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NRL REPORT 3531

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DEVELOPMENT OF A 1,000-MC INTERMEDIATE-FREQUENCY AMPLIFIER

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NRL REPORT 3531

DEVELOPMENT OF A 1,000-MC INTERMEDIATE-FREQUENCY AMPLIFIER

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September 19, 1949

Approved by:

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ABSTRACT

A major problem in the design of microwave superheterodyne receivers is that of obtaining good image rejection, a condition which may be accomplished by an increase either in the i-f or in the selectivity of the preselector. Since the latter is limited by mechanical difficulties and by the video bandwidth requirements of the receiver—a serious limitation in a receiver designed for the interception of extremely narrow pulses—the only alternative is to increase the i-f. In fact, in cases where the video bandwidth must be extremely wide, an unusually high i-f is imperative in order to prevent modulation pulse components from feeding directly through the high-gain i-f amplifier.

Efforts have been directed primarily toward determining circuitry to promote an optimum gain-bandwidth product for 1,000-Mc, wide-bandpass, grounded-grid, coaxial amplifiers. Initial bandpass measurements lead to the conclusions that (1) the development of a complete i-f amplifier for service use appears quite feasible when mechanically suitable tubes possessing a sufficiently high transconductance become available; (2) the gain-bandwidth product of a single stage increases with bandwidth; (3) the length of the interstage coupling between two cascaded stages is critical for obtaining an optimum over-all gain-bandwidth product; and (4) a frequency-modulated test oscillator is highly desirable, if not essential, for alignment of a multistage amplifier.

This report covers the work performed and the techniques employed in these studies. Recent developments in measurement techniques are briefly described and probable trends of future work are indicated.

PROBLEM STATUS

This is an interim report; work on the problem is continuing.

AUTHORIZATION

NRL Problem R06-07R
(NR 530-007)

DEVELOPMENT OF A 1,000-MC INTERMEDIATE-FREQUENCY AMPLIFIER

INTRODUCTION

One of the major problems encountered in the design of microwave superheterodyne receivers is that of obtaining good image rejection. Improvement in this respect may be accomplished by increasing the selectivity of the preselector, by increasing the i-f, or by increasing both. The selectivity of the preselector is limited by the mechanical difficulties of ganging an increasing number of tunable cavities at microwave frequencies and by the video bandwidth requirement of the receiver. This latter limitation becomes of great importance when designing a receiver for the interception of extremely narrow pulses. The only alternative for improving the image rejection then is to increase the i-f. In fact, in cases where the video band must be extremely wide, an unusually high i-f is imperative in order to prevent modulation pulse components from feeding directly through the high-gain i-f amplifier. Furthermore, in order to filter out the i-f signal from the video output at the second detector, the i-f should be considerably higher than the upper frequency limit of the video amplifier. Since the elimination of spurious frequency responses for intercepted microwave signals is very important in the radio countermeasures field, research was initiated on 1,000-Mc i-f systems.

Efforts to date have been directed primarily toward determining circuitry to promote an optimum gain-bandwidth product for 1,000-Mc, wide-bandpass, grounded-grid, coaxial amplifiers. A complete i-f strip has not been made because the tubes now available are mechanically unsuitable for use in Navy-type receivers and have low values of transconductance. Initial bandpass measurements indicated two important observations worthy of investigation: (a) in contrast to the essentially constant gain-bandwidth product of the ordinary grounded-cathode amplifier, the gain-bandwidth product of this amplifier appeared to vary with bandwidth and (b) the length of the interstage coupling cable appeared to be an important factor in determining the over-all gain-bandwidth product of two grounded-grid stages in cascade. This report covers the work performed and the techniques employed in these studies. Recent developments in techniques of measurements are briefly described and probable trends of future work are indicated.

THE TEST AMPLIFIER

The experimental work described in this report was performed on a 1,000-Mc grounded-grid coaxial amplifier employing the RCA coaxial triode type A-2302-A. Figure 1 shows the cathode and plate cavities of this amplifier, separated to display the method of assembly and of tube replacement. Figure 2, the cross-sectional view of the amplifier, shows internal details. The cathode and plate cavities operate on the quarter-wave mode and are capacitively tuned to frequency by small disc trimmers. These discs are mounted on

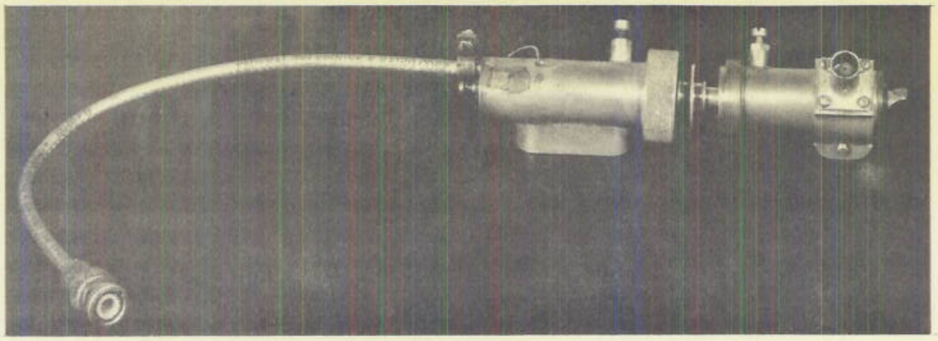


Figure 1 - Experimental 1,000-mc grounded-grid amplifier

threaded rod, thereby permitting external adjustment of the tuning capacity. In order to maximize the gain-bandwidth product of the loaded cavities, the cavities were designed to keep the trimming capacity near the minimum value required for alignment to 1,000 Mc.

Signal energy is introduced into the cathode cavity by means of an adjustable probe making direct contact on the inner cathode line. The axial position of the probe may be adjusted (within limits) to achieve a match between the signal source and the cathode line. As the cathode-loading is a function of the plate-loading, it is sometimes necessary to modify the size and shape of the cathode input probe to accommodate large load changes. This probe is accessible. It may be removed by taking out the four screws that fasten the r-f connector to the sliding assembly. The output coupling loop in the plate cavity is a continuation of the inner conductor of the output coupling cable and is soldered into position. For this reason the shape and size of the output coupling loop are not readily adjustable. The output coupling cable, although feeding a matched load, is not matched into the plate cavity. The cathode and plate d-c blocking condensers are built into the cavities. Internal cathode and plate decoupling networks were not employed in these test units. Decoupling will be required when several high gain stages are placed in cascade.

This amplifier does not lend itself to service use, the main objection being the difficulty of tube replacement. In order to remove the tube, the heater leads must be disconnected and the amplifier disassembled (as shown in Figure 1). Tubes of a more suitable mechanical design and with a higher transconductance are required for the development of a practical amplifier for service use.

TUNING PROCEDURE

The amplifier units were tuned to the operating frequency (approximately 1,000 Mc) in the following manner:

(1) After a sufficient warm-up period, the cathode tuning condenser and the cathode probe position were adjusted alternately until a good match was achieved between the 50-ohm signal source and the cathode line. Figure 3 shows the bench set-up used for measuring the SWR at the input to the cathode cavity. The LAE-2 signal generator, set in the C-W position, fed the cathode input connector of the amplifier through a slotted line. A flat 50-ohm load (VSWR = 1.6 db) consisting of a piece of RG-21/U

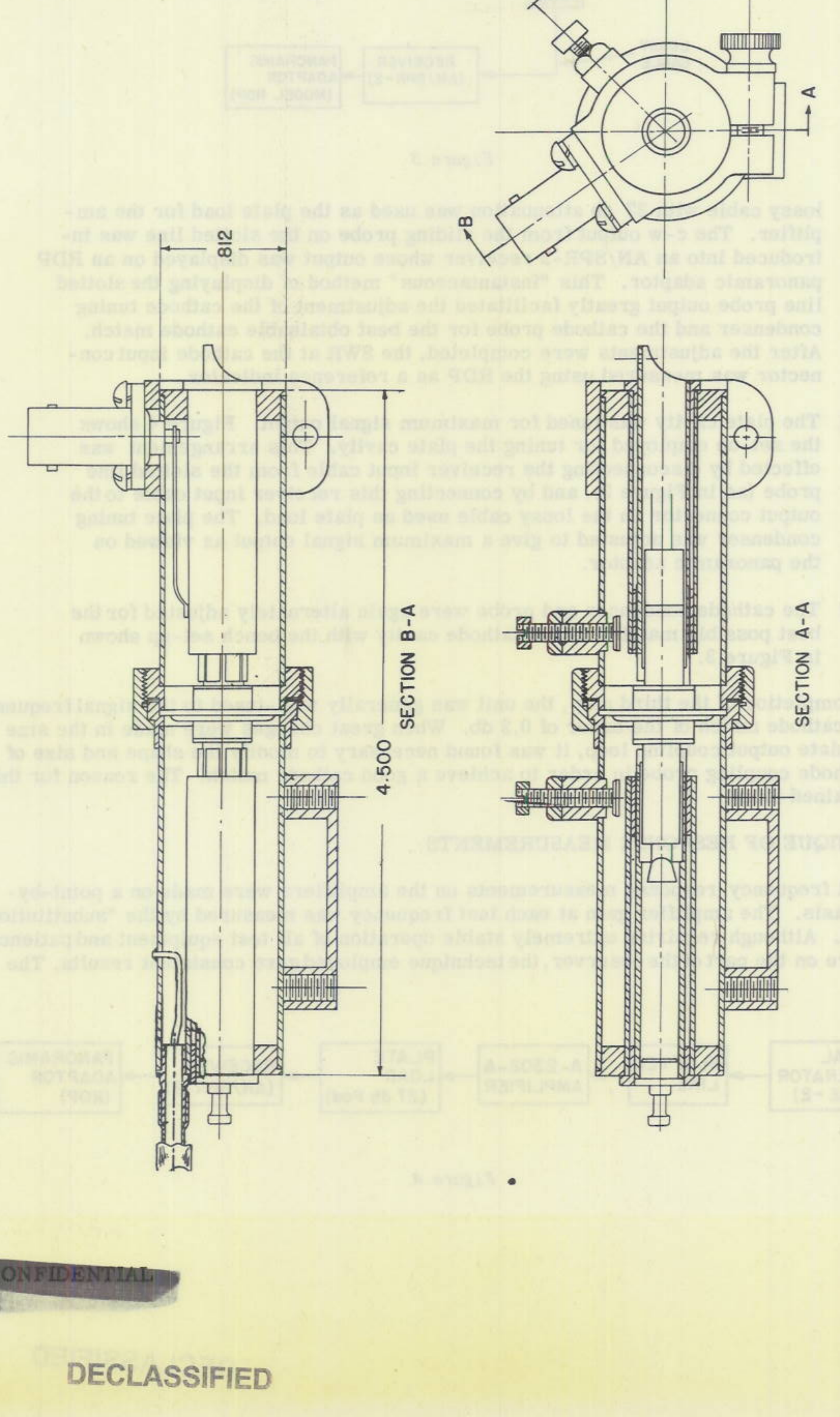


Figure 2 - Cross-sectional view of the 1000-mc amplifier

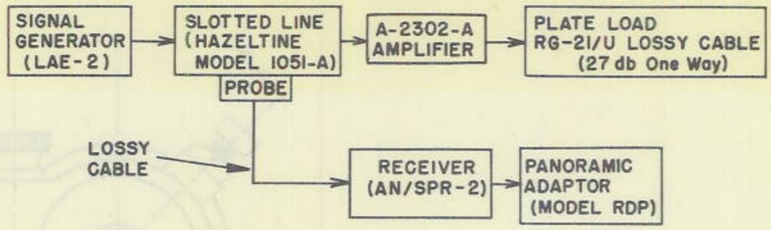


Figure 3

lossy cable with 27 db attenuation was used as the plate load for the amplifier. The c-w output from the sliding probe on the slotted line was introduced into an AN/SPR-2 receiver whose output was displayed on an RDP panoramic adaptor. This "instantaneous" method of displaying the slotted line probe output greatly facilitated the adjustment of the cathode tuning condenser and the cathode probe for the best obtainable cathode match. After the adjustments were completed, the SWR at the cathode input connector was measured using the RDP as a reference indicator.

(2) The plate cavity was tuned for maximum signal output. Figure 4 shows the set-up employed for tuning the plate cavity. This arrangement was effected by disconnecting the receiver input cable from the slotted line probe (as in Figure 3), and by connecting this receiver input cable to the output connector on the lossy cable used as plate load. The plate tuning condenser was adjusted to give a maximum signal output as viewed on the panoramic adaptor.

(3) The cathode condenser and probe were again alternately adjusted for the best possible match into the cathode cavity with the bench set-up shown in Figure 3.

Upon completion of the third step, the unit was generally well-tuned to the signal frequency with a cathode match of the order of 0.2 db. When great changes were made in the size of the plate output coupling loop, it was found necessary to modify the shape and size of the cathode coupling probe in order to achieve a good cathode match. The reason for this is explained later.

TECHNIQUE OF RESPONSE MEASUREMENTS

All frequency-response measurements on the amplifiers were made on a point-by-point basis. The amplifier gain at each test frequency was measured by the "substitution" method. Although requiring extremely stable operation of all test equipment and patience and care on the part of the observer, the technique employed gave consistent results. The

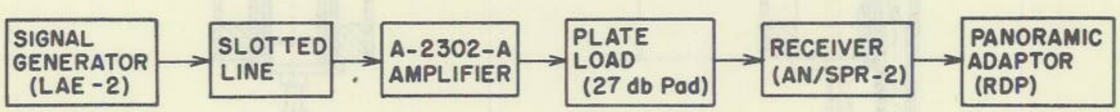


Figure 4

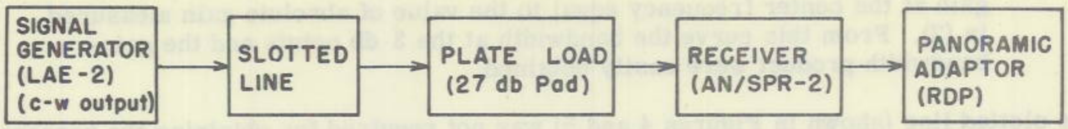


Figure 5

detailed procedure employed to obtain data for the response curves of the amplifier follows:

(1) After a sufficient warm-up period, the frequency-response data of the test equipment (without the A-2302-A amplifier) were obtained with the bench set-up shown in Figure 5. The signal generator, set for c-w operation and the receiver were adjusted to the previously determined center frequency of the A-2302-A amplifier. The attenuator of the LAE-2 and the gain controls of the SPR-2 and the RDP were adjusted to place the peak of the signal displayed on the RDP at a convenient reference level. This signal level, as read on the calibrated LAE-2 attenuator, was then established as the zero db reference level. Thereafter no gain control was touched. Signal-generator output was varied to check against saturation of the receiver or the panoramic adaptor. The signal-generator attenuator was adjusted to bring the signal peak to the reference level on the RDP scope at each of the test frequencies, and the attenuator and frequency readings were recorded. The equipment response relative to the zero db reference frequency was then obtained at each of the test frequencies by taking the difference between the attenuator readings at the test frequencies and the attenuator readings at the zero db reference frequency. To eliminate the response characteristic of the RDP from the measurements, the SPR-2 receiver was always tuned to place the signal at the same spot on the baseline of the RDP scope.

(2) The gain of the amplifier was measured at its center frequency by the substitution method. With the equipment set-up as shown in Figure 5 a convenient reference level on the RDP was established. The A-2302-A amplifier was then inserted between the slotted line and the plate load, as shown in Figure 4. The signal-generator attenuator was readjusted to give the same signal output on the RDP as before. The absolute gain of the amplifier is the difference of attenuator readings.

(3) The frequency response of the amplifier and associated test equipment connected as shown in Figure 4 was measured. The same method as that used for obtaining the equipment frequency response data in (1) was employed. In order to obtain measurements as far down as possible, on the "skirts" of the selectivity curve the SPR-2 and RDP gain controls were set initially for the maximum gain possible without displaying appreciable noise on the panoramic scope. The signal-generator output required was then quite low at the center frequency, providing a sufficient output reserve to make amplifier-response measurements down to gains of approximately -20 db.

(4) For each test frequency the equipment correction response data obtained in (1) were subtracted from the amplifier-response data obtained in (3). The amplifier-response data were then plotted, after subtracting a constant from the corrected amplifier-response data in order to make the amplifier

gain at the center frequency equal to the value of absolute gain measured in (2). From this curve the bandwidth at the 3 db points and the gain-bandwidth product were easily obtained.

The slotted line (shown in Figures 4 and 5) was not required for obtaining the response data. It was included in the bench set-ups so as to check more easily the cathode match before and after each data run. Later, it was recognized that this extra length between signal generator and cathode input connector promoted "long-line" effect. The slotted line should have been removed from the equipment set-up when obtaining response data.

AMPLIFIER BANDWIDTH ADJUSTMENT

Adjustment of the plate coupling loop is the principal factor determining the plate loading and stage bandwidth. The cathode cavity is so heavily loaded by the tube conductance that it has little effect upon determining the stage bandwidth in the normal case. With this assumption the equation expressing the stage bandwidth, Δf, as a function of plate loading, R_L, was derived as shown in the Appendix:

$$\Delta f = \frac{r_p + R_L}{\pi C_p R_L \left(2r_p + R_L\right)}, \text{ (Equation 16 in Appendix)}$$

where C_p is the effective plate-loading capacity. At resonance the cathode cavity is loaded by R_k where

$$R_k = \frac{r_p + R_L}{\mu + 1}. \text{ (Equation 10 in Appendix)}$$

Therefore, when changing the plate-loading factor, R_L, by modifying the shape or size of the output coupling loop, it is necessary likewise to change the cathode coupling to achieve a good match into the cathode cavity. The Appendix gives a more complete mathematical analysis of this grounded-grid amplifier.

VARIATION OF GAIN-BANDWIDTH PRODUCT WITH BANDWIDTH FOR A SINGLE STAGE

During the early experiments to determine the bandwidth limits available with the cavities employed, it was observed that the gain-bandwidth product of a single stage was not constant. In contrast to the conventional grounded-cathode amplifier, measurements indicated that the gain-bandwidth product of a stage increased considerably with increasing bandwidth. This observation stimulated the investigation of the manner of variation and of the limits of the gain-bandwidth product for a single stage.

The narrow bandwidth region was investigated first. The coupling loop illustrated in Figure 6 was most practical because of the extreme flexibility in varying the load on the amplifier. Change of load was effected by sliding the loop and its outer-conductor tubing (c) in barrel (d). Using this loop, bandwidths were obtained ranging from 2.1 to 17.5 Mc with corresponding gains ranging from 15 to 9 db. In attempting to obtain bandwidths narrower than 2.1 Mc, the amplifier became unstable. It appears that with greater decoupling of the B+ lead it may be possible to obtain narrower bandwidths.

Two types of coupling loops were used to investigate the broader bandwidth regions. A coupling loop like the one used in the early investigations (Figure 7) was employed in the investigation of the intermediate bandwidths. The broadest bandwidths were achieved

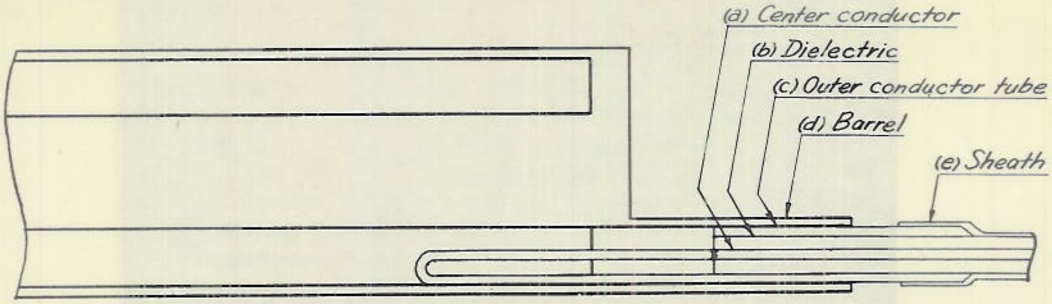


Figure 6 - Plate load coupling loops

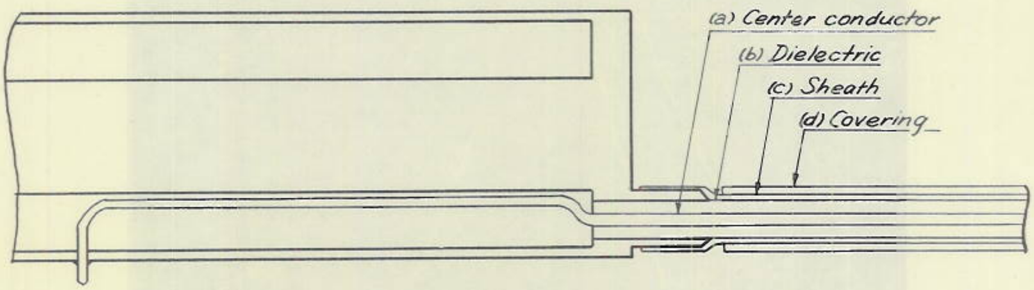


Figure 7 - Plate load coupling loops

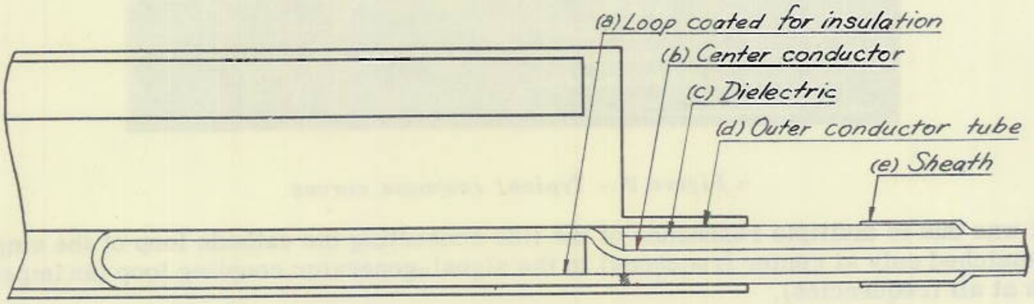


Figure 8 - Plate load coupling loops

by using the loop illustrated in Figure 8. With this loop, bandwidths up to 57.5 Mc were realized with a corresponding gain of 5.8 db. Greater bandwidths were obtainable, but response data for these cases could not be taken with the test equipment available. Typical experimental response curves are shown in Figure 9.

In the broad bandwidth region where the gain was low, long-line effect came into prominence, as shown by response curve A in Figure 10. This effect was present during all tests but was not recognized earlier because at the narrow bandwidths, the gain at resonant frequency had been greater and the frequency range covered had been considerably less. As a result, the measured bandwidths were less than those which could have been realized with short cables. Response curve B in Figure 10 illustrates the increase in bandwidth resulting from minimizing the long-line effect by shortening the cable from the signal generator to the cathode input connector (including removal of the slotted-line section). It should be noted that the gain is approximately the same for each case. The long-line

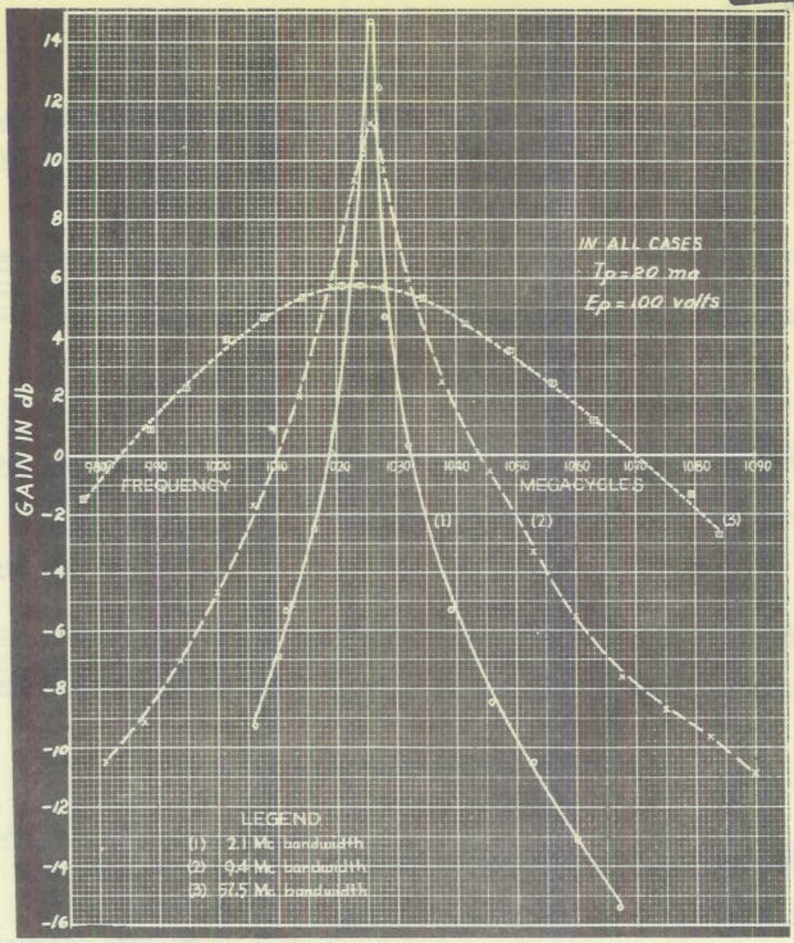


Figure 9 - Typical response curves

effect was due to multiple resonance of the line connecting the cathode loop of the amplifier (matched only at center frequency) to the signal-generator coupling loop (an imperfect match at all frequencies).

Figure 11 displays the experimental results of the gain-bandwidth product investigation. The gain-bandwidth product increased from 11.5 to 112 at bandwidths of 2.1 and 57.5 Mc, respectively. The voltage gain curve shows the manner of gain variation with bandwidth. This curve should prove useful in determining the optimum design of multistage amplifiers. Attention is again called to the 57.5-Mc bandwidth point in Figure 11. Here the long-line effect was minimized, resulting in a greater gain-bandwidth product than previously expected. If time had been available, all tests would have been repeated using the shortest possible length of cable between the cathode connector and the signal generator. However, it is believed that the data obtained do show the manner of variation of the gain-bandwidth product with bandwidth.

A mathematical analysis of this grounded-grid amplifier, based on certain simplifying assumptions, is given in detail in the Appendix. A brief comparison of circuits, starting with the grounded-cathode amplifier and leading up to the amplifier used in these experiments, will be useful.

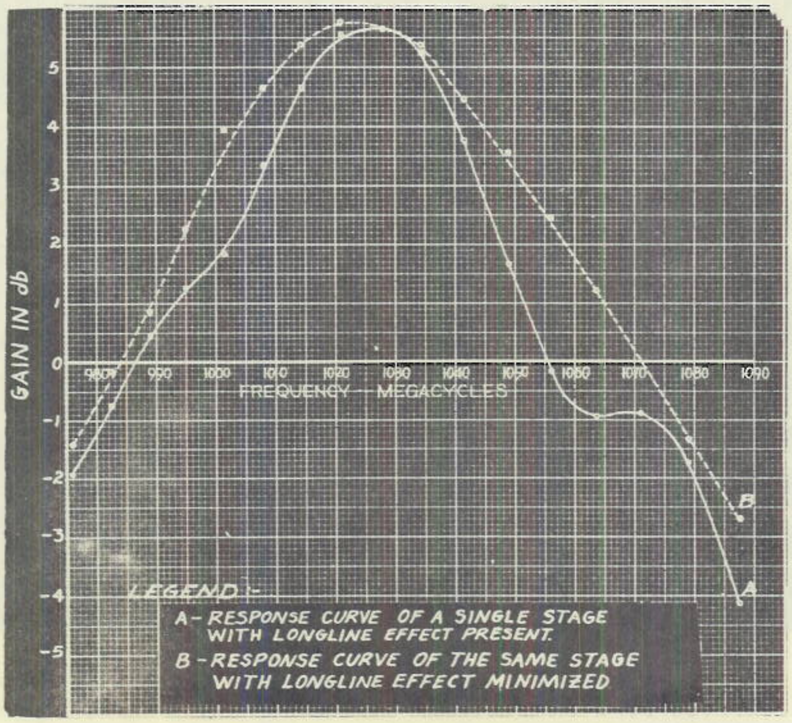


Figure 10 - Demonstration of long-line effect

The gain-bandwidth product of the familiar grounded-cathode amplifier, shown in Figure 12, is a constant. It does not depend upon the plate load, R_L. In the case of a grounded-grid amplifier having zero impedance between grid and cathode (Figure 13) the gain-bandwidth product is still constant. Plate load R_L does not enter into the $G_{k-p}(\Delta f)$ equation. Figure 14 shows a resistance, R_k, between the cathode and grid of a grounded-grid amplifier. Here it is seen that $G_{k-p}(\Delta f)$ is dependent upon the plate load, R_L. Furthermore, if R_k is matched to the resistance, R_c, looking into the cathode of the tube, then

$$R_k = R_c = \frac{r_p + R_L}{\mu + 1} .$$

Substituting this into the $G_{k-p}(\Delta f)$ equation, we get for R_k matched into the cathode,

$$G_{k-p}(\Delta f) = \frac{\mu + 1}{\pi C (R_L + 2 r_p)}$$

(R_k matched to R_c).

In this case the gain-bandwidth product is dependent upon the plate load, R_L.

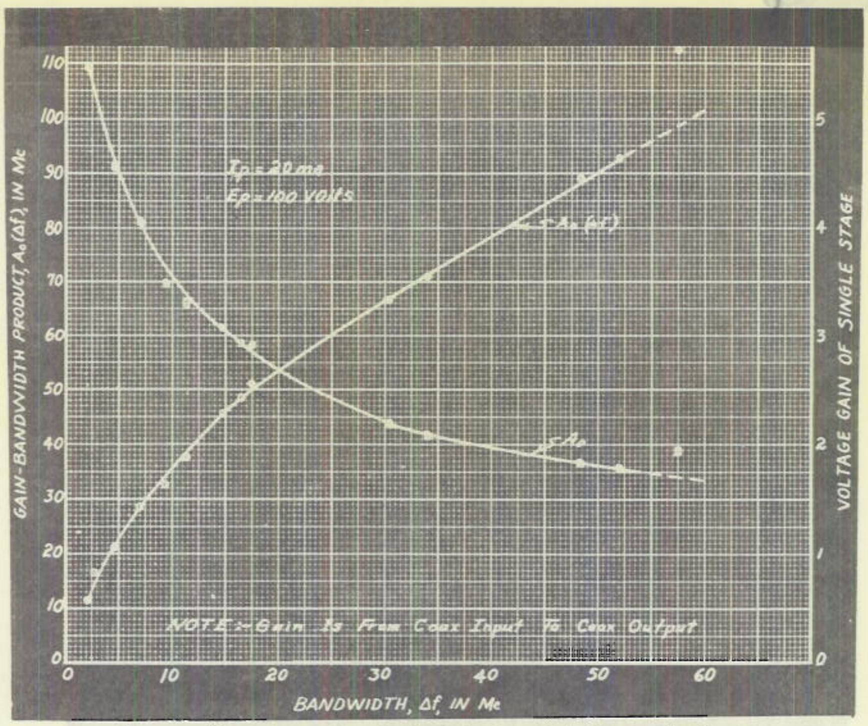
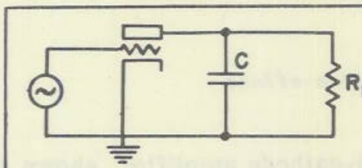
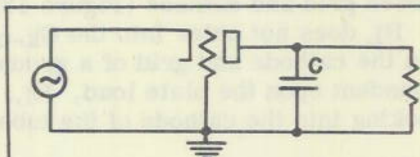


Figure 11 - Experimental gain-bandwidth product vs bandwidth and gain vs bandwidth



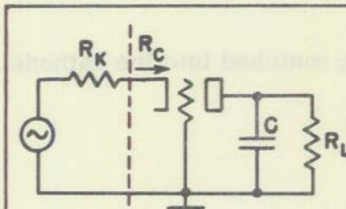
$$G_{g\text{-}p}(\Delta f) = \frac{\mu}{2\pi C r_p}$$

Figure 12



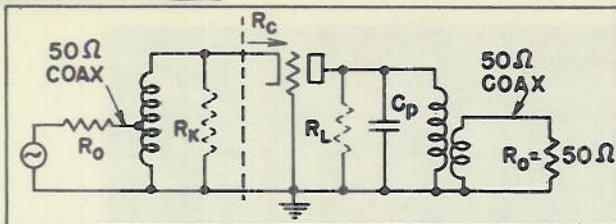
$$G_{k\text{-}p}(\Delta f) = \frac{\mu + 1}{2\pi C r_p}$$

Figure 13



$$G_{k\text{-}p}(\Delta f) = \frac{(\mu + 1)\left[r_p + (\mu + 1) R_k + R_L\right]}{2\pi C\,(r_p + R_L)\left[r_p + (\mu + 1) R_k\right]}$$

Figure 14



$$A_o(\Delta f) = \frac{\sqrt{\mu + 1}}{\pi C_p (2r_p + R_L)} \sqrt{\frac{R_L + r_p}{R_L}}$$

Figure 15

Figure 15 shows the equivalent "lumped" circuit of the grounded-grid amplifier under investigation. For this amplifier,

$$A_o = G_{k-p} \sqrt{\frac{R_k}{R_o}} \sqrt{\frac{R_o}{R_L}} = \frac{(\mu + 1) R_L}{r_p + R_L} \sqrt{\frac{r_p + R_L}{(\mu + 1) R_L}} = \sqrt{\frac{(\mu + 1)\ R_L}{r_p + R_L}} = \sqrt{G_{k-p}} \quad ,$$

where A_o = gain of the amplifier from input coax to output coax, and where $\sqrt{R_k/R_o}$ and $\sqrt{R_o/R_L}$ are the transformer ratios. It is interesting to note that the gain from coax to coax is the square root of the gain from cathode to plate. Since

$$\Delta f = \frac{r_p + R_L}{\pi C_p R_L (2r_p + R_L)}, \quad \text{(Equation 16 in Appendix)}$$

then

$$A_o(\Delta f) = \frac{\sqrt{\mu + 1}}{C_p \pi (2r_p + R_L)} \sqrt{\frac{r_p + R_L}{R_L}} \quad \text{(Equation 17 in Appendix)}$$

and the gain-bandwidth product is quite dependent upon plate loading.

An attempt was made to correlate the derived expressions with the experimental data. The static values of μ and r_p for the tube used in the series of experiments were obtained and inserted into the equations for Δf and for $A_o(\Delta f)$. From the physical dimensions of the cavity and the end loading capacity, the value of C_p was calculated* to be 4.13 micro-microfarads. The calculated values of Δf and $A_o(\Delta f)$ for various magnitudes of R_L are presented as curve C in Figure 16. It is interesting that curve C is so similar to the experimentally derived curve A when it is remembered that transit time effects long-line effect, regeneration at narrow bandwidths, cavity losses, and errors in estimation of μ, r_p, and C_p, as well as approximations underlying Equation 17, are not taken into account.

*The method used for calculating C_p is given on page 221 of "Radio at Ultra-High Frequencies, Vol II," published by RCA. Review. The equation is

$$C = 1/2 \left(C_t + \frac{C_o l}{\sin^2\theta}\right), \quad \text{where } C_o = \frac{33}{Z_o}, \quad C_t = \text{external capacity across the line},$$

l = length of line in centimeters, $\theta = \beta l$, and C = equivalent capacity of the loaded line.

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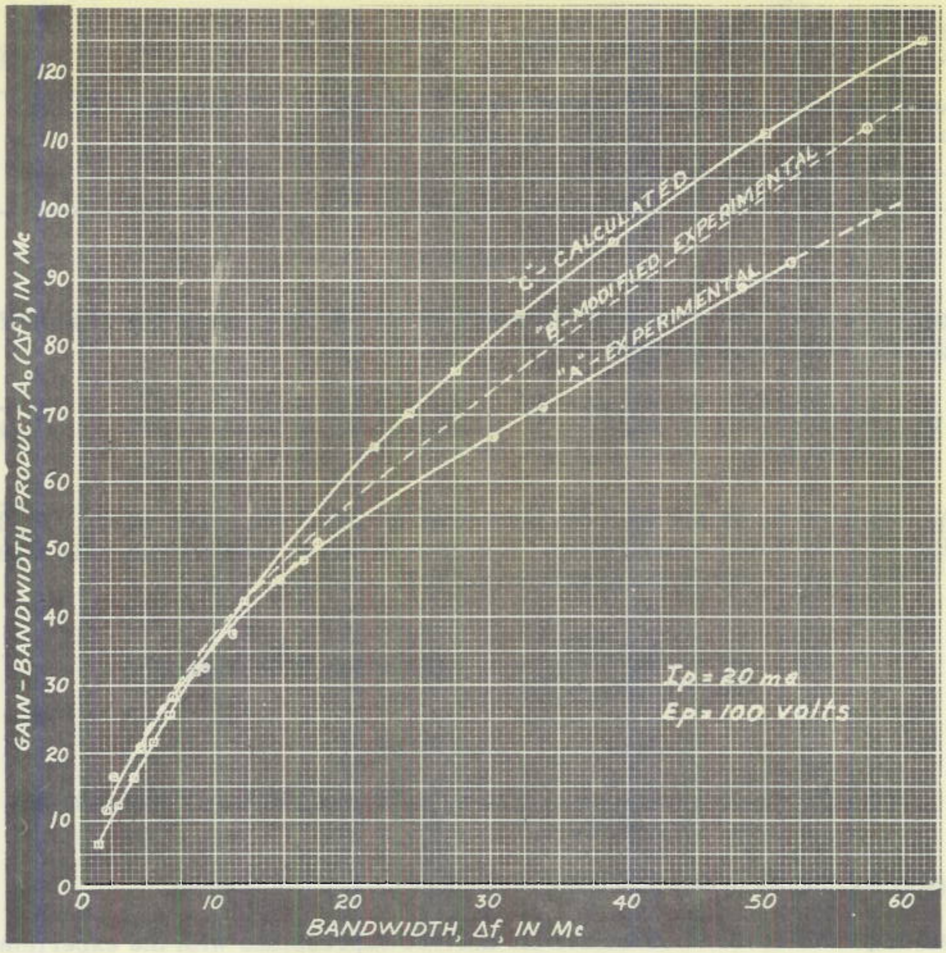

Figure 16 - Comparison of calculated and experimental curves for gain-bandwidth product vs bandwidth

As a check on the experimental data, C_p also was computed from the experimental data. Using $\mu = 17$, $r_p = 2360$, C_p was calculated to be 4.89 $\mu\mu$f at the $\Delta f = 57.5$ Mc experimental point on curve B. This value of 4.89 $\mu\mu$f compares reasonably close to the value of 4.13 $\mu\mu$f calculated from the physical dimensions of the cavity.

It is estimated that the experimental curve would have approximated curve B if the long-line effect had been minimized for all tests. A closer match between curves C and B is possible by juggling the values of μ, r_p, and C_p. It was felt that there was very little justification for assuming any other particular values of μ, r_p, and C_p. Furthermore, regeneration is quite evident at the narrow bandwidths. As an example, for a bandwidth of 2.1 Mc the coax-to-coax gain was measured to be 5.5. The maximum theoretical value without regeneration would be approximately 4.2 (i.e., A_o max. $\approx \sqrt{\mu + 1} = \sqrt{18}$).

The experimental "gain-vs-bandwidth" curve in Figure 11 is useful not only in obtaining an optimum design for a multistage amplifier, but also in adjusting the plate output coupling loop to obtain a desired stage bandwidth. After several measurements of the gain for different bandwidths of a particular stage, a rough gain-vs-bandwidth curve may be

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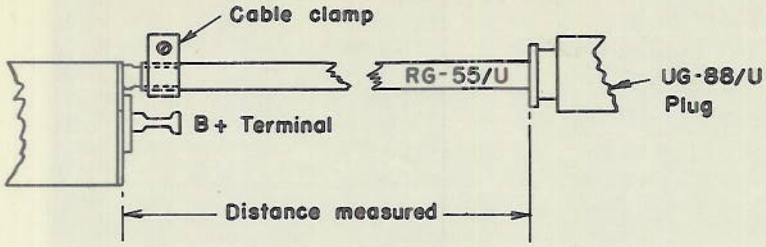

Figure 17

drawn. Using the curve, it is possible to predict the stage bandwidth by measuring the stage gain after changing the plate-coupling loop.

IMPORTANCE OF LENGTH OF THE INTERSTAGE COUPLING CABLE

The length of the interstage coupling cable between two cascaded stages was found to be an important factor determining the over-all gain-bandwidth product of the two stages. An investigation to determine the optimum length of coupling cable and to determine the criticalness of this length was carried out.

Two stages were individually tuned as close as possible to 1,001 Mc, and the response data for each stage were obtained. The two stages then were cascaded with a definite length of cable and the over-all response curve was experimentally determined using a plate input to each tube held constant at 20 ma and 100 volts. This procedure was repeated for sixteen lengths of interstage coupling cable ranging from 4-15/16 to 10-3/32 inches. These lengths were measured as shown in Figure 17. They do not include the effective length inside the plate cavity, the output connector, or the cathode input assembly of the second stage. It is believed that these additional lengths would be equivalent to approximately two inches of coax. Hence, the over-all coupling system effective coax length varied from about 7 to 12 inches, which corresponds to approximately $7\lambda/8$ to $3\lambda/2$ at 1,000 Mc.

In order to obtain a reference for comparison, the response curves of the two individual stages were added, as shown in Figure 18. This reference curve, designated "calculated," was plotted with the experimentally obtained response curve for the cascaded amplifiers labelled "experimental." The experimental and the "calculated" response curves for each length of coupling cable are presented in Figures 19 through 34. It is interesting to observe how the shape of the response curves changes with increasing cable length. Note the manner in which the shape of the high-frequency side of the experimental response curve changes (see Figures 19, 20, 21, 22, and 23) as the coupling-cable length is increased from 4-15/16 inches (about $7\lambda/8$) to 6-5/32 inches (about λ). The "bump" moves from the high-frequency side of the response curve toward the peak, broadening the response peak and making the curve more symmetrical. As the cable length is further increased to 8-1/8 inches (about $5\lambda/4$) the movement of the bump down the lower-frequency side of the curve decreases the 3 db bandwidth (Figures 24, 25, 26, 27, 28, 29, and 30). The response curve of the 8-1/8 inches-long coupling line is quite symmetrical and shows a low gain-bandwidth product. Figures 32, 33, and 34 show how the 3 db bandwidth increases as the length of coupling cable changes from an odd quarter-wavelength to an even quarter-wavelength.

The over-all gain-bandwidth product for each "measured" length of coupling cable is shown in Figure 35. The approximate centerlines of the maximum and the minimum peaks

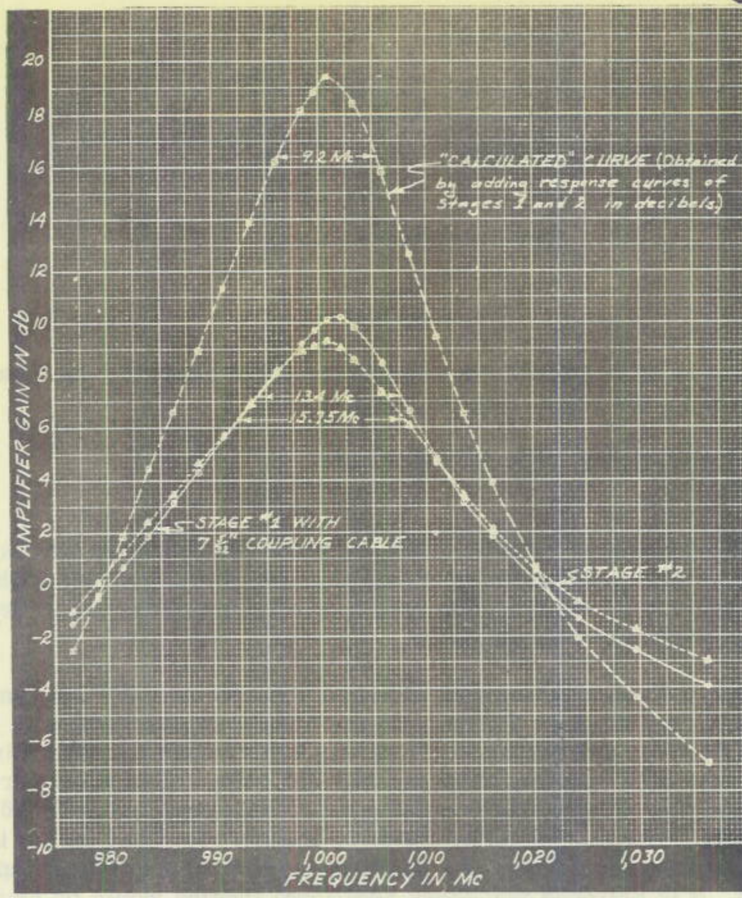

Figure 18 - Derivation of "calculated" response curve for a two-stage amplifier

of this curve were drawn to show that these coupling-cable lengths correspond to the most symmetrical broad and narrow response curves, respectively. Furthermore, these cable lengths differ by approximately $\lambda/4$.

The ratio of the experimental to the "calculated" gain-bandwidth product, and the ratio of the experimental to the "calculated" 3 db bandwidth for each coupling-cable length, are shown in Figure 36. A definite reason for the maximum ratios occurring at 6-17/32 inches rather than at 6-5/32 inches, the value corresponding to the most nearly flat-topped response, is not known at present. Nor can the drop at the 6-9/32 inch and the 6-13/32 inch points be explained unless certain errors not easily eliminated may have had an effect.

Although the experimental response curves, Figures 19 through 34, are believed to be quite accurate in each case for the amplifiers as they were operating, a comparison of gain-bandwidth products and bandwidths for these curves involves the inclusion of several undesirable factors.

First, the individual stages were not exactly tuned to the same center frequency, nor were their bandwidths identical. The difficulties involved in aligning the stages to exact

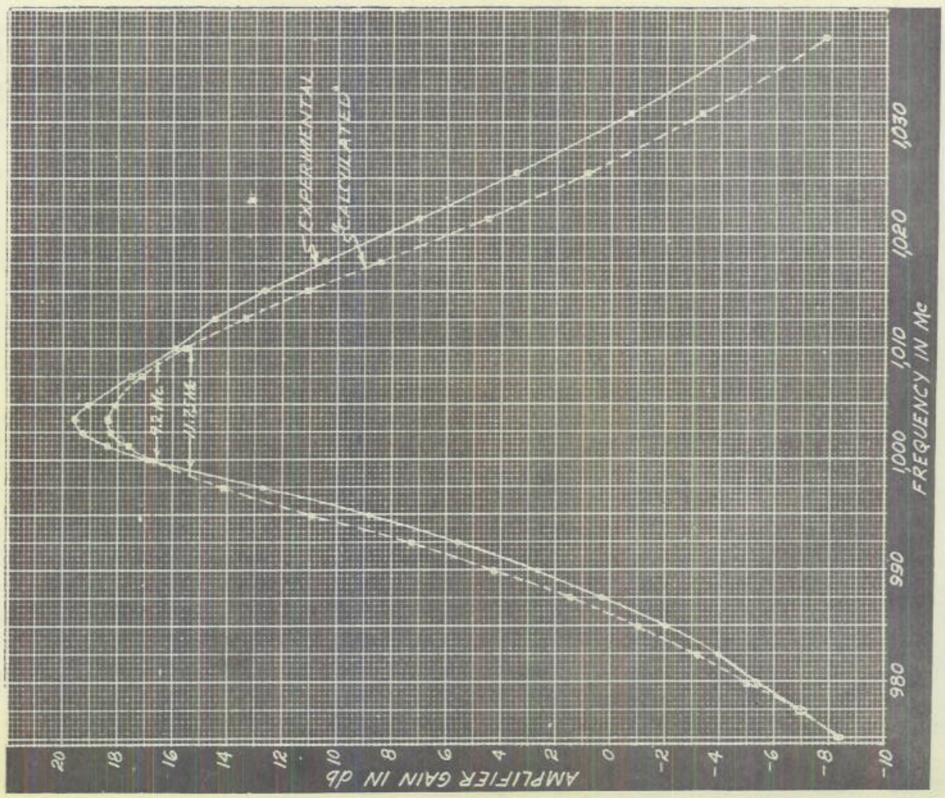

Figure 19 - Cascaded amplifier response curves for 4-15/16" coupling cable

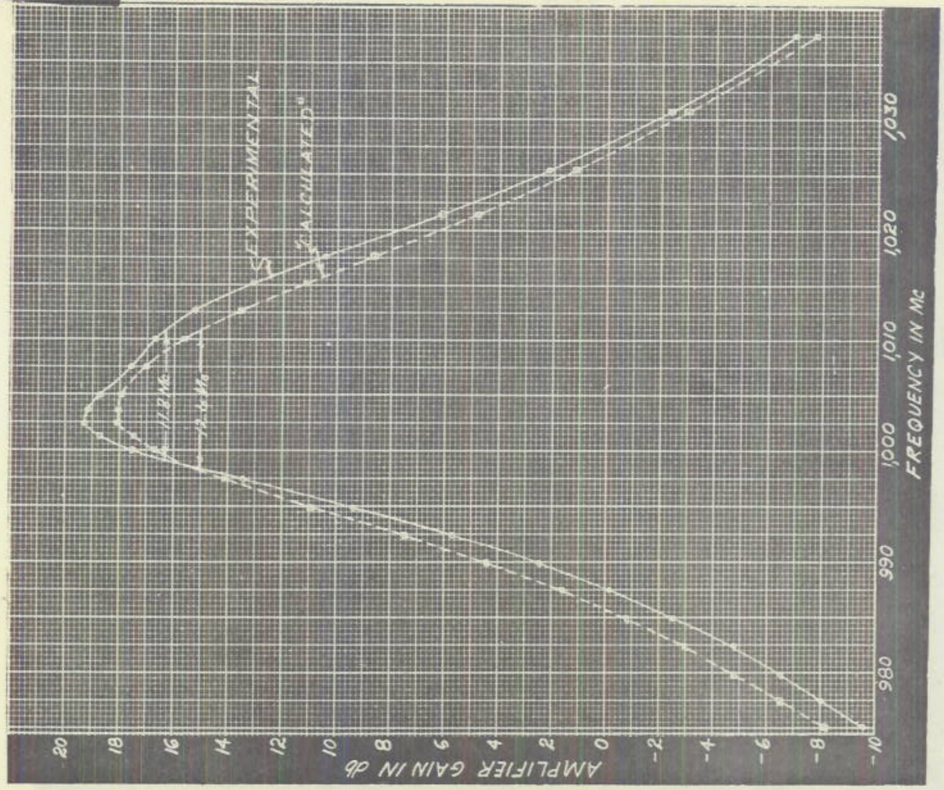

Figure 20 - Cascaded amplifier response curves for 5-13/32" coupling cable

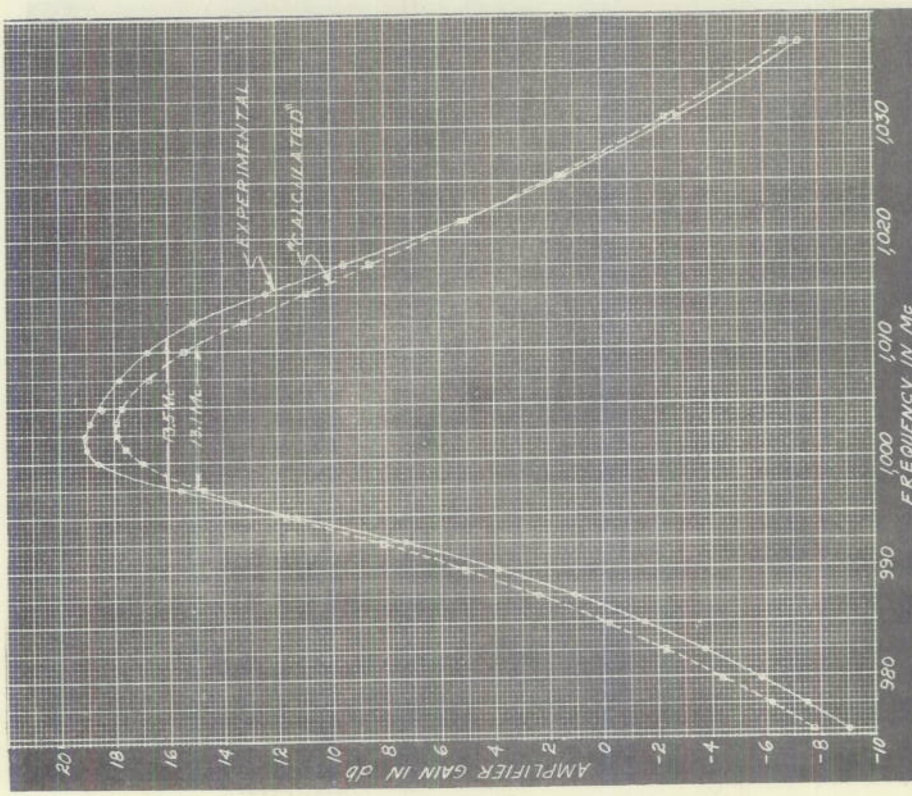

Figure 21 - Cascaded amplifier response curves for 5-21/32" coupling cable

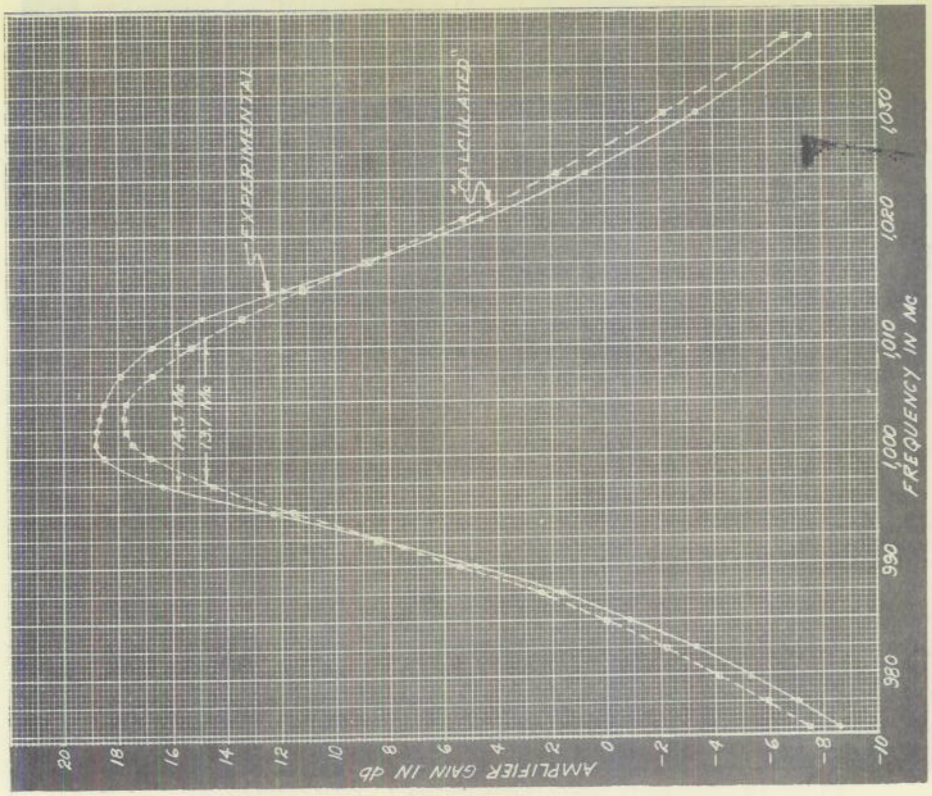

Figure 22 - Cascaded amplifier response curves for 5-29/32" coupling cable

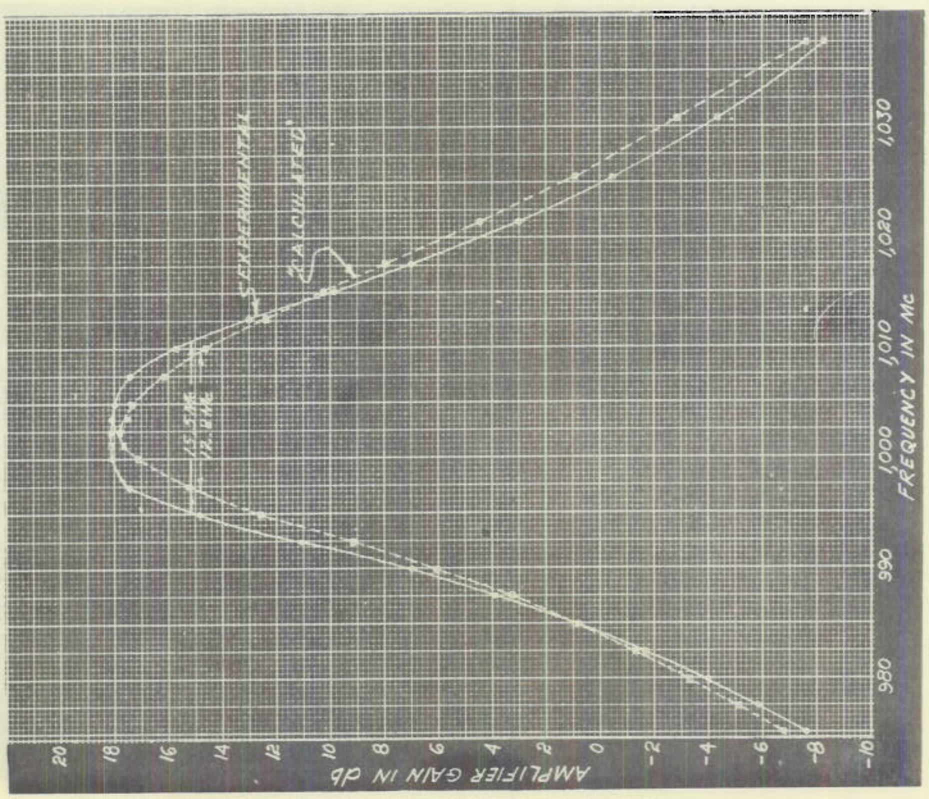

Figure 23 - Cascaded amplifier response curves for 6-5/32" coupling cable

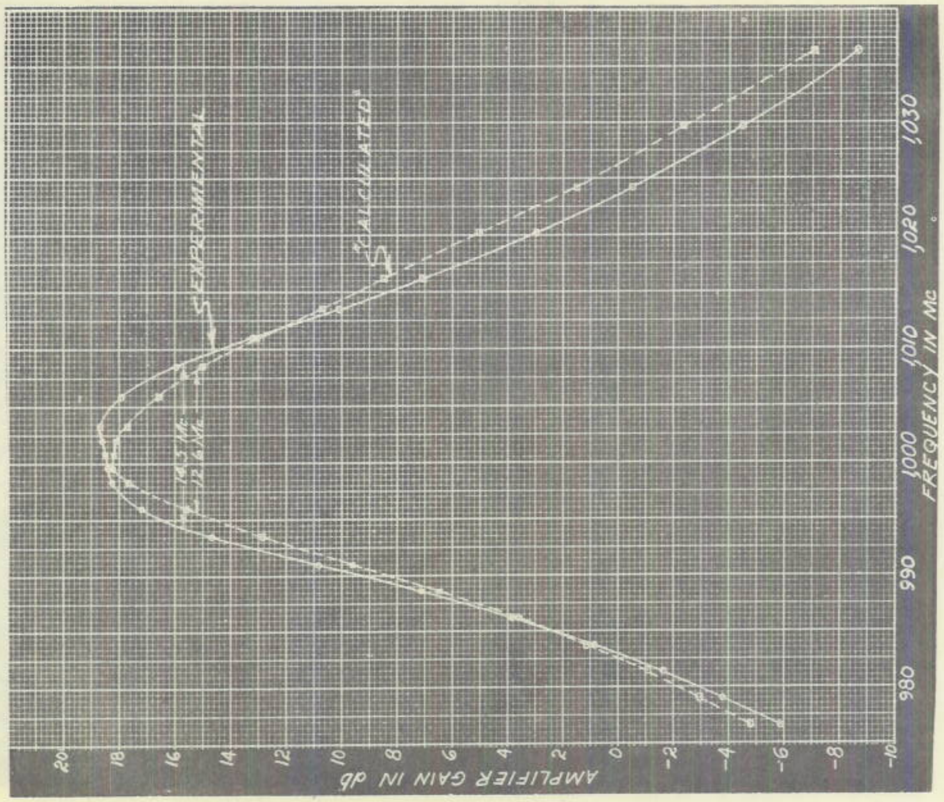

Figure 24 - Cascaded amplifier response curves for 6-9/32" coupling cable

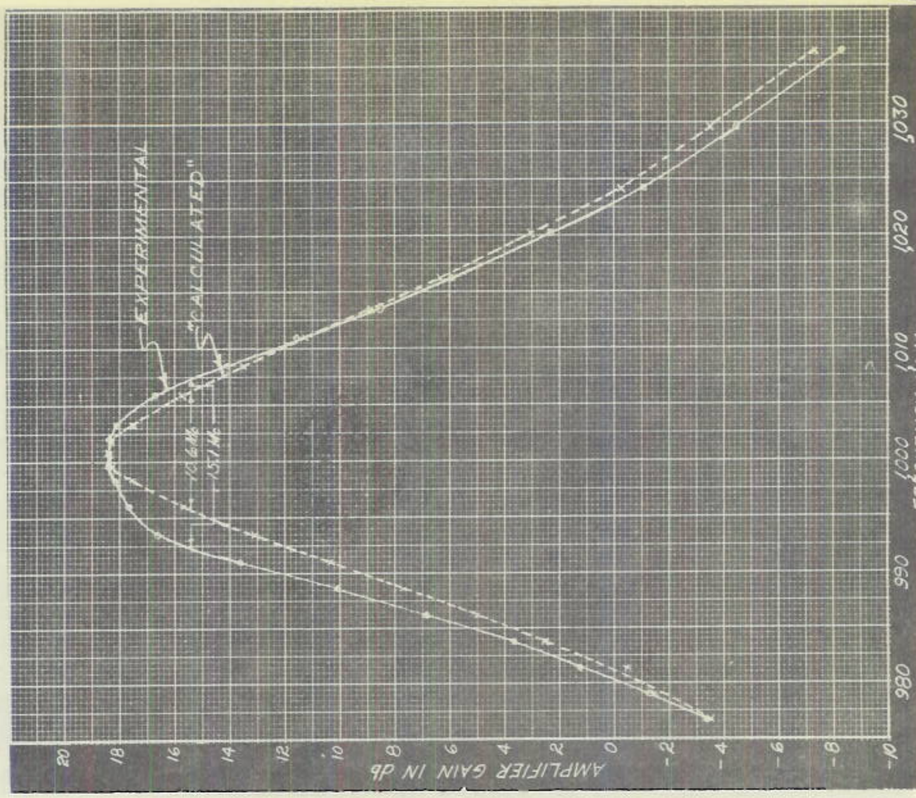

Figure 26 - Cascaded amplifier response curves for 6-17/32" coupling cable

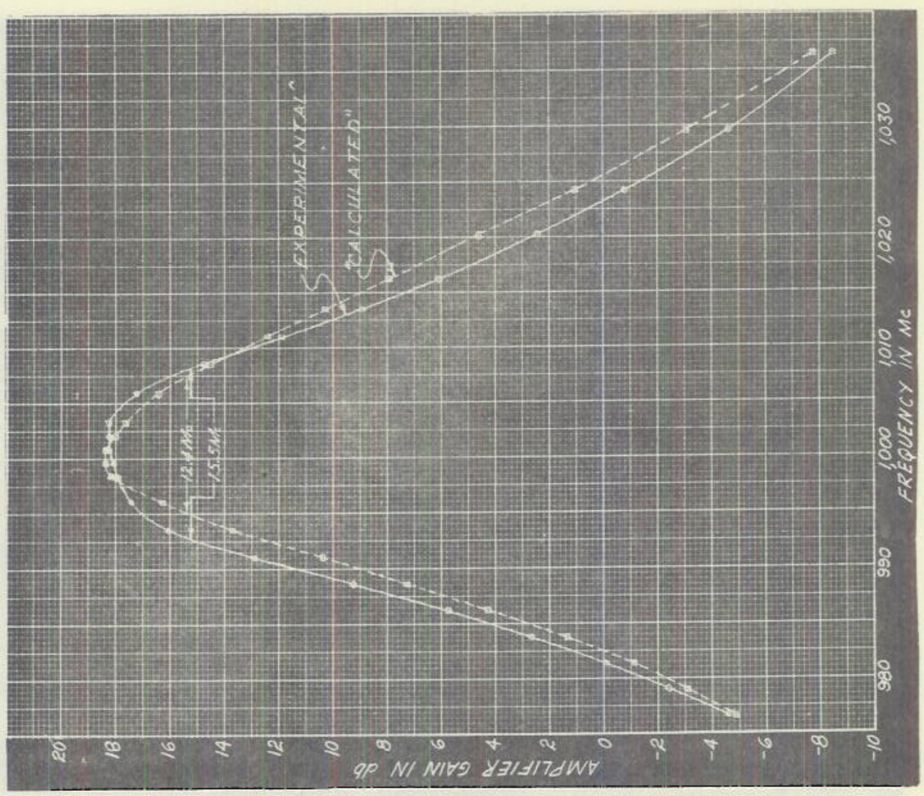

Figure 25 - Cascaded amplifier response curves for 6-13/32" coupling cable

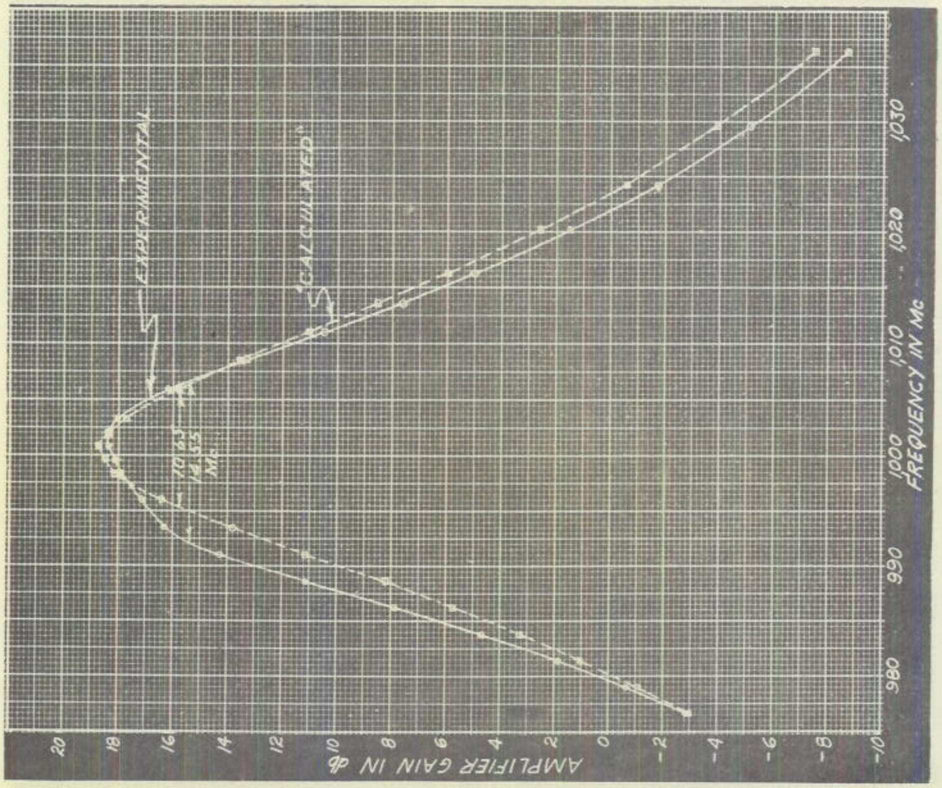

Figure 27 - Cascaded amplifier response curves for 6-21/32" coupling cable

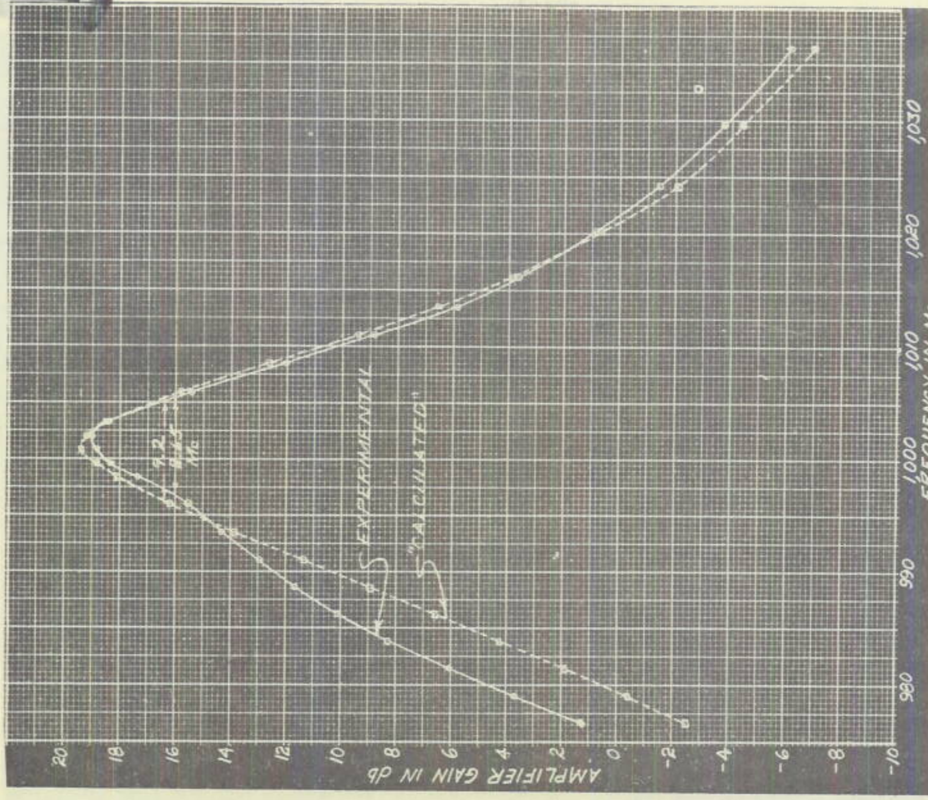

Figure 28 - Cascaded amplifier response curves for 7-5/32" coupling cable

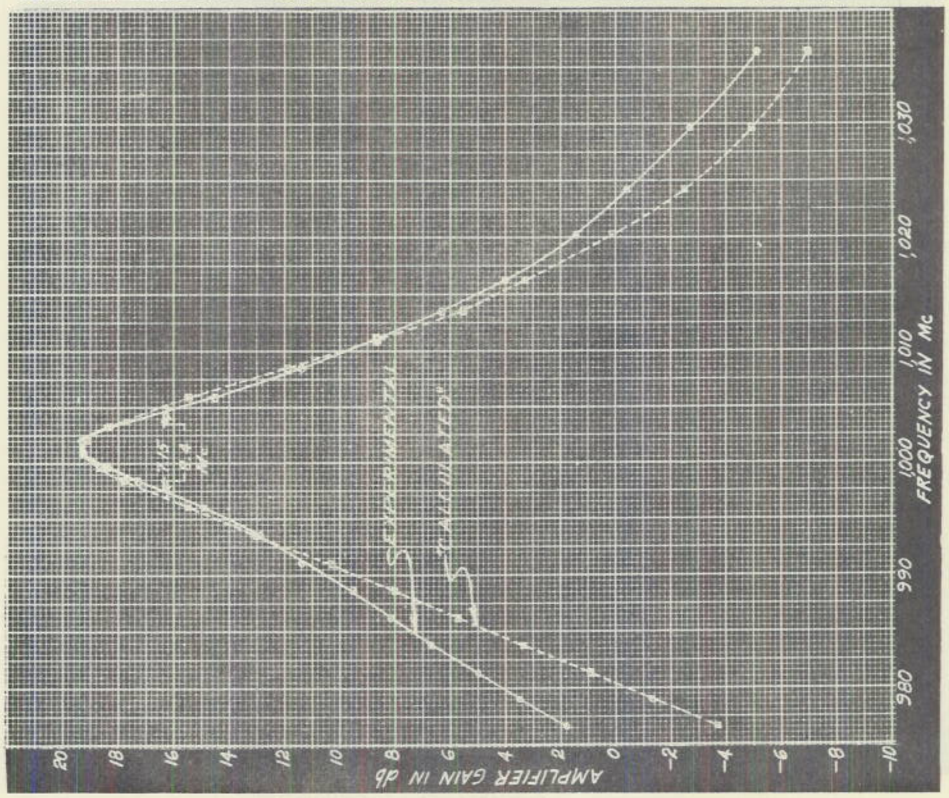

Figure 29 - Cascaded amplifier response curves for 7-21/32" coupling cable

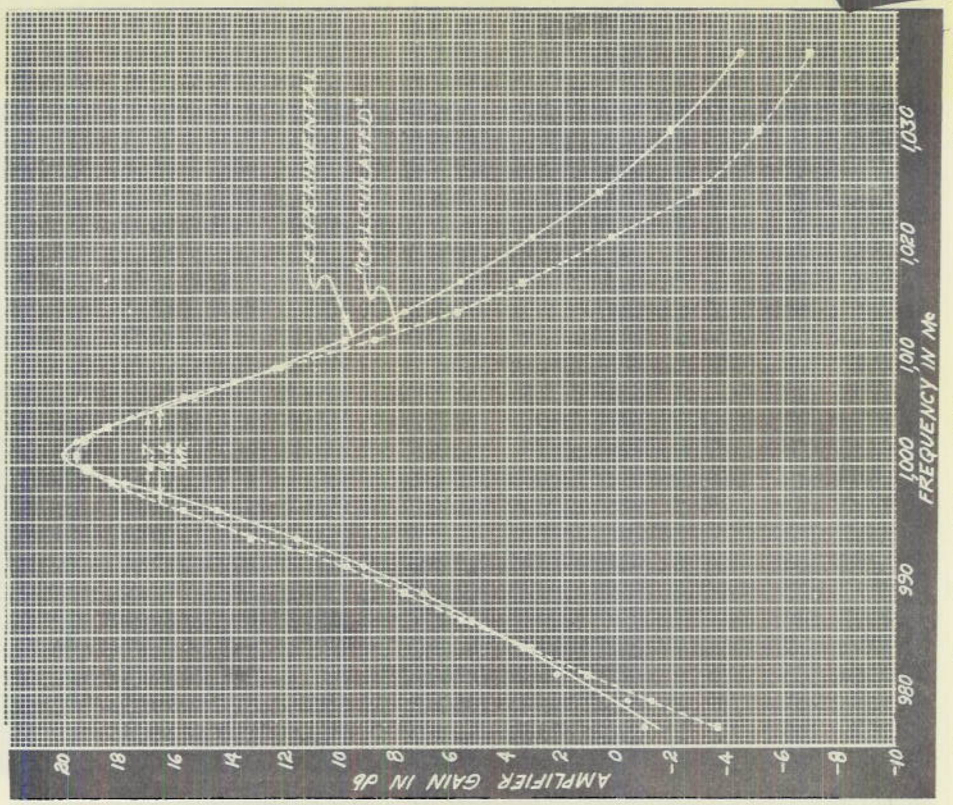

Figure 30 - Cascaded amplifier response curves for 8-1/8" coupling cable

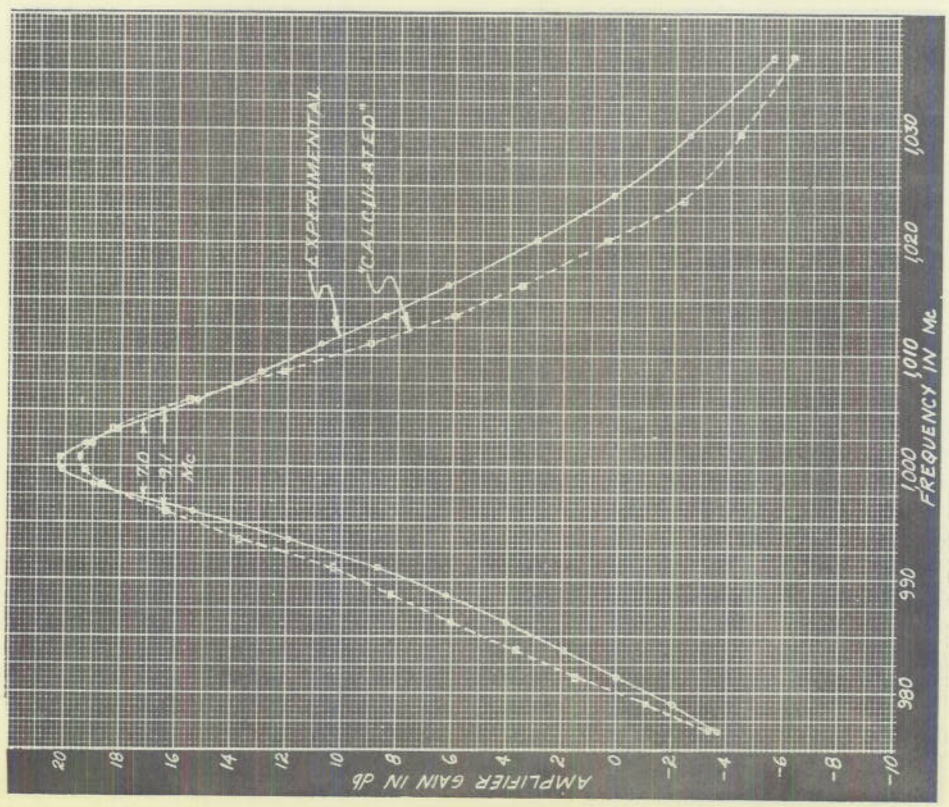

Figure 31 - Cascaded amplifier response curves for 8-5/8" coupling cable

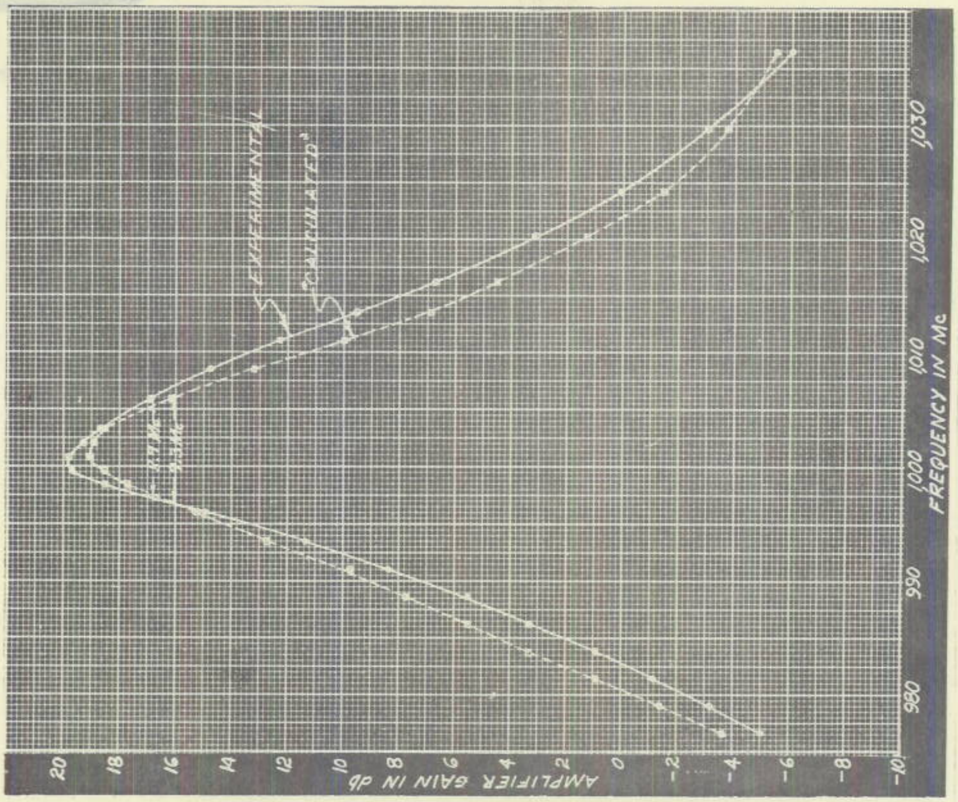

Figure 32 - Cascaded amplifier response curves for 9-3/32" coupling cable

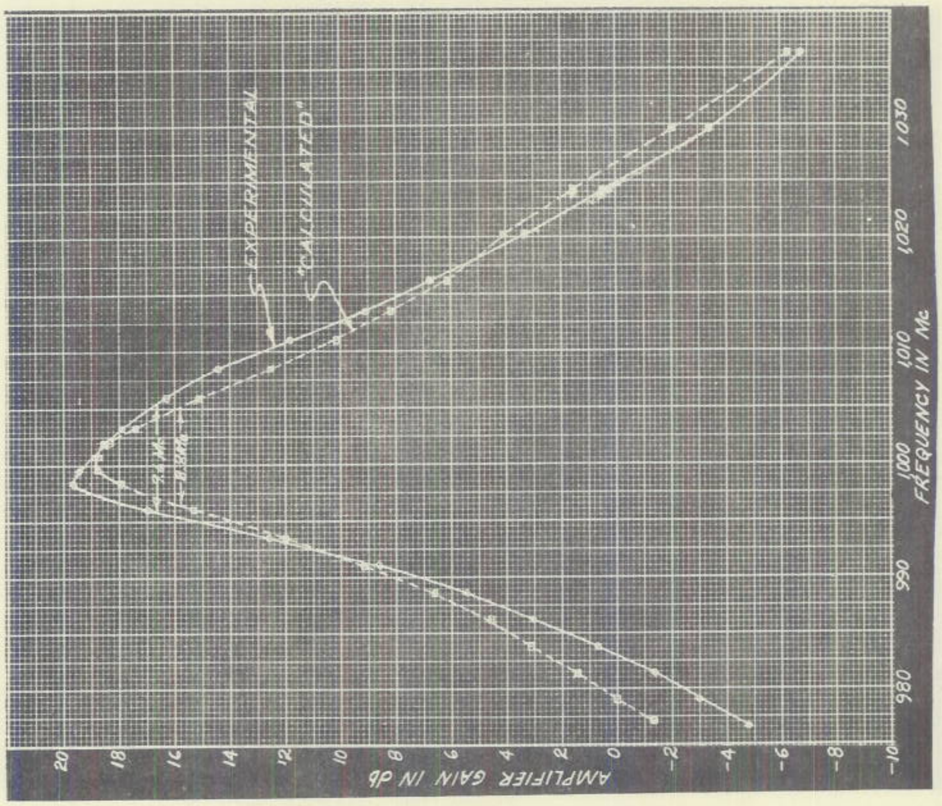

Figure 33 - Cascaded amplifier response curves for 9-19/32" coupling cable

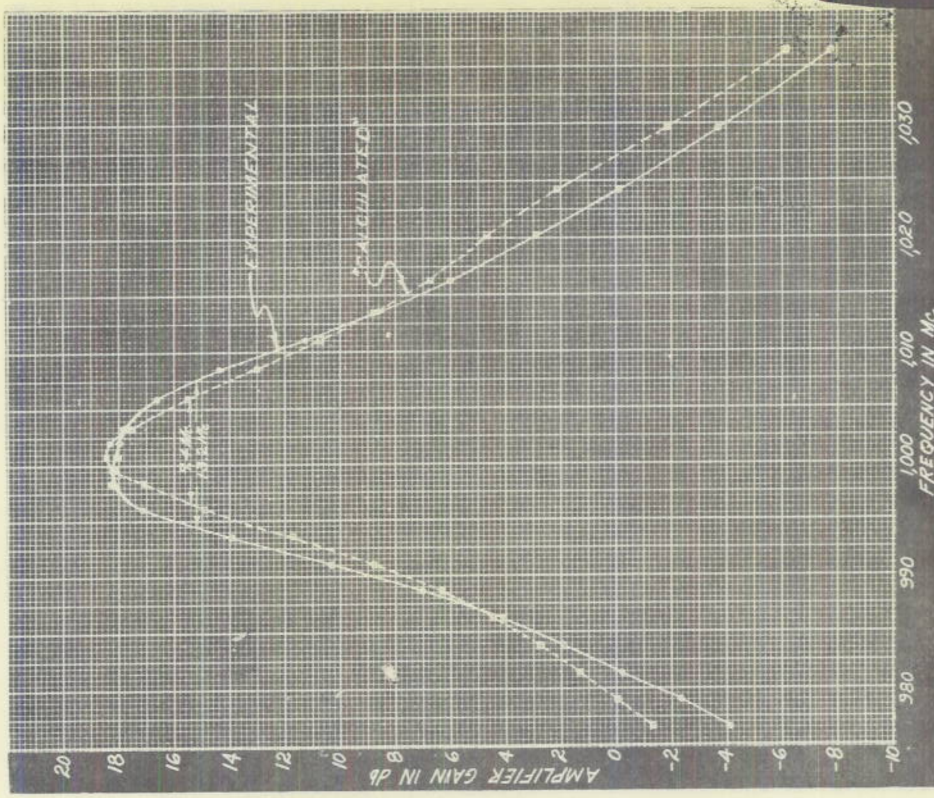

Figure 34 - Cascaded amplifier response curves for 10-3/32" coupling cable

frequency and in adjusting them to the same bandwidth were not worth the small increase in accuracy that would result.

Second, the lossy coaxial cable used as plate load for the last stage had a SWR of 1.6 db. In tuning the individual stages to center frequency, the reactance of this load terminating the individual stage reflected back to the plate cavity and was tuned out. But when the amplifiers are cascaded, the plate cavity of the first stage is loaded by the 50-ohm resistive cathode input of the second stage. The resonant frequency of the first plate cavity is now slightly different than when it was tuned to frequency with the interstage cable connected to the 1.6 db SWR plate load. As the plate of the second cascaded amplifier feeds the same 1.6 db SWR load as that used in the tuning process, its resonant frequency remains substantially constant upon cascading. Therefore, it appears quite probable that (a) the two stages, when cascaded, are not as closely tuned to the same frequency as is indicated by their individual response curves, and that (b) this difference in the resonant frequencies of the plate cavities of the two stages, when cascaded, varies with the length of the output coupling cable of the first stage.

Third, the effective length of the coupling cable is modified without changing the measured length if for any reason the position of the cathode probe of the second stage is changed. The necessity for changing the position of this cathode probe could be caused by aging of the tube during the long series of experiments or by any change in the plate output coupling loop of the second stage.

Fourth, during the series of experiments the gain-bandwidth products of the individual amplifiers varied appreciably because of variation in their 3 db bandwidths.

Even though these errors were present, the results of this investigation show quite conclusively that the effective length of the interstage coupling cable should approximate an integral number of half-wavelengths.

MISCELLANEOUS OBSERVATIONS

Effect of Variation of Plate Current

It was desired to determine the effect of variation of plate current upon the operation of a single stage. Operating with a plate current of 20 ma at 100 volts, a single stage was carefully tuned to 1,001 Mc. The response curve and the SWR curve looking into the cathode input assembly were taken. Then, with no adjustments on any part of the amplifier except the cathode bias, sets of data were obtained for plate currents of 25 ma and 15 ma.

Figure 37 shows graphically the results of this measurement. A decrease of plate current from 25 to 15 ma, (a) increased the frequency of maximum gain less than 1.0 Mc out of 1,000 Mc, (b) decreased the gain from 9.8 to 8.2 db, and (c) did not appreciably change the 3 db and 6 db bandwidths. The minimum cathode input SWR of 0.1 db occurred at 1,001 Mc with a plate current of 20 ma, the conditions under which the stage was aligned. With plate currents of 25 ma and 15 ma, the minimum SWR's were 1.6 db at about 1,000 Mc and 1.7 db at 1,002 Mc, respectively.

In another case the amplifier was aligned and data taken with a plate current of 25 ma. Then, with no other adjustments except that of changing the cathode-bias resistor, sets of data were taken for 20 ma and 15 ma. The results were comparable to those obtained in the first case.

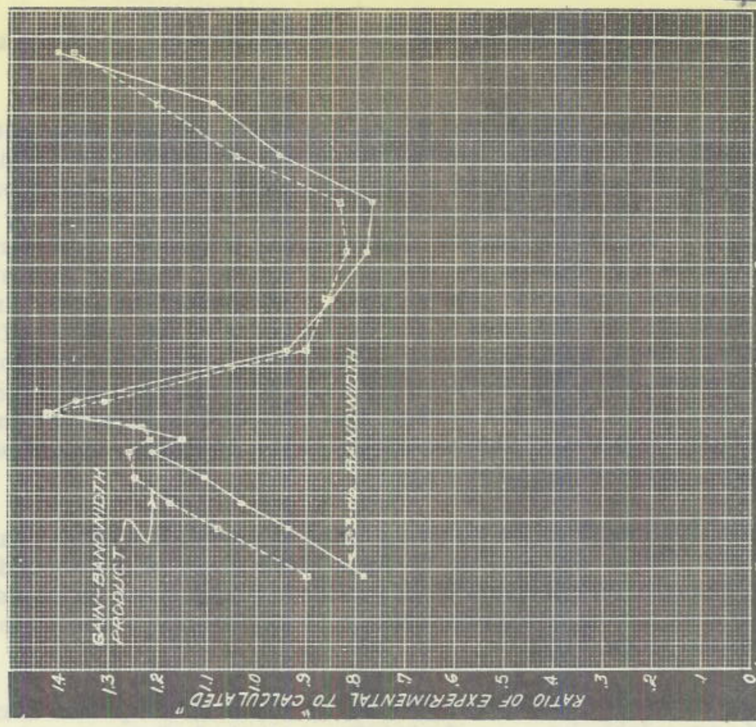

Figure 36 - Ratio of experimental to "calculated" gain-bandwidth product and 3-db bandwidth

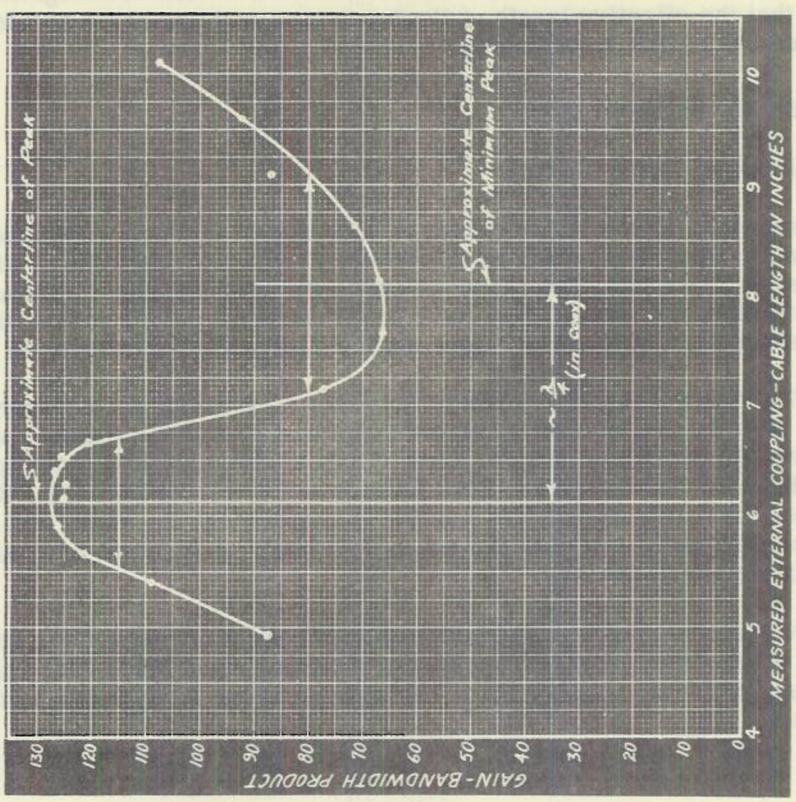

Figure 35 - Variation of gain-bandwidth product of two cascaded stages with coupling-cable length

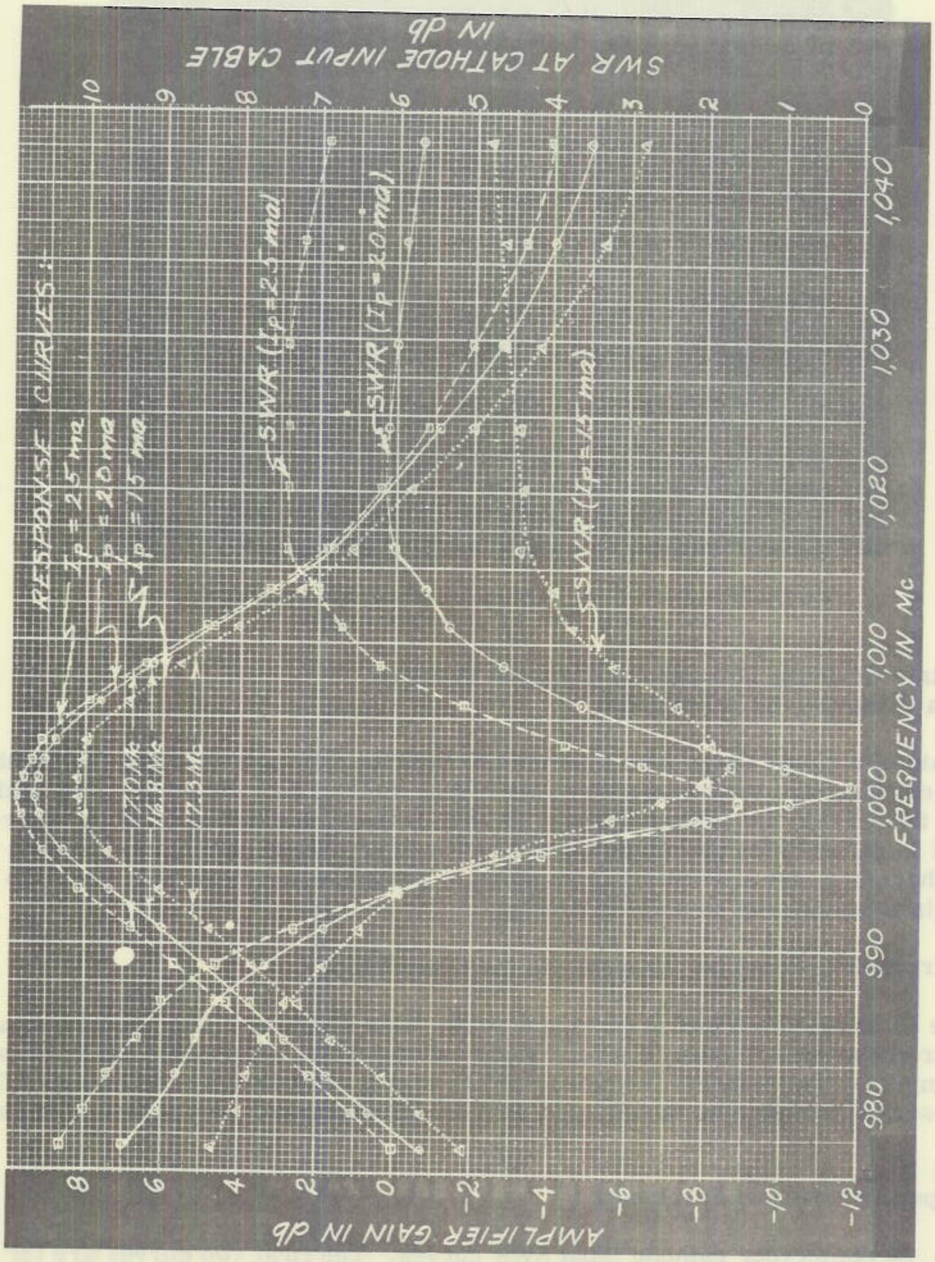

Figure 37 Effect of plate current variation
(single stage tuned with I_p = 20ma)

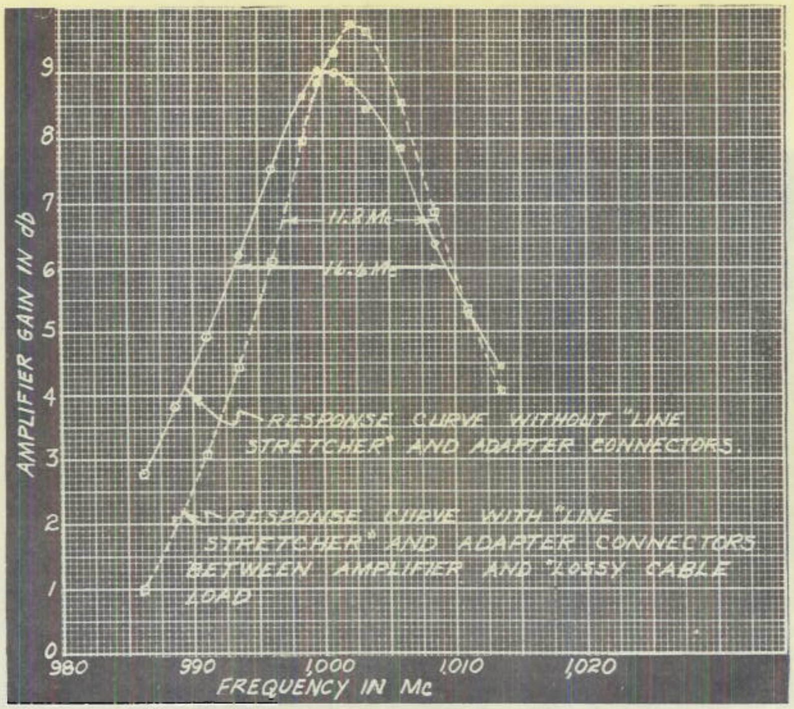

Figure 38 - Effect of frequency-sensitive elements in the plate load of a single stage

Cathode-Plate Coupling Through the Common Electron Stream of a Grounded-Grid Amplifier

The gain of a single stage was measured to be 9 db. By interchanging the input and output cable connectors to the amplifier, the signal was fed into the plate cavity and the output was taken from the cathode cavity. In this case an attenuation of 14 db was realized. The lack of isolation from the output circuit to the input circuit of this amplifier is evident. Furthermore, the attenuation from plate to cathode connectors was only 5 db greater than the gain from cathode to plate connectors.

Effect of Frequency-Sensitive Elements in the Plate Loads

Figure 38 shows the effect of placing a small "line stretcher" and two additional type "N" adaptor connectors between a single stage and the usual lossy cable load. In each case the amplifier was tuned to frequency before taking response measurements. The frequency-sensitive elements in this case narrow the bandwidth and increase the gain of the stage.

Cathode Input SWR versus Frequency

Figure 39 shows four cathode input SWR curves. The first two curves are for two separate stages having the gains and bandwidths indicated. The other two curves are for two-stage cascaded amplifiers made up by connecting the separate units, first in one order and then in the reverse order. Note that the input SWR of a particular cascaded amplifier resembles that of its first stage.

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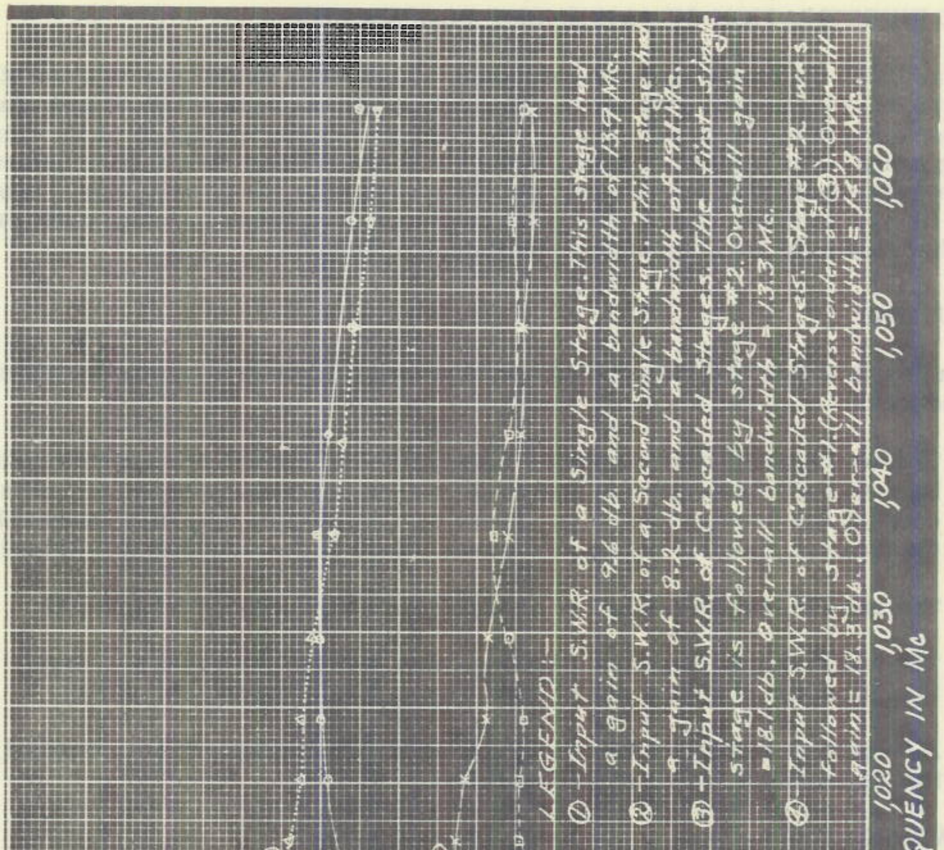

Figure 39 - Cathode input SWR (in db)

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FUTURE INVESTIGATION OF ALIGNMENT PROCEDURES

To facilitate the future study of multistage amplifiers and the investigation of alignment methods, a frequency-modulated signal generator has been designed and constructed. This generator sweeps from 950 to 1,050 Mc with the output constant to ±1 db over the sweeping range. It provides a means for the study of the many circuit parameters and alignment procedures that would be virtually impossible by the point-by-point procedure afforded by c-w methods. Adjustable plate output coupling loops in future plate cavities will further facilitate the alignment of multistage amplifiers.

CONCLUSIONS

As a result of this investigation on microwave grounded-grid coaxial amplifiers, it is concluded that:

(1) The development of a complete i-f amplifier for service use appears quite feasible when mechanically suitable tubes possessing a sufficiently high transconductance become available. These characteristics may be possessed by the Western Electric 416-A microwave triode or the Machlett EP-280 triode.

(2) The gain-bandwidth product of a single stage increases with bandwidth.

(3) The length of the interstage coupling cable between two cascaded stages is critical for obtaining an optimum over-all gain-bandwidth product.

(4) A frequency-modulated test oscillator is highly desirable, if not essential, for alignment of a multistage amplifier.

ACKNOWLEDGMENTS

Acknowledgment is made to Mr. E. A. Hanysz, formerly of the Radio Countermeasures Branch and now of the Research Division of the General Motors Corporation, for his investigation of the manner of variation of gain-bandwidth product with bandwidth for a single stage.

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* * *

APPENDIX

Mathematical Analysis of the Grounded-Grid Amplifier

Before making a mathematical analysis of the grounded-grid amplifier, certain assumptions are necessary to keep the derivations in their simplest forms. These are:

(a) No transit time effect exists.
(b) The coupling coefficients of the cathode and plate transformers are unity.
(c) The over-all amplifier bandwidth is dictated by the Q of the plate load.
(d) The values of μ, r_p, and C_p are invariant with change in bandwidth.

It is to be understood that the analysis is not intended to be a rigorous treatment of the grounded-grid amplifier but rather it is applicable to the particular amplifier under consideration. Assumption (c), for instance, is apropos of the investigation described in the text but is not a supposition that one would generally make. Assumption (b), perhaps the most extreme, was demanded because very little was known about the magnitudes and relative distributions of the inductances involved. However, these assumptions appear to be valid since the experimental results quite closely approximate those obtained from the mathematical analysis.

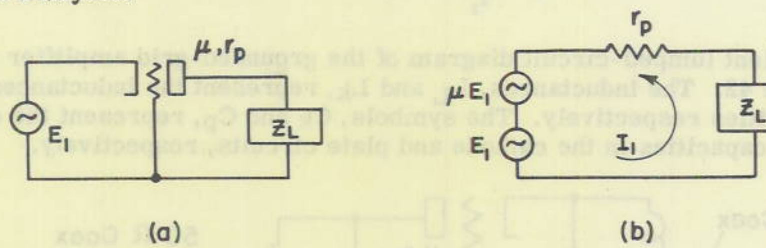

Figure 40

In Figure 40 (a) is shown the diagram of a grounded-grid amplifier with a load of Z_L in the plate circuit and a generator E_1 driving the cathode. The equivalent a-c circuit is shown in Figure 40 (b).

The sum of the voltages around the circuit of Figure 40 (b) equals zero. Thus,

$$E_1 (\mu + 1) - I_1 (r_p + Z_L) = 0. \qquad (1)$$

The output equals $I_1 Z_L$; from equation (1) we get

$$\text{Gain} = G = (\mu + 1) \frac{Z_L}{r_p + Z_L}. \qquad (2)$$

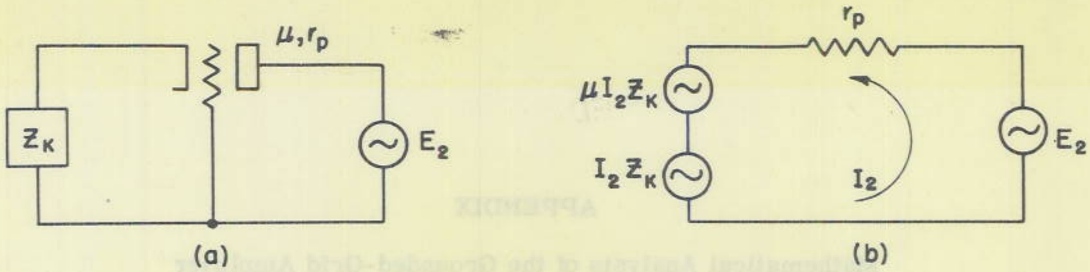

Figure 41

By solving for E_1/I_1 in equation (1) we find the input impedance to the tube, Z_C:

$$E_1/I_1 = Z_C = \frac{r_p + Z_L}{\mu + 1} \tag{3}$$

The impedance looking into the tube from the plate side can be derived from the circuit in Figure 41. Generator E_2 is driving the plate. Impedance Z_k is the cathode load. The sum of the voltages around Figure 41 (b) equals zero. Hence,

$$E_2 - I_2 (r_p + Z_k + \mu Z_k) = 0 . \tag{4}$$

The impedance looking into the tube from the plate side, Z_p, then becomes:

$$Z_p = \frac{E_2}{I_2} = r_p + (\mu + 1) Z_k . \tag{5}$$

The equivalent lumped-circuit diagram of the grounded-grid amplifier employed is shown in Figure 42. The inductances, L_L and L_k, represent the inductances of the plate and cathode cavities respectively. The symbols, C_k and C_p, represent the equivalent lumped-circuit capacities in the cathode and plate circuits, respectively.

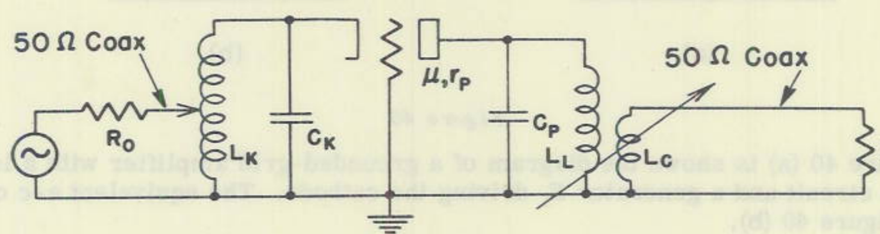

Figure 42

For any given load in the plate circuit, the proper alignment procedure is to "peak" the output with the plate-trimming capacitance and match the signal generator to the cathode circuit with proper cathode coupling and capacitive trimming. The maximum gain at center frequency in the plate is realized when the reactive component of the plate load reduces to zero. This condition results in an effective plate load of R_L. Or,

$$Z_L = R_L . \tag{6}$$

The equation for the gain from cathode to plate then becomes purely resistive, thus,

$$G_{k\text{-}p} = (\mu + 1) \frac{R_L}{r_p + R_L} . \tag{7}$$

The input impedance to the cathode also becomes purely resistive, and

$$Z_c = R_c = \frac{r_p + R_L}{\mu + 1} . \qquad (8)$$

When a circuit is properly matched, it can be opened at any point and the impedances looking in either direction at the open must be complex conjugates of each other. This must be the case at the cathode of the grounded-grid amplifier shown in Figure 42 if the open be made at the cathode. Hence,

$$Z_c = \overline{Z}_k . \qquad (9)$$

Or, since Z_c is purely resistive,

$$R_c = R_k = \frac{r_p + R_L}{\mu + 1} . \qquad (10)$$

Substituting equation (10) into equation (5),we find that the impedance looking into the tube from the plate side reduces to

$$Z_p = R_p = 2r_p + R_L. \qquad (11)$$

It is interesting to note that, although a matched condition exists in the cathode circuit of the plate amplifier, this same condition does not exist in the plate. For if it did, then

$$R_L = 2r_p + R_L, \qquad (12)$$

which condition cannot be realized unless r_p reduces to zero or R_L increases to infinity. The reason for the mismatch in the plate circuit is due to the existence of the tube which is a second generator in this circuit.

Assuming for simplicity that the coupling coefficients of the input and output transformers shown in Figure 42 are unity, and that proper match conditions have been realized, the voltage gains from input to cathode and from plate to output become $\sqrt{R_k/R_o}$ and $\sqrt{R_o/R_L}$, respectively. Thus the over-all gain of the grounded-grid amplifier from coax to coax is merely the gain from cathode to plate (equation 7) multiplied by the two transformer gain factors above. The gain reduces to the following simple expression which can be recognized as the square root of the cathode-to-plate gain:

$$A_o = \left[(\mu + 1) \frac{R_L}{r_p + R_L} \right]^{1/2} . \qquad (13)$$

The cathode cavity was designed to have a low characteristic impedance in order to reduce the effect of the cathode loading capacitance on the length of the cavity. Because of the heavy loading on the cathode (equation 8), the figure of merit of the cathode cavity is low. This characteristic is desirable since it facilitates broadband matching.

The figure of merit of the plate cavity is high in comparison with that of the cathode cavity, and thus it is assumed in this derivation that the over-all bandwidth of the amplifier is dictated by the plate-load bandwidth. The bandwidth is derived from the expression for the Q of the plate circuit:

$$Q = \frac{f}{\Delta f} = 2\pi f C_p R_t, \qquad (14)$$

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where

$$R_t = \frac{R_L R_p}{R_L + R_p} = \frac{R_L (2r_p + R_L)}{2(r_p + R_L)} . \tag{15}$$

Solving equation (14) for Δf, we have

$$\Delta f = \frac{r_p + R_L}{\pi C_p R_L (2r_p + R_L)} . \tag{16}$$

The gain-bandwidth product of the grounded-grid amplifier is the product of equations (13) and (16),

$$A_o \Delta f = \frac{\sqrt{\mu + 1}}{\pi C_p} \cdot \frac{1}{(2r_p + R_L)} \sqrt{\frac{r_p + R_L}{R_L}} . \tag{17}$$

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