NEW TECHNIQUES IN I. F. AMPLIFIER DESIGN

BY

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PREFACE

The work described in this report was carried out during the third term commencing January 15, 1951 and ending April 6, 1951. During this period, the author was assigned to the Research and Development Department of Bendix Radio, Division of Bendix Aviation Corporation, Baltimore, Maryland.

The project selected, involved work on a recently developed series connected I.F. amplifier. This equipment has more recently been referred to as direct coupling in H.F. amplifier.

After the invention at Bendix Radio of this amplifier, most of their work with this circuit has been directed along the line of adapting it for use with subminiature tubes and packaging. Bendix engineers were striving to get the maximum gain per cubic inch out of this amplifier. Mr. Crosby, the inventor, had received an invitation to give a paper, describing this circuit, at the I.R.E. Convention. My problem was to (a) design and build a stagger-tuned direct coupled amplifier, and (b) obtain the performance data needed by Mr. Crosby for his paper. Of primary interest was the effect of varying 'B+' voltage upon the apportioning of the strip voltage to the various stages. Additionally, it was suggested that performance measurements be made to determine the effect of different size bypass condensers.

For this construction a center frequency of 30 mc.,

was decided upon. This represents a compromise in that it permits the use of small components and yet does not introduce the feedback problems connected with higher frequencies. For use as a test amplifier, it was very important that the stagger-tuned band pass design be correct, and that all precautions be taken to prevent coupling between stages. Degeneration or regeneration occurring during any of the performance tests would lead to erroneous evaluation of the circuit.

A new circuit, based upon information gain from the performance associated with different size bypass condensers, was designed and tested. This circuit possessed many advantages among which was an improved gain bandwidth product.

Acknowledgements

The author gratefully acknowledges the cooperation and assistance of Edward Crosby, Jr., project engineer. Thanks are extended to C.C. Bath and C.P. Means for their assistance with the performance tests.

TABLE OF CONTENTS

LIST OF ILLUSTRATIONS

TABLE OF SYMBOLS AND ABBREVIATIONS

C	-	Total interstage capacity, includes inter-electrode, wiring and shunt tuning capacitances.
G_m	-	Grid to plate transconductance
B	-	Bandwidth measured at the 3 db. points
τ	-	Pulse duration
R	-	Total resistance loading the tuned circuit
L	-	Inductance necessary to resonate with 'C'.
f_0	-	Center frequency of I.F. amplifier
f_1	-	Center frequency of high stage of either a stagger-pair or stagger-triple
f_2	-	Center frequency of low stage of either a stagger-pair or stagger-triple
Δf	-	Over all bandwidth corresponding to 3 db. points
Δf_1	-	Bandwidth of stage having center frequency 'f_1'
Δf_2	-	Bandwidth of stage having center frequency 'f_2'
Δf_0	-	Bandwidth of center stage in a stagger-triple
$\Delta f_1'$	-	Bandwidth of 3rd., stage of staggered Quadruple
$\Delta f_2'$	-	Bandwidth of 4th., stage of staggered Quadruple
$\Delta f''$	-	Corrected over all bandwidth required when more than one pair or triple are used in cascade.
R_0	-	Total loading across coil to produce Δf_0
R_1	-	Total loading across coil to produce Δf_1
R_2	-	Total loading across coil to produce Δf_2
R_1'	-	Total loading across coil to produce $\Delta f_1'$
R_2'	-	Total loading across coil to produce $\Delta f_2'$
db	-	Decibel
$I.F.$	-	Intermediate frequency

H.F. - High frequency

D.C. - Direct current

B+ - Plate supply voltage

FCC - Federal Communications Commission

MIT - Massachusetts Institute of Technology

μf - Microfarad

$\mu\mu f$ - Micro microfarad

k - Thousand

G - Stage gain

NEW TECHNIQUES IN I.F. AMPLIFIER DESIGN

Introduction

Prior to the recent development of television, radar and missile guidance, broad band I.F. circuits were not generally needed. The principle coupling system for I.F. amplifiers was the familiar double-tuned transformer.

Work on the wide band requirements of both television and radar, brought to light the limiting factors in I.F. amplifier design. Measurements made on this problem revealed that regardless of the type of interstage coupling, the gain and bandwidth capabilities of any tube in an amplifier circuit are ultimately limited by its figure of merit. The figure of merit for a single stage amplifier is the product of voltage gain at band center by bandwidth. Stage gain can be expressed as $G_m R$, stage bandwidth as $\frac{1}{2\pi RC}$, and the product of these two quantities gives a figure of merit of $\frac{G_m}{2\pi C}$. 'C', in this case, includes tube interelectrode capacitances, wiring capacitances, and shunt tuning capacitances. The greatest gain for a given bandwidth is acquired when the ratio $\frac{G_m}{C}$ is a maximum. For a given tube this is accomplished by reducing 'C' to a minimum.

The series amplifier, whose characteristics are described in Chapter III was developed at Bendix Radio as a result of their attempts to subminiaturize existing I.F. amplifier equipment. After reducing the size of the components to a minimum, the next step in subminiaturization was a reduction in the number of the components, and if possible

provide increased reliability in the circuit performance.

In the high-gain I.F. amplifier, it appeared that considerable volume could be saved by the elimination of the plate-screen decoupling chain. These decoupling impedances are necessary to prevent the 'B+' bus from coupling the stages. The 'B+' bus is required because all of the stages in a conventional amplifier are shunted across a common plate supply. It was learned that connecting several stages, of an I.F. amplifier, in series across the power supply eliminates the 'B+' bus and the plate-screen decoupling chain.

CHAPTER I

DESIGN CONSIDERATIONS

In designing a wide band amplifier employing optimum tube performance, i.e., a minimum number of tubes for a given gain and bandwidth, the tube having the largest $\frac{Gm}{C}$ ratio is used. No shunt capacitances are employed as tuning, other than those which unavoidably occur in the tube and wiring. Any addition to this shunt capacitance lowers the gain for a given bandwidth or conversely lowers the bandwidth for a given gain.

The performance of a wideband amplifier depends on a great many factors, some of which are controllable and some of which are not. These factors include, desired amplitude-frequency response, desired phase-frequency response, transient or pulse response, stability with respect to regeneration, stability with respect to tube variation, noise figure, ease of alignment, etc.

(1) Factors determining I.F. amplifier gain requirements

When deciding upon the design requirements, consideration should first be given to the partition of the over-all gain between the I.F. amplifier and the video amplifier. In determining the over-all amplifier gain, the following arguments are considered in arriving at the proper design value.*

(1) For a given speed of pulse response, only half as much bandwidth is required in a video amplifier as in a centered I.F. amplifier.

* Microwave Receivers; Van Voorhis; McGraw Hill, p. 156

(2) Comparison of equivalent video and I.F. amplifier circuits, shows that higher gain bandwidth products are obtained from video stages.

The influence of the above factors upon the apportioning of more over-all gain in the direction of the video amplifier is over powered in favor of the I.F. amplifier, for the following reasons.*

(1) High-gain video amplifiers are inferior to high-gain I.F. amplifiers with regard to recovery time, transmission of long blocks of signals, and sensitivity to microphonics, hum, and power-supply variations.
(2) The output vs. input curve for a second detector is linear above a certain level and square-law below that level. For signals in the square-law range, 1-db of amplification preceding the second detector is equivalent to 2-db of amplification after the second detector. For this reason it is generally economical to provide enough gain in the I.F. amplifier to bring the weakest desired signal up to the level where the second detector action is linear.

(2) Bandwidth considerations

When considering the fidelity of pulse reproduction, the greater the bandwidth the better. Between the regions of increasing signal-to-noise ratio for small bandwidths and decreasing signal-to-noise ratio for large bandwidths there exists an optimum value which occurs for $B\tau = 1$. Here 'B' is the bandwidth measured at the 3-db points, 'τ' is the pulse duration.

Except for amplifiers of extremely large bandwidth, it is common to make the I.F. bandwidth closer to twice the reciprocal of the pulse length. This wider bandwidth costs little in minimum detectable signal, and affords the

* Microwave Receiver; Van Voorhis; McGraw Hill, p. 156

advantages of greater pulse fidelity and reduced criticalness in local oscillator tuning and in automatic frequency control.

(3) Center Frequency

The choice of center frequency is up to the designer in some cases, and in others, it is a fixed industrial standard or FCC regulation. The lower limit generally is fixed by bandwidth considerations. The selected frequency is invariably an engineering compromise based upon the particular requirements involved. In brief, the following arguments are considered for a low intermediate frequency.**

(a) For a given bandwidth the relative detuning caused by tube capacity variations is less for a low intermediate frequency. Hence, amplifiers of a given bandwidth and type exhibit less variability because of manufacturing tolerances and tube replacement if they have a low center frequency.
(b) The optimum noise figure attainable with a given tube type is lower at low frequencies.
(c) The input conductance of a tube due to cathode-lead inductance varies as the square of the frequency, and to the extent to which this loading is a problem, there is an advantage in a low intermediate frequency. An additional advantage of a low intermediate frequency follows from the fact that the susceptance of the grid-plate capacitive feedback path increases with frequency.

A high intermediate frequency has the advantage of better image rejection properties. In general the higher the intermediate frequency the smaller are the tuning coils and the bypass condensers. Also, at the higher intermediate frequencies, a wider separation of video and intermediate frequency is possible.

** Microwave Receiver; Van Voorhis; McGraw Hill, p.158

If superheterodyne reception is employed, the center frequency as a rule will be chosen on the basis of image response. To suppress images in a multi-channel receiver, entirely on the basis of I.F. response, the channel separation must either be greater than twice the I.F. center frequency or all of the channels should occupy a frequency band less than twice the center frequency in width. If no image problem exists, the choice may depend on the physical size, and the time constants of the de-coupling networks used in the amplifier.

<u>(4) Single-Tuned Band pass Amplifier</u>

Generally, the design starts with a presumed bandwidth and gain. If a pentode is used, a single stage is governed by the following design parameters. The gain at band center is $G = \frac{G_m}{2\pi C \Delta f}$ where 'Gm' is the transconductance, 'C' is the total shunt capacitance. The value of R required to give this bandwidth is $R = \frac{1}{2\pi C \Delta f}$.

The single-tuned inter stage coupling, fig. (1) was used in this amplifier, and its design will be considered at this time.

$C = C_{out} + C_{in} + C_{stray}$

$R = R + r_{p(t_1)} + r_{in(t_2)}$

L is the inductance resonating with C to the desired center frequency

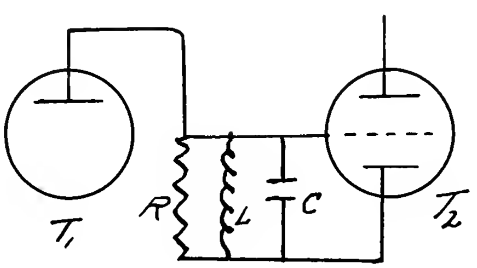


Fig. (1) Single-tuned inter-stage coupling

These single-tuned circuits may be employed in two ways in multi-stage amplifiers. The stages may be all identical, and all tuned to the same center frequency, or may be arranged in staggered groups in which the stages are not identical.

The first type, in which all stages are identical, is generally known as synchronous single-tuned amplifiers, and for medium bandwidth, and present type tubes, this type of amplifier is feasible. As long as adjacent channel selectivity is not a problem, it is a desirable amplifier. Since the response is relatively insensitive to detuning of individual stages, the couplings may be made fixed tuned if so desired. The smooth frequency response of synchronous tuned amplifiers gives an ideal transient response. Due to the relatively insensitive to detuning, gain control, which usually results in detuning of the stages so controlled, has little effect on the response. Until tubes with a higher Gm/C ratio can be obtained, this type of amplifier is not practical for very wide band use. Also, when adjacent channel selectivity requires sharp skirts, this type of coupling is not satisfactory.

The problems of radar, particularly those brought forth by very short pulse lengths, and the simplicity of single-tuned amplifiers, prompted an investigation of the possibilities of the staggered amplifier. Such an investigation was made by the receiver section of the M.I.T. Radiation Lab. The findings of these investigations are reported by

Dr. Henry Wallman, (9), staff mathematician for the receiver section. This report finds that by staggering frequencies, greater bandwidths may be obtained for the same gain, for a given figure of merit.

In the synchronous single-tuned amplifier, with identical stages, the shape of the response curve is unique. The staggered tuning system no longer gives a unique response. Its response may be curved at midband, or flat or may contain dips. Each type of response gives a unique stage gain times over all bandwidth product. For reasons of mathematical simplicity, as well as compromise between transient response and gain times bandwidth factor, Dr. Wallman has chosen a curve he calls 'transitional coupling'. This is a coupling which gives the greatest bandwidth with out the introduction of dips in the response. While dips result in greater bandwidth for a given gain, the transient response is less satisfactory.

In addition, the response curve of a staggered pair does not depend on the gain of individual elements, but only on the shape of the response curves of the individual elements. The figure of merit of the two stages may also be different. As long as the Q's, of the coupling circuits involved are the same, and the separation is correct, the response will be correct. This independence of stage gain with respect to over all response means that gain control of any given number of stages in a staggered amplifier has no effect on the response curve.

While the original description of the staggered pair and triple in Wallman's (9) report assumes the high Q case, Dr. Goldberg, (4), has shown the formulas will hold with sufficient rigor for the low Q case. In the low Q case however, a staggered pair, for instance, is not made up of two circuits of equal bandwidths. If equal bandwidths are used, the response will not have the desired shape. The two circuits, in the exact case must have equal Q's, which is stating that their bandwidths must be proportional to their center frequencies. The center frequency of the exact staggered pair will then be the geometric mean of the two resonant frequencies instead of the arithmetic mean. The following design formulas were derived by Dr. H. Goldberg from results in Wallman's report. These cover both the low Q and high Q cases and while not exact, are sufficiently close for design purposes.

<u>Staggered Pair</u>***

Center frequency 'fo', band width of pair

Two stages staggered at $f_0\left(1+\frac{0.35\Delta f}{f_0}\right)$ and $\frac{f_0}{\left(1+\frac{0.35\Delta f}{f_0}\right)}$

If the two frequencies are called f_1 and f_2 respectively, the bandwidths required at f_1 and f_2 are:

$$\Delta f_1 = 0.71\frac{f_1}{f_0}\Delta f \qquad \Delta f_2 = 0.71\frac{f_2}{f_0}\Delta f$$

The required effective shunt loading resistance is given by

*** Bendix Radio Report No. 75-827-7; H.Goldberg; p.14

$$R_1 = \frac{1}{2\pi C \Delta f_1} \qquad R_2 = \frac{1}{2\pi C \Delta f_2}$$

Staggered Triple

This triple is made up of two stages staggered at $f_0(1+0.43\frac{\Delta f}{f_0})$ and $\frac{f_0}{(1+\frac{0.43\Delta f}{f_0})}$ and one stage centered at f_0 .

For the staggered triple:

$$\Delta f_1 = 0.5\frac{f_1}{f_0}\Delta f \qquad \Delta f_0 = \Delta f \qquad \Delta f_2 = 0.5\frac{f_2}{f_0}\Delta f$$

$$R_1 = \frac{1}{2\pi C \Delta f_1} \qquad R_0 = \frac{1}{2\pi C \Delta f_0} \qquad R_2 = \frac{1}{2\pi C \Delta f_2}$$

Staggered Quadruple

Two stages are staggered at:

$$f_0\left(1+0.46\frac{\Delta f}{f_0}\right) \quad \text{and} \quad \frac{f_0}{(1+0.46\frac{\Delta f}{f_0})}$$

$$\Delta f_1 = 0.38\frac{f_1}{f_0}\Delta f \qquad \Delta f_2 = 0.38\frac{f_2}{f_1}\Delta f$$

$$R_1 = \frac{1}{2\pi C \Delta f_1} \qquad R_2 = \frac{1}{2\pi C \Delta f_2}$$

The remaining two stages are staggered at:

$$f_0\left(1+0.19\frac{\Delta f}{f_0}\right) \quad \text{and} \quad \frac{f_0}{(1+0.19\frac{\Delta f}{f_0})}$$

$$\Delta f_1' = 0.92\frac{f_1'}{f_0}\Delta f \qquad \Delta f_2' = 0.92\frac{f_2'}{f_0}\Delta f$$

$$R_1' = \frac{1}{2\pi C \Delta f_1'} \qquad R_2' = \frac{1}{2\pi C \Delta f_2'}$$

The bandwidth required for staggered pairs or triples and quadruples, in cascade, to give a desired over all bandwidth is given in the following table The factor given, is the factor by which the over all bandwidth must be multiplied to give the bandwidth of the component m-uple.

Staggered Pairs

No. of Pairs	1	2	3	4	5	6
Factor	1	1.246	1.400	1.516	1.600	1.690

Staggered Triples

No. of Triples	1	2	3	4	5
Factor	1	1.158	1.250	1.322	1.373

Staggered Quadruples

No.of Quadruples	1	2	3
Factor	1	1.117	1.182

Table (1) Multiplying factor to give the bandwidth of the m-uple.

Dr. Goldberg additionally points out that the advantages of the double-tuned transformer over the synchronous single-tuned is $1.55n^{1/4}$ for equal Q, and $2.2n^{1/4}$ for the loading on one side. The advantage over the cascaded staggered pair amplifier is independent of the number of stages and is 1.19 for the equal Q case and 1.69 for the loading on one side only.

When compared to staggered triples the advantage or disadvantage, depends on the number of stages. For the equal Q case, the ratio of the factor for the staggered triple system to the factor for the double-tuned system

is 0.875 $n^{1/2}$ Equal Q, 0.617$n^{1/2}$ loading one side only. For small values of n the advantage is with the double-tuned system.

In the above cases, it has been assumed that no additional shunt 'C' was added to the circuit in addition to the unavoidable shunt 'C'. This is possible in the single-tuned case; however, there arises practical difficulties in the way of doing this in the double-tuned case. Tuning may be accomplished by varying the inductance in the single-tuned case. In the double-tuned case, the inductance cannot be varied without simultaneously varying the coupling. In practical circuits, therefore, the double-tuned circuit must be capacity tuned. The addition of a tuning capacity will generally add the minimum C of the trimmer plus one half of the trimmer range to both the primary and secondary. In many cases this additional 'C' may easily wipe out the advantage of the double-tuned circuit by decreasing the effective figure of merit.

CHAPTER II

I.F. AMPLIFIER DESIGN AND ALIGNMENT

An I.F. amplifier employing one staggered pair and one staggered triple, with a center frequency of 30mc and an over all bandwidth of 5mc was designed, built and tested by the author. The band-pass characteristics of the amplifier, were designed from the expressions listed in Chapter I. The recently developed, direct coupling technique, described in Chapter III, was used in the construction of this amplifier. Since this amplifier was built primarily to study the direct coupling characteristics, other space saving and component saving expedients were not used in this amplifier.

A schematic diagram, showing the component values, necessary center frequencies and bandwidths, is shown in fig. (2). The bandwidth shrinkage factor caused by cascading one staggered pair and one staggered triple was found by solving for the geometric mean, between the factors given in table (1) for 2 staggered pairs and 2 staggered triples.

The following frequencies and bandwidths were used; center frequency 30 mc; over all bandwidth 5 mc.

for 2 pairs, correction required 1.246

for 2 triple, correction required 1.158

geometric mean $\sqrt{(1.246)(1.158)} = 1.20$

$$\Delta f \doteq 1.2(5) = 6.0\ mc, \quad f_0 = 30\ mc$$

Using the equations given on page 9, $f_o = 30$mc, and $\Delta f' = 6.0$mc, the following center frequencies and bandwidths, were assigned to the five stages.

Staggered Pair

$f_2 = 28.02$ mc $\quad \Delta f_2 = 3.98$ mc $\quad R_2 = 3{,}810\,\Omega$

$f_1 = 32.10$ mc $\quad \Delta f_1 = 4.55$ mc $\quad R_1 = 3{,}340\,\Omega$

Staggered Triple

$f_1 = 32.6$ mc $\quad \Delta f_1 = 3.26$ mc $\quad R_1 = 4{,}650\,\Omega$

$f_0 = 30.0$ mc $\quad \Delta f_0 = 3.0$ mc $\quad R_0 = 5{,}050\,\Omega$

$f_2 = 27.6$ mc $\quad \Delta f_2 = 2.76$ mc $\quad R_2 = 5{,}480\,\Omega$

A stage capacity of 10.5 uuf, which included the input and output capacity of a 6AK5, was used in calculating the load resistors necessary to provide the required bandwidth. While the design formula gives the loading resistors presumably necessary, it was found that the computed values did not give the desired results.

A correction must now be made to determine the ohmic resistance that should be placed across the coil such that the combined effect of this resistor, transit time loading, tube loading and coil losses give the calculated value. In addition, this discrepancy is further influenced by the

lack of definite information about the value of circuit 'C', which appears in the formula $R=\frac{1}{2\pi C\Delta f}$. As a result of this, the required loading is determined empirically, using the computed values as a guide.

(I) <u>Stage loading for staggered pair</u>

The adjustments involved, in a staggered pair consists of alternate suppression and adjustment of individual bandwidths to approximately correct values. The over all response of the pair is then examined. If the response is symmetrical, but double humped, and the center frequencies are of the proper value, the Q's are too high. The most straight forward method and in most cases, the quickest method, is to tune the two stages until the proper transitional response is observed. If the 3db bandwidth which results, is too narrow, the Q's are too high. The stages of the pair may now be loaded a bit more and again adjusted for transitional response. This technique is continued until the proper bandwidth is attained. Should there be other pairs in the amplifier, it should not be assumed that the loading for all other pairs are now determined. Differences in decoupling, mechanical layout, etc., often results in slight differences in loading.

(2) <u>Stage loading for staggered triple</u>

To determine the correct loading resistors for a staggered triple, the first, and most important step is the determination of the proper constants for the end stages. In general, experience has shown that best

results are obtained when the two end stages (the stages which are above and below the center frequency) are placed adjacent to each other. To determine these constants, the center frequency stage is loaded until its bandwidth is many times that of the triple. It is good practice to load this stage with a resistor which will make the stage gain one. This effectively removes the characteristics of the loaded stage from the amplifier. The two end stages are now adjusted to their proper resonant frequencies by peaking, using a modulated C.W. oscillator set to the proper frequencies for each stage. Having completed this adjustment, the response of the pair is examined. Experiments, carried out at Bendix Radio, shows that the process of examining each stage individually, and attempting to adjust the individual bandwidths is not trustworthy. Theoretically, the two end stages should give a response, when properly adjusted and resonated, which is double humped, with equal hump amplitudes, and a 13% dip at the center. The hump frequencies will be closer to the center frequency than the center frequency of either pair.

Cut and try methods are now used until the two end stages exhibit the proper shape. If the humps are not of equal heights, the higher one may be brought down by additional loading. After each readjustment of height, the stages must be retuned by the C.W. peaking method. The dip must be measured by using the attenuator on the signal generator, maintaining fixed output at the hump and center.

This dip value is critical and has a very great influence on the final performance of the triple. A fixed output is necessary to insure that the measurements are made on the same portion of the detector's response curve.

The next step is to add the middle frequency stage and retune the entire triple by peaking. The over all response should now be examined by sweeping through the frequency range and observing the response on an oscilloscope. Should the response appear unsymmetrical (Geometrically), and if the fault is not with the sweep oscillator (amplitude modulation in the sweep oscillator), the triple is regenerative. This means that the addition of the center stage has changed the loading of the end stages asymmetrically by regeneration.

If the response is symmetrical, the next step is to adjust the loading of the center stage. If the center stage is too broad, the triple will show a dip, and if too narrow, the response will be peaked and rounded. The proper loading is that which gives a flat top without a dip. If the above has been properly carried out, the triple may be aligned by the peaking process and it will be found that the bandwidth is very close to the calculated value.

In all cases proper response was obtained by the use of 5% carbon resistors. The circuit was loaded to the nearest RMA value of resistance. The advantage gained by using a combination of resistors to obtain an intermediate RMA value did not warrent the loss caused by the capacity increase.

A proper design, determined by the above process, is necessary in order to obtain all of the advantages characteristic of stagger-tuned amplifiers. Among these are proper response shape and ease of alignment. Once the design is correct, it is possible to align a stagger-tuned amplifier by peaking the different stages at their proper frequencies. This does not require the use of a sweep oscillator; however, the response of a correct design, checked with a swept oscillator will show the proper curve. Alignment by peaking individual stages is only possible if the final design is arrived at by the previously mentioned formulas. In an amplifier having a similar response arrived at by cut and try, juggling loading and stage frequencies, it will not be possible to align the amplifier without a swept oscillator. If it is necessary to juggle resonant frequencies and loading to attain the desired response curve, it implies improper design at least, and perhaps spurious response and regeneration.

CHAPTER III

CHARACTERISTICS OF DIRECT COUPLED H.F. AMPLIFIER

The decoupling and feedback problems introduced by using higher I.F. frequencies, soon over come the advantages gained in size and weight obtained by using such frequencies.

A search, made by Bendix Radio, for the amplifier circuit requiring the minimum number of components and also using the least power, resulted in the development of a direct coupled H.F. amplifier. This amplifier, invented by Mr. Edw. Crosby Jr., (2), is characterized by connecting the vacuum tubes in series with the 'B+' supply. This series connection, results in a very sizable reduction in power supply requirements, since the current drain is that drawn by a single amplifier tube. In addition to the saving in power, this technique permits the elimination of numerous coupling and decoupling components.

Consider, as an example, the I.F. amplifier design fig. (2), which was constructed and tested at the Bendix Radio Plant. The power requirements of this amplifier connected in cascade is 9 watts, 50 ma. plate and screen current and a 'B+' voltage of 180 volts. By connecting the vacuum tubes in series, the required current becomes 10 ma., and the 'B+' supply necessary is 325 volts, which represents 3.25 watts of power. In this amplifier the plate and screen are operated at the same D.C. voltage, which is 65 volts.

The measured Gm. of a 6AK5 operated with .5 volt bias, and with 65 volts applied to the plate and screen is 4100 10^{-6} mhos. A comparison, of this loss in Gm. from the 5000 10^{-6} mhos experienced with convential operation, appears to be a good bargain for the 6 watts of power saved and for the reduction in circuit components.

In order to present a clearer picture of the saving in components, a schematic diagram of a 4 stage, staggered-tuned direct coupled amplifier, fig. (3), is compared with a cascade amplifier of similar characteristics, fig. (4).

Tests made at Bendix Radio plant, shows that the series connection of heater elements, eliminates the need for the heater decoupling circuits between stages. An I.F. amplifier employing series connected heaters was operated satisfactory at 60 mc., without decoupling tuned circuits and without by-pass condensers in the heater circuit, fig. (5).

The amplifier built and tested by the author, did not use series connected heaters primarily because of the absence of a 31.5 volt heater supply.

The maximum number of tubes that can be operated in a series chain, is limited by insulation breakdown between cathode and heater. Heater-cathode voltages as high as 450 volts were used successfully in tests on this amplifier using 6AK5's.

Tests carried out on a 3 stage direct coupled amplifier and a 3 stage cascade amplifier shows the output of the direct coupled amplifier is influenced least by replacing

the original vacuum tubes by one's with lower Gm. This results largely from the direct coupled amplifier being a constant current device.

The recovery time experienced in the series amplifier is determined by the same factors as in conventional I.F. amplifiers; namely the circuit time-constants and the tube itself. The absence of coupling time-constants in the circuit causes a recovery behavior similar to a bifilar coupled conventional amplifier except for the series power supply impedances.

The best linearity, from this series chain, is obtained when fixed bias is used on all tubes. Limiting in the highest-level stage is experienced when one of the earlier stages is biased to near cutoff, in an attempt to supply AGC. The biased stage will begin to take more than its share of the applied strip voltage. The reduced voltage now available for the highest-level stage, and the larger signal applied to that stage, act together to cause limiting.

(1) <u>Bypass capacitor used as a circuit parameter</u>

The effect of using the bypass capacitor as a circuit parameter was investigated. It was found that by reducing this capacitor from 5000 uuf. to 30 uuf., and removing the loading resistor, a stage bandwidth of 9 mc., associated with a shift in center frequency of 1.6 mc., was observed.

For comparison purposes the stagger-tuned amplifier constructed by the author, was modified by removing the loading resistors and decreasing the value of the bypass condensers to 18 uuf., for the first two stages and 27 uuf.,

for the remaining three stages. These values of capacity were roughly determined from data previously taken on this amplifier. An attempt was made, by use of the small coupling condensers, to duplicate the bandwidth and center frequency values previously used in the stagger-tuned amplifier, fig. (2). The bandwidth of the pair whose bypass capacitors were 18 uuf. was 10.7 mc. The modified amplifier, synchronously tuned gave a bandwidth of 4.1 mc., and an over all gain of 46.6 db. This same equipment when operated as a stagger-tuned amplifier, with a bandwidth of 5 mc., has a gain of 88 db. In addition to the above effect, decreasing the bypass condenser had removed the isolation between stages, and because of the feedback now existing between stages it was impossible to operate this amplifier stagger-tuned.

In order to study the effect upon frequency characteristics of the bypass condenser without interference from the tube and wiring capacitances, a single stage amplifier tuned to 1000 cycles, followed by an isolating amplifier, with a gain of one, was constructed, fig. (6). From this amplifier, it was learned that replacing the bypass condenser by a resistor of equal ohmage merely reduced the output through degeneration. As might be expected, the addition of the resistor did not shift the resonant frequency of the tuned circuit.

Fig. (7), compares the bandpass characteristics obtained from a single stage, using bypass components of .05 uf, .015 uf and 3.2K ohms. The .05 uf condenser has 3.2 k. ohms impedance at center frequency.

Fig. (8), compares the bandwidth as a function of bypass capacity, of a 6AK5, connected in a series circuit, a 6 SL 7 (triode) series connected, and a 6AK5 with screen grid separately excited for D.C., but at A.C. ground.

The first bandwidth measurements on this amplifier were made by observing the frequency corresponding to a $+45^\circ$ and -45° phase shift in the output signal. These phase measurements were made very conveniently with the aid of a Bendix Radio, phase and frequency meter. A comparison made of the bandwidth obtained by use of the phase measurement technique and that obtained by the 3 db. points indicated an error varying with the size of bypass condenser used. This discrepancy increased with smaller values of bypass capacity. This meant that the impedance of the small bypass condensers, which was included in the output load, was changing considerably over the frequency range considered. This ruled out any further use of the phase meter to determine bandwidths on this amplifier.

In general the amplifier bandpass characteristics are caused by negative feedback to the screen grid. Further this feedback voltage which is developed across the bypass condenser is out of phase with the plate voltage, by some angle depending upon the value of the bypass condenser.

This phase angle associated with the screen grid feedback voltage is responsible for shifting the center frequency of the stage, as defined by the half power points.

The gain-bandwidth product of this circuit employing

screen feedback is inferior to that attainable with resistance loading, and in general it does not appear to be an economical method of obtaining bandwidth.

The following electrical circuit representing this amplifier is shown in fig. (9).

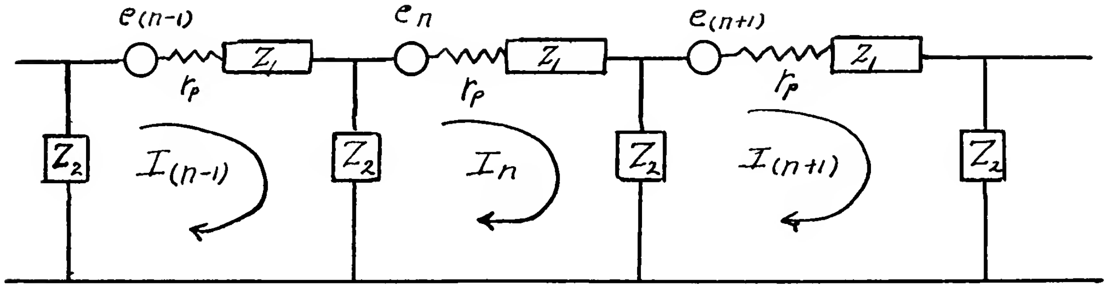


Fig (9)

Consider this amplifier as being made up of identical stages. The loop equation for the nth stage then becomes;

$$Z_2(I_n - I_{(n-1)}) + I_n(2Z_2 + Z_1 + r_p) + Z_2(I_n - I_{(n+1)}) = e_n$$

$$-I_{(n-1)} Z_2 - Z_2 I_{(n+1)} + I_n(4Z_2 + Z_1 + r_p) = e_n$$

$$-Z_2 - G^2 Z_2 + G(4Z_2 + Z_1 + r_p) = \frac{e_n}{I_{(n-1)}}$$

where G is stage gain $G^2 = \frac{I_{(n+1)}}{I_{(n-1)}}$

$$e_n = I_{(n-1)} Z_1 g_m r_p$$

$$-G^2 Z_2 + G(4Z_2 + Z_1 + r_p) = Z_2 + Z_1 g_m r_p$$

$$G^2 Z_2 - G(4Z_2 + Z_1 + r_p) + Z_2 + Z_1 g_m r_p = 0$$

$$G = 1 + \frac{r_p Z_1}{2Z_2} + \sqrt{\frac{r_p}{Z_2} + \frac{Z_1}{Z_2}(1 + g_m r_p) + \left(\frac{r_p + Z_1}{2Z_2}\right)^2}$$

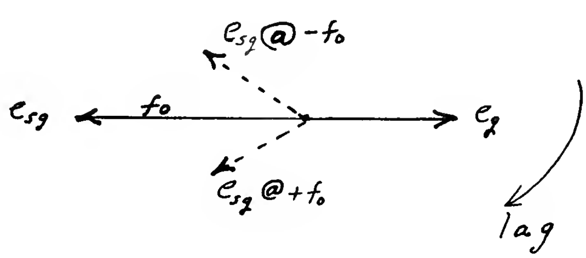
CHAPTER IV

SCREEN CONTROLLED BANDPASS AMPLIFIER

A new circuit was designed, based upon the information described in Chapter III. This circuit, Fig. (10), employs screen-grid feedback to affect broad banding, but it does not possess the faults previously encountered. It provides increase gain, over the single-tune circuit loaded to provide the same bandwidth. This circuit was first built and tested at 1000 cycles, then a two stage amplifier with a center frequency of 15 mc., was constructed.

Both at 1000 cycles and at 15 mc., the circuit shown, Fig. (10), gave a 10 db. per stage increase in gain over a single-tuned stage loaded to have the same bandwidth characteristics. In addition, this feedback circuit provides good isolation between stages, so that the individual stages may be stagger-tuned in an R.F. or I.F. amplifier.

There is no mutual inductance between the coils '1 ' and '2' of Fig. (10). With both coils tuned to the same frequency, at resonance the screen feedback voltage will be a maximum and will be 180° out of phase with the grid voltage. The following vector relation exists between the grid and screen-grid voltages.



By tuning coils '1' and '2' to different frequencies, the characteristic, critical and over coupled response of double tuned circuits may be obtained. The magnitude of the dip in such a response is proportional to the Q's and to the difference in the center frequencies.

This same technique can be extended one step further, by adding a third coil shown Fig. (11). A triple tuned response can be produced by tuning the three coils to different frequencies.

This circuit possesses the unique advantage of using the feedback impedance both to provide feedback voltage and signal voltage to the following stage. The improved gain bandwidth product is attributed to the addition of a feedback impedance in series with the conventional plate load.

Attempts to build a multi-stage amplifier with a center frequency above 15 mc., have not produced the gain bandwidth advantages experienced at the lower frequencies. Above 15 mc., part of the isolation between stages is lost due to grid loading, caused by Miller effect. Until some means of counteracting the Miller effect between grid and screen-grid is available, this amplifier is limited in its application to approximately 15 mc. A vacuum tube having a low, grid to screen-grid, transconductance would extend the upper frequency range of this amplifier considerably. Since the grid to screen-grid transconductance is not given in the tube manuals, it becomes necessary to experimentally check these characteristics in order to find such a tube. The

measured grid to screen-grid transconductance of a 6AK5, passing rated current is 850×10^{-6} mhos.

In addition, improved skirt selectivity may be obtained by designing the tuned circuits in such a way that the two coils become series resonant either above or below the pass band. Fig. (12) shows a response curve of a single stage which had been adjusted for series resonance below the pass band. By using two stages a response shown in Fig. (13) can be obtained. This type of skirt selectivity is customarily used in television picture I.F. amplifiers. The sharp attenuation in a picture I.F. response curve corresponding to the adjacent channel sound trap, and the 21.25 mc., sound trap are obtained by inserting frequency traps, shown Fig. (14), into the circuit. This trap appears as a high impedance off resonance and as a short circuit at resonance. By this technique of producing sharp skirt selectivity, the tuned circuit (trap) acts as a parasitic element in the vicinity of resonance, but can not contribute to the signal over the pass band.

The circuit, Fig. (10), produces the same skirt selectivity, and in addition, an improvement in gain is realized from the signal voltage across coils '2' and '4'.

The application of screen-grid feedback technique to improve the gain bandwidth factor of video amplifiers, appears very promising. Such a circuit is shown in Fig.(15). Since this amplifier does not encounter frequencies above 15 mc., the major problems resulting from Miller effect loading are absent.

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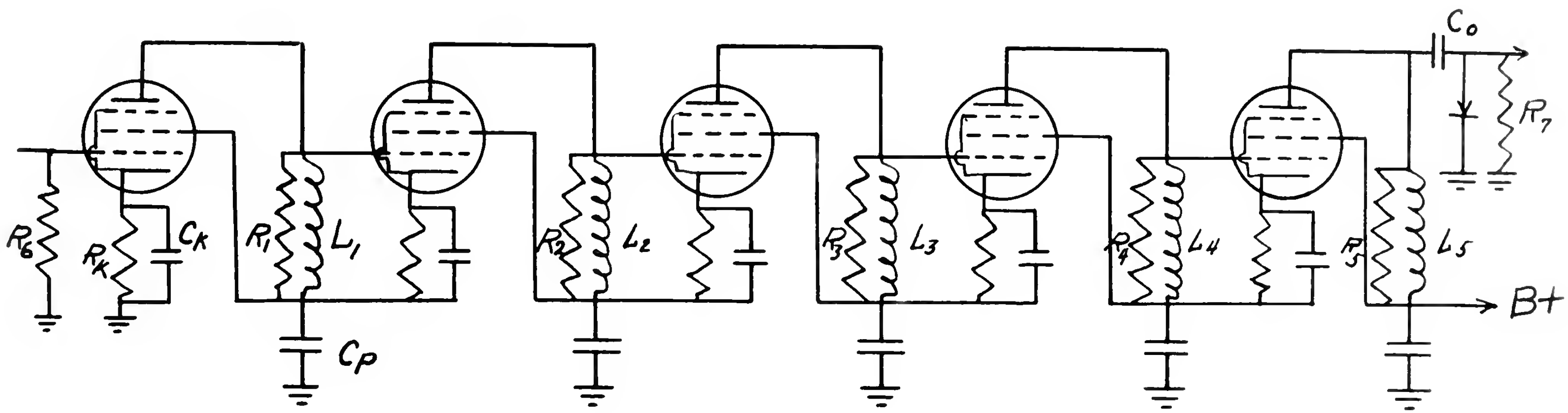


$f_1 = 28.02$ mc. $f_2 = 32.1$ mc.
Pair

$f_3 = 27.6$ mc. $f_4 = 32.6$ mc. $f_5 = 30$ mc
Triple

R_1	3000 Ω
R_2	3300 Ω
R_3	3300 Ω
R_4	4300 Ω
R_5	1800 Ω
R_6	47 Ω
R_7	200K Ω
R_K	56 Ω
C_K	5K μμf
C_P	5K μμf
C_o	50 μμf

Fig. (2) SCHEMATIC, FIVE STAGE DIRECT COUPLED AMPLIFIER

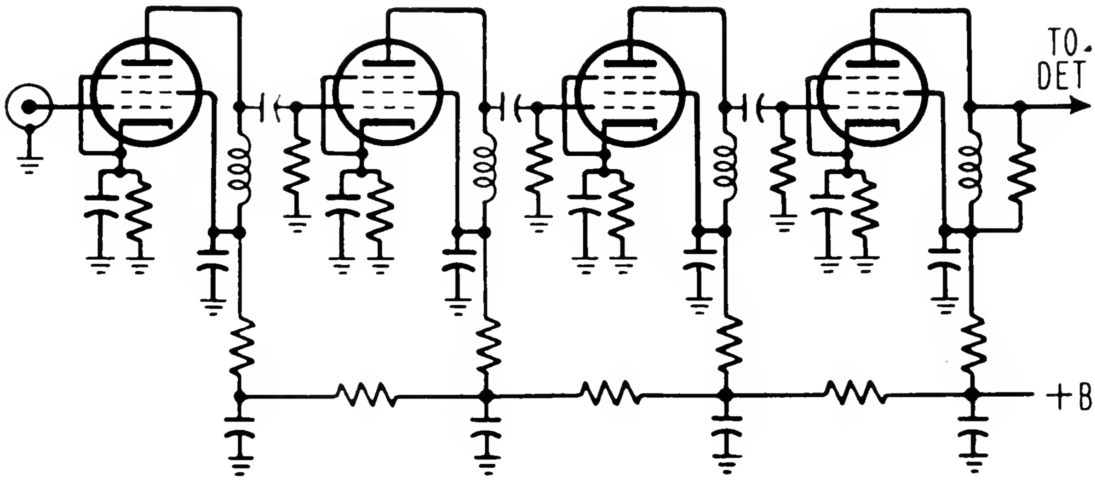


Fig.(4) SCHEMATIC,FOUR STAGE CONVENTIONAL BROADBAND I.F. AMPLIFIER

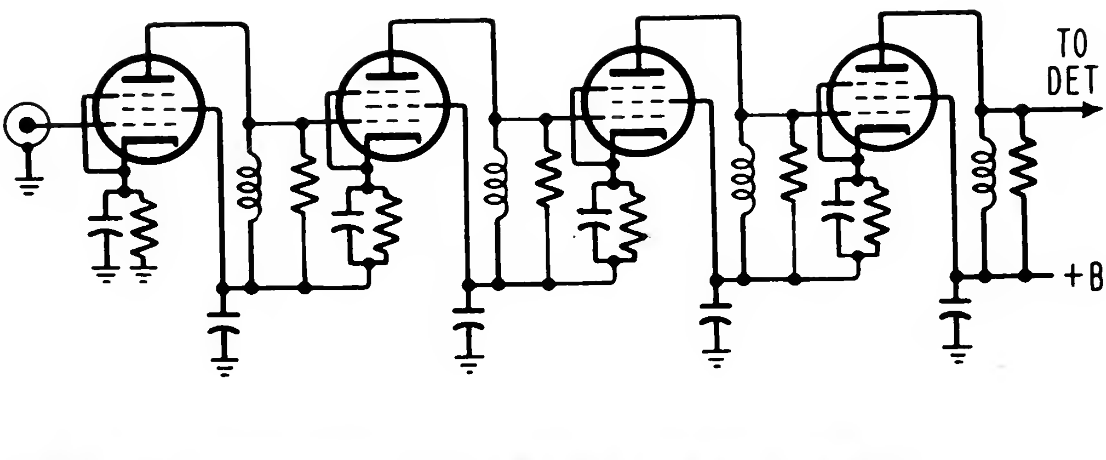


Fig.(3) SCHEMATIC,FOUR STAGE DIRECT COUPLED BROADBAND I.F. AMPLIFIER

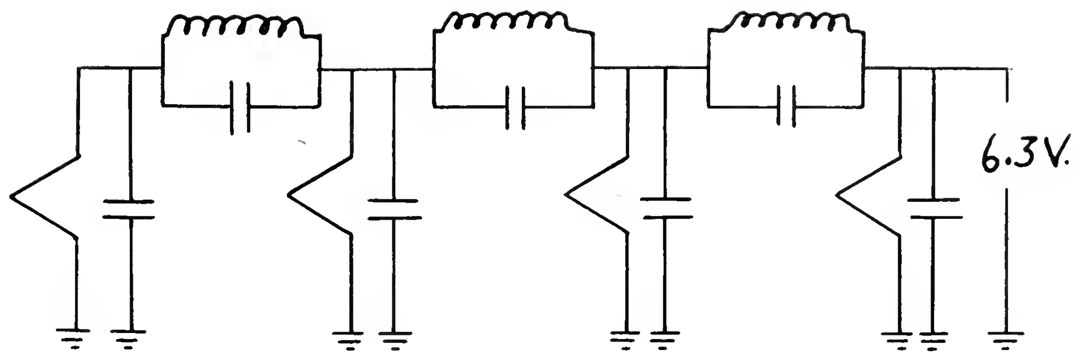


Fig.(5)A CONVENTIONAL HEATER DECOUPLING CHAIN

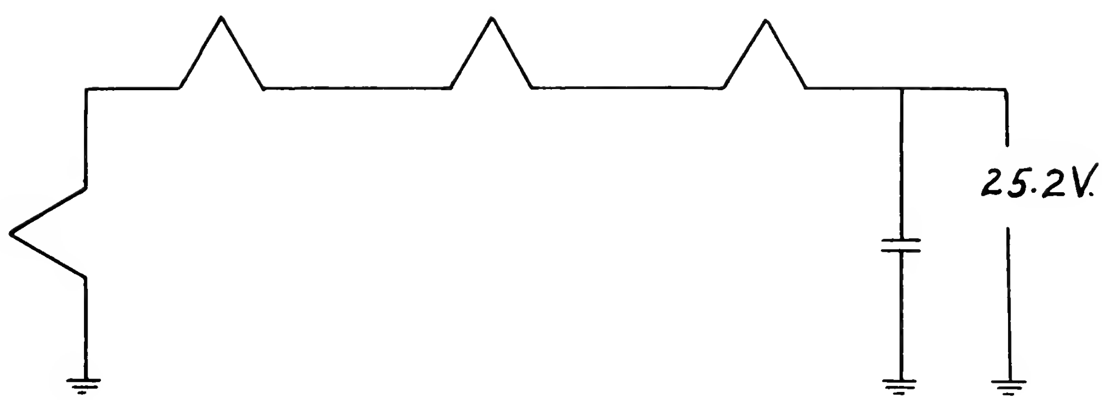
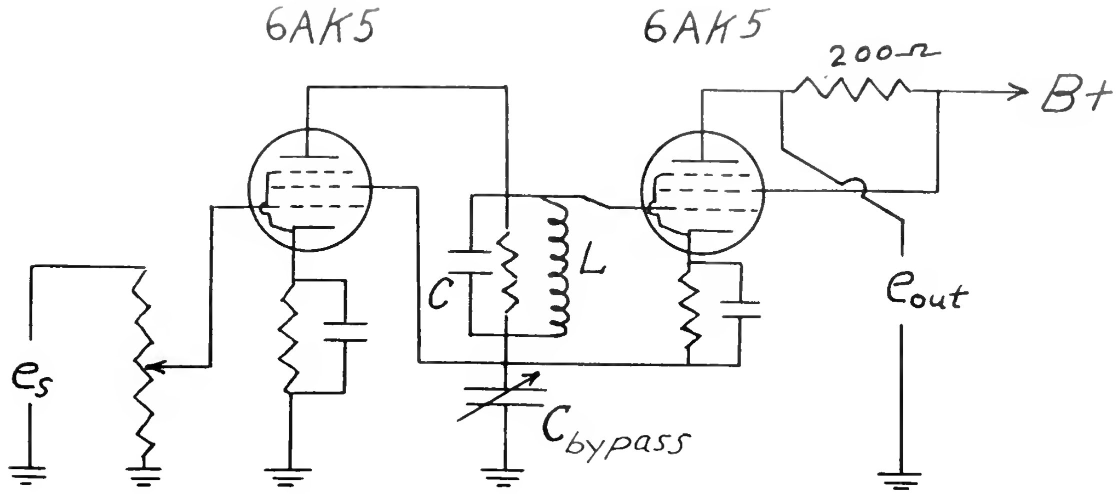
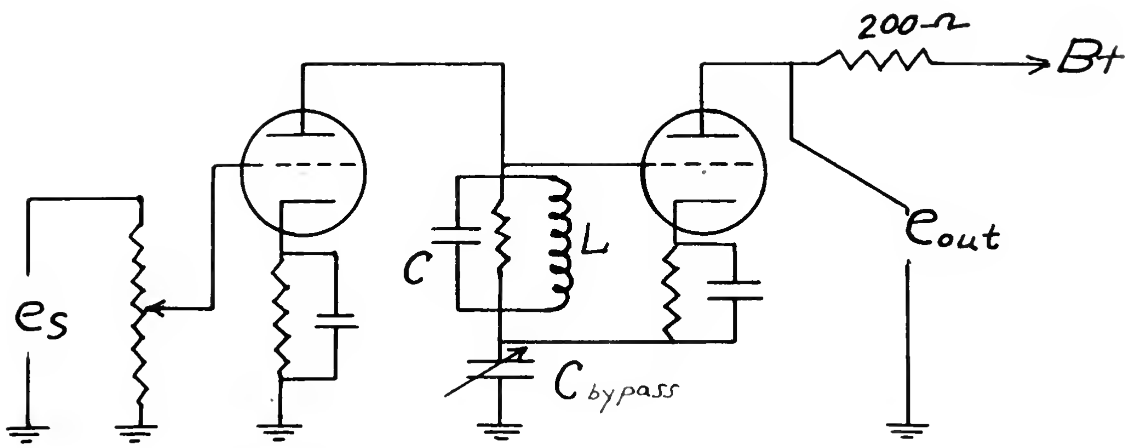


Fig.(5)B SERIES CONNECTED HEATERS



$L = 80$ mh.
$C = .35\ \mu f.$
$Q = 58$
$f_0 = 946$ c.p.s.

Fig.(6)A LOW FREQUENCY DIRECT COUPLED AMPLIFIER



$L = 80$ mh.
$C = .35\ \mu f.$
$Q = 58$
$f_0 = 946$ c.p.s.

Fig.(6)B TRIODE USED IN FEEDBACK AMPLIFIER

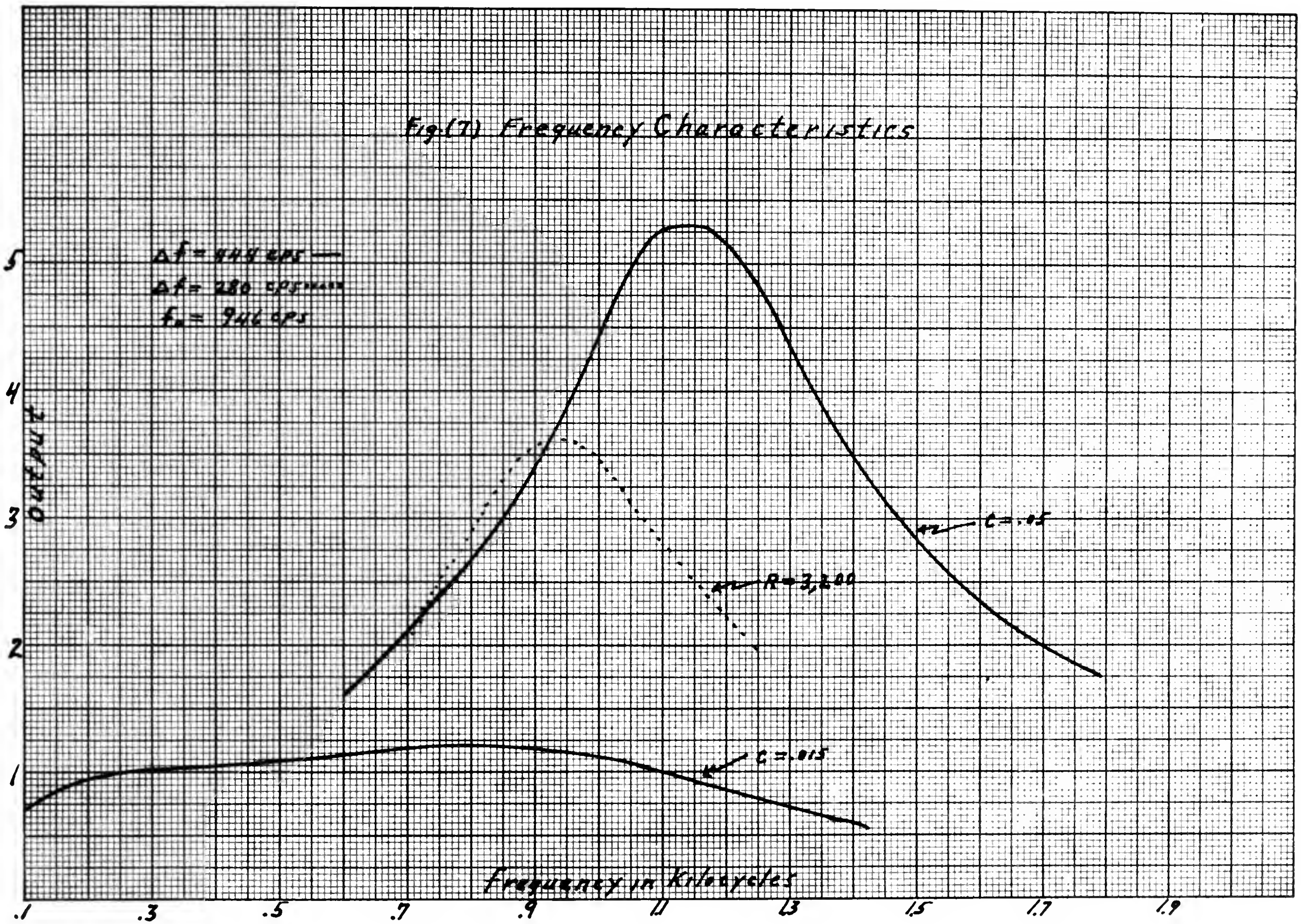


Fig.(7) Frequency Characteristics

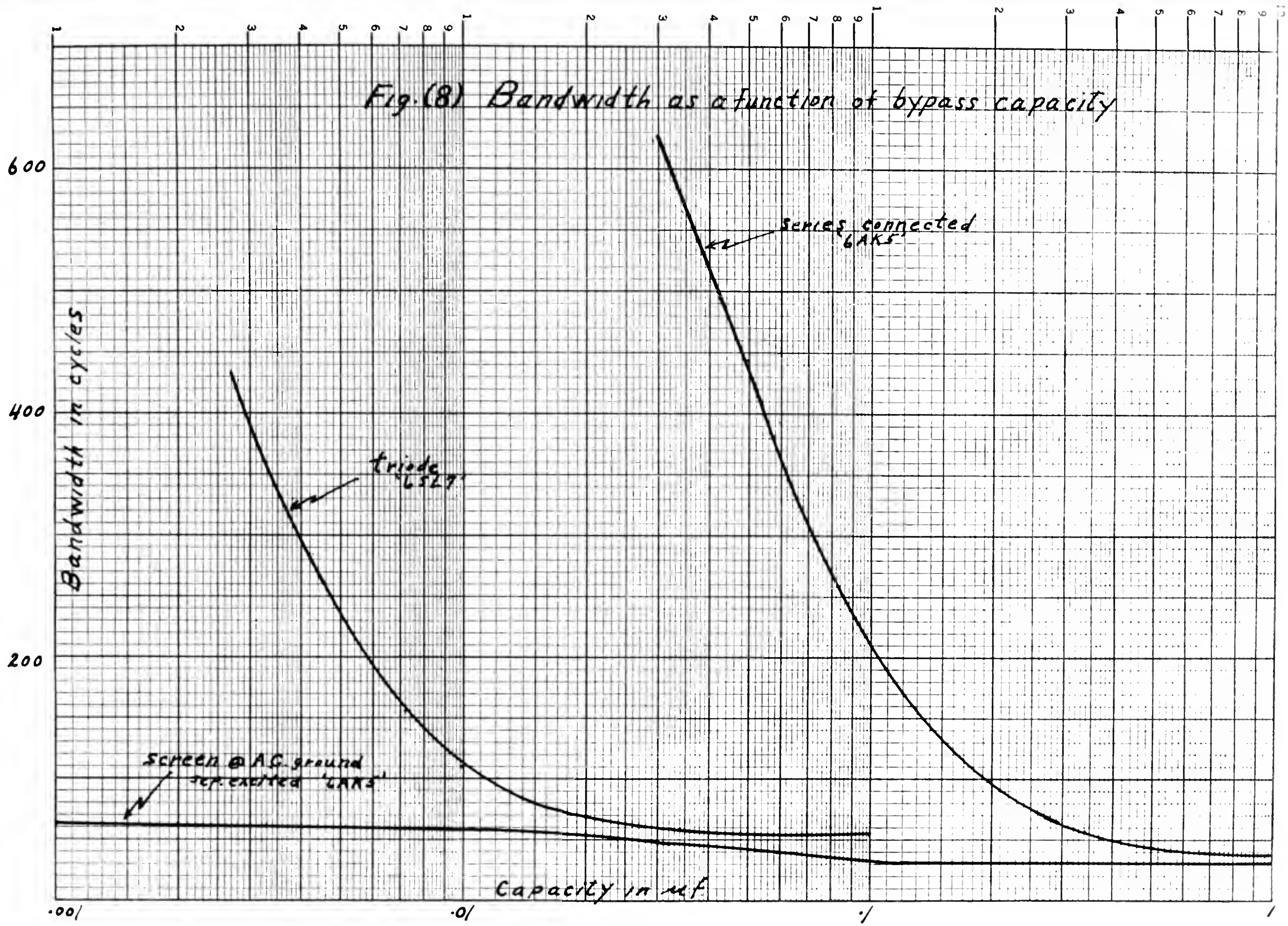


Fig.(8) Bandwidth as a function of bypass capacity
Series connected 6AK5
triode 6SL7
Screen @ A.C. ground sep. excited 6AK5
Bandwidth in cycles
600
400
200
Capacity in μf
.001
.01
.1
1

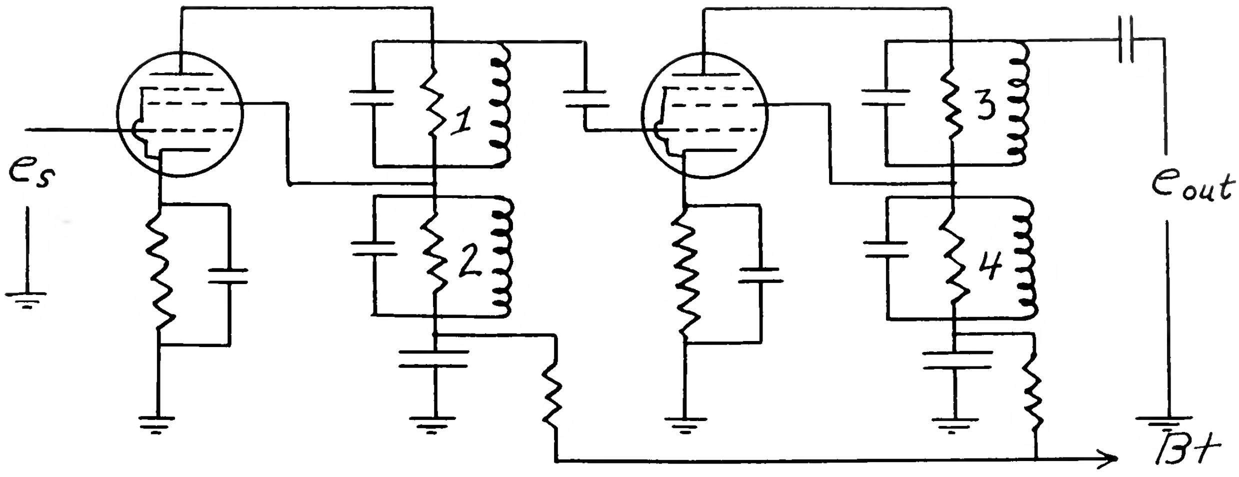


Fig.(10) TWO STAGE, SCREEN CONTROLLED BROADBAND AMPLIFIER

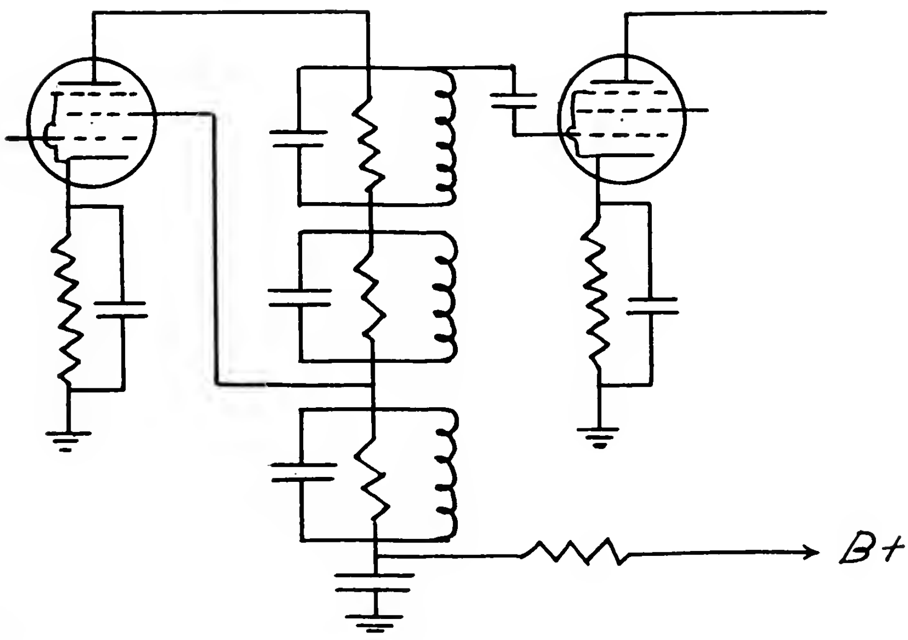


Fig.(11) TRIPLE-TUNED BAND PASS AMPLIFIER

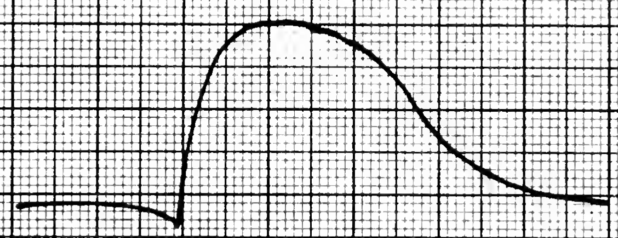


Fig.(12) SINGLE STAGE RESPONSE,COILS '1'AND'2' ARE SERIES RESONANT AT A FREQUENCY LOWER THAN THE PASS BAND

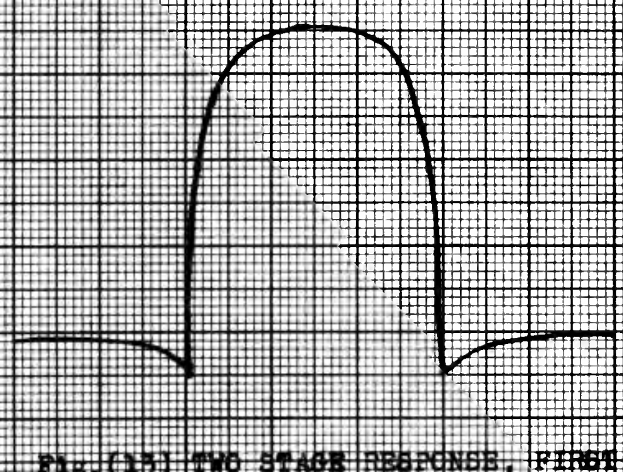


Fig.(13) TWO STAGE RESPONSE, FIRST STAGE IS SERIES RESONANT BELOW PASS BAND SECOND STAGE IS SERIES RESONANT ABOVE THE PASS BAND

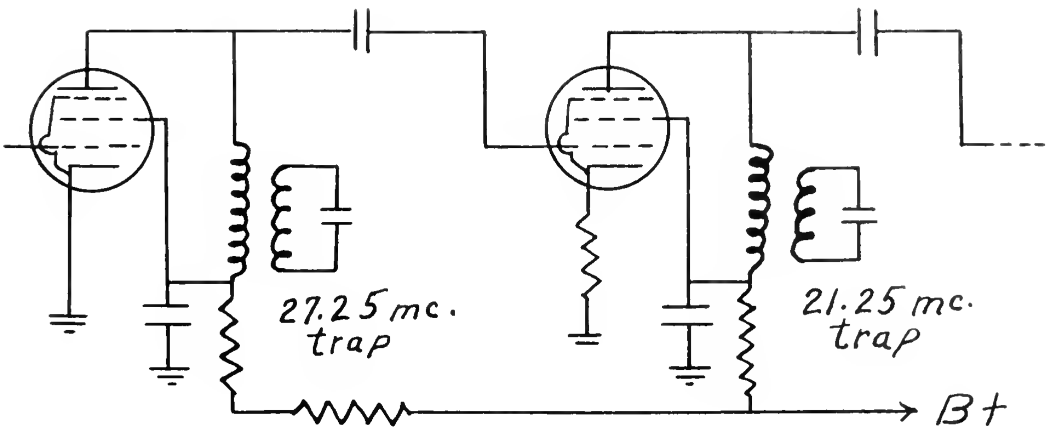


Fig.(14) TELEVISION PICTURE I.F. AMPLIFIER

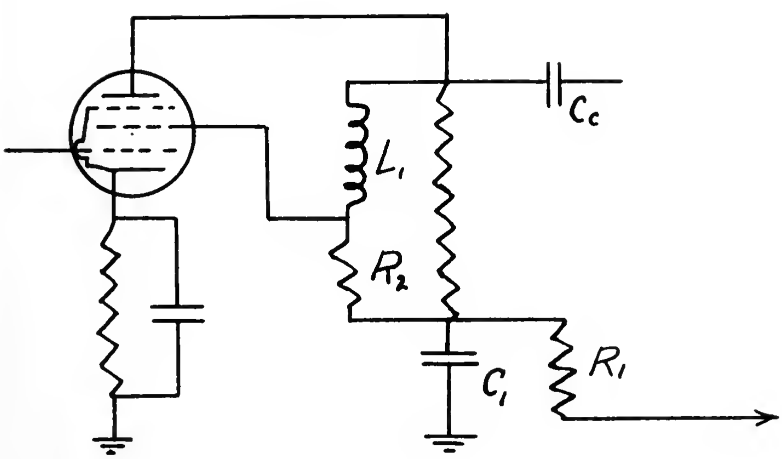


Fig.(15) SCREEN CONTROLLED VIDEO AMPLIFIER

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