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A WIDE-BAND LOW-PASS AMPLIFIER USING A PENTODE-TO-CATHODE FOLLOWER TUBE PAIR PER STAGE

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ABSTRACT

This report discusses an amplifier circuit useful for bandwidths approaching 100 mc and gains to 45 db. The use of cathode followers between pentodes simplifies the low frequency coupling problem and contributes to the high-frequency gain of the pentode stage by providing a smaller capacitive load. The consequent higher impedance in the pentode plate circuit increases the maximum linear plate voltage excursion by perhaps three times that of normal video amplifiers having equal bandwidths. The design lends itself to the use of triode-pentode tubes such as the 6U8. Several variations of this design have been built including one five-tube amplifier having 43 db gain with 1 db ripple to 70 mc and usable output to 89 mc.

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1. INTRODUCTION

This study resulted from the need for a wide-band noise source. An attractive method of obtaining the required noise level was to feed a broadband amplifier with a low-level noise generator. The resulting specifications on amplification, bandwidth, and output level for the amplifier were stringent enough to stimulate a study of cascade techniques. The inquiry raised two questions: (1) What network configurations would provide the greatest gain for a given ripple? and (2) what chassis construction techniques would minimize shunt capacities? One of the circuits reviewed is considered novel enough to warrant this description.

2. CIRCUIT DESIGN

Figures 1 and 2 show the circuit and its equivalent network. The circuit is a pentode grounded-cathode amplifier coupled by a three-pole network to a cathode follower. The cathode follower is direct-coupled by a three-pole network to the next stage. The amplifier has six poles; two of them are on the negative real axis. An approximate equivalent circuit for design purposes is obtained by neglecting the cathode follower grid-cathode capacitance C_3, and cathode ground capacitance C_4. Neglecting C_3 is tantamount to the assumption of independence of the two networks. Neglecting C_4 is justified if the pole on the negative real axis is far from the origin compared to the other poles.

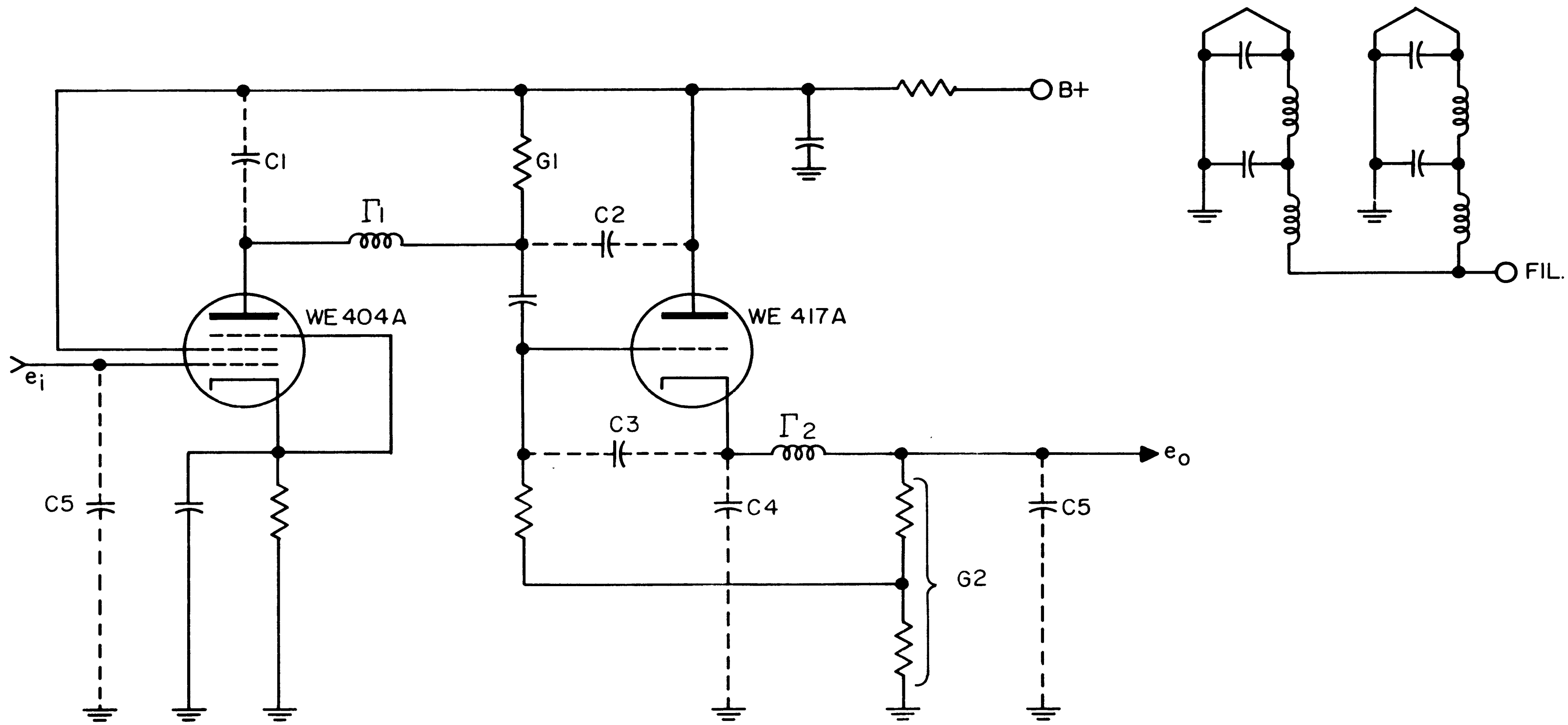

FIG. 1 AMPLIFIER STAGE

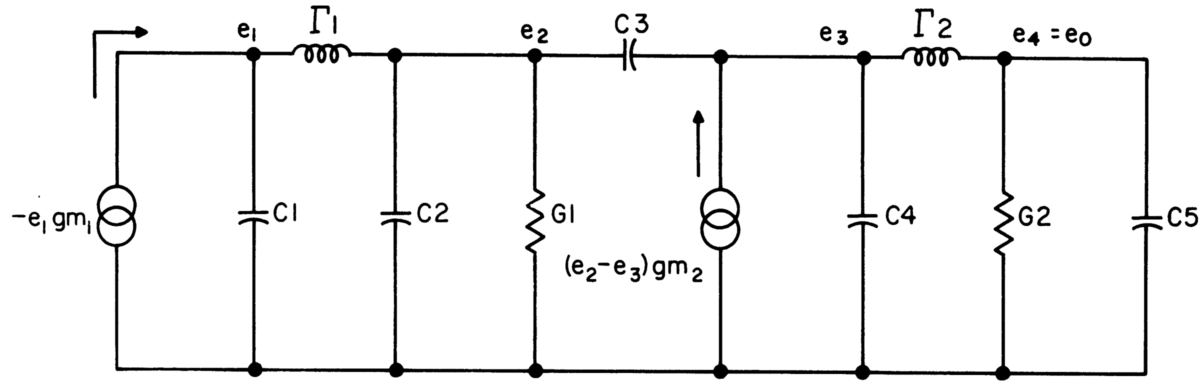

Figure 2 High Frequency Equivalent Circuit

The simplified amplifier has five poles; it will be established that, qualitatively, desirable element values tend to result if, in the complex frequency plane, the pentode can provide the real pole and the outer two complex conjugate poles, while the cathode follower provides the inner two complex conjugate poles. If the poles of the transfer function are located at

$$\begin{aligned} p_o &= \alpha_o \omega_{oc} \\ p_1 &= \omega_{oc}(\alpha_1 + j\beta_1) \\ \overline{p_1} &= \omega_{oc}(\alpha_1 - j\beta_1) \\ p_2 &= \omega_{oc}(\alpha_2 + j\beta_2) \\ \overline{p_2} &= \omega_{oc}(\alpha_2 - j\beta_2) \end{aligned} \qquad (1)$$

Then by solving the simultaneous equations obtained by equating the coefficients of the polynomial of the network denominator (of an all pole network) in terms of the element values to the coefficients generated by Eq 1, one finds the following approximate design equations.

$$\Gamma_1 = \frac{\alpha_o\ (\alpha_2^{\ 2} + \beta_2^{\ 2})\ \omega_{co}^{\ 2}\ C_2}{\alpha_o + 2\alpha_2} \qquad (2)$$

$$G_1 = (\alpha_o + 2\alpha_2)\ \omega_{co}\ C_1 \qquad (3)$$

$$\Gamma_2 = (\alpha_1{}^2 + \beta_1{}^2)\ \omega_{co}{}^2\ C_5 \qquad (4)$$

$$G_2 = 2C_5\ \omega_{co}\alpha_1 \qquad (5)$$

where the small Greek letters indicate pole locations normalized to 1 rad/sec. bandwidth and ω_{co} is the amplifier bandwidth (see Fig. 3).

It is found experimentally that when the values of ω_{co} approach the extreme bandwidth capabilities of the amplifier, Eq 5 should be more nearly

$$G_2 \approx C_5\omega_{co}\ \alpha_1 \qquad (6)$$

This is due to the fact that the interstages are not independent since C_3 is not truly negligible. Equations 2 to 5 determine the element values in terms of the locations of the poles of the transfer functions. Pole locations should be chosen corresponding to any of the classical functions approximating the ideal rectangular magnitude of gain characteristics. As an example, for a Tchebycheff five pole configuration, the pole locations are:

$$\alpha_k + j\beta_k = -\sinh a \cos \frac{\pi}{5} k \pm j \cosh a \sin \frac{\pi}{5} k$$

where

$$k = 0,\ 1,\ 2$$

$$a = \frac{1}{5} \sinh^{-1} \frac{1}{\sqrt{\epsilon}} \qquad (7)$$

$$\epsilon = \frac{\text{max. gain in pass band}}{\text{min. gain in pass band}} - 1$$

It will be recalled that the above analysis assumes C_3 and C_4 to be negligible. While the result is simple, the error will require considerable adjustment of the constructed amplifier.

A more exact solution requires the finding of the roots of the network determinant of the equivalent circuit (Fig. 2). The network determinant is

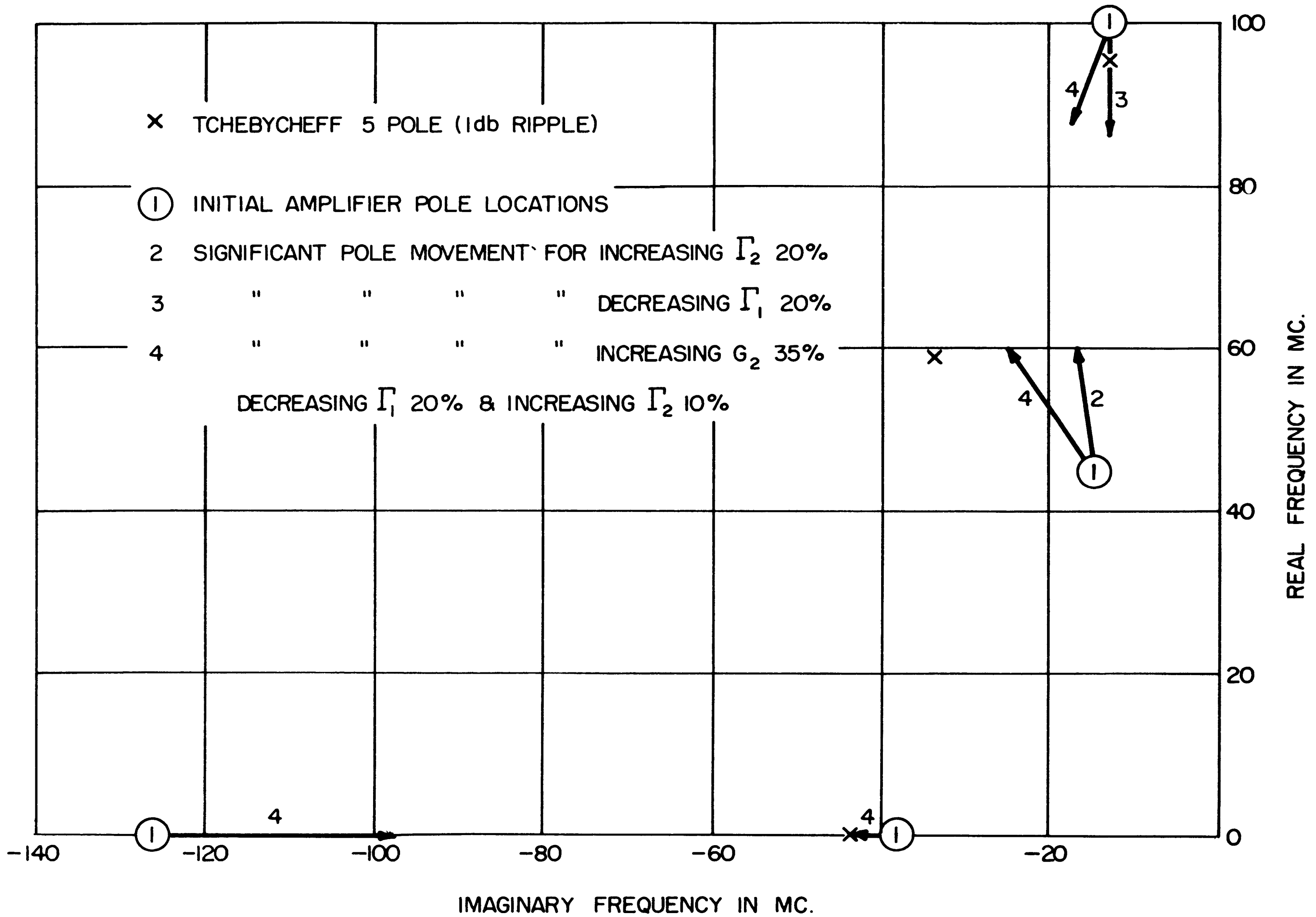

FIG. 3 NETWORK POLE LOCATIONS

$$\Delta = \begin{bmatrix} C_1 p + \dfrac{\Gamma_1}{p} & -\dfrac{\Gamma_1}{p} & 0 & 0 \\ -\dfrac{\Gamma_1}{p} & (C_2+C_3)p+G_1 + \dfrac{\Gamma_1}{p} & -pC_3 & 0 \\ 0 & -(pC_3+g_{m2}) & (C_3+C_4)p+g_{m2} + \dfrac{\Gamma_2}{p} - & -\dfrac{\Gamma_2}{p} \\ 0 & 0 & -\dfrac{\Gamma_2}{p} & pC_5+G_2 + \dfrac{\Gamma_2}{p} \end{bmatrix} \qquad (8)$$

The amplifier gain is then:

$$\text{Gain} \approx -\frac{g_{m1}\Delta_{14}}{\Delta} = -\frac{\Gamma_1\Gamma_2 g_{m1} g_{m2}(pC_3+g_{m2})}{p^2\Delta} \approx -\frac{\Gamma_1\Gamma_2 g_{m1} g_{m2}}{p^2\Delta} \qquad (9)$$

where:

$$\begin{aligned} p^2\Delta = \; & C_1C_5(C_1C_3 + C_1C_4 + C_3C_4)p^6 + C_1{}^2C_5 g_{m2} p^5 \\ & + \Big[g_{m2}C_1(C_1G_2 + C_5G_1) + C_1\Gamma_2(C_1C_3 + C_1C_4 + C_1C_5 + C_3C_4 + C_3C_5) \\ & + C_5\Gamma_1(2C_1C_3 + 2C_1C_4 + C_2C_4)\Big]p^4 \\ & + g_{m2}C_1(C_1\Gamma_1 + 2C_5\Gamma_1 + G_1G_2)p^3 \\ & + \Big[g_{m2}(G_1C_1\Gamma_2 + 2G_2C_1\Gamma_1 + G_1C_5\Gamma_1) \\ & + \Gamma_1\Gamma_2(2C_1C_3 + 2C_1C_4 + 2C_1C_5 + C_3C_4 + C_3C_5)\Big]p^2 \\ & + g_{m2}\Gamma_1(2C_1\Gamma_2 + G_1G_2)p \\ & + (g_{m2} + gp)\, G_1\Gamma_1\Gamma_2 \end{aligned} \qquad (10)$$

under the restrictions

$$g_{m2} \gg G_2$$

$$g_{m2} \gg G_1$$

$$g_{m2} \gg G_p$$

The network zero (from Δ_{14}) will be neglected since it is very far out on the negative real axis. For computational purposes one might rewrite this as:

$$
\begin{aligned}
p^2\Delta = {} & A_6 p^6 + A_5 p^5 + (a_4 G_1 + b_4 G_2 + C_4 \Gamma_1 + d_4 \Gamma_2)p^4 \\
& + (a_3 \Gamma_1 + b_3 \Gamma_2 + C_3 G_1 G_2)p^3 \\
& + (a_2 G_1 \Gamma_2 + b_2 G_2 \Gamma_1 + C_2 G_1 \Gamma_1 + d_2 \Gamma_1 \Gamma_2)p^2 \\
& + (a_1 \Gamma_1 \Gamma_2 + b_1 G_1 G_2 \Gamma_1)p \\
& + A_o \Gamma_1 \Gamma_2 G_1 \qquad (11)
\end{aligned}
$$

To investigate the roots of the determinant the following parameters might initially be assumed as reasonable for an 85-mc amplifier having about 0.34 db ripple.

$C_1 = 5\ \mu\mu f$		$G_1 = 1.5 \cdot 10^{-3}$ mhos	
$C_2 = 3\ \mu\mu f$		$G_2 = 1.0 \cdot 10^{-3}$ mhos	
$C_3 = 6\ \mu\mu f$		$1/\Gamma_1 = 1.0\ \mu$henry	
$C_4 = 5\ \mu\mu f$		$1/\Gamma_2 = 1.0\ \mu$henry	
$C_5 = 9\ \mu\mu f$			

$g_{m1} = 12{,}500\ \mu$mhos (WE 404A)

$g_{m2} = 24{,}000\ \mu$mhos (WE 417A)

The pole locations are found by Lin's method (Footnote 1) to be located roughly at the positions of the Tchebycheff function of Eq 7. These pole loci are compared in Fig. 3. The pole positions can be shifted to more desirable positions by an iterative procedure. To this end, the parameters of the network are perturbed slightly in order to find a linear operator relating each parameter adjustment and the corresponding pole movements. The plate load resistance of the pentode ($1/G_1$) establishes the gain and is assumed to be fixed

1. Gordon S. Brown, and Donald P. Campbell, Principles of Servomechanisms, John Wiley and Sons, 1948, page 89.

by the gain requirement. If parameters G_2, Γ_1, and Γ_2 are perturbed by ϵG_2, $\epsilon\Gamma_1$ and $\epsilon\Gamma_2$ and the roots of the network determinant solved, one concludes that, to a good working approximation:

$$\begin{aligned} \epsilon\alpha_1 &\approx K_1 \epsilon G_2 \\ \epsilon\beta_1 &\approx K_2 \epsilon \Gamma_2 \\ \epsilon\beta_2 &\approx K_3 \epsilon \Gamma_1 \end{aligned} \tag{13}$$

Consequently one can estimate the parameter correction required to move the poles to a desired location. Figure 3 shows the pole locations after one correction compared to the pole locations of the Tchebycheff frequency response. This agreement is probably sufficiently close considering the accuracies of the estimated capacitances and the effect of the sixth pole.

In moving the poles of the network, it is often an aid to note that:

$$\epsilon\alpha_0 + 2\epsilon\alpha_1 + 2\epsilon\alpha_2 + \epsilon\alpha_3 = 0 \tag{14}$$

for perturbations of G_1, G_2, Γ_1, and Γ_2. This is a consequence of the more general statement to the effect that the highest order term of the difference equation of a polynomial with perturbed roots is simply the sum of the real parts of the root perturbations. For this sixth order amplifier the highest perturbation term is the p^5 term, which does not include in its coefficient the parameters G_1, G_2, Γ_1, and Γ_2. Therefore the sum of the real parts of the root perturbations are invariant in G_1, G_2, Γ_1, and Γ_2.

To include the remarks on the network design, two statements are necessary concerning restraints imposed by the available tubes. The real part of the inner complex conjugate poles (β_1) is limited by the g_m of the cathode follower and the capacitance, C_5. This limitation appears to be a serious restraint on the bandwidth when G_1, G_2, Γ_1, and Γ_2 are adjusted for flat frequency response over a very wide bandwidth.

The second more critical restraint is imposed by the cathode follower grid-cathode capacitance, C_3. A slight increase in C_3 from the values used in this report results in the two real network poles joining and then becoming complex conjugates. One of the other pairs of poles appears to move to the right half plane. It is not known if there are stable pole locations for larger C_3 or if the pole locations could be made to approximate some useful configuration such as a Tchebycheff six-pole. This, in effect, makes a very sharp frequency limitation on the amplifier bandpass which must be observed before the amplifier can be adjusted. In the amplifier using WE 417A tubes, the grid has four grid pin connections. The two pins closest to the cathode were clipped off and the socket terminals for those pins were removed. There is strong evidence justifying these extreme measures.

The low-frequency response of the amplifier is determined by cathode and screen by-pass condensers and the coupling condenser between the pentode network and cathode-follower grid. The circuit of Fig. 1 (having dc coupling every other stage) has the advantage of requiring only half as many by-pass and coupling elements as more conventional amplifiers with like characteristics. In addition, the direct-current input impedance of the cathode follower is much higher than that of a high gain pentode. It appears that this circuit increases the low-frequency response by perhaps one order of magnitude.

3. CONSTRUCTION AND TESTING

In constructing the various amplifiers, all ground potential conductors were removed from the immediate vicinity of the network elements. A fiber chassis was used with no ill effects; indeed, removing the chassis removed a potential feedback loop.

In these amplifiers, the tube sockets were stripped of their metal mounting rings and shields and were sandwiched between fiber sheets. The ground, filament, and B+ buss wires were laid parallel to each other and about one inch above the tube sockets. The network elements were placed alongside the tube sockets and against the fiber sheets. To reduce feedback, special effort was given to the reduction of the area enclosed by the main current loops in the networks. With these techniques, stray capacitance was less than 2 $\mu\mu$fd at any network junction. Figure 6 is a photograph of a typical amplifier.

The amplifier could have been shielded by enclosing it in the center of a rather large box. However, to reduce the dimensions of the complete unit, the amplifier was set in a 3-inch slot in a conventional chassis. The proximity of this chassis to the amplifier had the effect of providing a regenerative feedback path rather than introducing shunt capacity. The resulting oscillations were eliminated by breaking up the loop of the slot with ground straps from the metal chassis to the ground buss wire of the amplifier. The positions for the several straps were not particularly critical, but were chosen by observing the sensitivity of the amplifier response to their application. In one amplifier this chassis feedback path was controlled sufficiently to increase the bandwidth by 15 mc by providing two more poles near the $j\omega$ axis (Fig. 5).

Figures 4 and 5 show typical frequency-response curves for two amplifiers having 43 db gain. The curve of Fig. 4 is that of two stages of the described amplifier plus one low-gain 3-pole stage added to obtain just 43 db gain. One db of ripple was obtained from 10 to 70 mc; no attempt was made to reduce the peak at 85 mc. Standard miniature coils were used.

The curve of Fig. 5 shows the response of a carefully adjusted earlier amplifier (using 6U8 tubes) having 2 db ripple and the apparent response of a 7-pole network. The extra two poles were introduced by the chassis feedback mentioned earlier as having provided 15 mc additional bandwidth. The smoothness

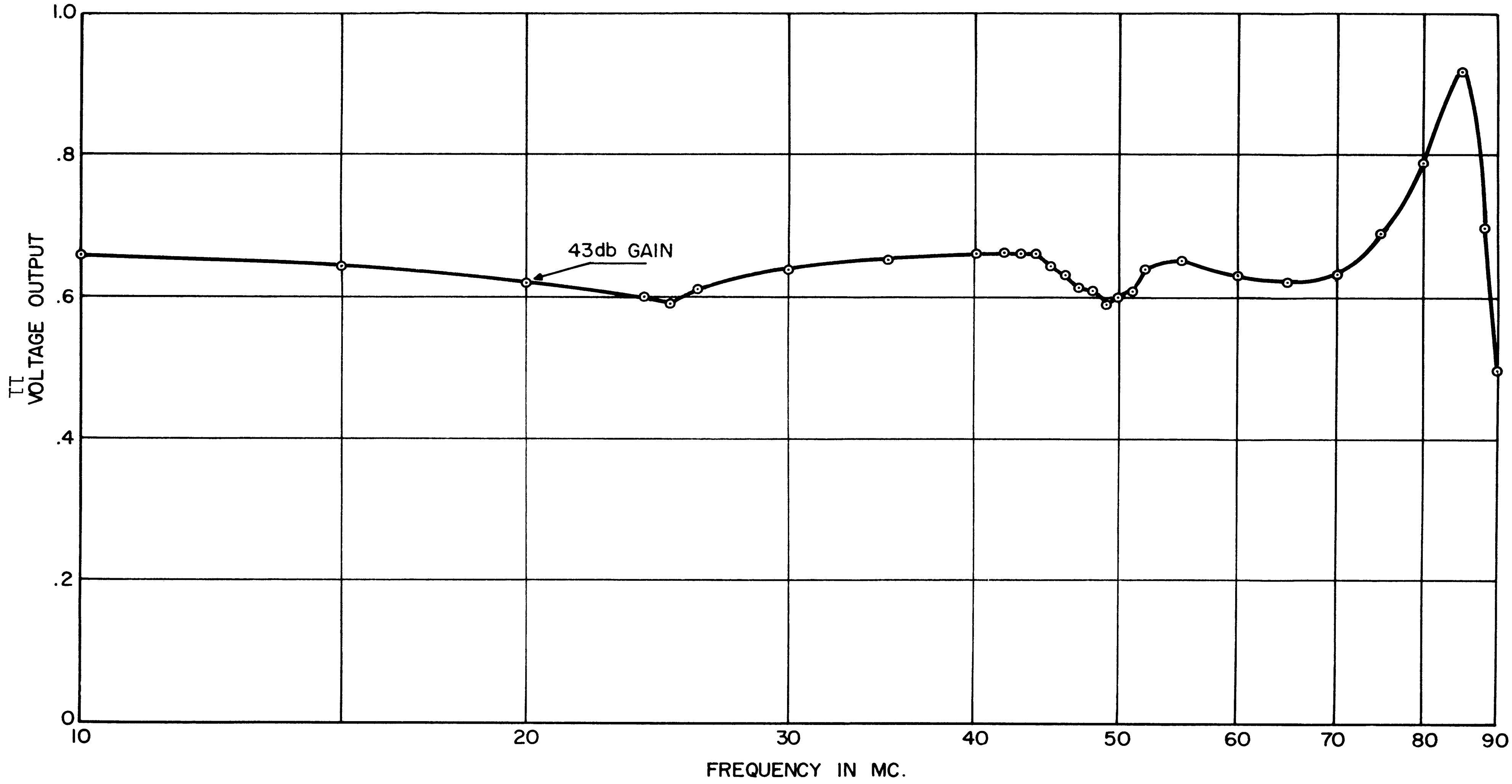

FIG. 4 AMPLITUDE RESPONSE OF AMPLIFIER USING WE-404 & WE-417A TUBES

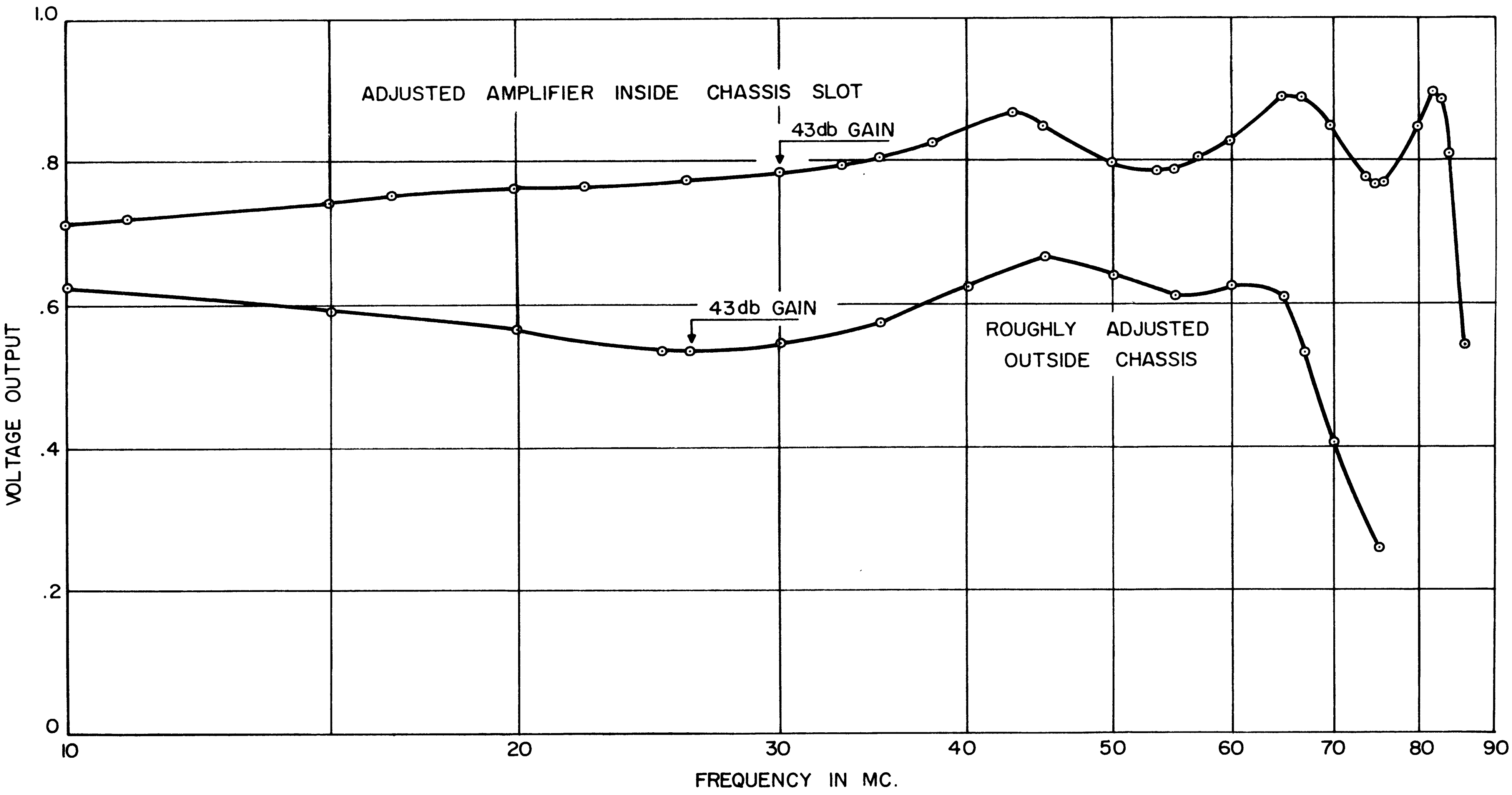

FIG.5 AMPLITUDE RESPONSE OF AMPLIFIER USING 6U8 TUBES

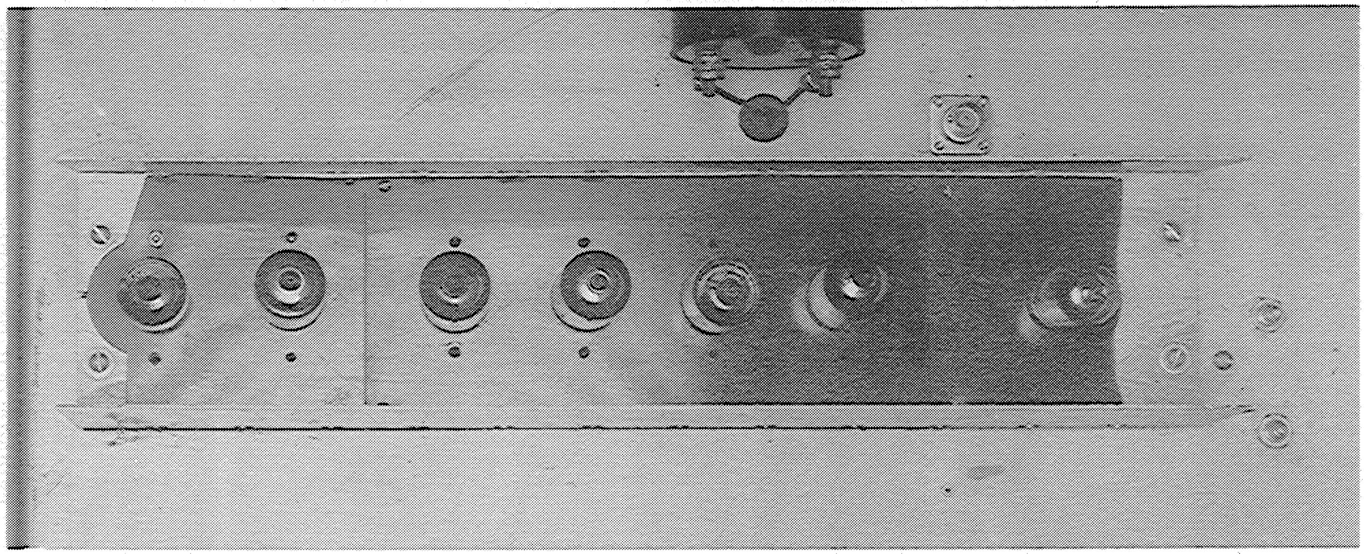

FIG. 6A. TOP VIEW OF AMPLIFIER CHASSIS.

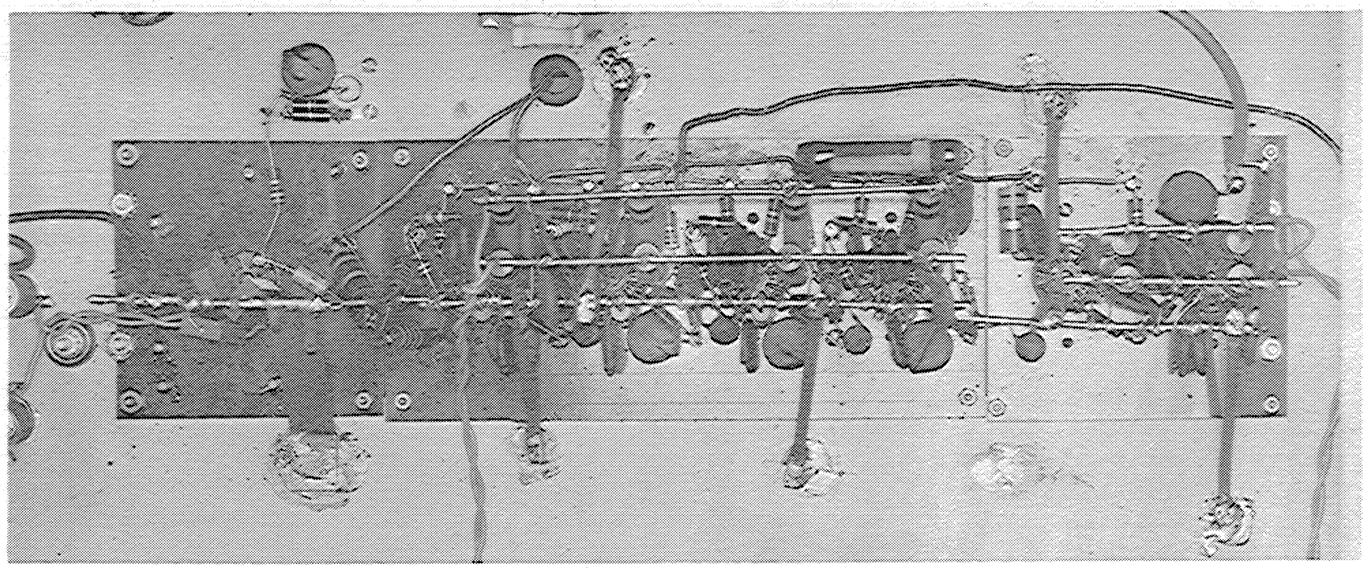

FIG. 6B. BOTTOM VIEW OF AMPLIFIER CHASSIS.

of this curve indicated very little undesirable feedback of the notch variety. In order to avoid notches, careful attention was given to the filament and B+ decoupling circuits. Resistive decoupling, rather than inductive decoupling, was found necessary in the B+ supply.

4. CONCLUSIONS

To compare this amplifier with other circuits in some general way would require a clever discussion of the restraints on the network polynomial. Specific comparisons are possible, however. This amplifier has a gain-bandwidth product of 240 mc per tube; the WE 404A tube with the measured stray capacitances of the amplifier of this report has a gain-bandwidth product of 150 mc. Consider an amplifier using a Tchebycheff three-pole network between cascaded pentodes. For the same ripple as the discussed theoretical amplifier, the gain of four such cascaded stages would be 53 and the nominal plate load resistor would be 217 ohms. The gain of two stages (four tubes) of the amplifier designed approximately in this report is 63 with a plate load resistor of 667 ohms. The amplifiers are roughly comparable in gain, but the amplifier of this report has the additional advantages of easier low-frequency coupling and roughly three times the available voltage excursion.

The amplifiers constructed were required to have nearly constant gain from 10 mc to 70 mc, and apparently the pole locations were adjusted near those of the Tchebycheff approximation. Better pole locations for a pulse amplifier might tend toward a more gradual high-frequency roll-off of gain. However, it should not be overlooked that many pulses will not have rise times sufficiently fast to excite the higher natural frequencies of these amplifiers. As an example, consider a pulse with a finite and constant rise time of 0.01 μsec. applied to one five-pole Tchebycheff filter of 85 mc bandwidth. The ringing amplitude associated with the outer poles will be about $1/5^{th}$ that of the ringing from a perfectly rectangular pulse.

In conclusion, it appears that the pentode-to-cathode-follower-pair amplifier is a useful amplifier for wide band applications, particularly for the final stage of a video amplifier. The advantages of this amplifier can be attributed to the high-transconductance, low-capacity triodes available. The design procedures outlined in this report will yield a specific amplifier, although the lack of a general analysis allows little to be said about the constraints which would guide the choice of the cathode-follower tube. The designer is forced to design by trial for each tube, and the solution, requiring as it does the repeated root-finding of the sixth order polynomial, involves appreciable time.

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