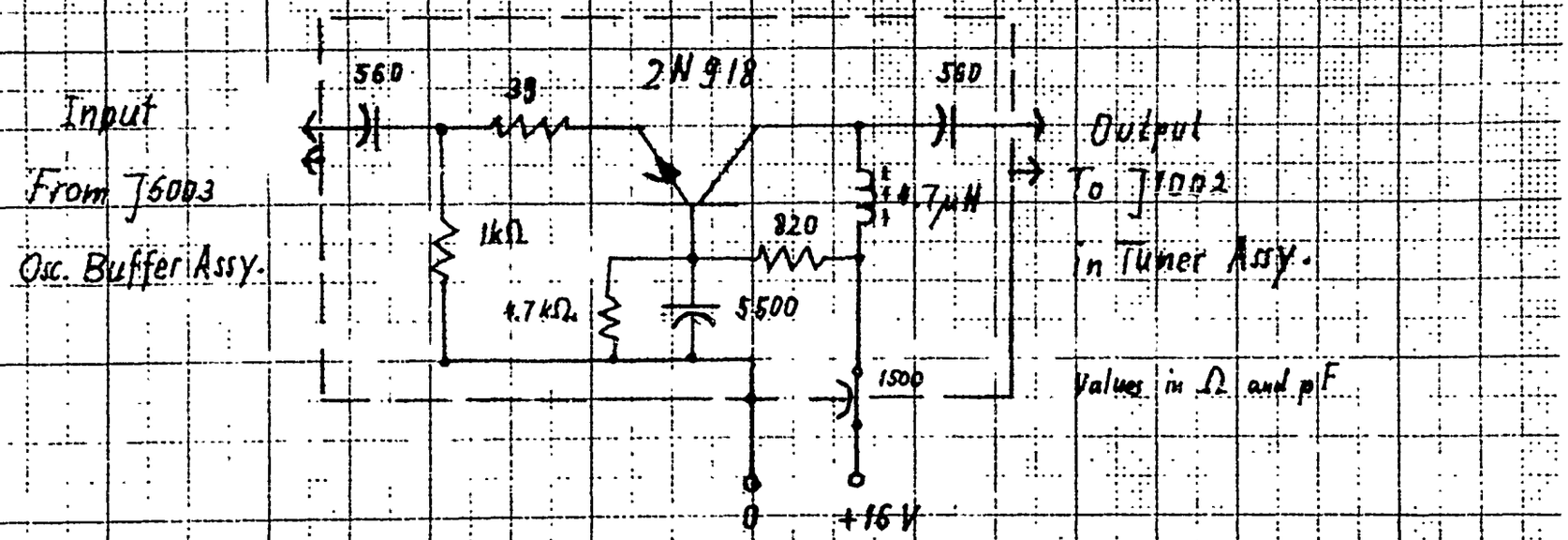


FIG 4 : ISOLATION AMPLIFIER

To be inserted between transmit driver and HF Tuner



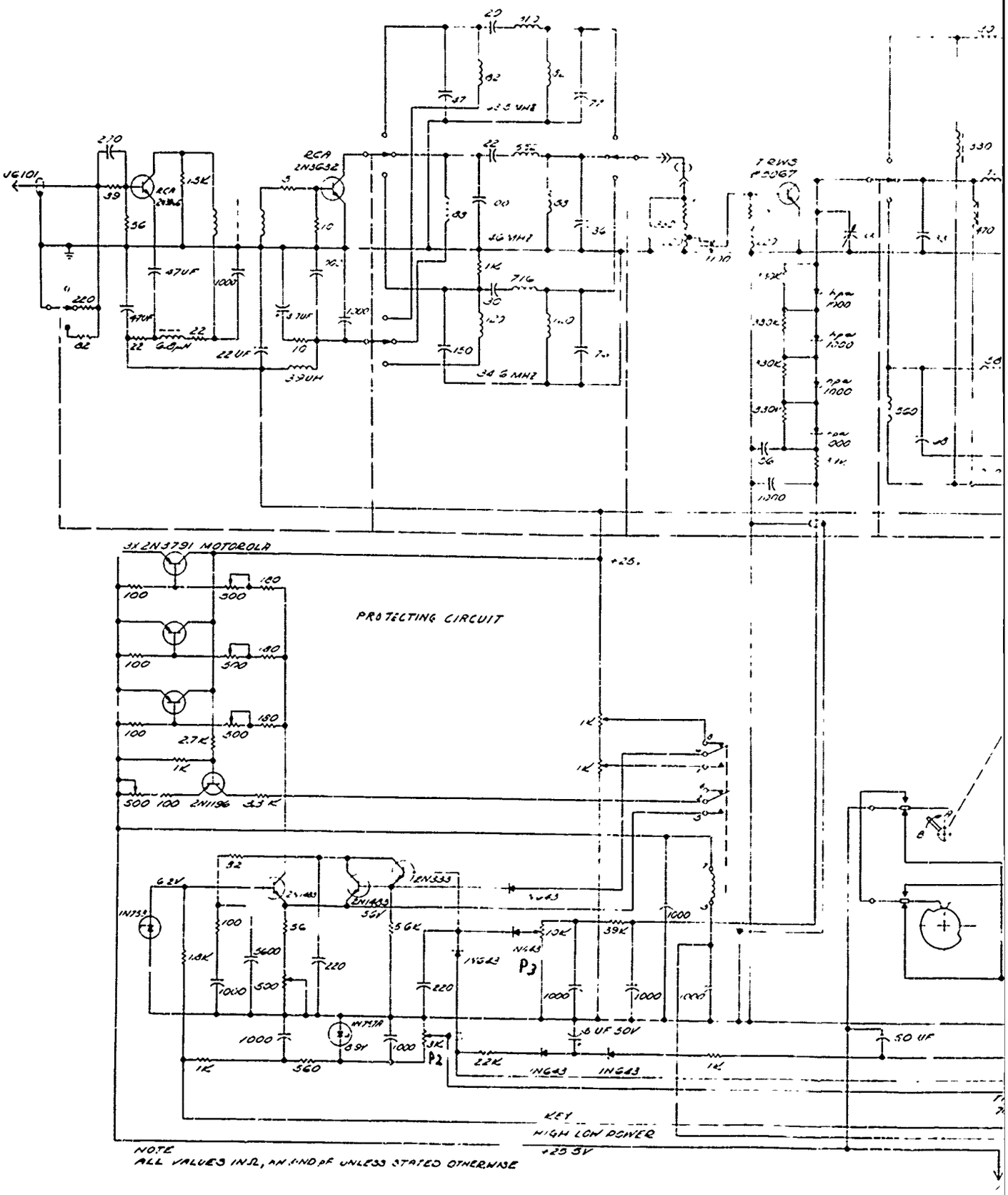
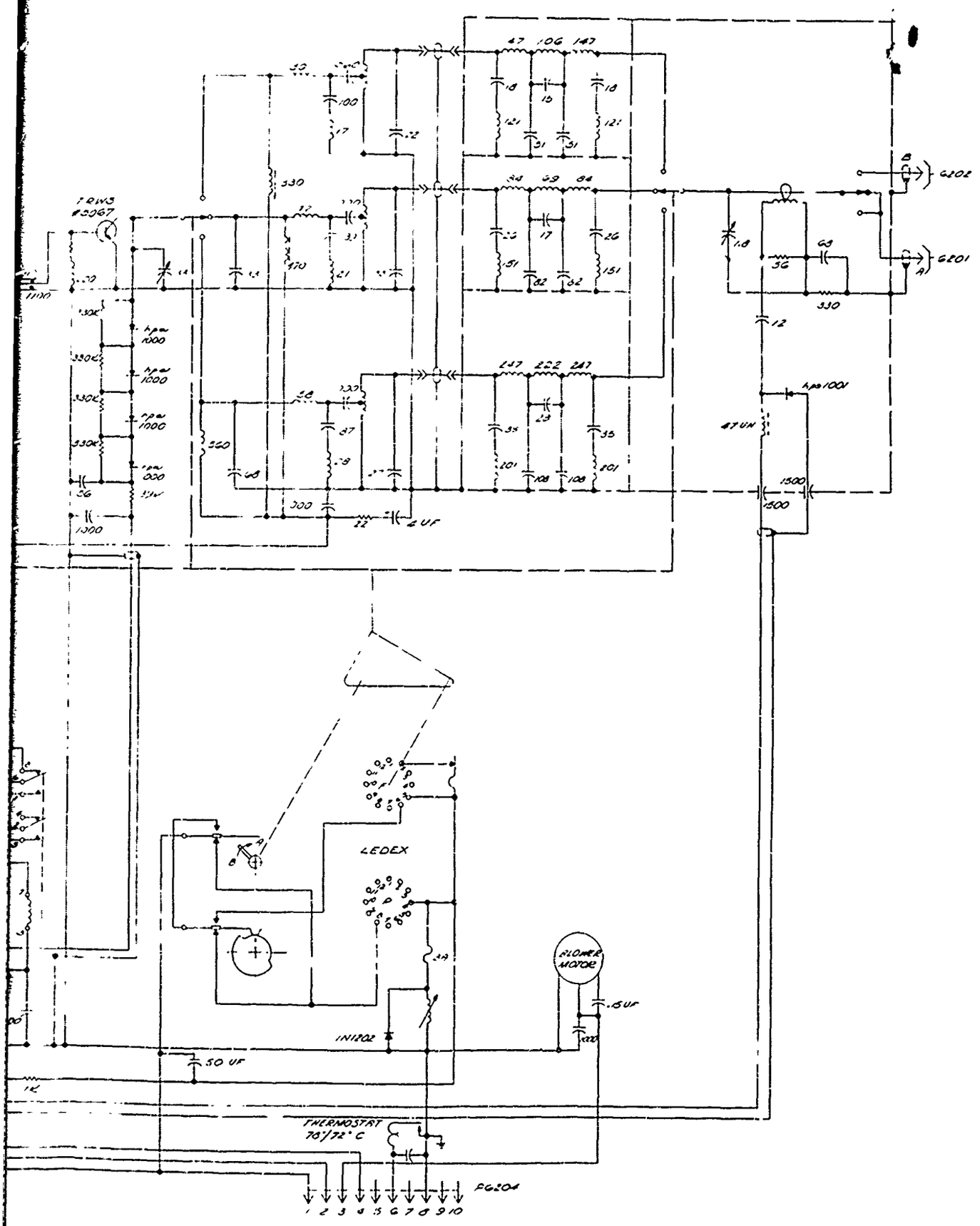
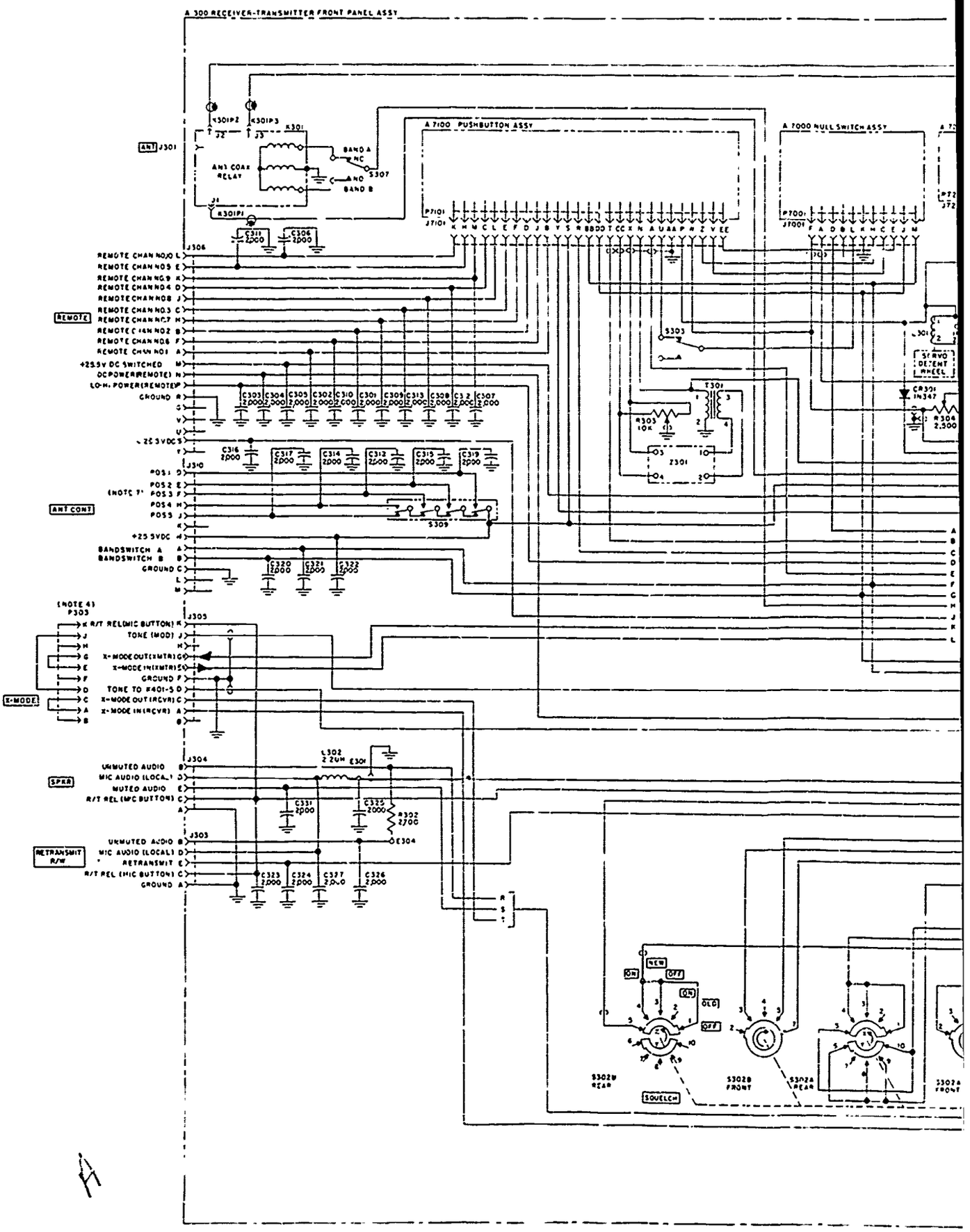


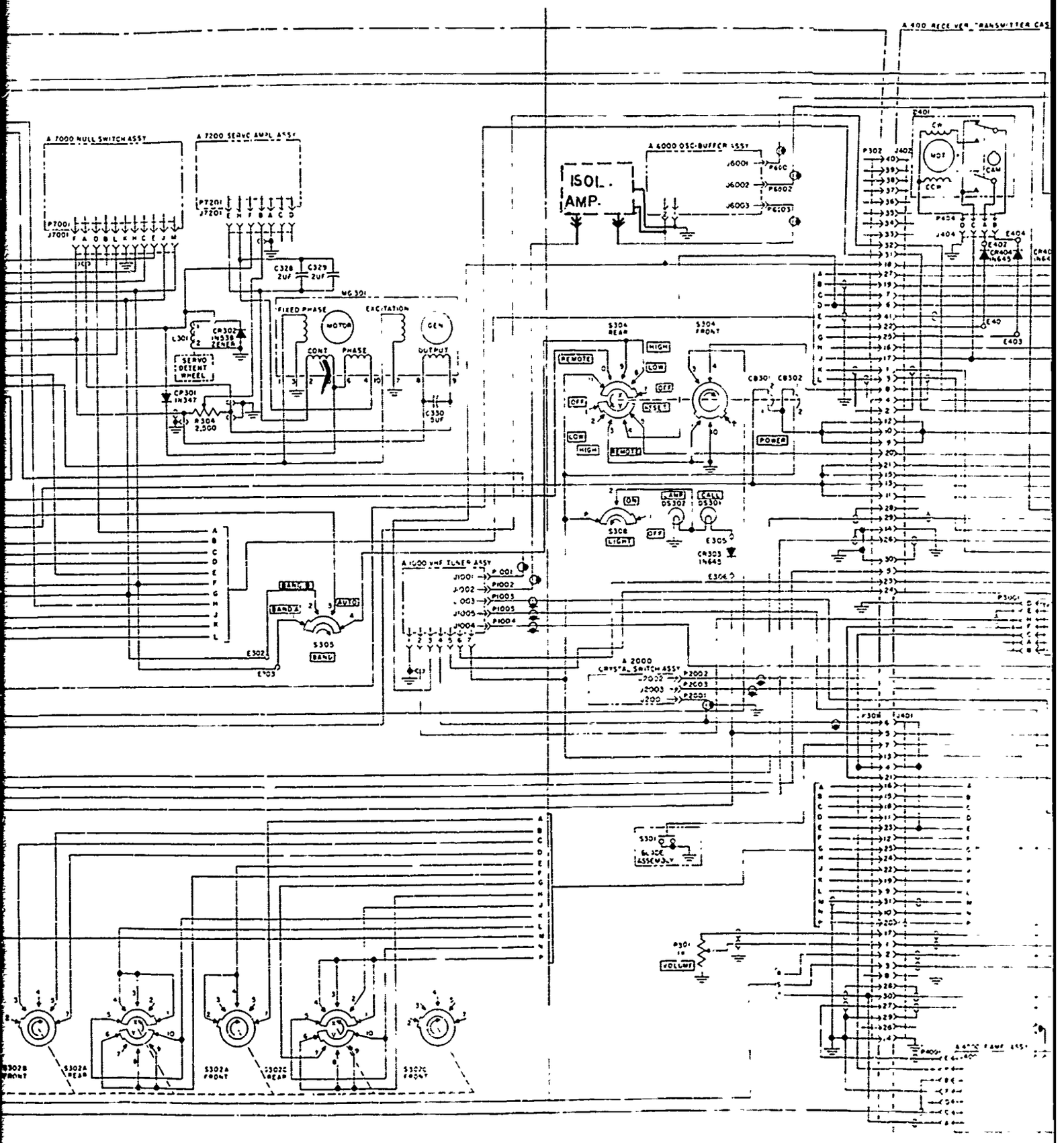
FIG. 5 A6200 PA ASSEMBLY

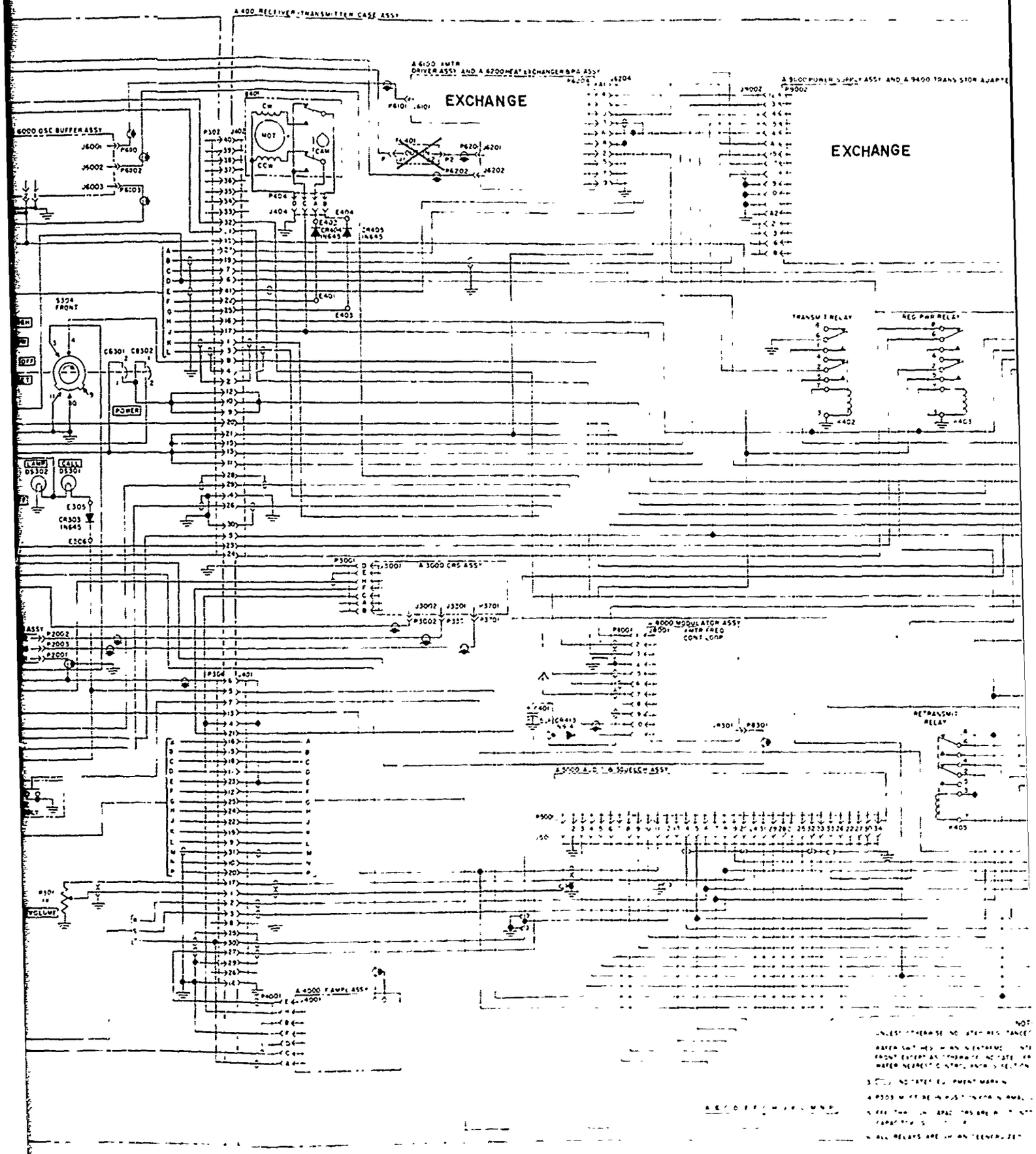


00 PA ASSEMBLY

B

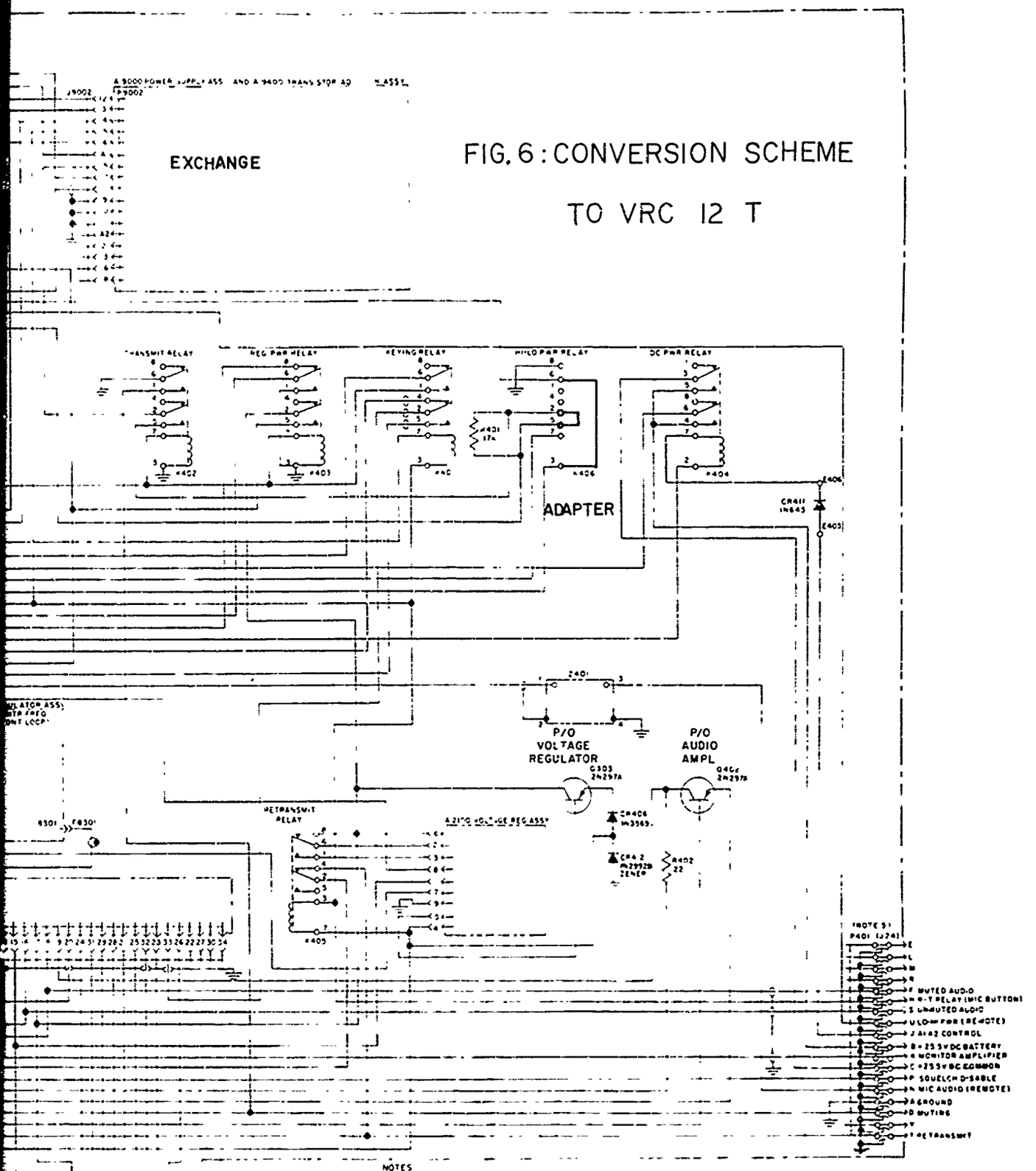

A 300 RECEIVER-TRANSMITTER FRONT PANEL ASSY
A 7100 PUSHBUTTON ASSY
A 7000 NULL SWITCH ASSY
ANT J301
ANT COAX RELAY
K301
K301P1
K301P2
K301P3
BAND A NC
BAND B NO
S307
C311 2000
C306 2000
J306
REMOTE CHAN NO.10
REMOTE CHAN NO.5
REMOTE CHAN NO.9
REMOTE CHAN NO.4
REMOTE CHAN NO.8
REMOTE CHAN NO.3
REMOTE CHAN NO.7
REMOTE CHAN NO.2
REMOTE CHAN NO.6
REMOTE CHAN NO.1
REMOTE
+25.5V DC SWITCHED
DC POWER (REMOTE)
LO-H. POWER (REMOTE)
GROUND
-25.5VDC
J310
POS 1
POS 2
POS 3
POS 4
POS 5
(NOTE 7)
ANT CONT
S309
+25.5VDC
BANDSWITCH A
BANDSWITCH B
GROUND
(NOTE 4)
P303
J305
R/T REL (MIC BUTTON)
TONE (MOD)
X-MODE OUT (XMTR)
X-MODE IN (XMTR)
GROUND
TONE TO K401-5
X-MODE OUT (RCVR)
X-MODE IN (RCVR)
X-MODE
J304
UNMUTED AUDIO
MIC AUDIO (LOCAL)
MUTED AUDIO
R/T REL (MIC BUTTON)
SPKR
L302 2.2UH
E301
C331 2000
C325 2000
R302 270Ω
J303
UNMUTED AUDIO
MIC AUDIO (LOCAL)
RETRANSMIT
R/T REL (MIC BUTTON)
GROUND
RETRANSMIT R/W
C323 2000
C324 2000
C327 2000
C326 2000
E304
T301
R303 10K
Z301
S303
SERVO DETENT WHEEL
CR301 1N347
R304 2,500
NEW
ON
OFF
OLD
S302B REAR
S302B FRONT
S302A REAR
S302A FRONT
SQUELCH


A 7000 NULL SWITCH ASSY
A 7200 SERVO AMPL ASSY
ISOL. AMP.
A 6000 OSC-BUFFER ASSY
A 400 RECEIVER TRANSMITTER CAS
A 1000 VHF TUNER ASSY
A 2000 CRYSTAL SWITCH ASSY
MG 301
FIXED PHASE
MOTOR
EXCITATION
GEN
CONT
PHASE
OUTPUT
SERVO DETENT WHEEL
S304 REAR
S304 FRONT
REMOTE
HIGH
LOW
OFF
RESET
POWER
S308 LIGHT
S305 BAND
BAND A
BAND B
AUTO
S301 BLOCK ASSEMBLY
P301 10 VOLUME
S302B FRONT
S302A REAR
S302A FRONT
S302C REAR
S302C FRONT


A 400 RECEIVER-TRANSMITTER CASE ASSY
A 6100 XMTR DRIVER ASSY AND A 6200 HEAT EXCHANGER & PA ASSY
EXCHANGE
A 9100 POWER SUPPLY ASSY AND A 9400 TRANSISTOR ADAPTER
EXCHANGE
6000 OSC BUFFER ASSY
MOT
CW
CCW
CAM
S304 FRONT
POWER
TRANSMIT RELAY
REC PWR RELAY
A 3000 CRS ASSY
A 8000 MODULATOR ASSY XMTR FREQ CONT LOOP
RETRANSMIT RELAY
A 5000 A.C. & SQUELCH ASSY
VOLUME
A 4000 F AMPL ASSY

C

FIG. 6 : CONVERSION SCHEME TO VRC 12 T



NOTES

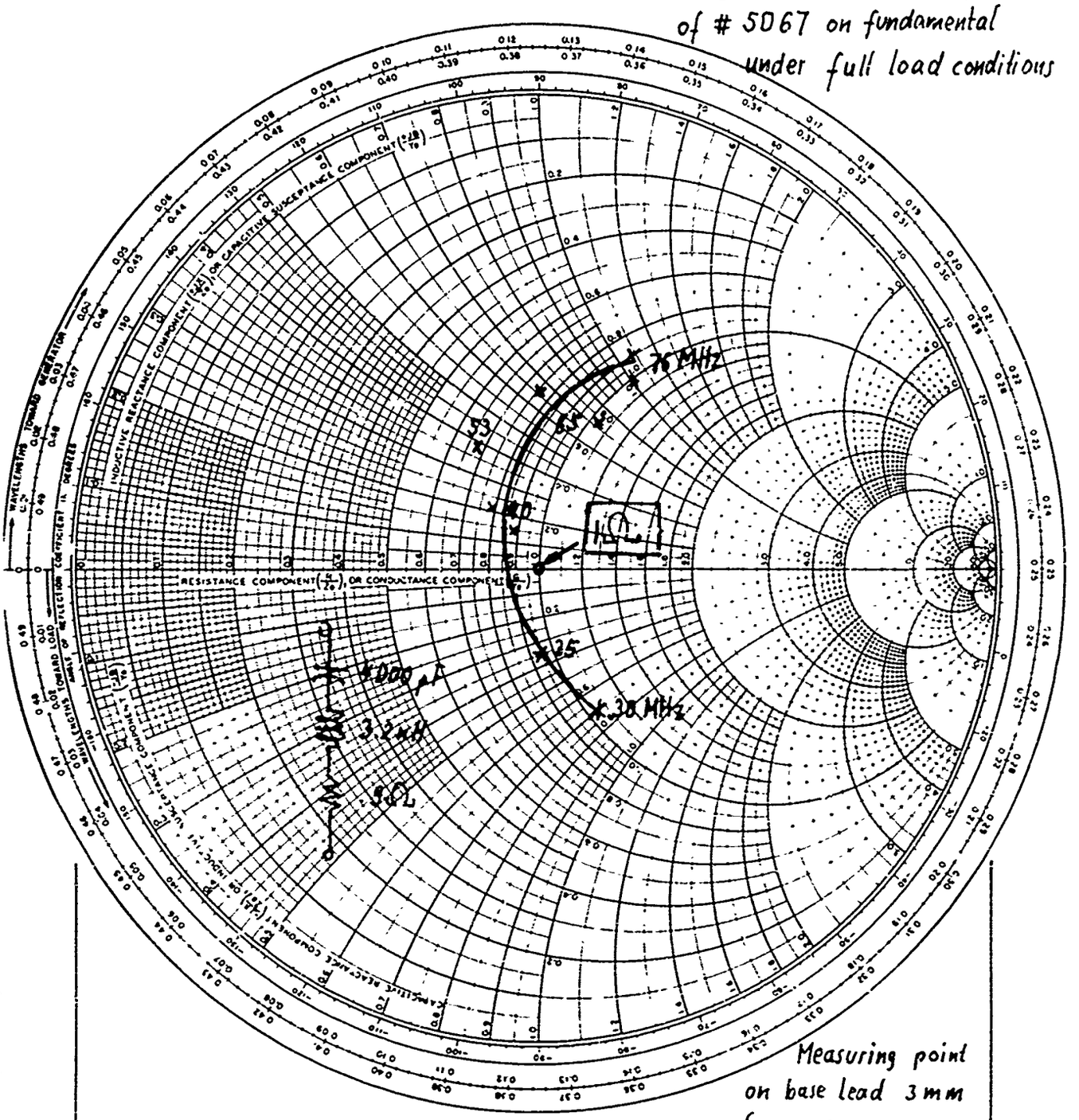
1. UNLESS OTHERWISE INDICATED RESISTANCES ARE IN OHMS CAPACITANCES ARE IN µf
2. WAFER SWITCHES SHOWN IN EXTREME CCW INTERPOSITION SEE POSITIONS AND ARE VIEWED FROM FRONT EXCEPT AS OTHERWISE INDICATED FRONT OF WAFER SIDE TOWARD CONTROL KNOB WAFER NEAREST CONTROL KNOB SECTION A SEGMENTS OF WAFERS ARE IDENTIFIED BY
3. ⬭ INDICATES EQUIPMENT MARKING
4. R303 MUST BE IN POSITION FOR NORMAL OPERATION
5. FREQUENCY CAPACITORS ARE [illegible] THE VALUE OF EACH CAPACITOR [illegible]
6. ALL RELAYS ARE [illegible] GENERAL ELEC

POS	BAND A	BAND B
1	32 [illegible] MC	53 [illegible] 55 [illegible] MC
2	33 75 [illegible] 25 MC	55 75 56 25 MC
2	35 [illegible] 36 [illegible] MC	55 30 59 [illegible] MC
2	36 75 [illegible] 25 MC	59 75 60 25 MC
3	37 30 [illegible] 60 MC	60 35 64 60 MC
3	64 65 [illegible] 35 MC	64 65 65 35 MC
4	42 40 [illegible] MC	64 65 [illegible] MC
4	4[illegible] 5 [illegible] 85 MC	[illegible] 5 [illegible] 85 MC
[illegible]	4[illegible] 52 95 MC	[illegible] 75 95 MC

SWITCHES ARE SHOWN AS FOLLOWS		
SWITCH		POSITION
S3[illegible]		OPEN
S3[illegible]	SQUELCH	OLD OFF
S3[illegible]		CLOSED
S3[illegible]	POWER	OFF
S3[illegible]	BAND	BAND A
S3[illegible]		BAND A
S3[illegible]	SHIFT	OFF
S3[illegible]		[illegible]

NAME	TITLE FIG. 7	DWG. NO.
SMITH CHART FORM 82B5[illegible] (2-46)	KAY ELECTRIC COMPANY, PINE BROOK, N.J. ©1949 PRINTED IN U.S.A.	DATE

IMPEDANCE OR ADMITTANCE COORDINATES Base Emitter Imped. of # 5067 on fundamental under full load conditions

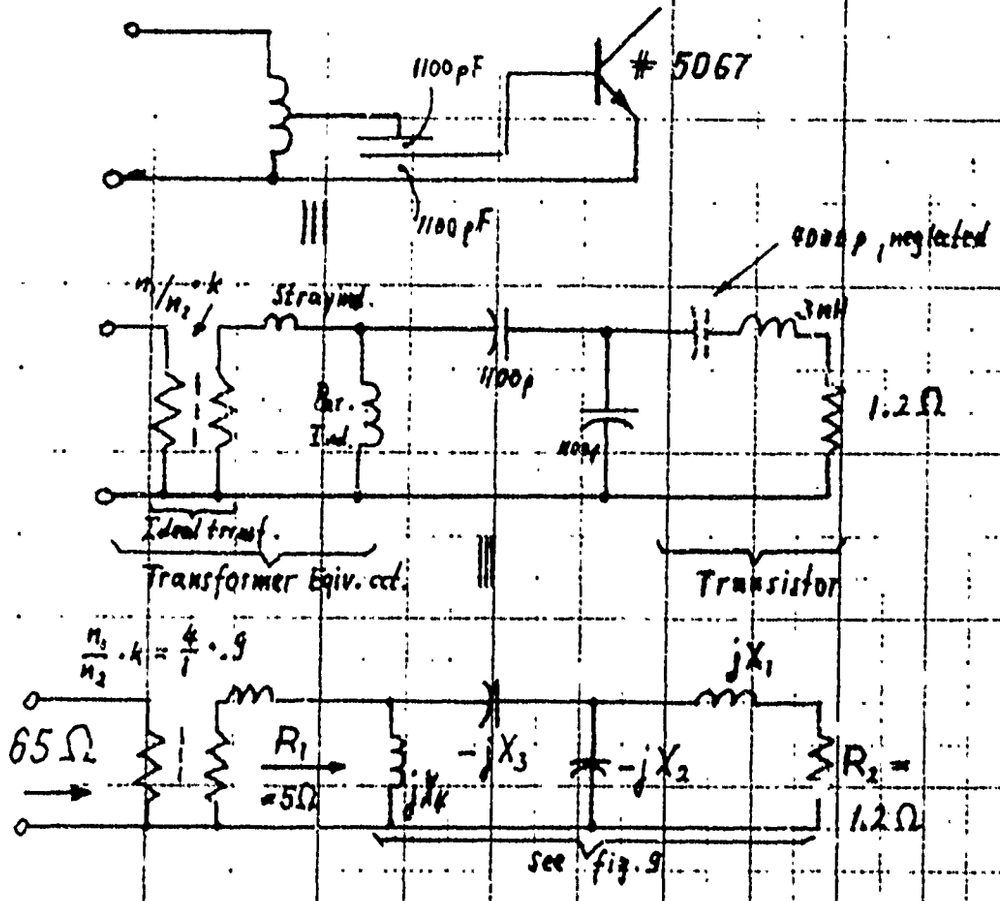


Measuring point on base lead 3 mm from case

RADIALLY SCALED PARAMETERS

TOWARD GENERATOR → ← TOWARD LOAD

CENTER



$R_1/R_2 = n = 4 \qquad R_2 = 1.2\ \Omega$

$X_1 = R_2 \cdot \sqrt{\sqrt{n} - 1}\ ; \quad L_1 = \frac{X_1}{2\pi f_u} = 3.5\ nH$

$X_2 = R_2 \cdot \sqrt{\frac{n}{\sqrt{n} - 1}}\ ; \quad C_2 = 1170\ pF$

$X_3 = R_2 \cdot \sqrt{(\sqrt{n} - 1)\, n} \quad C_3 = 1170\ pF$

$X_4 = R_2 \cdot \frac{n}{\sqrt{\sqrt{n} - 1}}\ ; \quad L_4 = 14\ nH$

FIG. 8 : INPUT MATCHING OF TRANSISTOR #5067

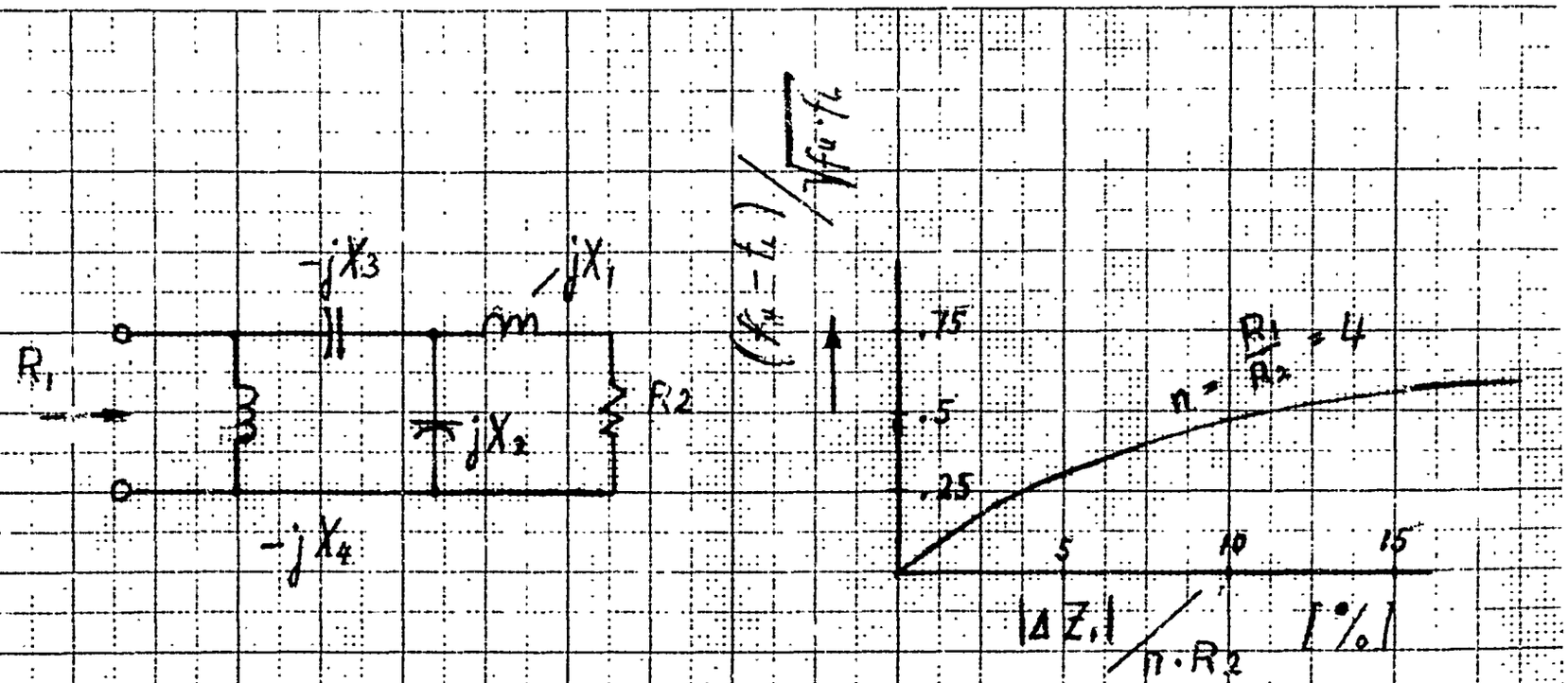


FIG. 9 : TRANSFORMING NETWORK AND ITS BANDWIDTH [1]

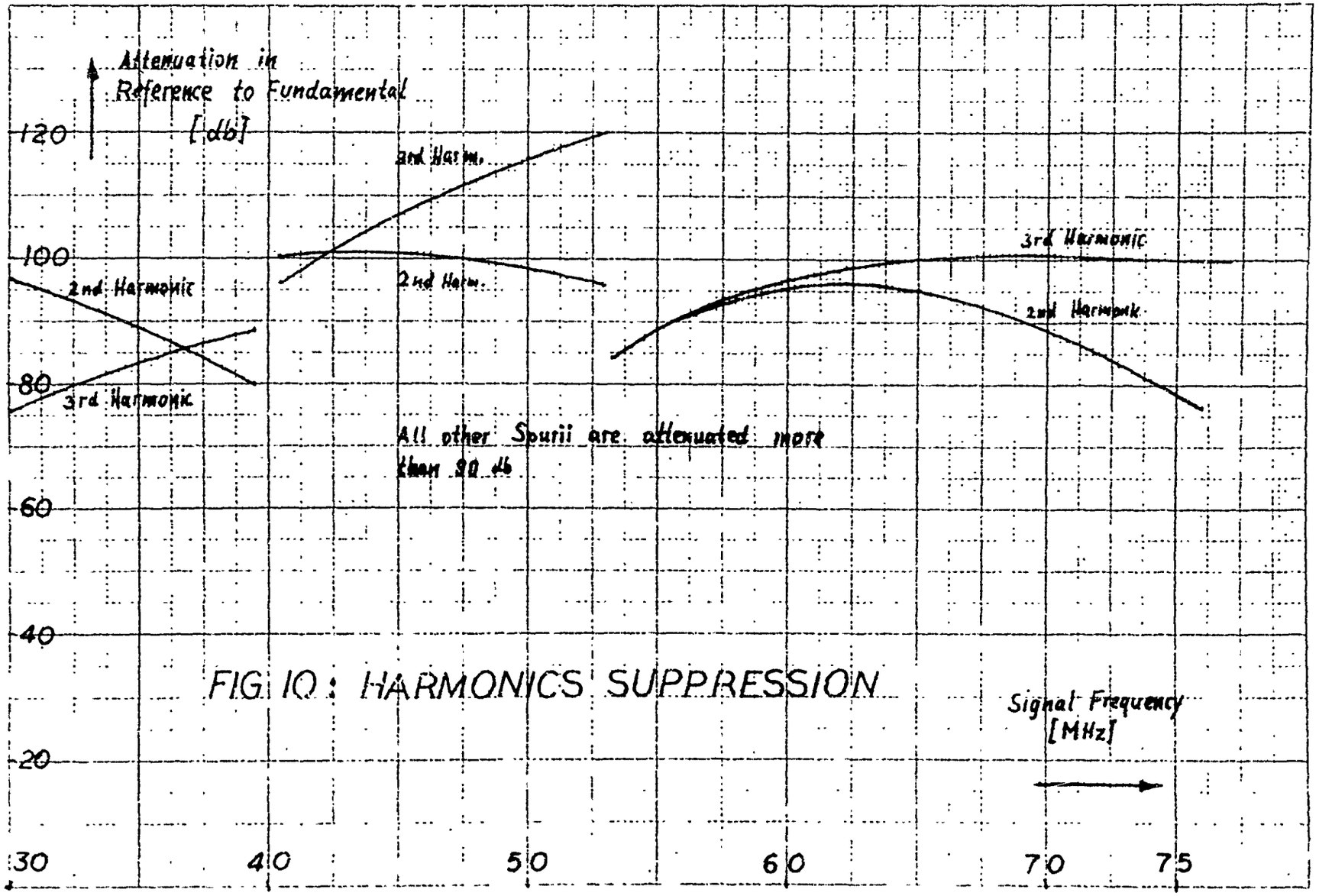
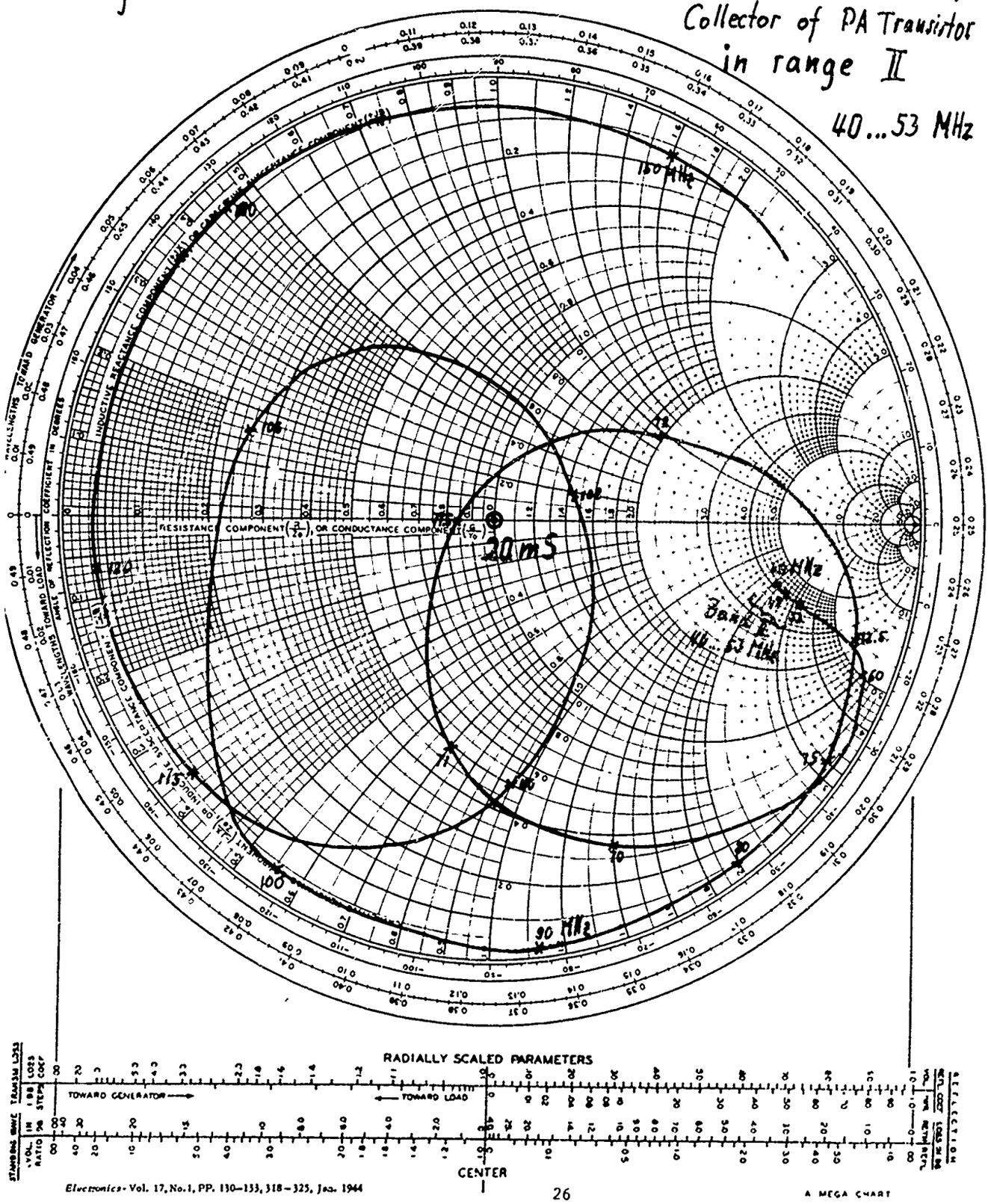


FIG 10: HARMONICS SUPPRESSION

NAME	TITLE	DWG. NO.
SMITH CHART FORM 8285PR (2-49)	KAY ELECTRIC COMPANY, PINE BROOK, N.J. ©1949 PRINTED IN U.S.A.	DATE

Fig. 11

IMPEDANCE OR ADMITTANCE COORDINATES Admittance "seen" by Collector of PA Transistor in range II 40...53 MHz



RADIALLY SCALED PARAMETERS

TOWARD GENERATOR ⟶ ⟵ TOWARD LOAD

CENTER

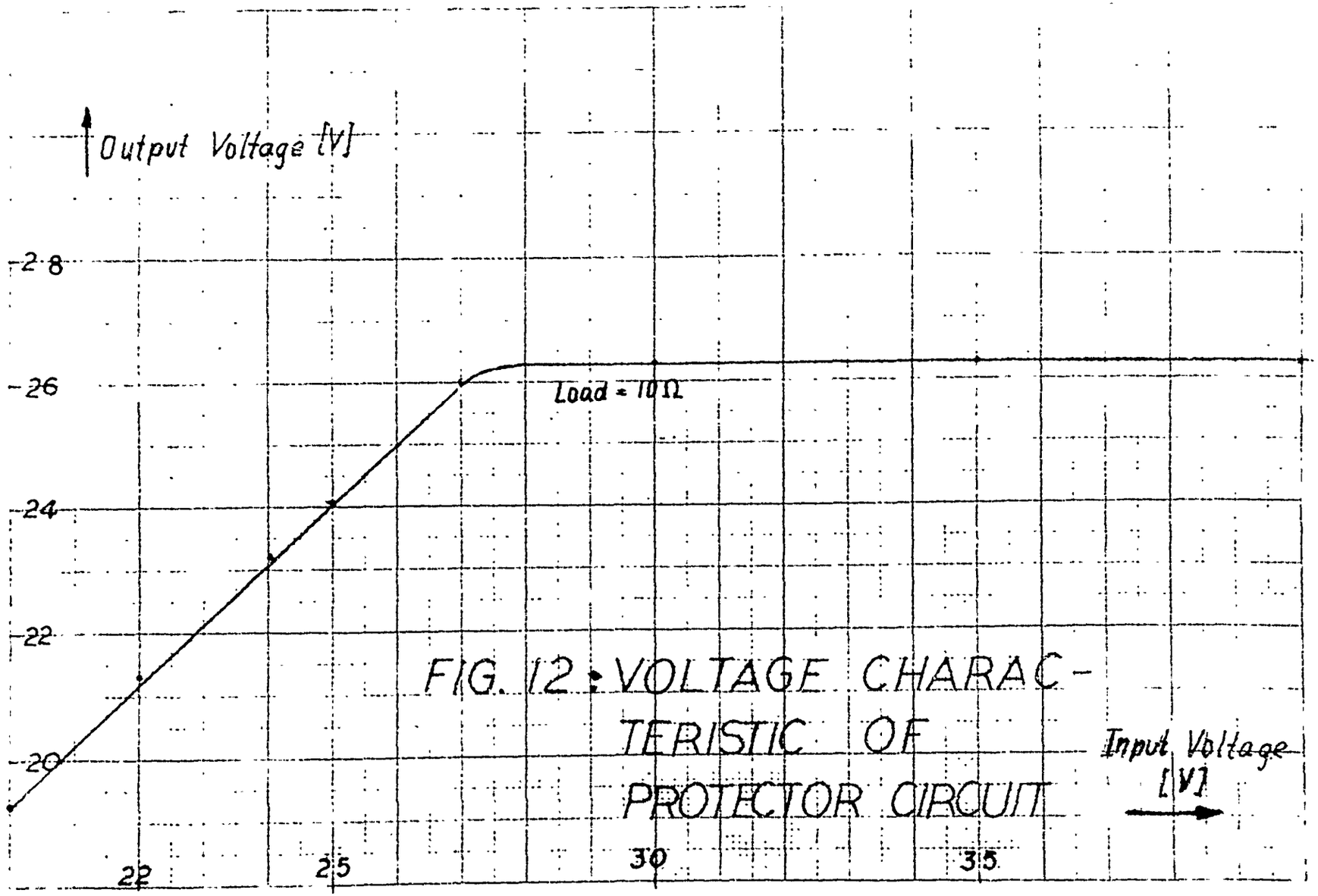


FIG. 12 : VOLTAGE CHARACTERISTIC OF PROTECTOR CIRCUIT

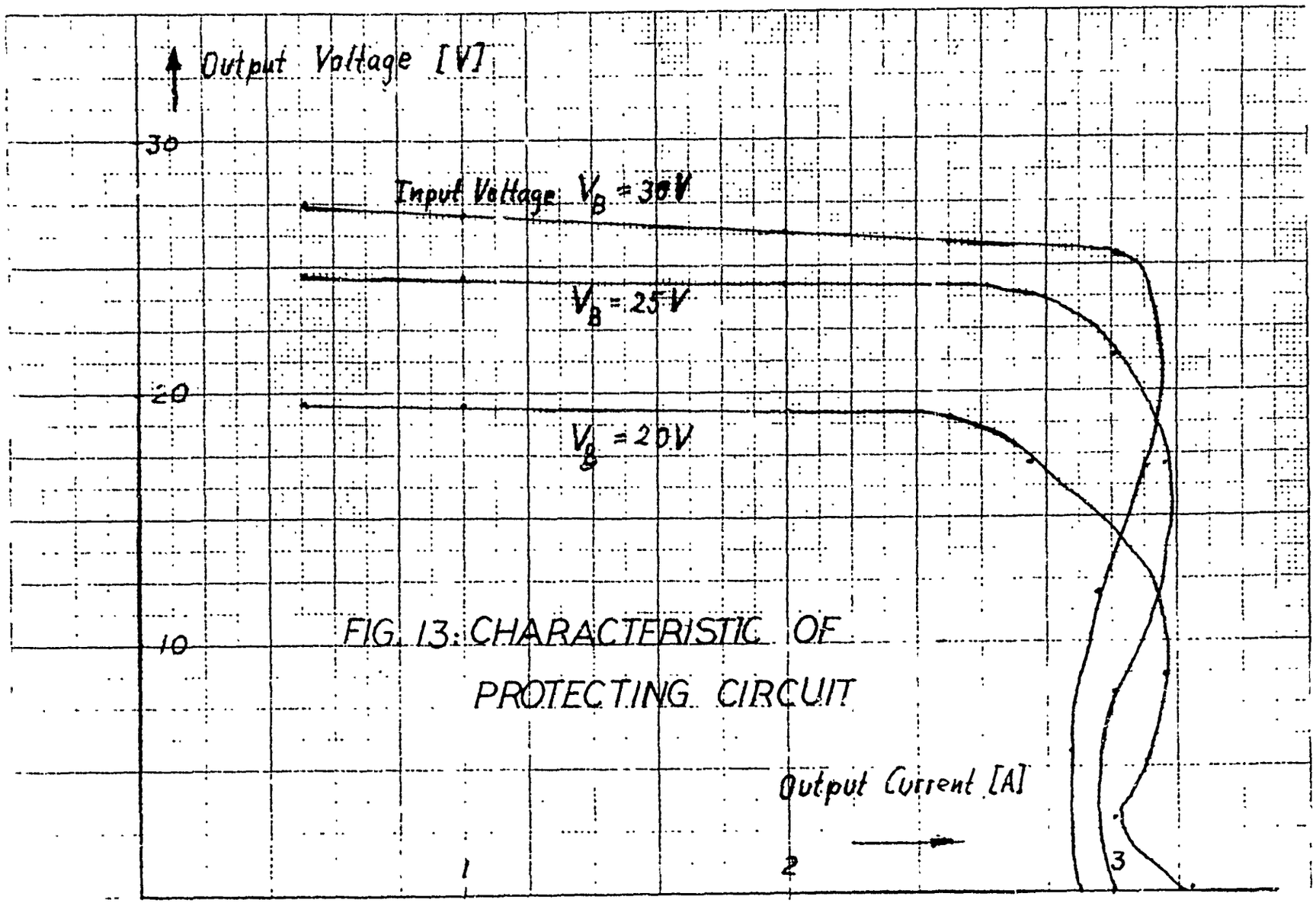
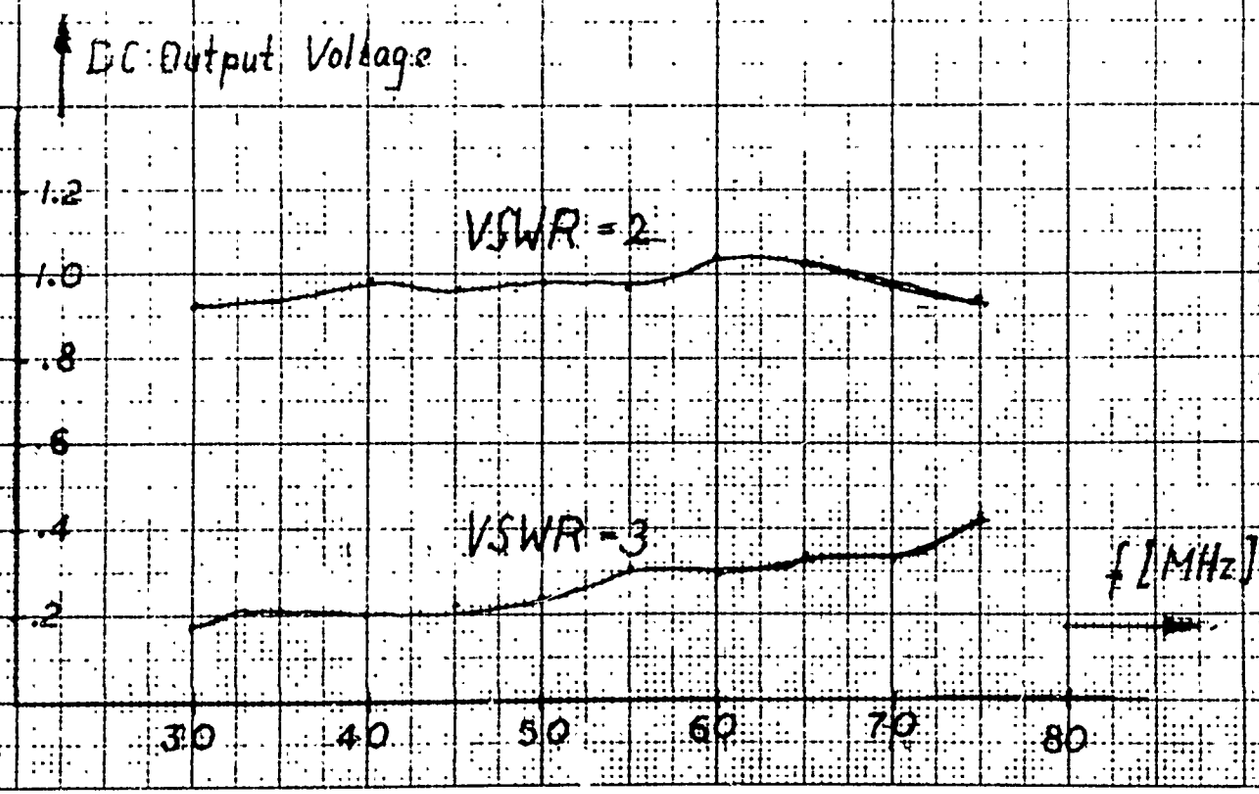


FIG. 13: CHARACTERISTIC OF PROTECTING CIRCUIT

FIG 14: Directional Coupler Performance at 40W to Load



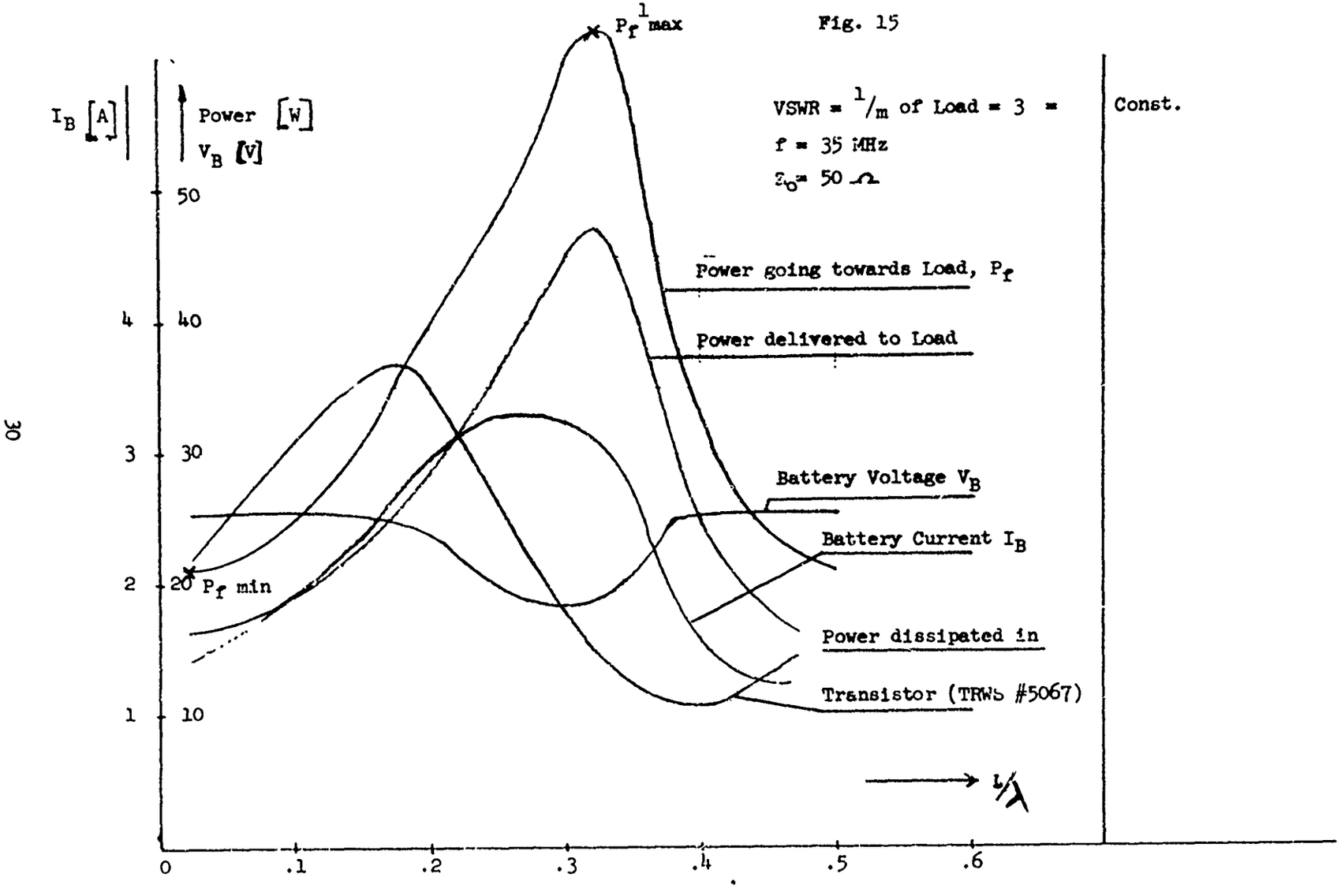

P_f^1 max
I_B [A]
Power [W]
V_B [V]
VSWR = $^1/_m$ of Load = 3 =
f = 35 MHz
Z_o = 50 Ω
Const.
Power going towards Load, P_f
Power delivered to Load
Battery Voltage V_B
Battery Current I_B
Power dissipated in
Transistor (TRWS #5067)
P_f min
l/λ
50
40
30
20
10
4
3
2
1
0
.1
.2
.3
.4
.5
.6

Fig. 15

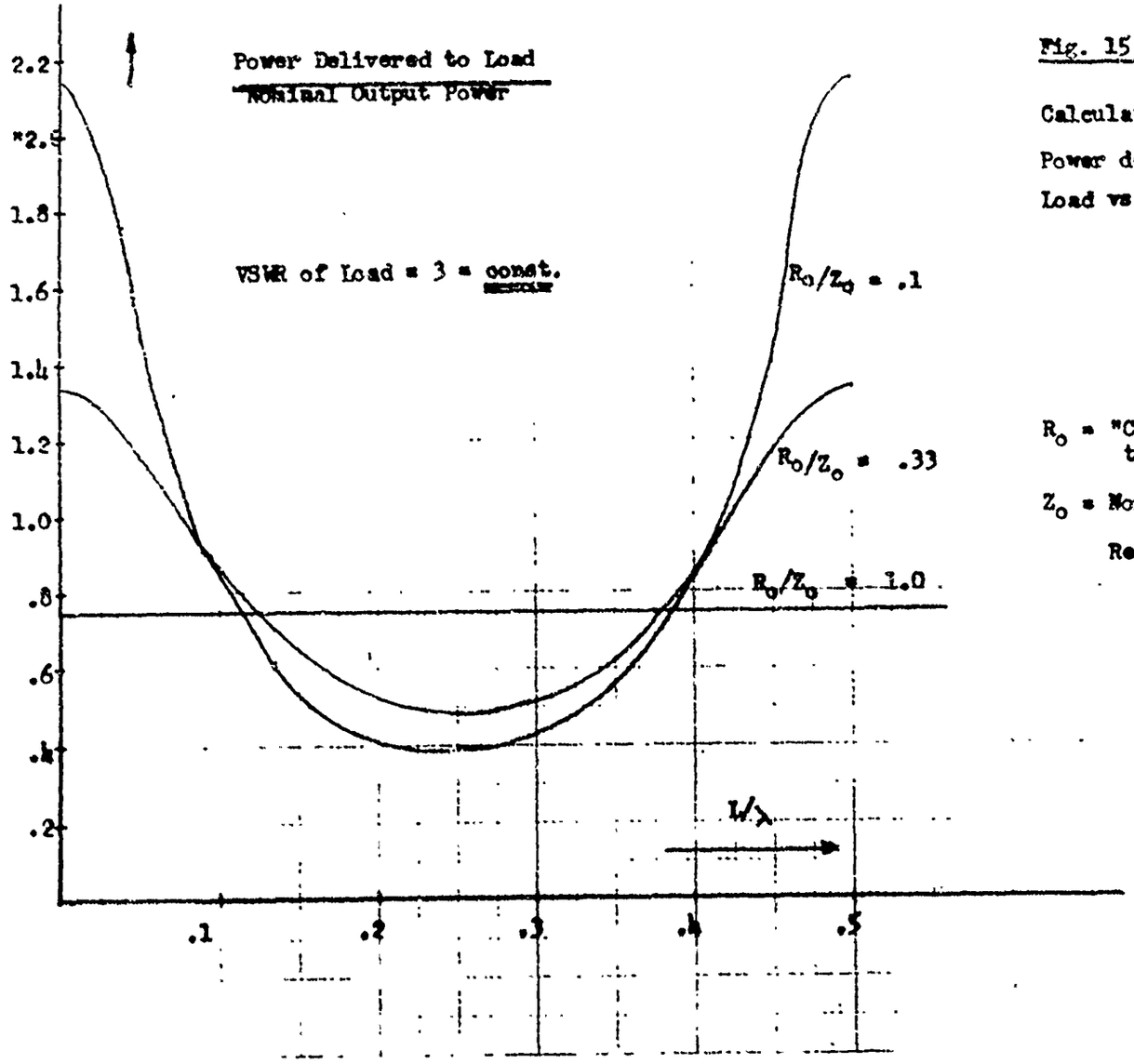


Fig. 15 a

Calculated Values of Power delivered to Load vs. Cable Length L.

R_o = "Characteristic Generator Resistance."

Z_o = Nominal Load Resistance = R_{Lo} = 50 Ω

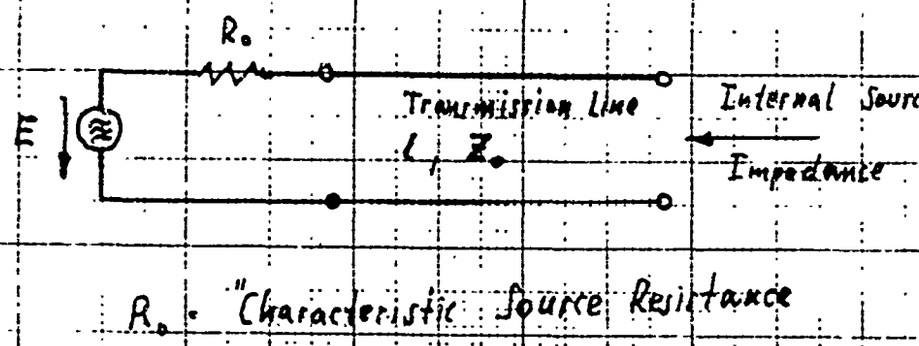
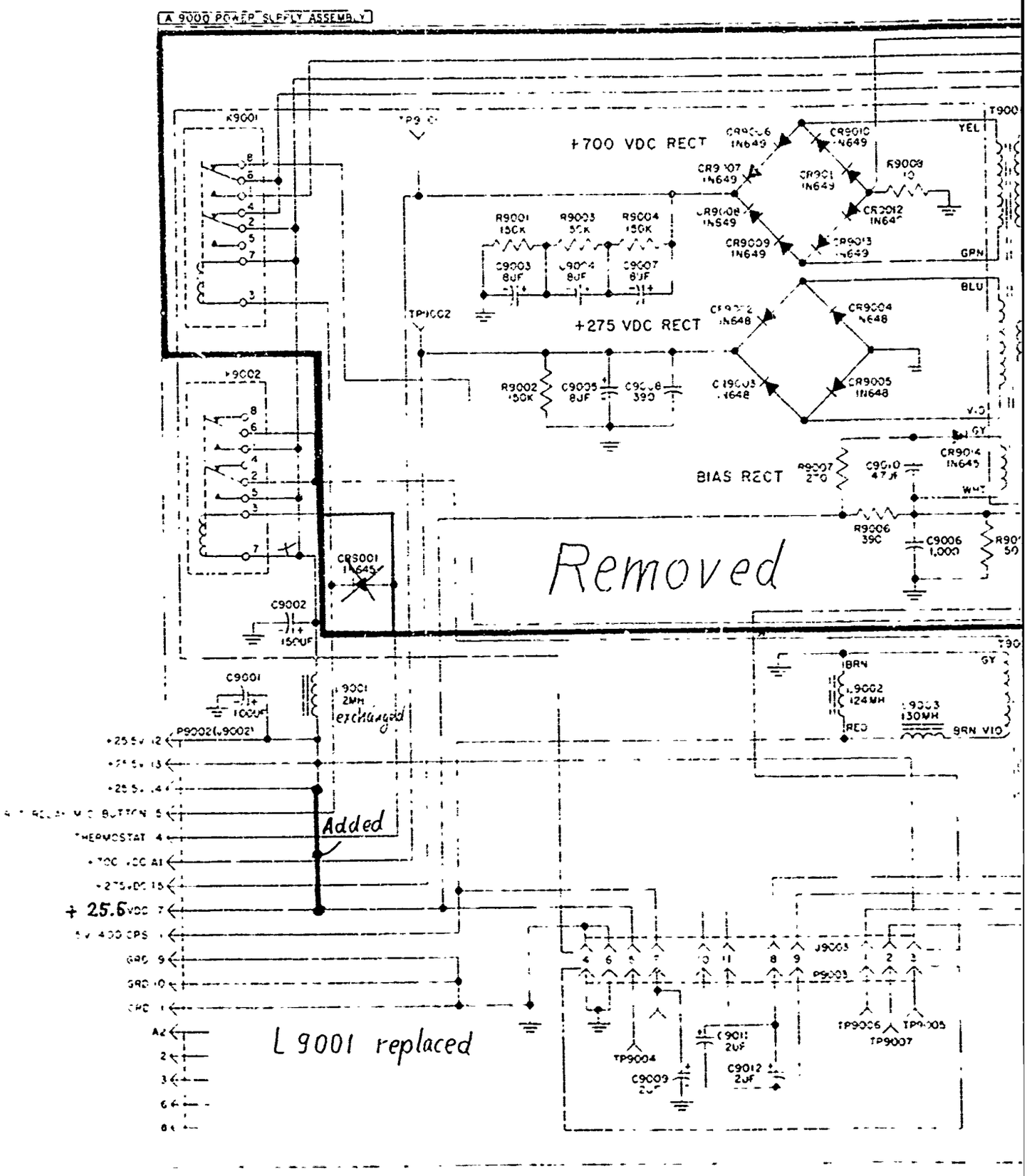


FIG. 16: EQUIV. CIRCUIT OF SOURCE


A 9000 POWER SUPPLY ASSEMBLY
+700 VDC RECT
+275 VDC RECT
BIAS RECT
Removed
Added
L 9001 replaced
+ 25.6VDC
THERMOSTAT
K9001
K9002
T900
YEL
GRN
BLU
GY
WHT
BRN
RED
ORN VIO
CR9001 1N649
CR9002 1N648
CR9003 1N648
CR9004 1N648
CR9005 1N648
CR9006 1N649
CR9007 1N649
CR9008 1N649
CR9009 1N649
CR9010 1N649
CR9012 1N649
CR9013 1N649
CR9014 1N649
R9001 150K
R9002 150K
R9003 150K
R9004 150K
R9006 390
R9008 10
C9002 150UF
C9003 8UF
C9004 8UF
C9005 8UF
C9006 1,000
C9007 8UF
C9008 390
C9009 2UF
C9011 2UF
C9012 2UF
L9001 2MH
L9002 124MH
L9003 130MH
P9002(J9002)
J9003
P9003
TP9004
TP9005
TP9006
TP9007
TP9002

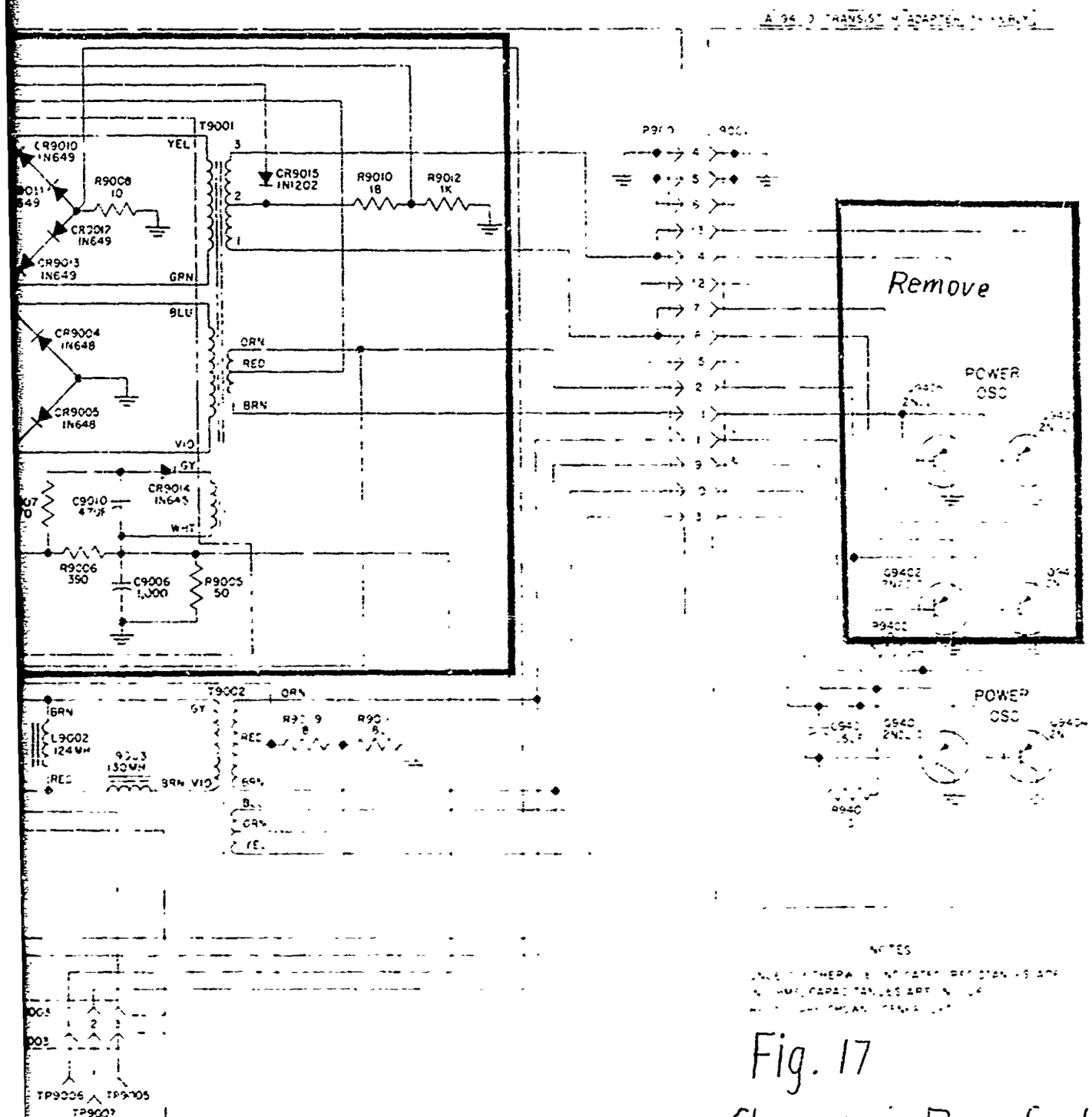


Fig. 17

Changes in Power Supply

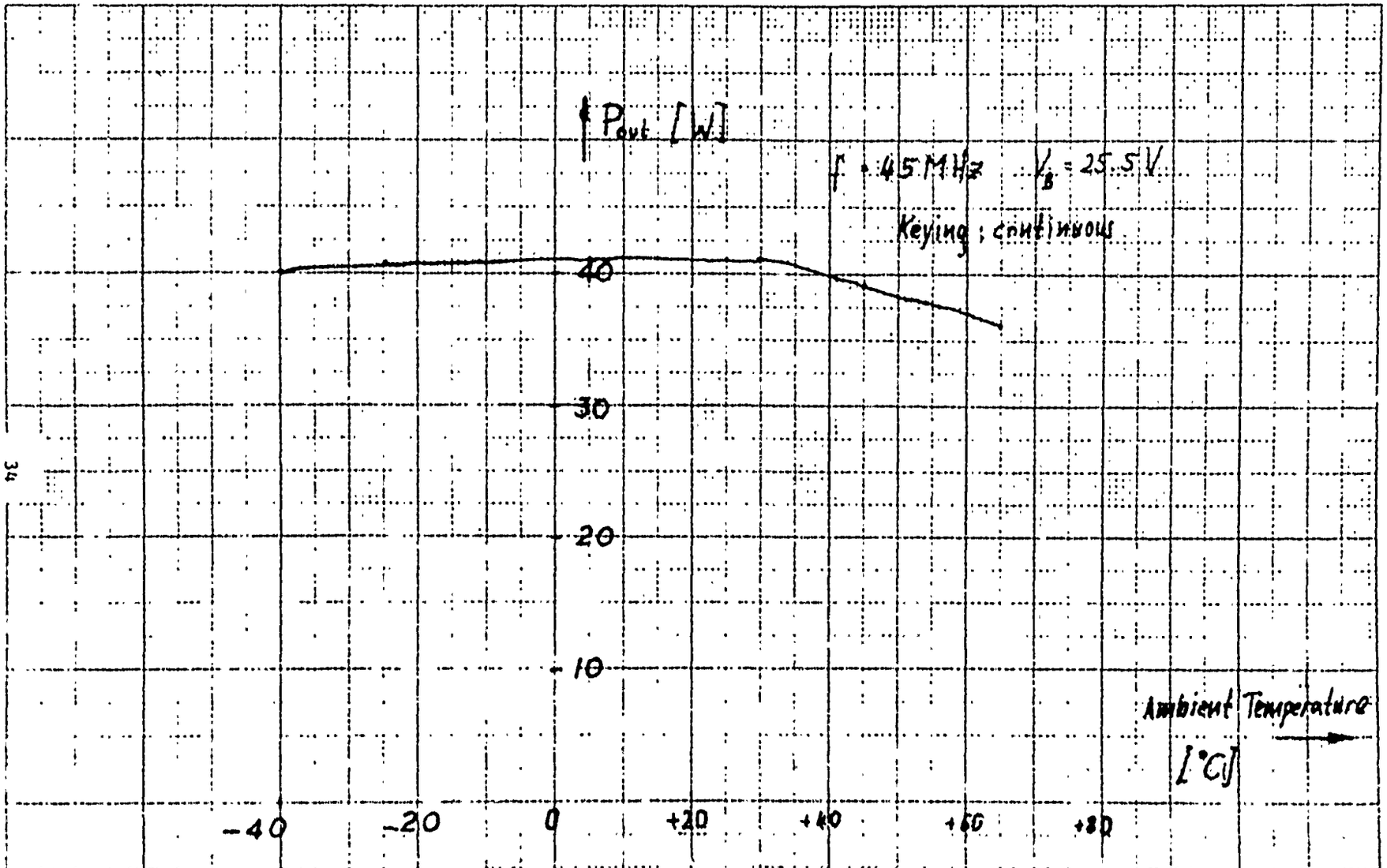


FIG. 18: PRELIMINARY THERMAL TEST

REPRINTED WITH PERMISSION BY ELECTRONIC DESIGN

6. APPENDIX B

Amplifier has 85% efficiency while providing up to 10 watts power over a wide frequency band. A switching transistor is the key in the design.

The efficiency of transistorized power amplifiers for cw operations can go as high as 85%, if the design is right. What's more, the amplifier can maintain its broadband characteristic.

The design approach centers on a transistor, used in a switching mode, that gives very high collector efficiencies.

To illustrate the advantages of this design, consider the output power stages of conventional vhf communication equipment. The output powers are usually around 50 watts. If the efficiency of the cw power amplifiers is 60%, a bulky heat sink is needed to handle the dissipated power. If the efficiency is increased up to 80%, the dissipated power is halved.

A common requirement of communication systems, that the second harmonic be suppressed at the output, also helps achieve the high efficiency, since it lowers the bandwidth demands on the amplifier. The output circuit of the power amplifier has to cover a frequency range of less than 1:2 for each position of the range-selecting switch. This assumption is valid because the low-pass or bandpass filters used to suppress the second harmonic cover a frequency range of 1:1.8 to 1:1.5.

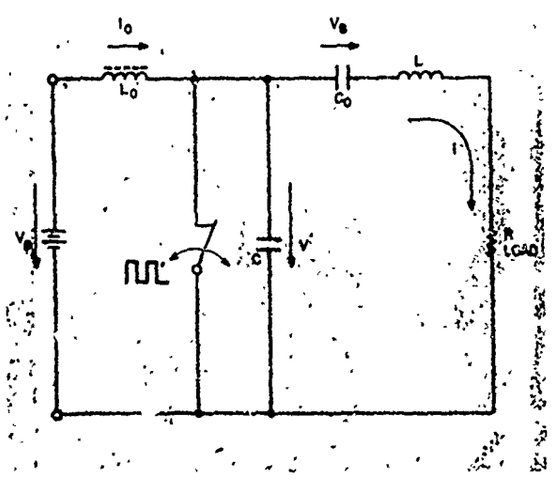
The criterion for achieving high collector efficiency is this: The impedance, seen from the base and looking into the matched circuit, should not be a short-circuit for the second harmonic over the whole frequency band.

The simplified equivalent circuit of the collector shows the transistor as a switch, being opened and closed periodically (Fig. 1). The choke L_o feeds a current to the collector of the transistor. The collector-emitter capacity is C, which, for simplicity, is assumed to be independent of voltage.

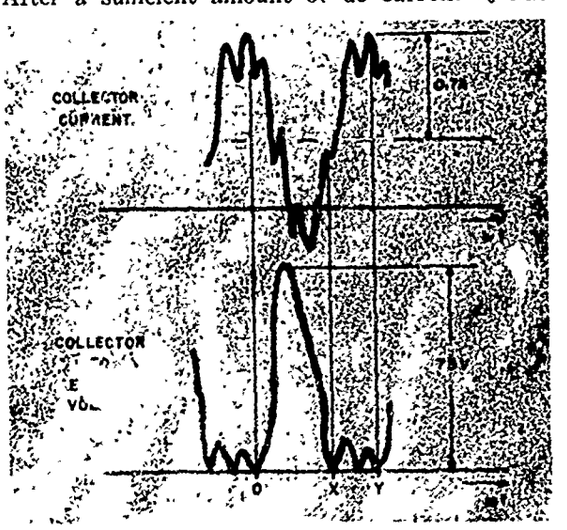
The dc blocking capacitor, C_o, charges to the battery voltage. For simplicity, an inductor L, in series with the load resistor R, has been substituted for a low-pass filter.

To understand the operation of the circuit, let us assume that the switch has just been closed. After a sufficient amount of dc current I_o has

Dieter R. Lohrmann, Development Engineer, USAECOM, Ft. Monmouth, N. J.



1. **Simplified equivalent circuit of transistor's collector circuit.** The ideal switch represents the transistor. It is operated by the signal fed to its base-emitter circuit by the dc current I_o. When the voltage across C crosses zero, the switch will be closed. The collector-emitter capacitance, C, is assumed to be independent of voltage. C_o, the dc blocking capacitor, charges to the battery voltage, V_B. The switch opens without any current.



2. **Traces of the collector voltage and collector current** indicate the open and closed periods. During the interval OX, the switch is open; during XY, it is closed. The collector current does not go to zero during the OFF period, because the collector capacitance is parallel to the switch. The transistor in the experiments is a type 2N3632, operated at 69 MHz. $I_C = 0.45$ A, $V_B = 25$ V. The output RF power is 9 watts with an input of 1 watt.

built up in L_o, the switch is opened. If the circuit of C, L and R is underdamped, the voltage across C will be a damped oscillation.

After the first half cycle, when the voltage crosses zero again, the switch is closed. This collector voltage, v, is traced in Fig. 2. During the interval OX the switch is open; during XY it is closed. The collector current, recorded in proper phase, is not zero during the switch-off cycle, because the collector's capacity is parallel to the switch. (Fig. 2)

It seems reasonable to assume that one of the reasons why this circuit yields high collector efficiencies is this one:

At the end of the OFF cycle the collector-emitter voltage returns to zero by itself, due to the transient. Hence the voltage will go down to zero, switching in the transistor, without need for any current. Thus only a little power will be dissipated in the switching slopes.

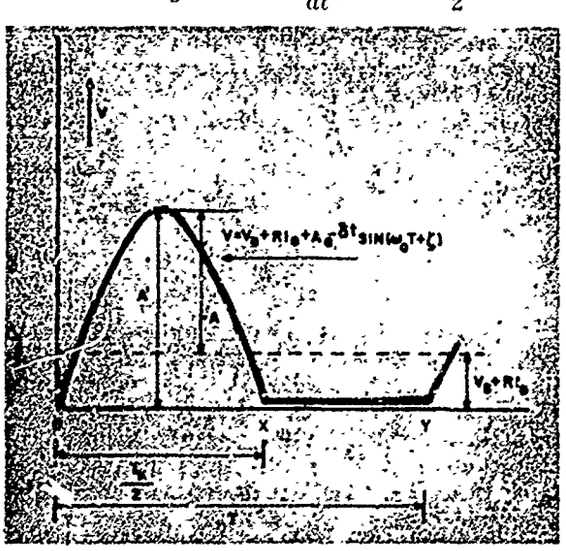
Linear equations described actions

Switching actions are usually described by a set of nonlinear differential equations. However, we can linearize them by using one set of linear equations for the ON interval and another for the OFF state.

The boundary conditions are provided by voltages and currents that must fit at the end of the preceding interval to those at the beginning of the next one. The equations for the open position (Fig. 3) are:

$$C\frac{dv}{dt} + i = I_o, \qquad (1)$$

$$v = V_B + Ri + L\frac{di}{dt}, \; 0 \le t \le \frac{T_k}{2}, \qquad (2)$$



3. The basic switching action results in a simplified mathematical model. One set of linear equations describes the OFF condition, another is used for the ON state. Voltages and currents are used as boundary conditions, to match the two intervals. The time between zero crossings of the collector voltage during the OFF period, T_k, is about equal to the period time of the transient in the collector voltage.

and for the switch closed:

$$v = 0, \qquad (3)$$

$$V_B + L\frac{di}{dt} + Ri = 0, \; \frac{T_k}{2} \le t \le T. \qquad (4)$$

These differential equations have very simple and well-known solutions. However, the equations for solving the boundary conditions become unwielding and nonlinear. Therefore we may use an approximate solution. The exact solution during the interval OA would be:

$$v = V_B + RI_o + Ae^{-\delta t}\sin(\omega_o t + \phi). \qquad (5)$$

For this, the approximate form is:

$$v \approx A'\sin\omega_k t, \quad 0 \le t \le \frac{T_k}{2} \qquad (6)$$

with $\omega_k = 2\pi/T_k$. Using Fourier analysis, we can show that the peak value of the fundamental voltage in the waveform is:

$$v = \frac{A'}{1+k} \cdot \mathrm{Sa}\left[\frac{\pi}{2}\left(1 - \frac{1}{k}\right)\right]. \qquad (7)$$

Here $k = T/T_k$ and $\mathrm{Sa}\, x = \sin x/x$. T_k is the time between the points of zero crossings of the collector voltage during the OFF period. It is approximately equal to the period time of a transient in the collector voltage—that is, in the case of Fig. 1 it is:

$$T_k = \frac{2}{\sqrt{\frac{1}{LC} - \frac{R^2}{4L^2}}}. \qquad (8)$$

The output power at fundamental frequency is:

$$P_{out} = \frac{v^2}{2}G_\omega, \qquad (9)$$

where G_ω is the real part of parallel admittance that the load circuit exhibits to the collector at the fundamental frequency.

Thus, in the case of Fig. 1:

$$G_\omega = Re\left(\frac{1}{R + j\omega L}\right) = \frac{R}{R^2 + \omega^2 L^2} \qquad (10)$$

Now A′ is still unkown, but it can be found from dc considerations. The choke in Fig. 1 cannot have a dc drop. Therefore the dc component of the waveform in Fig. 3 must be equal to the battery voltage, V_B. In other words:

$$V_B = \frac{1}{T}\int_0^T v\,dt = \frac{1}{T}\int_0^{T_k/2} A'\sin\omega_k t\,dt \cdot \qquad (11)$$

This yields

$$A' = \pi V_B\left(\frac{T}{T_k}\right), \qquad (12)$$

which is the peak collector voltage. If we put everything together, we have:

$$P_{out} = \frac{\pi^2}{2}\frac{V_B^2 k^2}{(1+k)^2}G_\omega \,\mathrm{Si}^2\left[\frac{\pi}{2}\left(1 - \frac{1}{k}\right)\right], \qquad (13)$$

where:

P_{out} = RF output power.

V_B = battery voltage.

T = 1/(signal frequency).

T_k = twice the length of cutoff interval—that is, twice the time between the zero crossings of the collector voltage. It is about equal to the cycle-time of a transient in the collector circuit. In our example:

$$T_k \approx \frac{2\pi}{\sqrt{\frac{1}{LC} - \frac{R^2}{4L^2}}}.$$

$k = T/T_k$.

G_ω = the real part of the load circuit's admittance exhibited toward the collector at the fundamental frequency. The collector's capacity is part of the circuit.

A' = the peak voltage at the collector.

If we let $T = T_k$, we get:

$$P_{out} = \frac{\pi^2}{8} V_B^2 G_\omega, \qquad (14)$$

and

$$A' = \pi V_B. \qquad (15)$$

As can be seen from these formulas, A′ tends to increase as the frequency decreases.

For all these calculations we assumed that the output circuit is less than critically damped. This means, in the case of Fig. 1, that:

$$\frac{1}{LC} >> \frac{R}{4L^2}. \qquad (16)$$

This condition is linked to the bandwidth of the load circuit, as shown in Fig. 4. The graph of the output power, P_{out}, over $V_B^2 G_\omega$ vs (signal frequency) (T_k), indicates a nonlinear dependence.

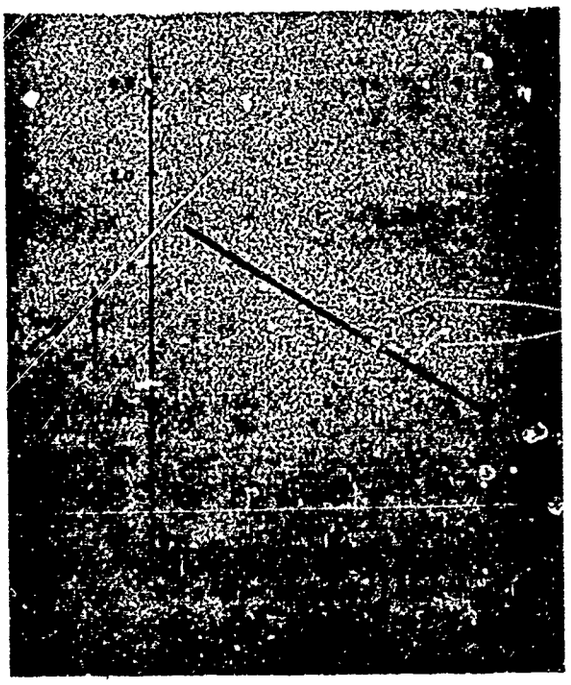
A filter (low pass) in the collector circuit slightly modifies the above equations. In principle, the same formulas apply. G_ω can be computed from the filter parameters. If $Z_T/R_\omega = 0.85$, then the eigenvalue of v will be $p_1 = (-0.168 + j0.895)(2\pi f_g)$, where f_g is the limiting frequency of the filter section—that is, the frequency at which the characteristic impedance turns imaginary. Hence $T_k = 0.895\ f_g$.

Practice modifies theories

Special requirements in an FM communication system demanded a power amplifier having about 10 watts output with a 10-dB amplification figure. It had to cover the frequency range of from 48 to 70 MHz with collector efficiencies up to 85%.

In the calculations thus far we neglected the following facts:

- The collector capacity of the transistor is dependent on voltage.
- There is strong coupling between the input and output circuits via the base-to-collector capacity, which is nonlinear. As a matter of fact, part of the input power might go directly to the output over this path.
- The switching will not happen instantaneously, and, most likely, *not* at exactly zero collector voltage.
- There are losses in the collector junction, which must be represented by a nonlinear resistor.
- In the practical circuit we have a low-pass filter section in the collector circuit and not as simple a circuit as the one assumed in Fig. 1 for the calculation.



4. The normalized output power is a nonlinear function of the normalized frequency. The ordinate is V^2G_ω, and the abscissa is the signal frequency times T_k.

We got more accurate results by replacing V_B with $V_B - V_{CE\,sat}$, in Eqs. 13 and 14. The latter factor is the collector-emitter saturation voltage. This voltage is normally given by the manufacturer only for dc. For RF it seems to be on the order of 2 to 5 volts for a good transistor. So let us assume 3 volts, since a good guess is better than none.

The constant factor in Eq. 13 has been found to be 0.85, rather than $\pi^2/8$, in the practical circuit. Thus Eq. 14 becomes:

$$G_\omega = \frac{1}{R_\omega} = \frac{P_{out}}{0.85\,(V_B - V_{CE\,sat})^2}.$$

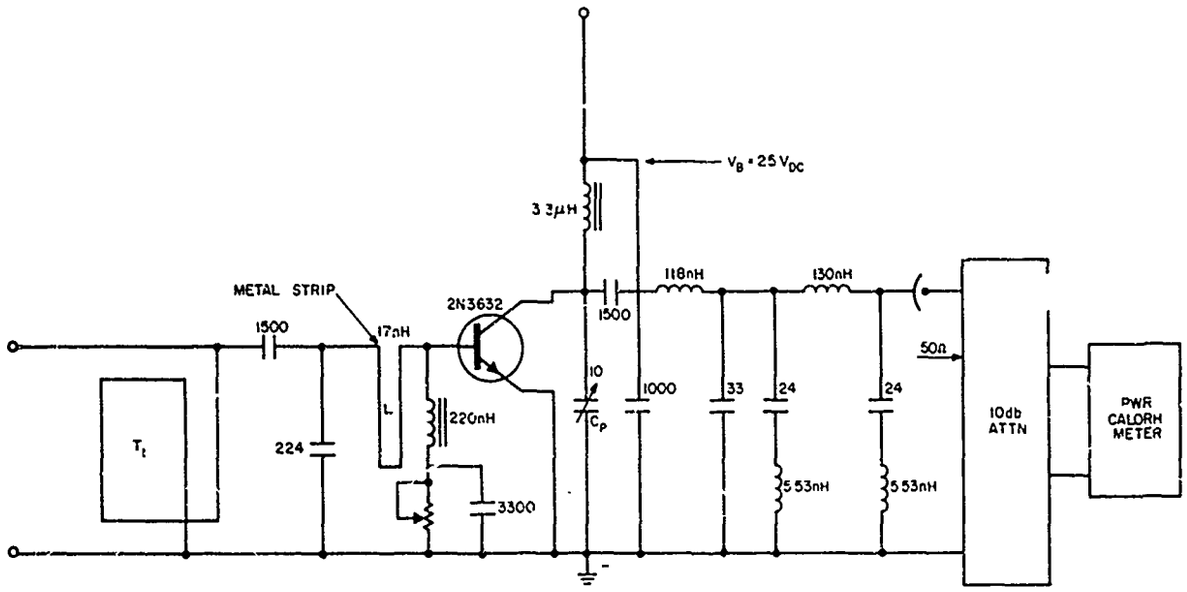
If P_{out} should be 10 watts at V_B = 25 V battery voltage, then G_ω = 24 mmho → 42 ohm is the load admittance the collector will have to see at the signal frequency. Once this R_ω is known, the first low-pass filter section can be designed. For the required amplifier, the frequency range desired was 50 to 75 MHz. Of course, the limiting frequency of the low-pass filter section must *not* be set to f_u = 75 MHz, since the mismatch is much too large at the limiting frequency.

It works well if one arranges:

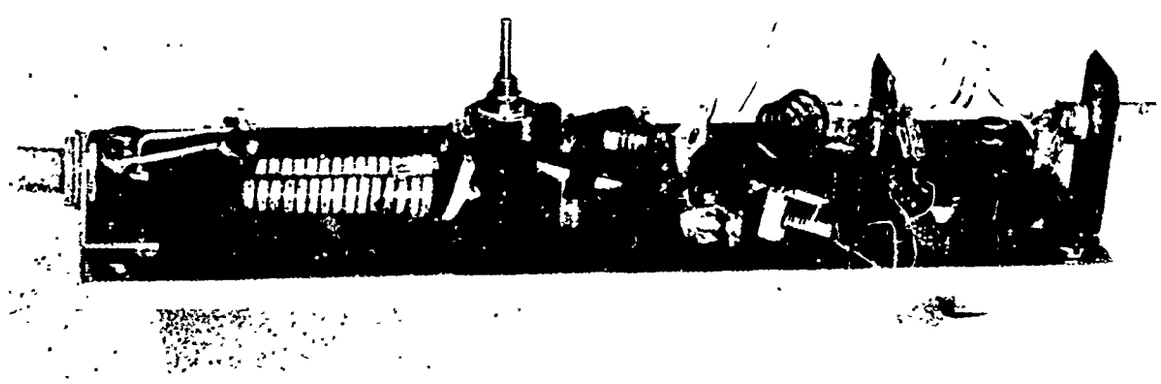
f_g = limiting frequency of low pass section $1.5\ f_u$ = 1.5 x 75 MHz.

The characteristic impedance of the low pass filter section is chosen as:

$$Z_T = 1.1 R_\omega = 1.1 \times 42\ \Omega = 46\ \Omega$$



5. **Broadband power amplifier** covers the frequency range from 48 to 70 MHz and has about 10 watts output power with 10-dB gain even if the dc path in the base circuit is eliminated. (The calorimeter monitors the output power.) To lessen the danger of parametric oscillation, feeding chokes should have the lowest possible values.



6. **Breadboard model** of the circuit in Fig. 5 illustrates construction gimmicks. The inductance of 17 nH, in the input matching circuit, is made of a piece of metal strip folded like a hairpin. It can be adjusted very conveniently with a slug, soldered across the two legs. Mica capacitors satisfy the need for very low series lead inductances.

The input capacity of the section is known from filter theory:

$$C = \frac{1}{2\pi f_g Z_\pi} = 31 \text{ pF}.$$

On the collector side, one has to deduct the collector's capacity from the value.

Again from filter theory:

$$L = \frac{Z_\pi}{\pi f_g} = 130 \text{ nH}.$$

The peak voltage between the collector and the emitter is expected to be in the midband.

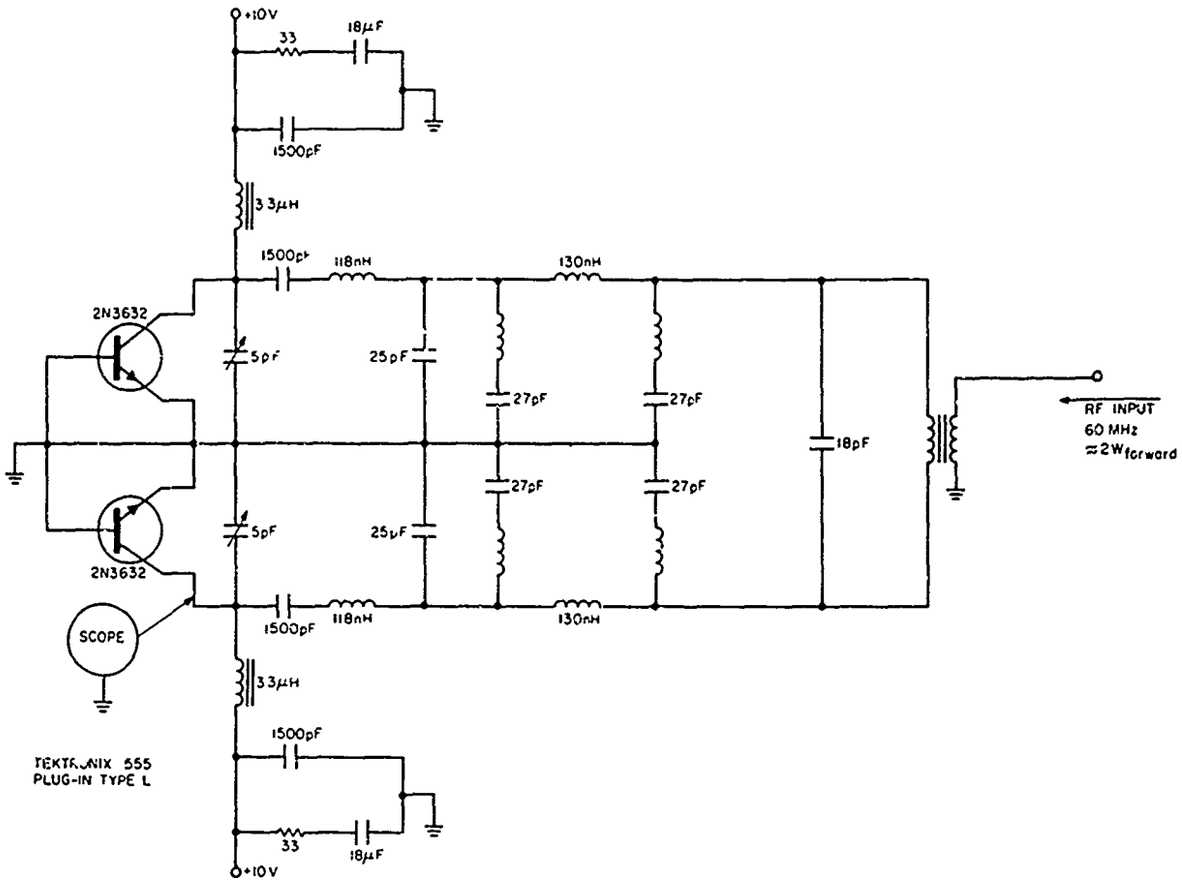
$$A' = \pi V_B = \pi 25 \text{ V} = 78 \text{ V},$$

which roughly agrees with measurements.

The pi filter section is followed by another one, to give additional harmonic suppression. It has an attenuation pole at a frequency that is twice that of the midband value.

Match input with transformers

To match the input, a broadband transmission line transformer T_1 is used (Fig. 5). It transforms from 50 down to 12.5 ohms, covering the entire range from 30 to 75 MHz. It is followed by



7. **Broadband push-pull amplifier** will brake out in parametric oscillations, even with its emitter and base leads grounded. (The oscilloscope picks up the parametric oscillations.) This proves the assumption that parametric oscillations are caused by a voltage-dependent capacitance in the collector and base circuit of the transistor, which gives rise to an undampening action. The capacitance acts like a varactor diode in a parametric amplifier.

a matching section made up of a 224-pF capacitor and an inductor of 17 nH. The capacitor must have very low series lead inductance. Therefore a mica capacitor of the button type was used. The inductor was made of a piece of metal strip folded like a hairpin. It can be adjusted very conveniently by soldering a slug across it (Fig. 6).

Under operational conditions, the equivalent input circuit for the transistor has been found to be a series resistor of about 5 ohms in the base lead, followed by a 1000-pF capacitor to ground.

Parallel to the 1000 pF capacitor is a resistor of about 10 ohms. About 200 pF of these 1000 are believed to be exhibited by the base-to-emitter junction. The remaining 800 pF is due to the Miller effect, caused by the collector's capacity. The circuit is valid between about 30 and 70 MHz.

The 224-pF capacitor, the 17-nH inductor and the equivalent input circuit of the transistor (5 ohms in series with about 1000 pF) form a broadband transformation network. It transforms the 5 ohms to 10 ohms and compensates for the 1000 pF.

The base is zero-biased. The biasing resistor is set to 10 ohms. At a 1-watt input driving power, the peak base-to-emitter reverse voltage is exceeded during the cutoff cycle, and reverse current is flowing. This reduces the dc current in the external base circuit. As a matter of fact, the transistor still gives 10 watts of output power if the dc path in the base circuit is eliminated. According to the manufacturer, the transistor is not damaged, provided that less than 25% of the permissible dc power for the base junction is dissipated by back conduction. The back conduction effect seems to increase the steepness of the switching slopes, thus improving the collector's efficiency.

The resistance in the base circuit was chosen as 100 ohms, for dc stability.

A 3.3-μH choke supplies the dc current to the collector. The collector circuit is formed by two low-pass pi filter sections that use the collector's capacity as part of the filter's input capacity. The trimmer in the collector line adjusts this output capacity. The second pi filter section has a pole at 141 MHz, to keep residual second harmonic power from being dissipated in the load.

A laboratory model, designed for 40 watts output power, is already being tested. The unit is similar to the one described below. It has the same high efficiency and a power gain of 10 dB.

Parametric oscillation is common problem

The tendency to break out in parametric oscilla-

tions seems to be a common problem to all high-efficiency switched power-amplifier stages.

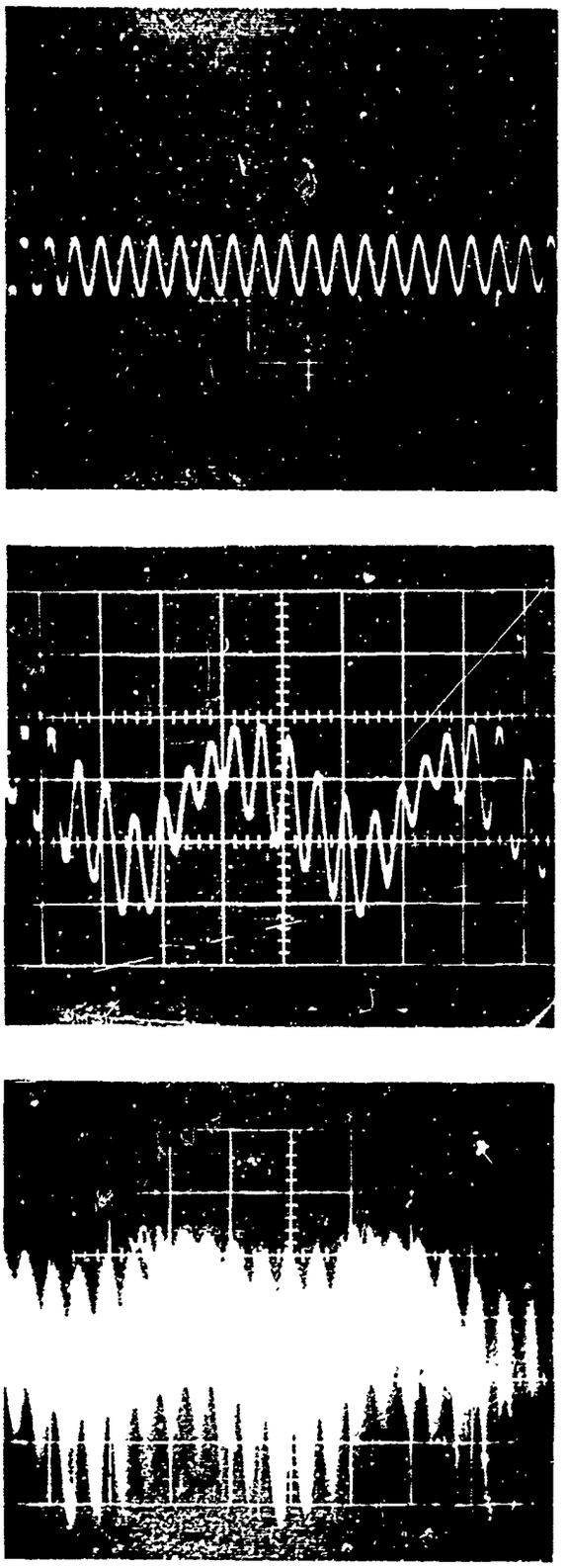
It is caused by the voltage-dependent capacitance in the collector and base circuits of the transistor. This capacitance is periodically varied by the signal itself. Therefore it acts like the varactor diode in a parametric amplifier (also called reactance amplifier). It is well known that such an amplifier may exhibit negative resistance under certain conditions, undampening the circuits. The circuit starts oscillating at some other frequency if this undamping action becomes too large. Therefore care must be taken to prevent the collector from sensing circuit resonances outside the passband, presumably at lower frequencies, down to dc. As a consequence, feeding chokes should have values as low as possible. On the power supply line, there must not be any resonant circuits at low frequencies. They can be eliminated by a resistor in series with a capacitor across the line.

The largest change of capacity of the collector junction occurs near zero collector voltage. Therefore linear stages, which do not use that portion of the characteristic, are much less likely to oscillate because of this effect; switched stages should be more troublesome.

A small collector capacity is more favorable than a large one for a given power output. Switching transistors with low losses—that is, with high efficiency—are more sensitive than those having higher losses. A small series resistor in the collector line helps, but it reduces the collector efficiency.

To prove this hypothesis of parametric oscillations, we took a broadband push-pull amplifier and grounded all emitter and base leads with short, broad-ribbon leads (Fig. 7). Then we fed signal power into the output terminals of the amplifier. Fig. 8 shows the voltage waveforms at the collector. At a 0.4-watt input (the normal output power of the amplifier in normal operation is 25 watts) we still had a sine voltage at the collector (Fig. 8, top). But at 1.7-watts input, an oscillation of lower frequency suddenly appeared (Fig. 8, middle). At 2.7 watts the oscillation became noise-like (Fig. 8, bottom). Since these oscillations tend to kill the transistors, the experiment was made with a 10-volt battery. The dc current in the transistors was zero until the oscillations started; then it jumped up to about 0.5 A, though the bases were grounded.

The differential equations describing these effects are basically of the Mathieu and Hill types, and the solutions have stable and unstable regi[illegible]. ■ ■



8. Collector waveforms show effects of increased input and the [illegible] At an input of 0.4 [illegible] at 1.7 watts (middle) [illegible] appears which becomes [illegible] 2.7 watt input (bottom) The dc current [illegible] A from zero as oscillations started, despite the grounded bases

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Security Classification

DOCUMENT CONTROL DATA - R & D

(Security classification of title, body of abstract and indexing annotation must be entered when the overall report is classified)

1. ORIGINATING ACTIVITY *(Corporate author)*
U. S. Army Electronics Command
Fort Monmouth, New Jersey 07703

2a. REPORT SECURITY CLASSIFICATION
UNCLASSIFIED

2b. GROUP

3. REPORT TITLE
HIGH-EFFICIENCY TRANSISTOR CW RF POWER AMPLIFIER

4. DESCRIPTIVE NOTES *(Type of report and inclusive dates)*
Technical Report

5. AUTHOR(S) *(First name, middle initial, last name)*
Dieter R. Lohrmann

6. REPORT DATE
May 1967

7a. TOTAL NO. OF PAGES
40

7b. NO. OF REFS
5

8a. CONTRACT OR GRANT NO.

b. PROJECT NO. 1E6 34301 D246

c. Task No. 01

d. Subtask No. 22

9a. ORIGINATOR'S REPORT NUMBER(S)
ECOM-2836

9b. OTHER REPORT NO(S) *(Any other numbers that may be assigned this report)*

10. DISTRIBUTION STATEMENT
Distribution of this document is unlimited.

11. SUPPLEMENTARY NOTES

12. SPONSORING MILITARY ACTIVITY
U. S. Army Electronics Command
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Fort Monmouth, New Jersey 07703

13. ABSTRACT

Research work on VHF transistor power amplifiers of the switched type demonstrates semi broadband designs yielding collector efficiencies as high as 80% at 40W output in the frequency range from 30 to 76 MHz. A laboratory model of such an amplifier was built. It was designed to fit the AN/VRC-12.

DD FORM 1 NOV 65 1473 REPLACES DD FORM 1473, 1 JAN 64, WHICH IS OBSOLETE FOR ARMY USE.

(1)

Security Classification

Security Classification

14. KEY WORDS	LINK A ROLE	LINK A WT	LINK B ROLE	LINK B WT	LINK C ROLE	LINK C WT
VHF transistor						
power amplifiers						
switching transistor						
matching circuit						
transformers						

Security Classification