Lawrence Berkeley National Laboratory

Recent Work

Title

GREATER GAIN-BANDWIDTH IN TRIGGER CIRCUITS

Permalink

<https://escholarship.org/uc/item/0n6472b5>

Author

Brown, Melvin.

Publication Date

1958-09-25

UNIVERSITY OF
CALIFORNIA

*Radiation
Laboratory*

GREATER GAIN-BANDWIDTH
IN TRIGGER CIRCUITS

TWO-WEEK LOAN COPY

*This is a Library Circulating Copy
which may be borrowed for two weeks.
For a personal retention copy, call
Tech. Info. Division, Ext. 5545*

DISCLAIMER

This document was prepared as an account of work sponsored by the United States Government. While this document is believed to contain correct information, neither the United States Government nor any agency thereof, nor the Regents of the University of California, nor any of their employees, makes any warranty, express or implied, or assumes any legal responsibility for the accuracy, completeness, or usefulness of any information, apparatus, product, or process disclosed, or represents that its use would not infringe privately owned rights. Reference herein to any specific commercial product, process, or service by its trade name, trademark, manufacturer, or otherwise, does not necessarily constitute or imply its endorsement, recommendation, or favoring by the United States Government or any agency thereof, or the Regents of the University of California. The views and opinions of authors expressed herein do not necessarily state or reflect those of the United States Government or any agency thereof or the Regents of the University of California.

UCRL-8473
Instruments

UNIVERSITY OF CALIFORNIA

Radiation Laboratory
Berkeley, California

Contract No. W-7405-eng-48

GREATER GAIN-BANDWIDTH IN TRIGGER CIRCUITS

Melvin Brown

September 25, 1958

Printed in USA. Price 75 cents. Available from the
Office of Technical Services
U. S. Department of Commerce
Washington 25, D. C.

GREATER GAIN-BANDWIDTH IN TRIGGER CIRCUITS

Contents

Abstract.	3
1. Introduction	4
2. Dynamic Plate-Load (DPL)	5
3. DC Analysis	5
4. Transient Analysis of the DPL	14
5. Experimental Results	22
6. Physical Concept of DPL	25
7. Further Considerations	25
8. Conclusion.	26
9. Acknowledgments	26

GREATER GAIN-BANDWIDTH IN TRIGGER CIRCUITS

Melvin Brown

Radiation Laboratory
University of California
Berkeley, California

September 25, 1958

ABSTRACT

The relation between switching speed of a trigger circuit and the gain-bandwidth (GBW) of an amplifier is discussed. A special series connection of two tubes - referred to as a dynamic plate-load amplifier - is then presented and analyzed. It is shown that the dynamic plate-load amplifier may have three times the advantage in GBW over a conventional amplifier using the same tube type. This GBW improvement recommends its utilization in fast trigger circuits.

GREATER GAIN-BANDWIDTH IN TRIGGER CIRCUITS

Melvin Brown

Radiation Laboratory
University of California
Berkeley, California

September 25, 1958

1. INTRODUCTION

The speed of transition of a trigger circuit from one state to another is dependent upon (1) the rate at which an active device charges or discharges the circuit capacitances and (2) the rate at which the passive circuit elements such as the capacitors charge and discharge through their associated resistors.

Generally, in a circuit where the active device, such as a vacuum tube, is permitted to go into cut-off, the time constants of the passive elements determine the switching speed. However, it may be shown that to obtain the maximum speed of operation, it is necessary that the active device be operating at all times.¹

This paper is concerned with vacuum tubes as the active device and presupposes that they will be conducting at all times so as to retain a minimal transconductance. The method usually employed to insure this manner of operation is to clamp the plate and (or) grid swings with biased diodes.

Now it is well known that the speed of switching of an active device is directly dependent on its gain-bandwidth product, abbreviated GBW, as an amplifier.² Attempts to increase the GBW of a circuit by directly paralleling tubes have been unsuccessful because as the transconductance increases, the shunt capacitance increases in the same proportion. However, the special parallel connection of tubes in a distributed amplifier does increase the GBW. Unfortunately, the time delay between the input and output of distributed amplifiers prevents their utilization in regenerative switching circuits.

¹J. Millman and H. Taub, Pulse and Digital Circuits (McGraw-Hill, New York, 1956), p. 152.

²Frederick E. Terman, Electronic and Radio Engineering (McGraw-Hill, New York, 1955) p. 288.

Herein is described a vacuum-tube configuration in which tubes are connected so as to produce a GBW per stage of up to three times that of a conventional amplifier without introducing the unwanted time delay associated with the distributed amplifier.

2. DYNAMIC PLATE-LOAD CIRCUIT

The circuit to be described has been presented by Valley and Wallman as an output circuit for direct-coupled amplifiers without reference to its transient response.³ It has been used recently in a binary which has resolution in excess of 40 Mc.⁴

The basic configuration is shown in Fig. 1 and will be referred to as a dynamic plate-load circuit (DPL).

The analysis will be in two parts: dc (low frequency) analysis where the tube and circuit capacitances are neglected (Section 3), and transient analysis where the capacitances are included (Section 4). The gain and output-impedance equations of the dc analysis agree with those presented by Valley and Wallman. However, since the method of analysis differs from that in Valley and Wallman, and the results of the dc analysis are used to provide the transient characteristics of the circuit, it is included here for completeness.

3. DC ANALYSIS

The equivalent circuit without the shunt capacitances of the DPL is shown in Fig. 2. The tubes, which for simplicity of analysis are identical, are replaced by equivalent voltage generators.

Of interest in the dc analysis is the gain and output impedance of the circuit. The method of approach will be to set up loop equations based upon the equivalent circuit of Fig. 2 and to solve for the desired quantities by determinants.

³Vacuum Tube Amplifiers, G. E. Valley, Jr., and H. Wallman, Eds. (McGraw-Hill, New York, 1948), p. 456.

⁴Michiyuki Nakamura, Forty-Megacycle Scaler, Rev. Sci. Instr. 28, 1015 (1957).

MU-16139

Fig. 1. Schematic of the dynamic plate-load circuit (DPL).

MU-16141

Fig. 2. Equivalent circuit of DPL.

In Loop 1 we have

$$i_1 [2r_p + R_k] + i_2 r_p = \mu [v_{g1} + v_{g2}] \quad (1)$$

where

$$v_{g2} = -i_1 R_k \quad (2)$$

Substituting Eq. (2) into Eq. (1) and rearranging terms, we obtain

$$i_1 [2r_p + (\mu+1) R_k] + i_2 r_p = \mu v_{g1} \quad (3)$$

In Loop 2 we have

$$\begin{aligned} i_1 r_p + i_2 [r_p + R_L] &= \mu v_{g2} \\ &= \mu (-i_1 R_k) \end{aligned} \quad (4)$$

Rearranging Eq. (4) we obtain

$$i_1 [r_p + \mu R_k] + i_2 [r_p + R_L] = 0 \quad (5)$$

Solving for i_2 by Cramer's rule from Eqs.(3) and (5), we find

$$i_2 = \frac{\begin{vmatrix} 2r_p + (\mu+1) R_k & \mu v_{g1} \\ r_p + \mu R_k & 0 \end{vmatrix}}{\begin{vmatrix} 2r_p + (\mu+1) R_k & r_p \\ r_p + \mu R_k & r_p + R_L \end{vmatrix}} \quad (6)$$

Expanding Eq. (6) and simplifying, we obtain

$$i_2 = - \frac{\mu v_{g1} (r_p + \mu R_k)}{r_p^2 + R_k r_p + (\mu+1) R_k R_L + 2r_p R_L} \quad (7)$$

Now the dc gain is

$$\frac{v_o}{v_{g1}} = \frac{i_2 R_L}{v_{g1}} \quad ; \quad (8)$$

Hence, substituting Eq. (7) in Eq. (8) and rearranging, we obtain

$$\frac{v_o}{v_{g1}} = - \frac{\mu (r_p + \mu R_k)}{\frac{r_p}{R_L} (r_p + R_k) + (\mu + 1) R_k + 2r_p} \quad (9)$$

Now for comparison, the gain expression of a conventional amplifier is

$$\frac{v_o}{v_g} = \frac{\mu R_L}{R_L + r_p} \quad (10)$$

Since the conventional amplifier will be referred to often, for the sake of clarity its well known physical and equivalent circuit are shown in Fig. 3.

It is now desirable to plot Eqs. (9) and (10) with R_L and R_k as varying parameters for a dual triode 6BQ7 (Fig. 4).

Both identical halves of a 6BQ7 are used in the DPL, while one half of the same tube is used in the conventional amplifier. It is seen in Fig. 4 that, as R_k gets larger and R_L is held constant, the gain advantage of the DPL increases. However as R_k gets larger it increases the grid bias on the upper tube and tends to cutoff the tube, or at least it will decrease the anode current. This decreased anode current reduces the g_m and μ of the tube. In order to offset the reduced anode current, the B+ voltage may be increased until the maximum anode-voltage rating is reached. Thus in the case of the 6BQ7 an anode current of 7.5 ma and an R_k of 560 ohms were the maximum values which satisfied the above conditions.

The tube parameters that were substituted into Eqs. (9) and (10) were directly measured at an anode current of 7.5 ma. These parameters are: $r_p = 6100$ ohms, $g_m = 7000$ μ mhos, and $\mu = 43$.

The curves of Fig. 4 have been verified by direct measurement. The results show that for low-load resistances the DPL has considerably more gain than the conventional amplifier. In particular for the case we will be interested in later, with an R_L of 1000 ohms and an R_k of 560 ohms the DPL has 2.75 times the dc gain of the conventional amplifier.

We now desire to obtain the output impedance of the DPL. It may be shown that to obtain the output impedance of a two-port device it is

MU-16142

Fig. 3. Conventional amplifier: (a) Physical circuit, (b) Equivalent circuit.

MU-16144

Fig. 4. Comparison of the dc gain of the DPL with that of a conventional amplifier for a 6BQ7 tube.

necessary only to open-circuit the input and to drive the output with a voltage source.⁵ The output impedance is then the ratio of the driving voltage to the current leaving this source. The equivalent circuit to obtain the output impedance is shown in Fig. 5.

The equation for Loop 1 of Fig. 5 is

$$\begin{aligned} i_1 (2r_p + R_k) + i_2 (r_p) &= \mu v_{g2} \quad (\mu v_{g1} = 0) \\ &= \mu (-i_1 R_k). \end{aligned} \quad (11)$$

Rearranging Eq. (11), we obtain

$$i_1 (2r_p + (\mu + 1) R_k) + i_2 r_p = 0. \quad (12)$$

For Loop 2, we have

$$\begin{aligned} i_1 r_p + i_2 r_p &= \mu v_{g2} - v_2 \\ &= \mu (-i_1 R_k) - v_2, \end{aligned} \quad (13)$$

which can be rearranged to

$$i_1 (r_p + \mu R_k) + i_2 r_p = -v_2. \quad (14)$$

Solving Eqs (12) and (14) for i_2 by Cramer's rule, we obtain

$$i_2 = \frac{\begin{vmatrix} 2r_p + (\mu + 1) R_k & 0 \\ r_p + \mu R_k & -v_2 \end{vmatrix}}{\begin{vmatrix} 2r_p + (\mu + 1) R_k & r_p \\ r_p + \mu R_k & r_p \end{vmatrix}} \quad (15)$$

Expanding Eq. (15) and simplifying, we have

⁵Thomas L. Martin, Jr., Electronic Circuits, (Prentice-Hall, Englewood Cliffs, New Jersey, 1956), p. 92.

MU-16140

Fig. 5. Equivalent circuit of DPL to obtain the output impedance.

$$i_2 = -\frac{v_2 [2r_p + (\mu + 1) R_k]}{r_p [r_p + R_k]} \quad (16)$$

Now the output impedance is

$$Z_{22} \equiv \frac{v_2}{i_2} \quad (17)$$

Hence, from Eq. (16) and (17) we obtain

$$Z_{22} = \frac{r_p (r_p + R_k)}{2r_p + (\mu + 1)R_k} \quad (18)$$

A plot of Eq. (17) with R_k as the independent parameter is shown in Fig. 6. This curve has been verified by direct measurements. With an $R_k = 560$ ohms, the output impedance, Z_{22} , is 1100 ohms. This is approximately 18% of the output impedance of the conventional amplifier, which is 6100 ohms.

4. TRANSIENT ANALYSIS OF THE DPL

For simplicity of analysis, the current-generator equivalent circuit will be used for the transient analysis.

The physical circuit is shown in Fig. 7a where the various shunt capacitances are included. The equivalent circuit of Fig. 7a is shown in Fig. 7b. Note the identities between the equivalent capacitances in the legend of Fig. 7b and the physical capacitances of Fig. 7a. Also it should be observed that G_3 is a parallel combination of the current-generator shunt conductance and the load conductance.

In the analysis which follows, C_{in} of Fig. 7b includes the Miller capacitance of the lower tube. This capacitance as well as the generator internal impedance R_g is neglected in the following analysis, because we are assuming that $R_g C_{in}$ is a much smaller time constant than any other in the system.

The method of analysis will be to write the nodal equations and then to obtain the system transfer function. The location of the poles of the system then will be determined by setting the denominator of the transfer

MU-16143

MU-16138

Fig. 7. (a) Schematic of DPL showing capacitances.
 (b) Equivalent circuit of DPL with $C_1 = C_{PG} + C_{GP} + C_{Stray}$,
 $C_2 = C_{GK} + C_{Stray}$, $C_3 = C_{KF} + C_{PK} + C_{Stray} + C_L$,
 and $G_3 = (R_L + r_P) / (R_L \cdot r_P)$.

function equal to zero. It will be shown that one of the poles is dominant. The product of this pole, and the dc gain will give the gain-bandwidth of the system.

The nodal equations are

$$v_1 [p(C_1 + C_2) + G_1 + G_2] - v_2 [pC_2 + G_2] = -g_m v_{g1} \quad (19)$$

and

$$-v_1 [pC_2 + G_2] + v_2 [p(C_2 + C_3) + G_2 + G_3] = g_m v_{g2} \quad (20)$$

Now noting that

$$g_m v_{g2} = -g_m [v_2 - v_1] \quad (21)$$

and substituting Eq. (21) into Eq. (20), we obtain

$$-v_1 [pC_2 + G_2 + g_m] + v_2 [p(C_2 + C_3) + G_2 + G_3 + g_m] = 0. \quad (22)$$

Solving for v_2 by Cramer's rule with Eqs. (19) and (22), we obtain

$$v_2 = \frac{\begin{vmatrix} p(C_1 + C_2) + G_1 + G_2 & -g_m v_{g1} \\ -[pC_2 + G_2 + g_m] & 0 \end{vmatrix}}{\begin{vmatrix} p(C_1 + C_2) + G_1 + G_2 & -[pC_2 + G_2] \\ -[pC_2 + G_2 + g_m] & p[C_2 + C_3] + G_2 + G_3 + g_m \end{vmatrix}} \quad (23)$$

Expanding Eq. (23) and rearranging, we obtain

$$\begin{aligned} v_2/v_{g1} = & [-g_m(pC_2 + G_2 + g_m)] / \{p^2[C_3(C_2 + C_1) + C_1C_2] \\ & + p[C_1(G_2 + G_3) + C_2(G_1 + G_3) + C_3(G_1 + G_2) + C_1g_m] \\ & + G_1(G_2 + G_3 + g_m) + G_2G_3\} \end{aligned} \quad (24)$$

Now the location of the poles of the transfer function v_2/v_{g1} may be obtained by setting the denominator of Eq. (24) equal to zero. Hence we have

$$p^2 + p \left[\frac{C_1(G_2 + G_3) + C_2(G_1 + G_3) + C_3(G_1 + G_2) + C_1 g_m}{C_3(C_2 + C_1) + C_1 C_2} \right] \\ \frac{G_1(G_2 + G_3 + g_m) + G_2 G_3}{C_3(C_2 + C_1) + C_1 C_2} = 0 \quad (25)$$

It is now desirable to illustrate a typical calculation involving Eq. (25). The values of the capacitances of Fig. 7a were measured with a Tektronix Model 130 LC meter and are, in μmf :

$$C_{GP} = 1.7$$

$$C_{PK} = 1.9$$

$$C_{GK} = 3.3$$

$$C_{KF} = 3.6$$

$$C_{PG} = 1.9$$

These values include the socket capacitances. Care was exercised in using the guard voltage of the LC meter so as to isolate the above capacitances from each other.

In addition there are stray capacitances due to the physical components and the output connections. These values were also measured and are added to the above capacitances to obtain the values for the equivalent circuit. Hence, from the identities of Fig. 7b, the equivalent circuit capacitances are, in μmf :

$$C_1 = C_{PG} + C_{GP} + C_{\text{Stray}} = 1.9 + 1.7 + 2.0 = 5.60$$

$$C_2 = C_{GK} + C_{\text{Stray}} = 3.3 + 1 = 4.3$$

$$C_3 = C_{KF} + C_{PK} + C_{\text{Stray}} + C_2 = 3.6 + 1.9 + 5.2 + C_L \\ = 10.7 + C_L$$

Now choosing $R_k = 560$ ohms and $R_L = 1000$ ohms, the conductances of Fig. 7b are, in μmhos :

$$G_1 = \frac{1}{r_p} = \frac{1}{6100} = 164$$

$$G_2 = \frac{1}{R_k} = \frac{1}{560} = 1790$$

$$G_3 = \frac{R_L + r_p}{R_L r_p} = \frac{10^3 + 6.1 \times 10^3}{6.1 \times 10^6} = 1170$$

Normalizing the above conductances by dividing by 10^{-3} and the capacitances by multiplying by 10^{12} and substituting them into Eq. (25), we obtain

$$p^2 + \frac{p(1.95 C_L + 84.7)}{9.95 C_L + 140.3} + \frac{3.73}{9.95 C_L + 140.3} = 0. \quad (26)$$

For the one remaining substitution a normalized load capacitance, C_L , of 25 farads is chosen and Eq. (26) is solved for its roots. Thus we obtain $p_1 = 0.031$ and $p_2 = 0.313$.

Since p_1 is one-tenth as far as p_2 from the origin of the p plane then p_1 may be considered a dominant pole. Substituting other values of C_L into Eq. (26) will show that p_1 is always at least ten times closer to the p -plane origin than p_2 . Therefore p_2 will be neglected, and p_1 will be considered the only pole that affects the bandwidth of the system. It may be shown that in a one-pole system, the numerical value of the pole is identical with the 3-db bandwidth in units of radians per second.⁶

Now denormalizing p_1 and dividing by 2π to convert to cycles per second, we obtain

$$BW = \frac{p_1 \times 10^9}{2\pi} = 4.94 \text{ Mc.}$$

The dc gain may be determined from Fig. 4 or Eq. (9) for $R_L = 1000$ ohms. Thus the gain is 16.8. Hence GBW_{DPL} is $16.8 \times 4.94 = 83.0$ Mc.

⁶Martin, loc. cit., p. 117.

Now for a conventional amplifier we have

$$BW = \frac{G_{Total}}{2\pi C_{Total}} \quad , \quad (27)$$

where

$$G_{Total} = \frac{R_L + r_p}{R_L r_p}$$

and

$$C_{Total} = C_{Tube} + C_{Strays} + C_{Load}$$

Substituting into Eq. (27) the same load resistance and capacitance as in the DPL example above and using measured values of C_{Tube} and C_{Strays} , we obtain

$$BW = \frac{1.17 \times 10^{-3}}{2\pi (1.9 + 4.1 + 25) \times 10^{-12}} = 6.01 \text{ Mc.}$$

The dc gain for the conventional amplifier obtained from Fig. 4 or Eq. (10) is 6.05. Hence, we have

$$GBW_{Conv. Ampl.} = 6.05 \times 6.01 = 36.4 \text{ Mc.}$$

Similar computations of GBW were made for other values of load capacitance for both the DPL and conventional amplifier and the results are plotted in Fig. 8. The curves of Fig. 8 show the advantage of the DPL over the conventional amplifier. The GBW improvement is a function of load capacitance and varies approximately from 80% with $C_L = 15\mu\text{f}$ to 200% with $C_L = 55\mu\text{f}$. The lack of substantial improvement of the DPL at very low load capacitances (around 5 μf) may be accounted for by considering all the additional tube and stray capacitances present in the DPL in contrast to those present in the conventional amplifier. However, at larger load capacitances the tube and stray capacitances are masked out, and the DPL shows its advantage. Furthermore, in physical trigger circuits the load capacitance is generally greater than 5 to 10 μf , and the above theoretical limitation is often not pertinent.

MU-16145

Fig. 8. Theoretical comparison of GBW of the DPL and conventional amplifier for a 6BQ7 tube.

The question of the optimum value of R_k to obtain maximum GBW is of interest. Simple calculations show that an increase of R_k decreases the bandwidth less than it increases the dc gain; therefore R_k should be as large as possible. The upper limit on the value of R_k has been discussed in the dc analysis (Section 3).

5. EXPERIMENTAL RESULTS

Experimental confirmation of Fig. 8 has been obtained by observing the pulse response of the DPL and conventional amplifier whose parameters were given in the previous section. The same tube was used for both tests. In each case the anode current was 7.5 ma and the signal input bias voltage was -0.5 v. The dc gain of the DPL measured 16.5 and the load resistance of the conventional amplifier was adjusted to give the same gain. The pulse response to a 0.2 v negative step input pulse as the load capacitance is varied is shown in Fig. 9.

The pulse source was a mercury-relay pulse generator with an internal impedance of 125 ohms and a rise time of a fraction of a μ sec. The photos were taken on a Tektronix 517 A oscilloscope which has a rise time of 7 μ sec. These photos were magnified 24 times to permit careful measuring of the 10 to 90% rise time.

Translating rise time to bandwidth by using $BW = 0.35/T_R$ and multiplying the result by the dc gain, we obtain the experimentally measured GBW plotted in Fig. 10. This may be compared to the theoretical results of Fig. 8.

The maximum error between experimental and theoretical GBW occurs at the lowest values of load capacitance and is approximately 16% for the DPL. As the load capacitance is increased the error decreases to about 6% at $C_L = 55\mu\text{f}$. Examination of Eq. (25) will show that the tube and stray capacitances contribute heavily to the value of GBW at low load capacitances. Therefore it would appear that there were errors in the measurement of the tube and stray capacitances for the equivalent circuit. It would further seem that large load capacitances mask out these tube and stray capacitances, and thus their contributory error to the GBW is minimized. Substitution of other values of tube and stray capacitances into Eq. (25) will show that

ZN-2047

Fig. 9. Rise time of DPL (left) and conventional amplifier (right) as load capacitance is varied (sweep speed: 50 m μ s/cm; pulse amplitude: 3.3 v). Values of C_L are (top to bottom) 5 μ f, 15 μ f, 30 μ f, and 55 μ f.

MU-16146

Fig. 10. Experimental GBW of DPL and conventional amplifier for a 6BQ7 tube.

the above interpretation is valid. The GBW product is relatively insensitive to even a 25% change in the tube and stray capacitances when the load capacitance is large.

6. PHYSICAL CONCEPT OF THE DPL

It is desirable to obtain some physical picture of the DPL to explain its advantage. At the risk of oversimplification an attempt will be made to do so.

One way to understand the advantage of the DPL over the conventional amplifier is by considering the results of the dc analysis. The low output impedance of the DPL is in parallel with the external load resistance and capacitance. This low impedance constrains the parallel combination to have a short time constant.

Another (or possible similar) way of looking at the DPL is to consider the ratio of internal voltage-generator impedance of the DPL to its load impedance. In the DPL, the voltage drop across the internal generator impedance is less than in a conventional amplifier. Hence more voltage is available at the output terminals, providing more gain and consequently a greater GBW. The essential advantage appears to rest in the lower output impedance of the DPL.

Because the plate load of the lower tube is an active device (another tube), and it is this active device that accounts for the lower output impedance, the name dynamic (or active) plate load suggested itself.

7. FURTHER CONSIDERATIONS

Shunt peaking may be employed in the DPL as in a conventional amplifier. However, in the DPL the inductance is placed in series with the cathode resistance, R_k , instead of in series with the load resistance. It has been found experimentally that the same relative advantage in GBW may be obtained in the DPL as in a conventional amplifier by shunt peaking.

The tube used for comparison above was medium- μ dual-triode 6BQ7. If such a tube is used in one side of a regenerative trigger circuit, the formerly neglected Miller capacitance should be considered as part of the load capacitance of the other tube. If pentodes are used, the neglect of the Miller capacitance is, of course, fully justified.

It has been found experimentally and by calculations for the DPL that varying the load resistance, R_L , has little effect on the GBW with a given load capacitance, C_L . That is, if the gain is halved by decreasing R_L , the BW is doubled, provided the load capacitance remains the same. However, as may be seen from Fig. 8, GBW is not constant with varying load capacitance for a given R_L . This means that if C_L is halved, the GBW is not doubled at a fixed R_L .

Hitherto no mention has been made of the time delay in the DPL. Inspection of the physical circuits of the DPL and the conventional amplifier indicates no appreciable increase in time delay. A comparison of the circuitry of the DPL and a distributed amplifier (another method for improving GBW) emphasizes the relative insignificance of time delay in the DPL.

8. CONCLUSION

The dc analysis shows that the DPL with a typical value of $R_k = 560$ ohms has an output impedance about 18% that of a conventional amplifier using the same tube type. Furthermore, the dc gain at low load resistances (around 1000 ohms) varies from two to three times that of the conventional amplifier, depending upon the value of R_k .

The transient analysis shows that for one tube type (6BQ7) the DPL may have a factor of three improvement in the GBW over a conventional amplifier. Empirical confirmation of the transient analysis checks to within 18%, and the areas of error between the results have been discussed.

The broadbanding factor in the DPL appears to be R_k , and it should be made as large as possible, the maximum value being determined by the tube type and operating potentials.

All results tend to recommend the utilization of the DPL as the active device in regenerative trigger circuits.

9. ACKNOWLEDGMENTS

Discussions during the course of this work with Prof. Donald O. Pederson of the Electrical Engineering Department of the University of California have been most valuable. Thanks are also due Michiyuki Nakamura

of the University of California Radiation Laboratory for introducing the author to the possibilities of the DPL in trigger circuits.

This work was done under the auspices of the U. S. Atomic Energy Commission.

This report was prepared as an account of Government sponsored work. Neither the United States, nor the Commission, nor any person acting on behalf of the Commission:

- A. Makes any warranty or representation, express or implied, with respect to the accuracy, completeness, or usefulness of the information contained in this report, or that the use of any information, apparatus, method, or process disclosed in this report may not infringe privately owned rights; or
- B. Assumes any liabilities with respect to the use of, or for damages resulting from the use of any information, apparatus, method, or process disclosed in this report.

As used in the above, "person acting on behalf of the Commission" includes any employee or contractor of the Commission to the extent that such employee or contractor prepares, handles or distributes, or provides access to, any information pursuant to his employment or contract with the Commission.