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Burtnyk, N.

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I. F. AMPLIFIERS AND GAIN CONTROL FOR A MATCHED TWIN-CHANNEL RECEIVER

N. BURTNYK

OTTAWA

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ABSTRACT

Development of I.F. amplifiers and a gain control for a twin-channel receiver is discussed with particular reference to a specific design in AN/GRD-501 HF/DF equipment. Final choice of circuit design was rigidly governed by desired performance specifications as well as ease of maintenance.

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I.F. AMPLIFIERS AND GAIN CONTROL FOR A MATCHED TWIN-CHANNEL RECEIVER

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INTRODUCTION

In a conventional twin-channel direction-finding receiver, the signals fed into the two channels are displayed on a cathode-ray tube to produce a bearing trace indicating the direction of arrival of the wave. The sharpness and direction of this indicated bearing depend directly on the phase and gain characteristics of the amplifiers in the two channels. Gain mismatch will result in a direct bearing error, while phase mismatch will produce a blurring effect by opening the trace into an ellipse. These two characteristics may be associated with each other, as in the tracking of two tuned circuits where phase mismatch is accompanied by gain mismatch, or they may be independent of each other, as in the variation of tube transconductance where gain mismatch is not necessarily accompanied by phase mismatch. This requirement of maintaining phase and gain balance between the two channels across their pass bands poses the major problem in the design of I.F. amplifiers for a twin-channel direction-finding receiver. Furthermore, this requirement must be met over the specified dynamic range of the receiver for both manual and automatic control of receiver gain, since a direct error results from any gain mismatch in the system. Other requirements, such as bandwidth and image rejection, however, must also be considered seriously in the choice of the final design.

SPECIFICATIONS GOVERNING DESIGN

The initial specifications which were set as a target for this receiver and directly affected the design of the I.F. system are the following:

1) 3 to 30 mc/s operating frequency range,
2) variable bandwidth selection from 1 kc/s to 9 kc/s,
3) input signal dynamic range of 90 db,
4) 80-db spurious response rejection, including images,
5) remote control operation up to 6000 feet between aerial site and operational hut,
6) CRT bearing display at output,
7) maximum instrumental error of $\pm \frac{1}{2}°$.

It was immediately apparent that a low intermediate frequency was desirable to make phase mismatch and attenuation in the remote coaxial lines negligible, as well as to simplify the narrow bandwidth requirement. The image rejection requirement, on the other hand, indicated the need of a relatively high intermediate frequency. Dual conversion was accepted as a solution, and intermediate frequencies of 2 mc/s and 175 kc/s were chosen. The output from the second intermediate frequency amplifiers thus became very suitable for direct display on a cathode-ray tube using electrostatic deflection. Some consideration was given to the use of magnetic deflection, but it was rejected because of the very low frequency required and the resulting bandwidth restrictions.

It should be noted that in this report first and second intermediate frequencies are referred to with respect to location rather than frequency, as a convenience.

Coaxial line losses necessitate that both stages of conversion be located at the remote end. Hence the remote or first I.F. amplifier should consist of a 2-mc/s stage, a second converter or mixer, and some impedance changer, such as a cathode follower, to match the remote coaxial lines. The local or second I.F. amplifier should contain the bandwidth control device and the 175 kc/s amplifier stages necessary to provide sufficient gain to drive the cathode-ray display.

METHODS OF COUPLING IN I.F. AMPLIFIERS

The types of interstage coupling methods that were considered are:

i) synchronous single-tuned,
ii) double-tuned,
iii) stagger-tuned.

Each of these types has its advantages and disadvantages with respect to:

i) efficiency,
ii) simplicity of construction,
iii) gain stability,
iv) criticalness of adjustment.

Generally, synchronous single-tuned circuits may be considered of maximum simplicity and minimum efficiency, while the others, though of considerably greater efficiency, are also of greater complexity.

Consideration of gain and bandwidth requirements showed that five synchronous single-tuned stages, including the second mixer, would provide the desired results in the 175-kc/s portion of the receiver. High-Q critically-coupled double-tuned stages would have a marked advantage in available gain, for a given overall bandwidth, over synchronous single-tuned stages, but since constructional simplicity and non-criticalness of adjustment are of utmost importance, double-tuned circuits were considered unfavourable.

A staggered triple, on the other hand, because of its relatively high efficiency, has about the same gain for the required bandwidth as four synchronous single-tuned stages. The calculations are shown in Appendix A. This offered a considerable advantage in reducing the total number of stages, so the system was fully investigated. Calculations, as shown in Appendix B, revealed that for a 3-db bandwidth of 10 kc/s and a gain of 120 db at band center, there was an equal loss of about 1.7 db at the pass-band limit with either system for a 20% error of Q's in one channel relative to the other. Furthermore, it was revealed that the Q's required for the synchronous single-tuned circuits would be controlled almost completely by external damping, while for two of the three stages of the staggered triple the Q's required would be so high that the initial Q of a practical coil would contribute to at least half the required circuit damping. Thus the merit of synchronous-tuned circuits, compared with staggered-tuned circuits, lies in the practical realization of Q tolerances governed by performance specification requirements.

Prior to selecting a suitable interstage coupling method for the 2-mc/s amplifier, its requirement characteristics were established. Briefly, they were:

i) Gain must be sufficient to make second mixer noise negligible compared with R.F. and first mixer noise.
ii) Rejection of image frequency must be at least 80 db.
iii) Bandwidth must be sufficiently great that the overall receiver bandwidth is unaffected by this portion of the receiver.
iv) A gain control device must be included to prevent non-linearity due to excessive drive in succeeding stages.

The above gain requirement indicated that one stage of 2-mc/s amplification would be sufficient. The image rejection and bandwidth requirements, on the other hand, necessitated steep skirts on the selectivity curve. It was concluded, therefore, that two double-tuned transformers coupled below the critical point would be the favourable choice. Triple-tuned circuits, because of their criticalness of adjustment, were too difficult to control and therefore were undesirable.

CHOICE OF AGC SYSTEM

The foremost problem in the choice of an AGC system for a matched twin-channel receiver is linearity of control between the two channels. Other desired characteristics of AGC, such as rise time and decay, are similar to those in a conventional communications receiver and are of lesser immediate importance. The source of control signal was established as a bias produced by rectification of the I.F. output signal.

Various AGC systems that were proposed and investigated were:

i) damping of tuned circuits by cathode followers,
ii) variation of bias on amplifier stages,
iii) use of a staggered-frequency twin-channel AGC system,
iv) injection-level control to mixers.

The operation of the first method was such that the dynamic impedance of the cathode follower, which was controlled by bias variation on its grid, was used to damp the circuits of the synchronous single-tuned stages of the amplifiers. It soon became evident, however, that the resulting linearity was no better than that which could be attained by applying AGC bias directly to the amplifier grid circuits. Moreover, the bandwidth of the tuned circuits varied with AGC action, though originally this was of no consequence as the overall bandwidth was to be controlled by a separate bandwidth control device. A further disadvantage of over-loading in the cathode follower was revealed when the R.F. voltages across the tuned circuits were of the order of several volts. Thus direct grid bias control appeared to be a simpler method than the cathode-follower damping system requiring a smaller number of tubes and decreased power dissipation.

Various means of insuring linear control, using grid bias, were tried with no satisfactory results. Selecting tubes with the best transconductance curve match reduced the error to $\pm 1\frac{1}{2}°$, which was not acceptable. A method of equalizing direct currents in respective stages of the two channels was tried but gave no improvement. Several other methods of control, such as biased negative feedback diodes and variable magnetic-core-coupling transformers were investigated, neither producing favourable results.

A block diagram of the staggered-frequency twin-channel AGC system is shown in Fig.1 and the operation is briefly thus:

The mixers of the two channels are fed by two separate injection frequencies about 40 kc/s apart (i.e., 1805 kc/s and 1845 kc/s) to convert the 2 mc/s signal input into 155 kc/s and 195 kc/s outputs, respectively. These are fed into a common broadband amplifier centered at 175 kc/s with a 50 kc/s bandwidth, then separated by sharply tuned filters at 155 kc/s and 195 kc/s and converted to 175 kc/s by mixing each with signals from a 20 kc/s oscillator. The AGC bias is applied to the broadband amplifier common to both channels, thus producing no tracking error.

Several serious problems were encountered, however, of which the main ones were:

i) A high degree of selectivity and stability was necessary in the tuned filters to reject the unwanted frequency sufficiently and yet maintain the required pass band. Furthermore, the pass bands of these filters as well as those of other circuits in the staggered frequency portions of the system had to be matched accurately in phase and gain, a requirement that could be met much more easily in channels at the same frequency.

ii) Harmonics of the 20-kc/s signal generated by the final mixer caused spurious noise that considerably thickened the bearing trace on the cathode-ray tube display. Even when reduced to a minimum by shifting to 20.6 kc/s where the 8th and 9th harmonics were both

10 kc/s off-resonance, the result was unsatisfactory.

iii) A noticeable phase change resulted from the pull over the locking range of the staggered second local oscillators, thus producing bearing ellipses. This effect might have been overcome by more complicated circuitry, but the two methods tried were unsatisfactory.

No further work was carried out on this system as a new manual gain control method, with a direct AGC application was introduced at this stage of the development. It will be discussed in the following section in conjunction with the manual gain control systems.

MANUAL GAIN CONTROL

As in the choice of an AGC system, linear tracking was the major problem in a manual gain control system. The requirement of instantaneous control does not apply, however, so a different approach may be taken.

The scheme proposed originally consisted of a resistive attenuator operated by a multi-contact rotary switch. This introduced attenuation in 2-db steps which were matched in the two channels by selection of accurate components. This attenuator was located in the 2-mc/s amplifier to protect the second mixer from overloading; hence remote control operation was required. A servo system or rotary-solenoid-operated system was used to operate this control. The main disadvantages of such a system were lack of smooth control, the high-stability and high-accuracy resistor component requirement, and the problems of matching the low impedance attenuator to the adjoining circuits.

Development of a matched, double-tuned inductance-coupled piston attenuator was the next step in the course of the design. Initial experimental tests revealed the relative ease of attenuation law matching of two ganged attenuators within the specified error limits, so the principle was immediately adapted to the receiver. Co-planar coils on a $\frac{1}{2}$ inch diameter form in a 2 inch I.D. tube were used with an attenuation range of approximately 80 db.

Tuned coils were used in the attenuator for two reasons. First, the insertion loss was reduced to approximately zero since the coil spacing could be varied right up to critical coupling without any tracking error. Secondly, it replaced a double-tuned transformer in the 2-mc/s amplifier without any deterioration of image rejection.

Generally, the system was entirely satisfactory electrically, the main disadvantages being its bulkiness and relatively slow operation.

The final design of the manual gain control system was based on a design introduced by T.R. O'Meara [1] of the University of Illinois. It achieves linear control of mixer gain by variation of oscillator injection level, and it features circuit simplicity, and smooth single knob control. The system, as shown in Fig. 2 operates as follows.

The sensitivity control is used to vary the cathode bias on the second oscillator buffer, thus varying the injection level feeding the mixers. With both mixer grids operating on cathode resistor bias the result is that the level of R.F. voltage on each grid controls the output level linearly.

One disadvantage which arose was the decrease in overall gain of the mixer stage caused by the maximum injection level that maintains linear control. This level was about ten times lower than the injection level used on conventional mixers, so that additional gain in one of the succeeding I.F. stages was required. The mixer noise output, however, did not change; thus the contribution to overall noise by the second mixer was increased. Furthermore, this noise contribution did not decrease when sensitivity was decreased, so that the minimum noise spot of the output trace was limited by the mixer noise.

This system was equally adaptable for AGC control by feeding the AGC bias to the grid of the second oscillator buffer, and was therefore used.

BANDWIDTH CONTROL

The target specification of a variable bandwidth control covering the range 1 kc/s to 9 kc/s was recognized as an extremely difficult problem and a compromise providing a choice of three fixed bandwidths was accepted. A 9kc/s bandwidth was made available as the resultant bandwidth of the cascaded circuits of the receiver. 3 kc/s and 1 kc/s bandwidths were produced by inserting high-Q double-tuned transformers with the proper degree of coupling at the input to the second I.F. amplifier.

FINAL DESIGN

The final design of the I.F. amplifiers is shown in Figs. 3, 4, and 5. Some of the desired features are the relatively low Q and easily controllable single-tuned circuits, and the low L/C ratios which make the circuits almost independent of tube capacitance variations.

Commoning line inputs are provided at all grids as a maintenance aid. There are preset controls for approximately equalizing the gains of the two channels, but the exact balance is automatically corrected by a reference injection system in the receiver. This consists of an accurate R.F. goniometer whose field coils feed the two channels of the receiver at the input. The aerials are short-circuited and a synthesized R.F. signal is fed into the search coil of the goniometer. The search coil is rotated to a position where the resulting trace on the cathode-ray tube corresponds with the position of the trace produced by the outside signals. The search coil position is read as the indicated bearing, thus represented by the ratio of the signals in the two channels at the input to the receiver rather than at the output, and is independent of channel gain balance. The blurring due to phase mismatch is not removed, however, but rather than contributing a direct error to the observed bearing, its effect is to reduce the accuracy of reading the major axis of the ellipse compared with a straight line trace.

It should be noted that matched AGC is required even when using this injection system, as the AGC bias level may change considerably when switching from the outside signal to injection signal.

The overall gain of the I.F. amplifiers from first I.F. grid to output is about 116 db with balanced output for driving a CRT display.

The manual gain control system and the AGC operate over a 50-db range. Both control-signal level and injection-signal level are limited to a maximum of 0.45 volts for linear operation of the mixer.

The bandwidth control transformers are matched to the remote line by the resistive attenuating pads, which also equalize the gains for the three bandwidths. The coils are made of universally wound Litz wire units in ferrite pot cores with working Q's of the order of 150. Q adjustments are provided to assist in alignment of the two channels for phase and gain. Proper shielding between these transformers is of utmost importance, as the high impedances involved make them very susceptible to cross-coupling.

The final design of the 2-mc/s amplifier consists of a double-tuned stage fed by the single-tuned circuit in the plate of the first mixer. This meets the 80-db image rejection requirement with circuit Q's of about 50 and still results in a bandwidth which is large compared with the overall bandwidth of the receiver.

APPENDIX A

GAIN COMPARISON OF A STAGGER-TUNED AMPLIFIER AND A SYNCHRONOUS SINGLE-TUNED AMPLIFIER

Because of the relatively high degree of stability required, the amplifier stages were to be operated with cathode resistor feedback giving working transconductances of about 2000 micromhos. The capacitance across the tuned circuits was fixed at 300 pf, to make the effect of tube capacitance variation practically negligible. This resulted in a gain bandwidth product per stage of

$$\frac{gm}{2\pi C} = \frac{2000\times10^{-6}}{2\pi\times300\times10^{-12}} = 1.06 \text{ mc/s}.$$

For a required overall bandwidth of 10 kc/s, the ratio

$$\frac{\beta}{G\beta} = \frac{0.010}{1.06} = 0.0094.$$

Using Wightman's [2] graphical method for comparing synchronous-tuned circuits and stagger-tuned N-uples, it was concluded that a staggered triple or four cascaded, synchronous, single-tuned circuits give about equal gains of 120 or 130 db for a normalized bandwidth of about 0.01.

APPENDIX B

EFFECT OF Q ERROR ON OVERALL BANDWIDTH OF FOUR SYNCHRONOUS-TUNED CIRCUITS AND A STAGGER-TUNED TRIPLE

1. FOUR SYNCHRONOUS SINGLE-TUNED STAGES

Shrinkage factor [3] for four cascaded stages is 0.44. Therefore, for a given overall bandwidth of 10 kc/s, bandwidth per stage is 22.7kc/s.

For Δf = 5 kc/s, $\frac{\Delta f}{\beta} = \frac{5}{22.7} = 0.22$.

From universal selectivity curve,

$$\frac{\text{gain at } f_0 - \Delta f}{\text{gain at } f_0} = -0.82 \text{ db}.$$

For a 20% increase in Q of each circuit, β decreases to 18.2 kc/s

$\therefore \frac{\Delta f}{\beta} = \frac{5}{18.2} = 0.27$, and

$$\frac{\text{gain at } f_0 - \Delta f}{\text{gain at } f_0} = -1.26 \text{ db}.$$

Net change in gain at $f_0 - \Delta f$ when Q' = 1.2 Q, is
-1.26 + 0.82 = 0.44 db/stage.

For 4 stages this becomes -1.76 db, if all circuits are 20% high, and desired working Q of each circuit is about 7.7.

2. STAGGER-TUNED TRIPLE

a) Center circuit has $\beta = 10$ kc.

When $\Delta f = 5$ kc, $\dfrac{\text{gain at } f_0 - \Delta f}{\text{gain at } f_0} = -3$ db.

For a 20% increase in Q, $\beta = 8$ kc.

$\dfrac{\Delta f}{\beta} = 0.625$ and $\dfrac{\text{gain at } f_0 - \Delta f}{\text{gain at } f_0} = -4.15$ db.

Net change in gain = $-4.15 + 3 = -1.15$ db.

b) Side circuits centered at $f_0 \pm 4.3$, $\beta = 5$ kc

For $\Delta f_1 = 4.3$ kc, $\dfrac{\Delta f_1}{\beta} = 0.86$; $\dfrac{\text{gain at } f_0}{\text{gain at } f_0 \pm 4.3} = -6$ db.

For $\Delta f_2 = 0.7$ kc, $\dfrac{\Delta f_2}{\beta} = 0.14$; $\dfrac{\text{gain at } f_0 - \Delta f}{\text{gain at } f_0 - 4.3} = -0.36$ db.

For $\Delta f_3 = 9.3$ kc, $\dfrac{\Delta f_3}{\beta} = 1.86$; $\dfrac{\text{gain at } f_0 - \Delta f}{\text{gain at } f_0 + 4.3} = -12$ db.

For maximum loss in gain at 5 kc detuning, decrease Q of side circuits by 20%. $\therefore \beta = 6$ kc.

$$\frac{\text{gain at } f_0}{\text{gain at } f_0 \pm 4.3} = -4.9 \text{ db.}$$

$$\frac{\text{gain at } f_0 - \Delta f}{\text{gain at } f_0 - 4.3} = -0.22 \text{ db.}$$

$$\frac{\text{gain at } f_0 - \Delta f}{\text{gain at } f_0 + 4.3} = -10.5 \text{ db.}$$

If gain at f_0 is kept as the reference, and since each side circuit increases gain at f_0 by $6 - 4.9 = 1.1$ db, the relative gain at all other frequencies is reduced by 2.2 db.

Increase in gain at $f_0 - \Delta f$ by side circuit tuned to $f_0 - 4.3 = 0.36 - 0.22 = 0.14$ db.

Increase in gain at $f_0 - \Delta f$ by side circuit tuned to $f_0 + 4.3 = 12 - 10.5 = 1.5$ db.

Decrease in gain at $f_0 - \Delta f$ by center circuit is 1.15 db.

Therefore net relative gain change at $f_0 - \Delta f$ is $-2.2 + 0.14 + 1.5 - 1.15 = 1.7$ db.

Working Q of center circuit is 17.5.

Working Q of side circuits is 35.

ACKNOWLEDGMENT

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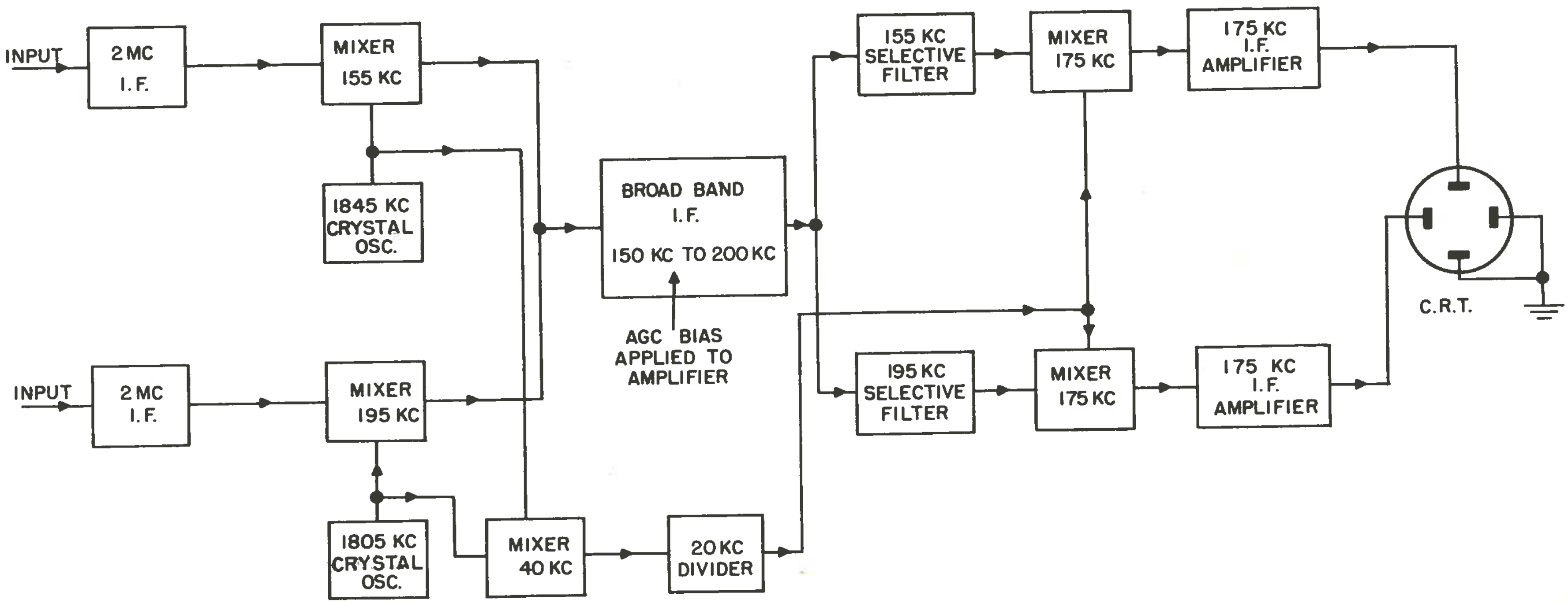

FIG. 1 BLOCK DIAGRAM OF STAGGERED-FREQUENCY TWIN-CHANNEL A.G.C. SYSTEM

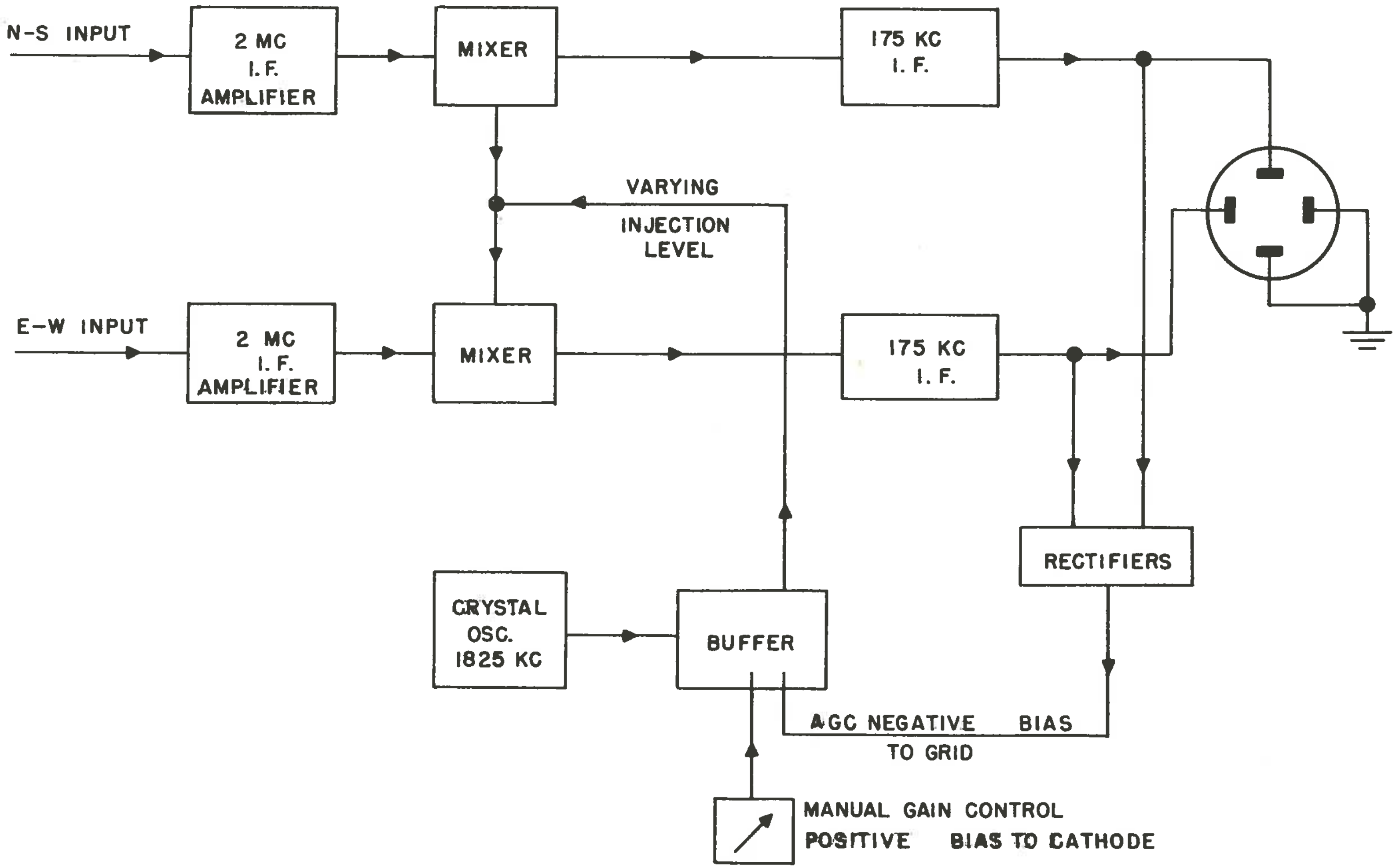

FIG. 2 BLOCK DIAGRAM OF A MATCHED TWIN-CHANNEL GAIN CONTROL SYSTEM

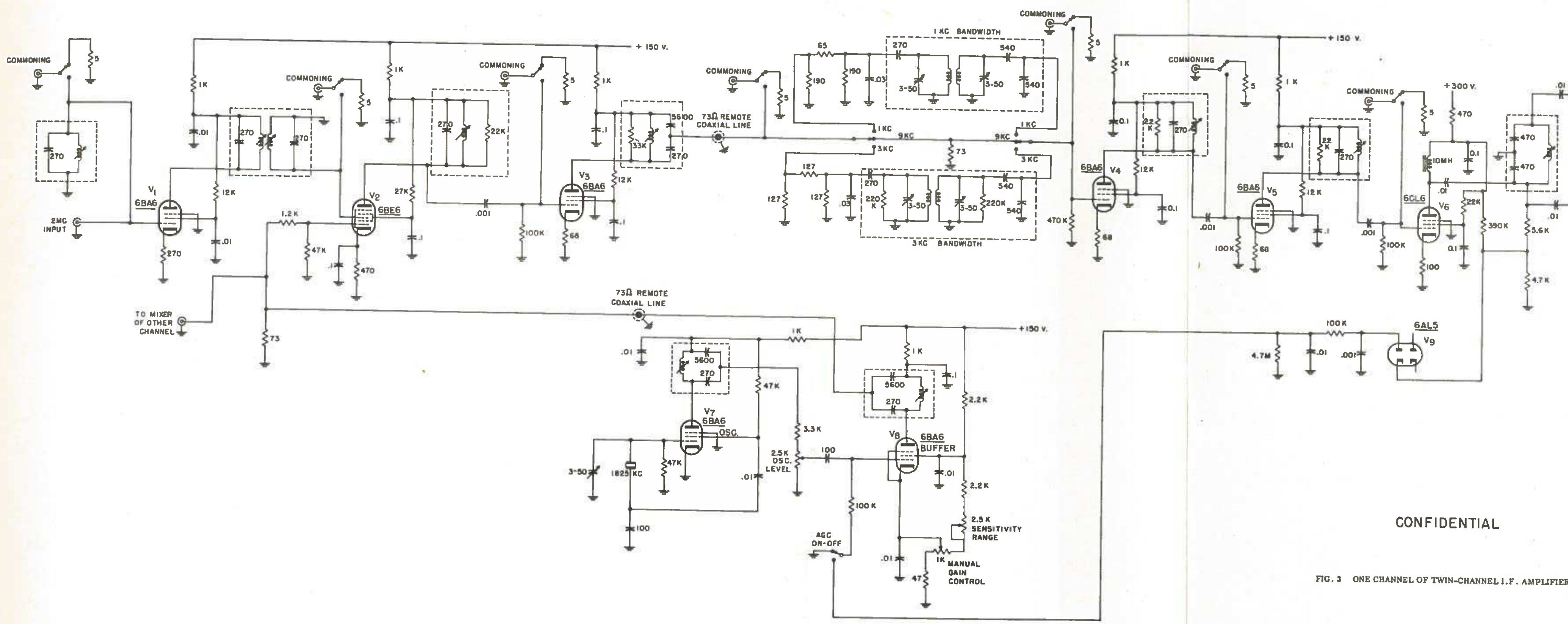

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FIG. 3 ONE CHANNEL OF TWIN-CHANNEL I.F. AMPLIFIERS

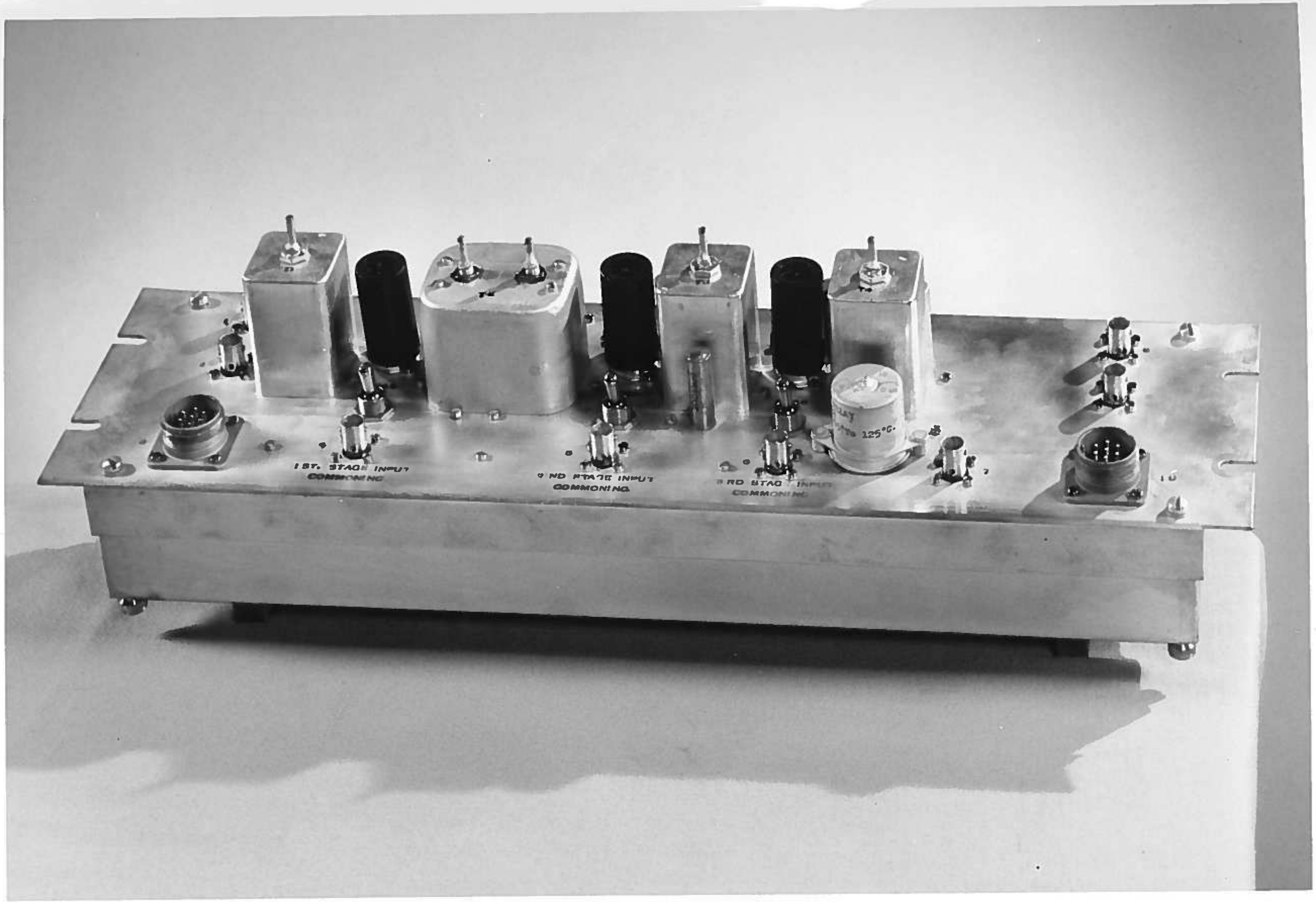

FIG. 4(a) FIRST I.F. AMPLIFIER

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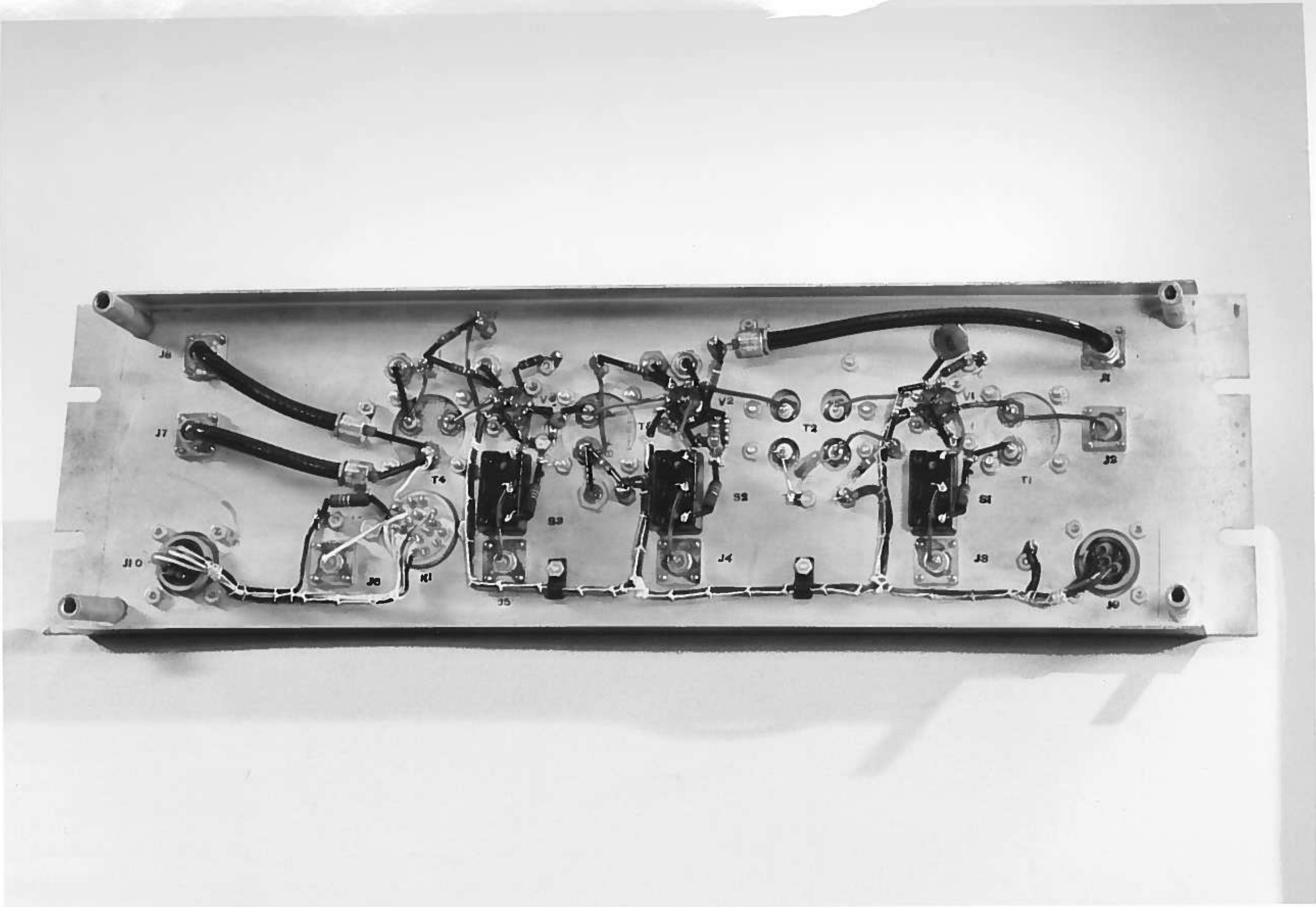

FIG. 4(b) FIRST I.F. AMPLIFIER

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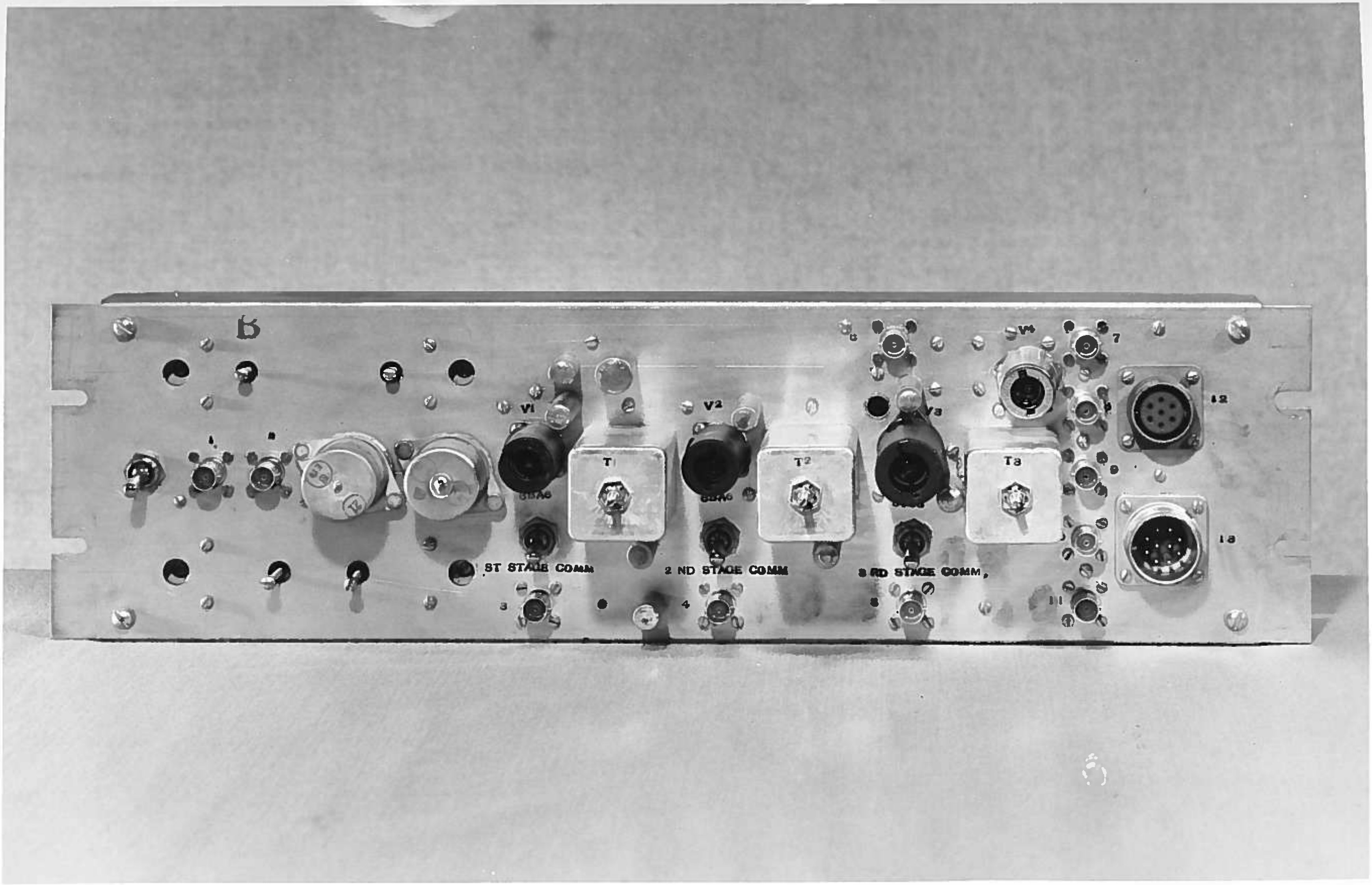

FIG. 5(a) SECOND I.F. AMPLIFIER

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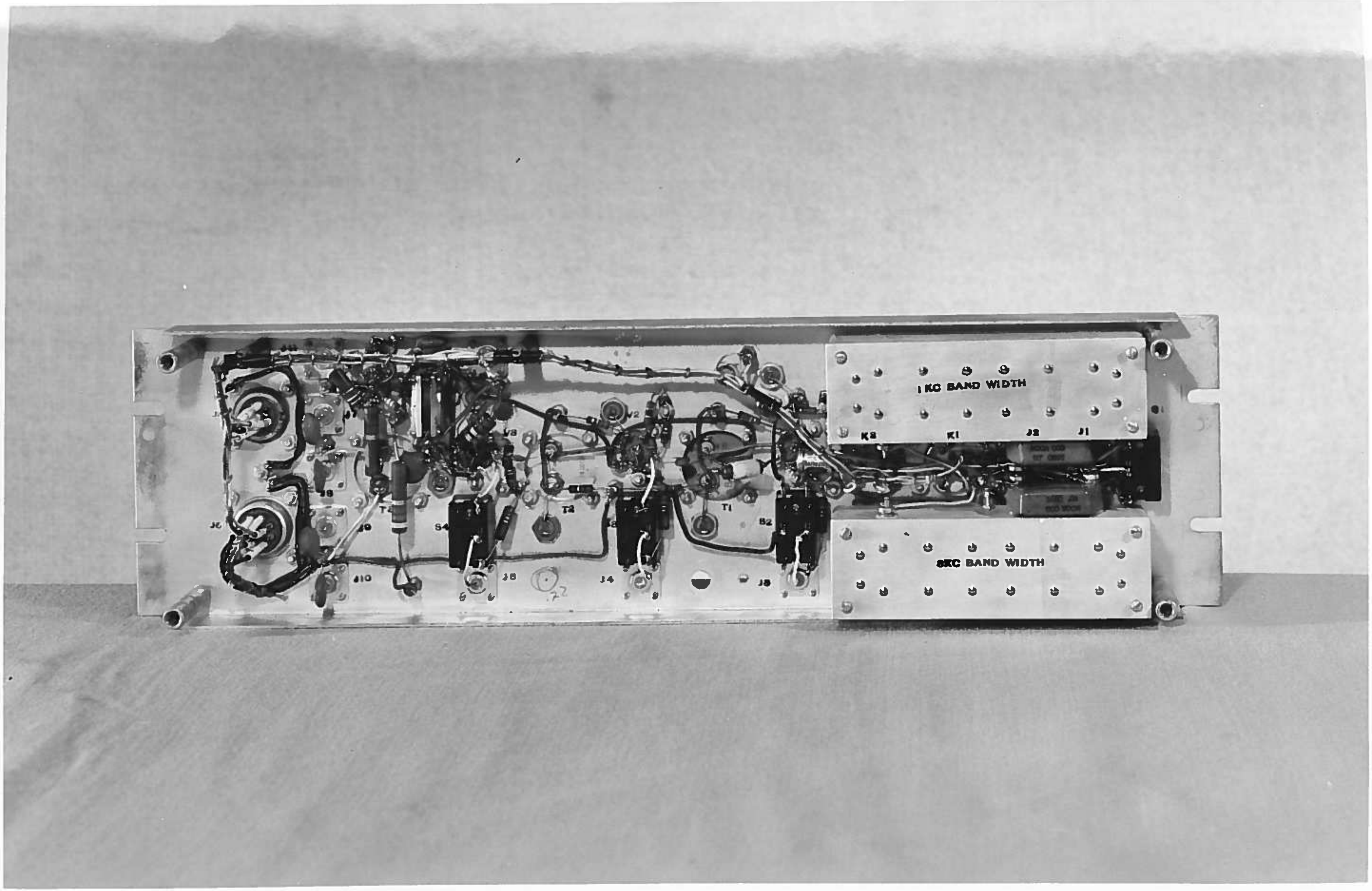

FIG. 5(b) SECOND I.F. AMPLIFIER

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