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A WIDE-BANDWIDTH MEDIUM-FREQUENCY RECEIVER

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CONTENTS

Abstract	ii
Problem Status	ii
Authorization	ii
INTRODUCTION	1
Effect of Bandwidth on Transmitted Information	1
Bandwidth Limitations in Transmitters and Receivers	1
Bandwidth Requirements of a Receiver	2
Tuned-Radio-Frequency Receivers	2
DESIGN OF THE TUNED-RADIO-FREQUENCY RECEIVER	3
Resonant Frequencies and Dissipation Factors	3
Tuning Capacitors and Inductors	3
Mechanical Design	9
Design Details and Performance	10
Video Section Details	14
DISCUSSION OF LIMITATIONS	15
ESTIMATED RECEIVER TUNING RANGES	15
THE RECEIVER AS A SUPERHETERODYNE	16
SUMMARY	18
ACKNOWLEDGMENTS	18
APPENDIX - Properties of an Inductance-Capacitance-Tuned Parallel Circuit	19

ABSTRACT

Radio-frequency signals of relatively wide bandwidths are being transmitted in the communications-frequency spectrum which extends approximately from 15 kc to 30 Mc. A countermeasures receiver must have sufficient bandwidth to accept all of the transmitted sideband frequencies if it is to present all of the information that was transmitted.

An experimental tuned-radio-frequency amplifier and detector has been developed with a fractional bandwidth of ten percent of the center frequency. It is tunable through a frequency range of 500 kc to 2.5 Mc and maintains practically constant gain. The receiver skirt selectivity measured at the 60 and 6 db points is approximately 3.75 to 1. The receiver is composed of flat-staggered quintuple r-f stages, in which both capacitance and inductance are tuned, and a push-pull detector followed by two dc-coupled cathode-follower video filter stages.

PROBLEM STATUS

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

AUTHORIZATION

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A WIDE-BANDWIDTH MEDIUM-FREQUENCY RECEIVER

INTRODUCTION

Effect of Bandwidth on Transmitted Information

Increasing use of wide-bandwidth radio-frequency signals in the communications spectrum has increased the need for wider bandwidth countermeasures receivers. All of the information transmitted cannot always be detected if the bandwidth of the receiver is less than that of the signal. One example of a signal that has a bandwidth wider than the passband of standard communications receivers is the navigational radio aid, pulsed LORAN signals. For LORAN signals in the region of 2.0 Mc, the recommended total receiver bandwidth [1] is 50 kc at 6 db down, and 150 kc at 60 db down, i.e., a 60 to 6 db skirt selectivity of 3 to 1.

Many of the modulation techniques used at microwave frequencies could conceivably be adapted for use at communications frequencies. In general, the wider the available bandwidth the more information can be sent or the faster it can be sent. For example, if a signal transmitted at 2.5 Mc with a spectrum of 10 kc can carry 10,000 items of information per second, it is conceivable that a signal of 250 kc spectrum could carry 250,000 items of information per second.

Bandwidth Limitations in Transmitters and Receivers

The limit to the bandwidth of a transmitted signal is determined by numerous factors. An important consideration is the difficulty of designing efficient wide-band amplifiers and practical antennas in the communications frequencies. The propagation characteristics of frequencies differing by small percentages are sometimes very dissimilar, especially for long-distance transmission, and thus limit the maximum usable bandwidth. Although the available bandwidths of signals in the communications spectrum are governed by federal regulations and international agreements, the basic characteristics are not established by man and are not so limiting.

Several of the factors limiting the bandwidth are directly related to the center frequency of transmission. The Q factors of transmitter tank circuits and of resonant antennas do not vary to the same extent as do the frequencies of resonance; in other words, the available bandwidth of a transmitter is naturally approximately proportional to the center frequency. The percentage available bandwidth, that is, the maximum ratio of signal bandwidth to center frequency, is taken to be ten percent for the purpose of the experimental receiver described here.

Intercept receivers now in use have insufficient bandwidth to accommodate signals of ten percent bandwidth throughout their range of center frequencies. The RAO-9 receiver has a fixed intermediate-frequency bandwidth of approximately 10 kc, the radio-frequency section bandwidths ranging from less than 10 kc at 550 kc to more than 200 kc at 30 Mc. This receiver has been modified for increased bandwidth by the addition of a second i-f

[1] Musselman, G. H., "LORAN," MIT Radiation Laboratory Series Number 4, Chapter 12, McGraw-Hill, New York (1948), p. 388

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amplifier and detector.[2] The Hammerlund Super-Pro receiver is more versatile in this respect as it has a variable bandwidth of 3 to 16 kc.

Bandwidth Requirements of a Receiver

A receiver with continuously adjustable bandwidth might be the best answer to the problem. Such a receiver could, after having been tuned to a signal, be adjusted for best signal-to-noise ratio by narrowing the bandwidth until it was wide enough for all the signal components but not wide enough to accept signals of frequencies different from those of the desired signal. Some progress has been made in the development of a variable bandwidth i-f amplifier. An experimental unit has been designed and constructed to operate at a center frequency of 300 kc with bandwidth variations of 8 to 1. The unit is composed of three sections of modified-lattice bandpass filters with variable high- and low-frequency cutoffs, and is controlled by changing the mutual inductance between inductors. The amplifier has a drawback in that when it is adjusted for sharp cutoff characteristics for frequencies near the passband limits it has insufficient rejection for signals falling outside of the high- and low-cutoff frequencies.

One of the drawbacks arising from the use of a variable bandwidth i-f amplifier in a wide-bandwidth superheterodyne receiver is that there must still be an r-f amplifier and an oscillator. The r-f amplifier must have enough bandwidth to pass the highest frequency components that are passed by the i-f amplifier when in its widest position, and if spurious responses due to image signals are to be made small, the r-f amplifier would have to be relatively sharp with good skirt selectivity.

A natural approach is to eliminate the wide bandwidth i-f amplifier and design the r-f section of the receiver to have the desired skirt selectivity; this entails, of course, the disadvantage of having more receiver bandwidth than is necessary for signals of less than maximum bandwidths. The r-f section should also have a large dynamic range of operation so that it would not easily overload and produce cross-modulation interference.

Tuned-Radio-Frequency Receivers

Advantages of a trf receiver when used as a wide-bandwidth amplifier are the elimination of the conversion, h-f oscillator, and i-f amplifier stages, and thus the interference problems associated with their use. An advantage which is not characteristic of trf receivers in general but which occurs in this receiver is the wide tuning range resulting from the use of both capacitance and inductance tuning giving a tuning range of 5 to 1 for medium frequencies.

Disadvantages of the trf receiver are: the beat-frequency detection of cw signals is inconvenient to arrange without an i-f amplifier; the r-f amplifier is more complex than the r-f amplifiers used in standard superheterodyne receivers, and if a multiband receiver were to be designed, the video sections and detectors would not be usable for more than one band in which it was desired to retain the entire constant fractional bandwidth. This latter disadvantage is comparable, however, to that of the variable-bandwidth superheterodyne receiver, in which separate i-f amplifiers would be needed for each band.

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[2] Bachelor, W. B., and Bullock, G. M., "Increased Bandwidth for the RAO-9 Communications Receiver," NRL Memo Report 159 (Confidential), May 1953

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DESIGN OF THE TUNED-RADIO-FREQUENCY RECEIVER

Resonant Frequencies and Dissipation Factors

The r-f amplifier filter stages of the receiver are arranged as a flat-staggered quintuple (Fig. 1). For such an amplifier-filter having a bandwidth band-center ratio, δ, of less than 0.3, the frequencies of resonance and the dissipation factors, or bandwidths of the individual parallel-tuned stages, are as follows: [3]

two stages staggered at $f_0 \pm 0.48\,\beta$ have dissipation factors of $0.31\,\delta$;

two stages staggered af $f_0 \pm 0.29\,\beta$ have dissipation factors of $0.81\,\delta$;

one stage centered at f_0 has a bandwidth β, where

f_0 = band-center frequency, β = over-all bandwidth, and $\beta/f_0 = \delta$.

When $\delta = 0.1$, the dissipation factors are so small that the Q factors can be taken as the reciprocal of the dissipation factors, and the impedance of each stage can be taken as the product of either reactance at resonance and the Q factor.[4] These calculated values and a sample tabulation of the resonant frequency of each stage, considering the center frequency to be 500 kc and the maximum capacity at the center frequency stage to be 250 $\mu\mu$f, are given in Table 1.

Electrically, the circuits are arranged according to resonant frequencies from input to output as follows: Number 1, medium high frequency; Number 2, medium low frequency; Number 3, high frequency; Number 4, low frequency; Number 5, center frequency. The center-frequency, low-Q circuit was placed at the output end in order to reduce the effect of detector loading on the passband shape. The next lowest-impedance circuits were placed at the input end, where signals have small amplitude, because of the reduced susceptibility to stray electrostatic pickup of a low-impedance circuit.

Tuning Capacitors and Inductors

Each parallel-tuned circuit is composed of a powered-iron-slug tuned, single-layer inductor, and a modified straight-line-capacitance variable capacitor. The inductance changes at approximately the same rate as does the capacitance. Each has a variation of approximately 5 to 1; thus, the frequency of resonance also changes by a factor of 5 to 1.

Several benefits are realized by the inductor and capacitor tuning system: (1) the frequency tuning range is equal to the square of that which it would be with the inductor or capacitor alone; (2) the impedance at resonance is independent of the resonant frequency—(see Appendix); (3) the bandwidth is proportional to the frequency setting—(see Appendix); (4) the shape of the selectivity curve is the same for all frequency settings.

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[3] Wallman, "Vacuum Tube Amplifiers," Radiation Laboratory Series Number 18, Section 4-8, New York, McGraw-Hill (1948), p. 186

[4] Terman, F. E., "Radio Engineers Handbook," New York, McGraw-Hill (1943), p. 144

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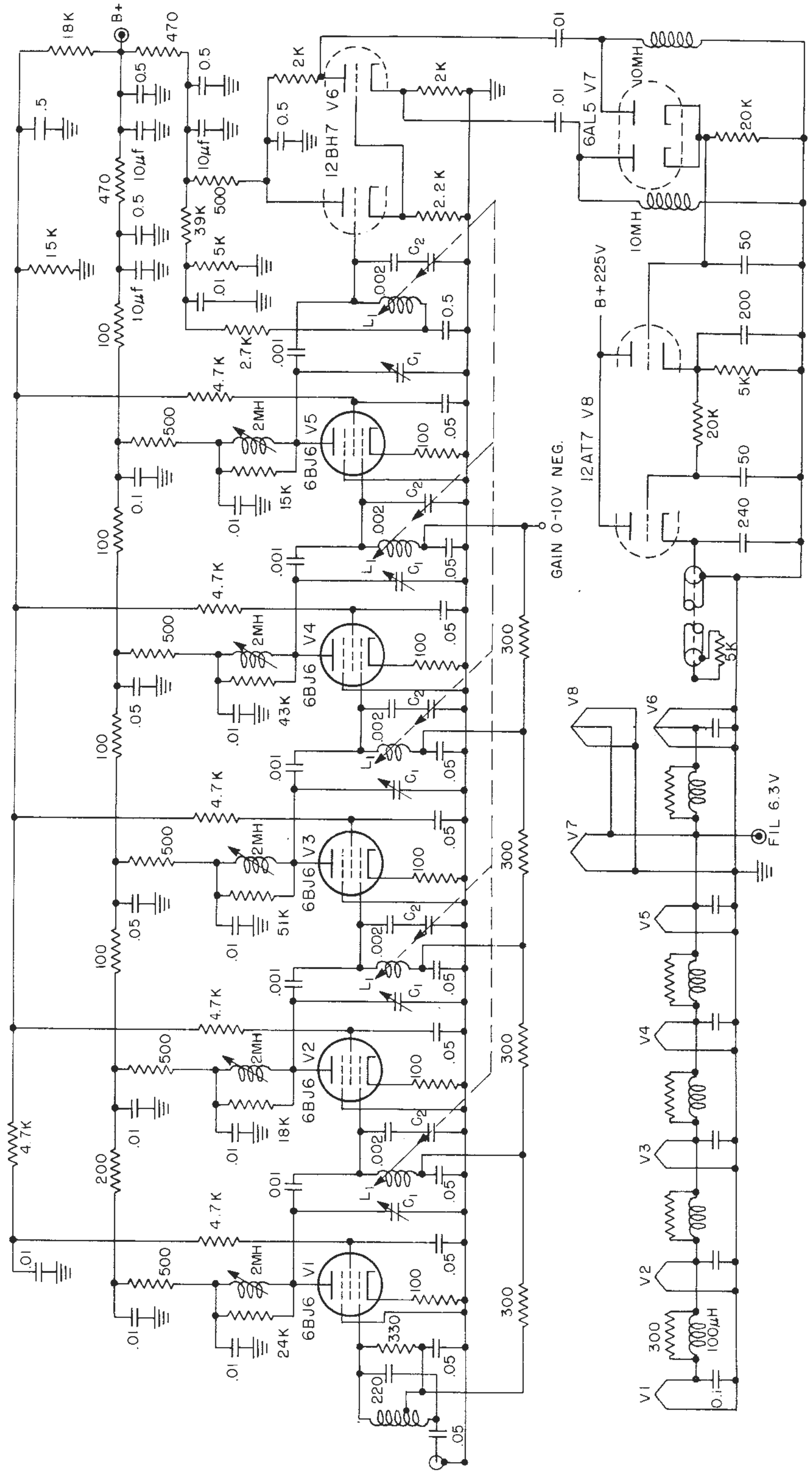

Fig. 1 - Schematic of 500 kc to 2.5 Mc receiver

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TABLE 1
Flat-Staggered Quintuple of $\delta = 0.1$ and $f_0 = 500$ kc

Stage Resonant Frequency (kc)	Approximate Q Factor	Approximate Impedance at Resonance (ohms)
476	32	41,000
485.5	12	16,000
500	10	13,000
514.5	12	16,000
524	32	41,000

Although it is desirable to have the change in capacitive reactance equal to the change in inductive reactance, considerable tolerance can be allowed so long as all the inductors are similar in inductive variation and all capacitors are similar in capacitive variation. If such is the case, an error in the tracking of the bank of capacitors with the bank of inductors would merely shift the response curve in frequency and change the L-to-C ratio thus changing the impedance and gain, but would not appreciably change the shape of the response curve. Constant gain in the receiver is not considered highly important, although a receiver with rapid change in gain versus frequency can be an inconvenience.

The five stages are isolated from each other and from the antenna by pentode remote-cutoff type-6BJ6 vacuum-tube amplifiers, and the final stage is isolated from the detector section by an untuned cathode follower of type 12BH7. The gain of the receiver is controlled by applying negative voltages to the control grids of all five pentodes. Since five pentode r-f amplifier stages with the plate impedances used in this tuner can produce much more gain than is needed, the grid bias can be set for relatively low stage gain even when the maximum over-all gain about 85 db is desired. The high plate resistance resulting from this low transconductance setting further helps in isolating stages from one another, and avoids large changes in input loading of the higher Q circuits when tube gain is changed. The plate resistance at maximum gain of the 6BJ6 amplifiers is approximately 3 megohms, a high value when compared to a load impedance of 0.04 megohm in the highest impedance stages.

Unbypassed resistors are placed in the cathode-to-ground circuits to make the tube input impedances more nearly constant. The value of these resistors was determined by experiment. For a constant high gain setting of the receiver, values of resistance were changed until the passband shape as observed on an oscilloscope using a frequency sweeping signal was closest to the passband shape obtained for low gain settings. The value used was 100 ohms.

Though the tuning range is wide compared to what it would be when using either capacitance or inductance tuning separately, it is limited by the maximum percentage bandwidth desired, as well as by the maximum variation that can be obtained with the capacitors and inductors. The video filter must be wide enough to pass the highest frequency components in the sidebands of the r-f signal when tuned to the highest center frequency, and at the same time, not so wide as to pass low r-f components when the receiver is tuned to the lowest frequency. The transient response of the video filter to pulsed signals will be improved if the filter has gradual, rather than sharp, cutoff characteristics. A practical tuning-range ratio was chosen to be 5 to 1. At that setting, the ratio of the lowest r-f components passed by the r-f section at a point 60 db down on the low-frequency side to the video high-frequency half-power point, was approximately 3.3. Front and rear view photographs of the receiver are given in Figs. 2 and 3.

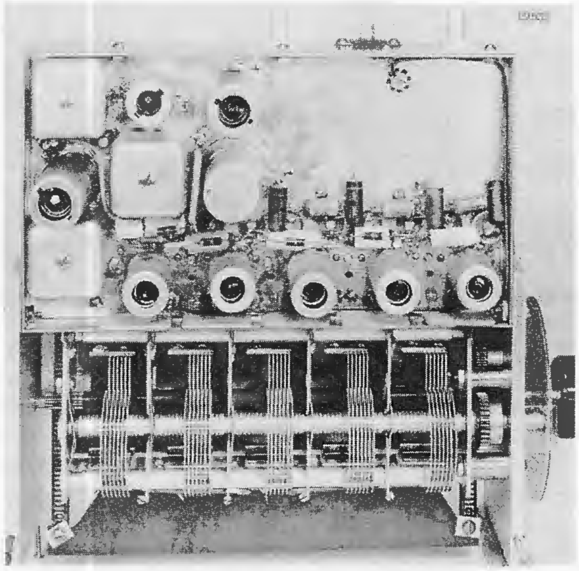

Fig. 2 - Receiver, front view

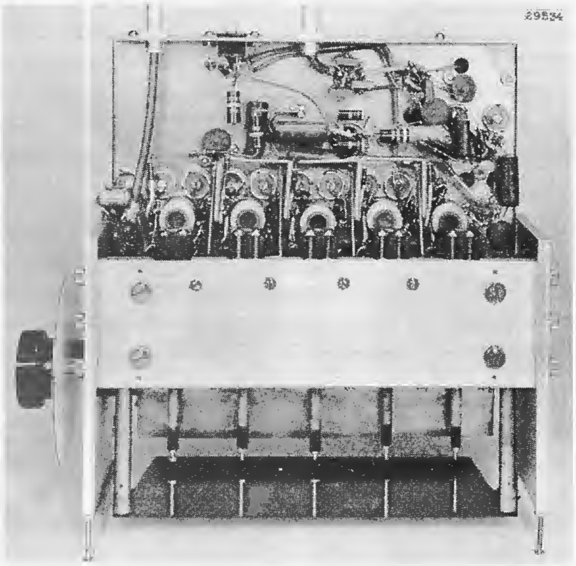

Fig. 3 - Receiver, rear view

The capacitors used in the tuner have maximum and minimum values of 230 and 10 $\mu\mu$f (Fig. 4). Each capacitor section was modified by removing plates from the rotor and stator until the maximum capacitance was 230 $\mu\mu$f. After 30 $\mu\mu$f are allowed for distributed capacitance and for trimming capacitors for each stage, the total range is greater than the design figures of 250 and 50 $\mu\mu$f, which values resonate with inductances of approximately 400 and 80 μh in the frequency range of 500 kc to 2.5 Mc.

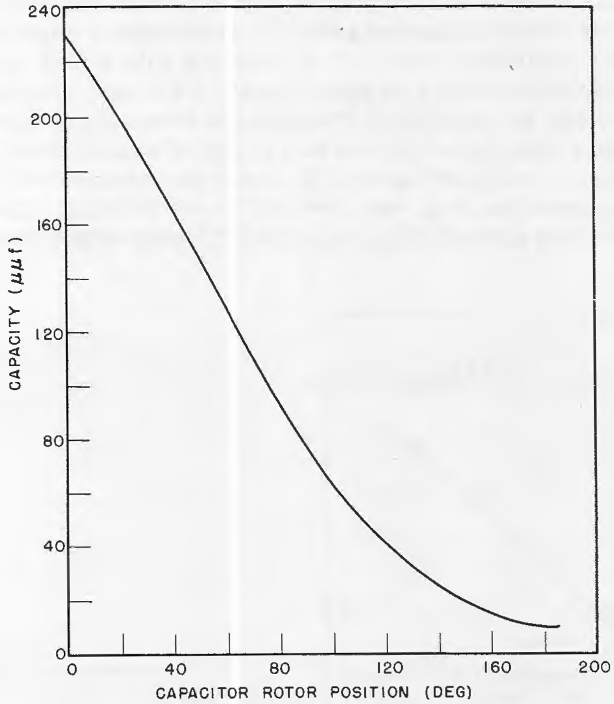

Fig. 4 - Capacitor variation

The inductors are wound in single-layer depth with Litz wire—seven strands of number 41 gauge—on 5/16-inch bakelite tubing to a length of approximately two inches. The tuning slugs, 1/4 inch in diameter and 1-5/8 inches long, are made of powdered iron. The slugs are a product of the Stackpole Carbon Company and are identified by the company as part number SK-1, grade S-49 powdered iron. The commercial slug, part number SK-1, was modified by the addition of a 1/8-inch length of powdered iron to each slug. The completed tuning inductors have Q factors ranging from a minimum of 67 to a maximum of 73 throughout the tuning range. Since the highest circuit Q needed is only 32, the greatest change from the average loaded value of 32 is only two percent over the tuning range.

A check of the temperature coefficient of inductance of the inductor for temperatures ranging from 25° to 90°C showed an approximately linear relationship with values averaging +0.032 percent per degree centigrade. No attempt has been made to temperature-compensate the tuned circuits. If the temperatures of the different inductors are alike, the relationship of their inductances will remain approximately constant, and the shape of the over-all passband likewise will remain approximately constant. The L to C ratios and the center frequencies of the passband would change with temperature.

Figure 5a is a plot of the positions of the tuning inductor slug (with respect to the inductor winding) required to resonate the inductance with capacitance, when the frequency of resonance is taken as being inversely proportional to the circuit capacitance for the various positions of the capacitor rotor. The meaning of the graph and its value may be better shown by explaining how it was obtained. An accurate plot of the capacitance values of the tuning capacitor versus rotor position was obtained (Fig. 4). Then, a circuit was assumed (Fig. 5b) to consist of a series capacitance of 2000 $\mu\mu f$ and shunt capacitance of 30 $\mu\mu f$ together with the tuning capacitor, and the combination capacitance was calculated for all positions of the capacitor rotor. Then starting with a high value of circuit capacitance corresponding to a capacitor setting of approximately 220 $\mu\mu f$, the frequency of resonance with an inductor was set at 500 kc, and other frequencies inversely proportional to the circuit capacitance values were calculated for the full range of capacitance. These values of circuit capacitance and frequency were set up on a Q meter to resonate with a tuning inductor. The positions of the tuning inductor slug (with respect to its winding) required to resonate the circuit were measured and plotted (Fig. 5a) against tuning-capacitor rotor positions.

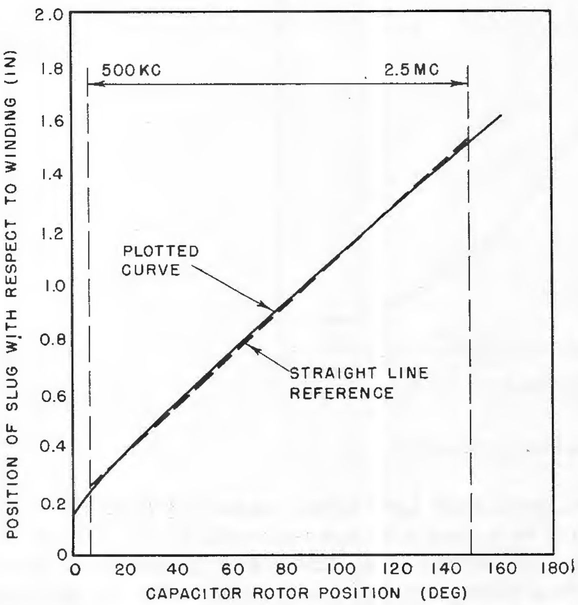

Fig. 5a - Similarity of inductive to capacitive variation

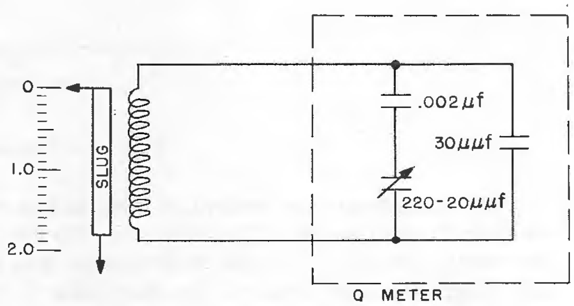

Fig. 5b - Circuit used to calculate Q-meter settings for graph of Fig. 5a

It is seen that if the reactance variation of the inductor and circuit capacitance are similar, the plot would show as a straight line and the two moving elements could be coupled through a linear mechanical system. The plot of Fig. 5a shows the best characteristics of the several test plots obtained for the frequency range desired. This plot was used to determine the ratio of linear movement of the inductor slug to rotational movement of the capacitor rotor, and the starting positions of each. While allowance for changing the starting positions of the inductor slug and capacitor rotor can readily be built into the mechanical linkage, provision for changing the relative motion between the two moving parts is not so easily made and was not attempted. A design figure of 0. 0094-inch slug movement per degree rotation of the capacitor rotor was obtained from the plot of Fig. 5a and was found to be satisfactory in the completed tuner. A slug movement of 1. 25 inches and a capacitor rotation of 143 degrees were required to cover the range of frequencies. The completed tuner has a maximum slug motion of 1.4 inches.

The resonant frequencies of the completed receiver, as functions of the manual tuning-control angles (see next section) are given in Fig. 6. In order to show the bandpass region clearly, the 3-db-down points rather than the center frequencies are plotted (curves A and B). Two reference curves are included in the graph. They show that the change in resonant frequencies with respect to the change in dial settings is equal to a constant in the case of curve C, and is proportional to the resonant frequency in the case of curve D. A cw signal at any tunable frquency would be in the passband of two receivers—a constant-small-bandwidth receiver with the tuning characteristics shown in curve C, and a constant-percentage-bandwidth receiver with the tuning characteristics shown in curve D—for unchanging periods of times as their respective dials were moved at constant rates. Tuning characteristics of a particular type could be obtained by designing the tuning elements in such a way that the total circuit capacitance and total circuit inductance changed (with motion of the dial) in inverse proportion to the desired frequency versus dial-setting characteristics.

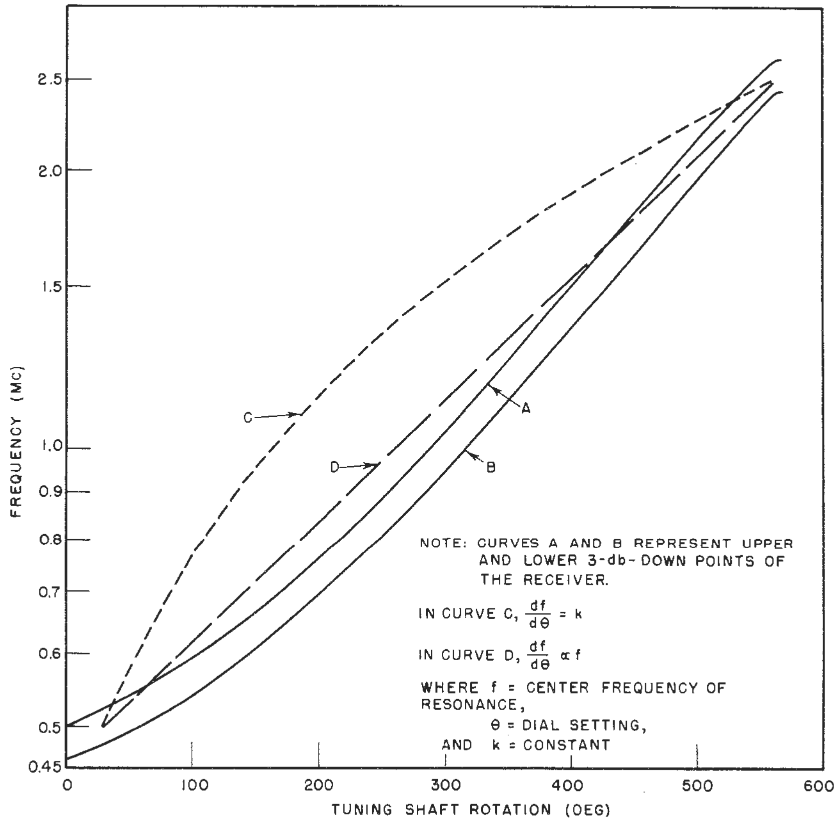

Fig. 6 - Frequencies of the upper and lower half-power points versus tuning-shaft angle

Mechanical Design

The inductor slugs and capacitor rotors are mechanically coupled through a pinion, rack, and gear system. In order to avoid binding of the slug carriage, racks are provided at both ends of the carriage and are driven by a common pinion gear. The pinion-gear shaft extends through the end plates and is used as the manual tuning control. The rack is coupled to the capacitor rotor shaft through an idler gear which can be adjusted to fit snugly with both the rack and the rotor gears, thus eliminating backlash in the motion between the inductor and capacitor moving parts.

Precision of movement is maintained by ball-bearing pinion and idler gear supports and close-tolerance sliding cylindrical mounts for the racks. Aluminum end plates of 1/4-inch thickness are accurately spaced by a 3/8-inch thick, 2-5/8-inch wide, 9-5/16-inch long aluminum connecting plate. Parts of the mechanical construction are shown in Figs. 2 and 3. Nonconducting linkages are used in the pinion and slug support bar to improve the electrical isolation of the amplifiers.

Design Details and Performance

Three trimming adjustments are employed in the receiver. High-frequency adjustments are controlled by the shunting capacitance trimmers, which compensate for the different input capacitances of the sections and allow for the partial correction of the dissimilarities of inductors and capacitors at their minimum settings. The powdered-iron tuning slugs are mounted with screw fittings and are used at medium frequencies as the main alignment adjustment. A more satisfactory alignment system would be possible if the tuning capacitors were also adjustable, but the capacitor sections in this tuner are rigidly mounted to a common shaft. A third adjustment for the correction of errors in the low-frequency portions of the tuning range is provided by shunting inductors.

The shunting inductors are in parallel with the tuning inductors and have the effect of reducing the total circuit inductance. They are adjustable by means of ferrite slugs through a range of approximately 2 to 1. Their values are approximately ten and five times larger than the inductance of the main tuning inductor when it is tuned to maximum inductance. Thus, the total effective inductance of the parallel-tuned circuit at 500 kc can be changed by means of the ferrite slugs through values of 5/6 to 10/11 that of the main tuning inductor alone. The change of circuit inductance compared to the average of the maximum and minimum inductance values was computed to be approximately nine percent, and five identical shunting inductors were constructed. However, in the final receiver adjustment it was necessary to reduce the inductance of the two shunting inductors which are in the above-center circuits of the quintuple in order to have an adequate range of adjustments when the receiver was tuned to the low-frequency end.

When they are suspended in air the shunting inductors are self-resonant at frequencies near 3.5 Mc, but when they are installed in the tight quarters of the tuner they show self-resonance at approximately 1.3 Mc. Their impedance becomes capacitive at frequencies above self-resonance and their maximum capacitance at 2.5 Mc (measured on a Q meter) is 5 $\mu\mu$f. Capacitance trimmers could have been used to compensate for low-frequency mistracking by using a large-value variable capacitor in series with the tuning capacitor so that the effective capacitance of the series combination would be less than that of the tuning capacitor alone. Such a variable capacitor of small volume appeared more difficult to obtain than a small volume variable shunting inductor.

The shunting inductor is especially valuable in that it permits the use of low dc resistance paths in both plate and grid circuits. Low resistance values are needed in the plate supply circuit to keep the plate voltage high and constant and maintain high plate resistance for all values of plate current. Low resistance is needed in the grid-return path in order to minimize the time constant of the interstage coupling capacitor and grid-return resistance, and thereby improve the overload characteristics. In this receiver, the shunting coil is in the plate circuit, the tuning inductor is in the grid circuit, and they are coupled through 1000 $\mu\mu$f capacitors. In a wide-bandwidth receiver there is a great possibility of encountering large amplitude signals in the passband when the receiver is tuned to a smaller amplitude signal. When the receiver is set for maximum over-all gain (85 db) the grid of the r-f cathode follower is driven beyond cutoff for receiver input signals larger than 1400 μv peak to peak, under which condition the peak-to-peak voltage at the cathode of the V-6 cathode follower is 60 volts.

The period during which the receiver remains blocked after the overloading signal is removed should be short. Ignition pulses are typical of a frequently encountered intermittent signal which can disable a receiver by blocking action though the pulse length is short compared to the pulse period. A receiver with fast recovery can return to normal operating condition shortly after the blocking signal is removed. The minimum time constant here is determined by the dc coupling circuits in the plate, control-grid, and screen-grid circuits rather than by the r-f coupling circuits. The control-grid decoupling resistors and capacitors have values of 330 ohms and 0.05 μf, and the plate decoupling values are 100 ohms and 0.1 μf. The overload characteristics of the completed receiver for two conditions of gain and signal output level are illustrated in Fig. 7. The photographs show the signal after detection and cathode-follower filter action in the receiver at the output terminal. In traces (a), (b), and (c) a 400 μv signal at 1.85 Mc modulated 95 percent by a 2000 cps sine-wave signal is applied at the input terminals in parallel with a series of synchronized pulse packets which have pulse lengths of 50 μsec and radio frequencies near 1.85 Mc. The amplitude of the cw r-f signal is just smaller than saturation amplitude. Trace (a) shows the demodulated 2000 cps wave and the pulse at equal amplitude. In trace (b) the pulse input amplitude was increased 30 db, and in trace (c), it was further increased 20 db making the pulse approximately 50 db into saturation condition. In traces (a), (b), and (c) the demodulated pulses were slightly stretched to make them more visible in the photographs, The receiver gain was approximately 85 db.

In traces (d), (e), and (f) the receiver gain was approximately 45 db, the input voltage was 200 μv, the radio frequencies of the mcw and pulsed signals were approximately 500 kc. The sine-wave modulation frequency was 2000 cps and the pulse length was 100 μsec. To compensate for the reduced gain of the receiver, the gain of the oscilloscope amplifiers was increased. Peak limiting of the receiver output was used to avoid overloading the oscilloscope amplifiers. In trace (d), the pulse envelope and cw modulation have equal amplitudes and in traces (e) and (f) the pulse amplitude was increased 40 and 60 db with respect to that trace (d). Both traces (c) and (f) show that the 2000-cps modulation is reduced in amplitude following the large pulses, and that full recovery of the receiver to normal operation requires approximately three milliseconds and that recovery is almost 75 percent complete after 0.5 millisecond under these conditions.

When high-permeability core material is used the effects of saturation of the core must be considered. In this receiver, the main tuning inductors are placed in the grid circuits where normally there is no direct current flowing. Changes in dc plate current can change the permeability of the trimming-inductor slugs and thus have much less effect on the total inductance of the tank circuits than they would have if the main tuning inductors were in the plate circuit and carried dc.

The four traces of Fig. 8 show the tuner response to square wave pulse packets at three frequencies. Trace (a) is plotted by the oscilloscope at a sweep speed of two microseconds per centimeter when the receiver is tuned to the signal frequency of 2.5 Mc. Trace (b) results from sweep speeds of five microseconds per centimeter with the input signal at

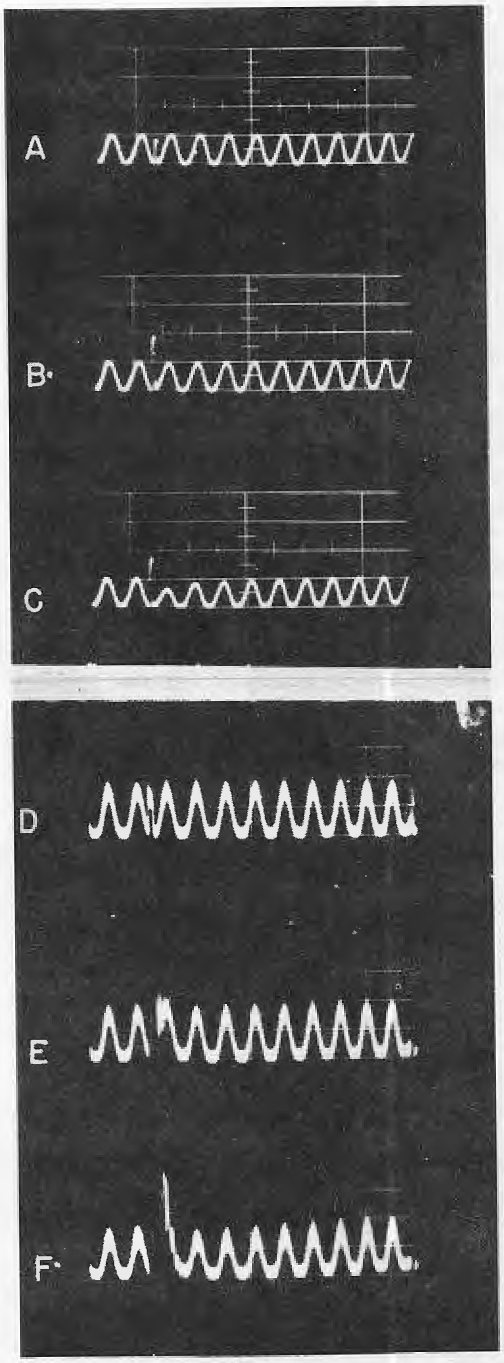

Fig. 7 - Overload characteristics

1.0 Mc. In trace (c) the sweep speed is ten microseconds per centimeter and the frequency is 500 kc. Trace (d) shows the pulse shape when the sweep speed and pulse length are the same as in trace (c) but the receiver was tuned to the signal at 2.5 Mc. The horizontal scale of the oscilloscope is divided into centimeters so that the approximate pulse length shown in trace (d) is 40 microseconds.

A graph of the relative responses of the receiver at 500 kc, 1000 kc, and 2.5 Mc is given in Fig. 9. For purposes of comparison, the frequencies were plotted as ratios of the center frequency for each case, and the maximum gains at the center-frequency settings were made equivalent by adjusting the output indicator gain control. The actual over-all gain values were 65 db at 500 kc, 64 db at 1000 kc, and 62 db at 2.5 Mc. These values occurred when the grid voltage was five volts negative with respect to ground, and the plate supply voltage was 225 volts.

Maximum gain is considered to be 85 db and the range of gain control is more than 85 db. When the receiver is set for 85 db gain, signals higher by a factor of at least 1.5 in frequency than the attuned frequency are attenuated no less than 90 db. Signals with frequencies lower than those of the receiver passband frequencies are similarly attenuated except when the second harmonics fall in the passband. Because of the harmonic output of standard signal generators a meaningful figure for the nonlinearity of the receiver is difficult to obtain. A reliable measure of the attenuation of signals whose harmonics fall in the passband of the receiver was obtained when the signal was at 500 kc. A narrow-bandpass filter centered at 500 kc was inserted between the signal generator and the receiver which attenuated the harmonics of 500 kc not less than 80 db. With the filter in place and the receiver gain set at 85 db, a second-harmonic output (at 1.0 Mc) of 0.1 volt peak to peak occurred when the input signal amplitude, at 500 kc, was 60 db greater than that which gave the same receiver output when the receiver was tuned to the fundamental. Higher order harmonics existed at a level below -90 db.

When the 500-kc input signal amplitude was increased so that the receiver output was 1.0 volt peak to peak there was 50 db difference in the 500-kc input signal amplitude for receiver frequency settings of 500 kc and 1.0 Mc. As before, higher harmonics were more than 90 db down.

Specific conditions were set up for a cross-modulation test. The receiver gain was 85 db. A 500-kc filtered signal was applied to the receiver input in parallel with a signal whose amplitude was adjusted to give cross-modulation components with maximum output amplitudes half that of the 500-kc signal (approximately 0.2 volt peak to peak cross-modulation output). When the two signals were in the passband of the receiver, cross modulation was maximum and the condition was called zero attenuation. Table 2 gives the relative attenuation values obtained for other frequencies of the interfering signal.

TABLE 2
Two-Signal Cross Modulation at 500 kc

Frequency of Interfering Signal (kc)	Relative Attenuation of Interfering Signal (db)
350	85
412	60
445	40
466	20
485-522	0
550	20
575	40
615	60
670	80
725-3000	>90

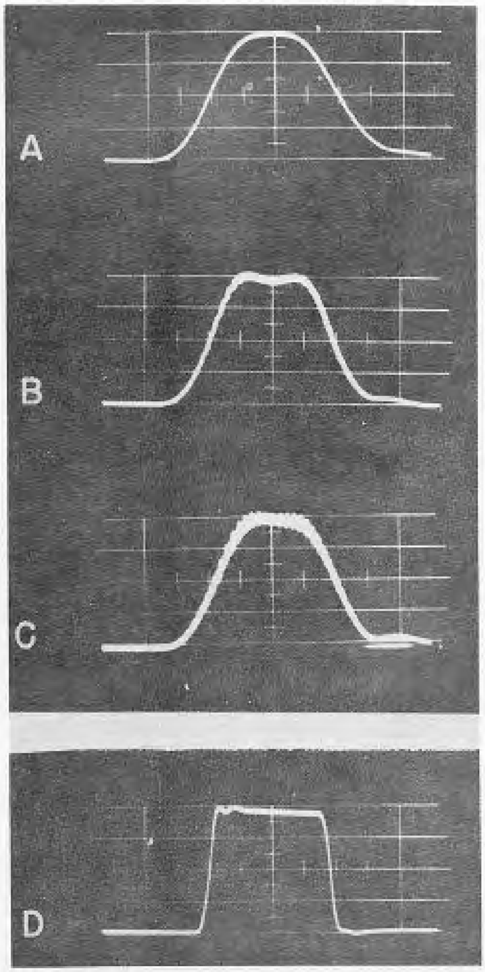

Fig. 8 - Square-wave response

In the test just described, the outputs of the two signal sources (filter output of the 500-kc signal and normal output of the standard signal generator for the variable frequency) were isolated by a resistance attenuation pad so that the combined signal-source impedance was essentially a constant resistance.

The results of these tests indicate that the linearity of the receiver at full gain is acceptable for operation under ordinary conditions. Tests performed at other frequency settings of the receiver gave similar results for cross modulation although harmonic generation tests could not be made without constructing filters for each frequency setting. The r-f linearity is best when the output signal is small, but detector linearity and efficiency are poor for small signals. An r-f and a video gain control on the receiver would enable the operator to adjust the receiver for best reception under the conditions encountered.

An r-f input circuit was developed which had an input resistance of approximately 50 ohms and inductive reactance ranging from 18 ohms at 500 kc to 55 ohms at 2.5 Mc. The circuit (Fig. 1) was composed of a ferrite cup-core input transformer with a secondary-to-primary turns ratio of approximately two-to-one, a 220-$\mu\mu$f coupling capacitor, and a 330-ohm secondary load.

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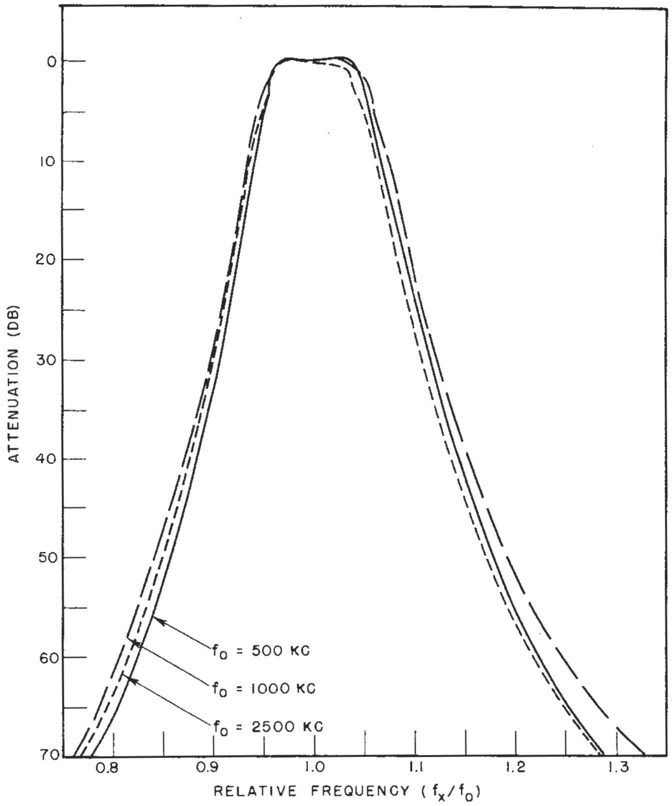

Fig. 9 - Receiver response characteristics

During development of the receiver a grounded-grid input connection was tried but abandoned when it was found that the receiver passband shape was adversely affected by changes in gain settings of the first stage because of the lower plate impedance.

Video Section Details

The video section is composed of a push-pull detector and two cathode-follower dc-coupled stages (Fig. 1). Load resistances of 5000 ohms in the cathode-follower circuits provide nearly linear operation through a large range of amplitudes. Direct-current coupling of the video signals eliminates phase and amplitude changes in the low frequencies. The output of the final tuned circuits is coupled to the grid of an r-f cathode-follower which in turn drives a resistance-loaded phase inverter. The cathode follower minimizes changes in the reflected load by isolating the detector section from the tuned circuit section.

The use of a push-pull detector simplifies the video filter design. The video filter must be wide enough to pass the highest frequency components when tuned to the highest center frequency. For ten percent bandwidths the maximum r-f bandwidth is 250 kc and the video should pass signals of at least half this frequency range, or approximately 125 kc. When the r-f circuits are tuned to 500 kc a signal carried at 475 kc would be the lowest frequency signal to receive practically full amplification, and to avoid intermodulation with video components it should be attenuated after detection. The use of a push-pull detector assists in the attenuation of the fundamental, and leaves the main r-f signal at twice carrier frequency, or 950 kc in this example, which is far enough removed from the video passband to be easily attenuated with simple RC filters. Another advantage derived from the use of a push-pull detector and a simple RC video filter with a gradually sloping attenuation characteristic is the improved transient response of such a filter over that obtained with a sharp cutoff filter.

DISCUSSION OF LIMITATIONS

The levels of impedance for the tuned circuits of this first experimental receiver appear to be too high. The same circuit Q values could be obtained if the L-to-C ratios were smaller. The effects of vacuum-tube variation and electrostatic coupling would be lessened, reducing the isolation and shielding problems. It is believed that in this receiver the full capacitance available in the original five-gang capacitor of about 450 $\mu\mu$f per section could be used to advantage. As was stated previously, the five pentode amplifiers have ample gain to permit lower load impedances.

Tracking of the inductances and capacitances would be more easily accomplished if the capacitors could be changed, one with respect to the other, by changing the relative positions of the rotors. The use of separate capacitors would allow more complete isolation between stages. When using a ganged capacitor with a common ground shaft there exists a path for r-f currents through the shaft. These extraneous currents can couple the first stage to the last stage thus making external shielding useless beyond a certain minimum point. The design of staggered single-tuned amplifiers such as are used in this receiver is based on perfect isolation of the stages. When coupling exists through any path except the isolating tubes the frequencies of resonance and the resonant impedances can change. The feedback could be regenerative or degenerative—either type affects the shape of the passband. The gain of the receiver is limited to 85 db because at higher-gain settings the passband shape is affected by uncontrollable couplings. At the maximum gain setting possible within the dissipation limits of the r-f amplifiers, feedback is sufficient to cause oscillation.

ESTIMATED RECEIVER TUNING RANGES

It is estimated that receivers of this type could be designed to operate through the frequency range of 20 kc to 40 Mc in five units. Difficulties in obtaining adequate inductance variation by permeability tuning in the high-frequency regions, and impracticably large variable capacitors required in the very low frequency region place some limitation on the frequency coverage of the high and very low frequency tuners. However, in the very low frequency region where large inductance changes are practical, the full range of 5 to 1 might be obtained by permitting the L/C ratio to change somewhat as a result of reducing the capacitor variation, and increasing the inductance variation through the use of higher permeability core materials. Estimates of band coverages are given in Table 3.

TABLE 3
Estimated Receiver Tuning Ranges

Frequency Region	Frequency Range (kc)	Tuning Range Ratio
Very Low	20 to 100	5 to 1
Low	100 to 500	5 to 1
Medium	500 to 2500	5 to 1
High	2500 to 10,000	4 to 1
High	10,000 to 40,000	4 to 1

THE RECEIVER AS A SUPERHETERODYNE

The receiver could be adapted as a superheterodyne receiver by the addition of an i-f amplifier and of a sixth LC section to the r-f tuner for the local oscillator. The i-f amplifier might have continuously variable bandwidth, or might have various fixed bandwidths made available with switched RCL circuits and electromechanical filter sections such as crystal and mechanical resonators. The versatility of the 500 kc to 2.5 Mc receiver would be good provided i-f bandwidths of 50 kc, 10 kc, 2 kc, and 400 cps were available. The widest i-f bandwidth, 50 kc, is approximately the bandwidth of the experimental receiver r-f section when the receiver is tuned to 500 kc. The 50 kc bandwidth would be sufficient to permit panoramic adapter presentation over a bandwidth somewhat larger than 50 kc. The remaining narrow bandwidths are those required of a standard communications receiver. A block diagram of a possible receiver incorporating these suggestions is given in Fig. 10.

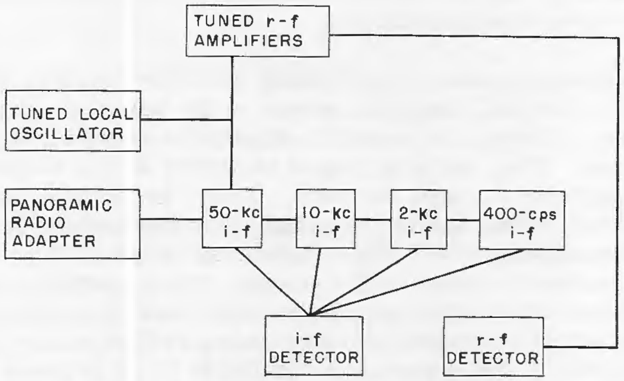

Fig. 10 - Block diagram of speculative multibandwidth receiver

If all five r-f amplifiers were used as the r-f section of a superheterodyne receiver, the gain of the r-f amplifiers would have to be low in order to reduce cross modulation of signals in the r-f stages. An experimental setup for checking cross modulation in the r-f section was arranged in which a standard HRO communications receiver was used as the local oscillator and i-f amplifier. The r-f section of the experimental receiver was changed from that shown in the circuit diagram of Fig. 1 by adding a 500-ohm resistor in series with the 2000-ohm resistor in the cathode of the r-f cathode follower. The 500-ohm resistor was placed between the 2000-ohm resistor and ground, and a 0.01 μf capacitor coupled the signal developed across the 500-ohm resistor to the HRO receiver.

In the cross-modulation test the lower frequency signal of two signals applied to the input of the r-f tuners was of constant amplitude and was unmodulated. The upper frequency signal was 100 percent modulated with a 1000-cps sine wave. The HRO receiver was tuned to the lower frequency reference signal. Signals were placed at the edges of the experimental passband at points not more than 3 db down. The amplitude of the higher frequency signal was increased until 1000 cps cross-modulation components appeared on the lower frequency signal as observed on an oscilloscope using the audio output of the HRO receiver. These components had amplitudes corresponding to about 20 percent modulation of the lower frequency reference signal. Tests were made with gains of 20 db and 60 db and with 10-μv reference signal input to the r-f tuner. The gain of the HRO receiver

[redacted]

was adjusted to give an audible output signal without overloading. The results of these tests are given in Table 4. The low gain required of the r-f tuners to obtain linear amplification would make it inconvenient to operate the receiver as a trf and superheterodyne simultaneously.

TABLE 4
Cross Modulation Within the R-F Passband

Center Frequency of R-F Tuners (kc)	Gain of R-F Tuners (db)	Frequency of Signals (kc)	Input Signal Amplitude
880	20	860 900	10 μvolt 10 μvolt + 75 db
1920	20	1840 2000	10 μvolt 10 μvolt + 78 db
880	60	860 900	10 μvolt 10 μvolt + 40 db
1920	60	1840 2000	10 μvolt 10 μvolt + 55 db

The best center frequency for the i-f amplifiers must be selected by means of a compromise between the need for minimum r-f interference and the need for maximum image-frequency rejection. On the one hand a low intermediate frequency minimizes the interference caused by r-f signals which fall in the passband of the i-f amplifier when the r-f amplifier is tuned to its lowest frequency (500 kc). The only attenuation of these low-frequency r-f signals is that of the r-f amplifiers. On the other hand, the image-frequency rejection provided by the r-f amplifiers is improved when the image frequency is well removed from the r-f passband—the condition corresponding to a relatively high intermediate frequency (assuming the local oscillator frequency is above, and the i-f passband is below the r-f passband).

An inspection of Fig. 9 shows that 60-db attenuation of the image signals, when the r-f amplifier is tuned to a center frequency of 2.5 Mc, could be obtained with the image signal at a relative frequency of 1.22, a frequency of 3.05 Mc. This corresponds to an intermediate frequency of 275 kc. The graph also shows that when the receiver is at 500 kc the relative attenuation of r-f signals at 275 kc would be much more than 60 db.

Nonlinearity in the r-f amplifiers produces first-order difference frequencies between signals which fall in the r-f passband. When the receiver is tuned to 2.5 Mc the bandwidth is approximately 250 kc so that two signals located at the three-db-down frequencies could generate a difference frequency of 250 kc which would be amplified in an i-f passband centered at 250 kc. Thus, an i-f amplifier with higher center frequency is needed to make the receiver less sensitive to that type of interference. A third factor favoring the use of a high i-f is the relative ease compared to lower frequencies, of filtering the r-f components from the video signals after detection.

A compromise i-f design center frequency for this receiver is estimated at 350 kc. At this i-f potentially interfering signals of 350 kc separation centered about 2.5 Mc would be at 15-db down points on the r-f passband, or if one signal were on one edge of the r-f passband, the other would be at a point approximately 30 db down on the opposite side.

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Image signals would fall on a point more than 60 db down on the r-f selectivity curve. When the receiver was tuned to 500 kc, r-f signals at 350 kc would be on a point more than 60 db down on the r-f selectivity curve.

Instead of using all five r-f sections as the superheterodyne front end, the signal could be taken after, for example, two r-f stages. The first two filters in this receiver are symmetrically tuned about the center frequency and are of relatively low impedance and low gain. The problem of cross modulation would be much reduced because of the lower signal amplitudes at that point, and the gain of the r-f stages could be increased, thus making possible the simultaneous use of the wide and narrow bandwidth detectors. The main disadvantages are the poorer image rejection resulting from reduced selectivity of the first two low-Q-factor stages and reduced attenuation of the radiation from the local oscillator.

SUMMARY

The final amplifiers in many r-f transmitters operate with Q factors in the region of ten, and have the capability of amplifying signals occupying fractional bandwidths of ten percent. The fact that most communications signals require no more than a fixed bandwidth of 10 kc reduces, but does not eliminate the possibility of using the full transmitter bandwidth for the purpose of employing more channels of information or of using one channel at a faster rate.

The receiver described is an example of a method of obtaining relatively broad bandwidths in the communications frequencies. It has a fractional bandwidth of approximately ten percent throughout the tuning range of 500 kc to 2.5 Mc. Because of the combination capacitive and inductive tuning system used, the over-all gain and selectivity characteristics are almost constant. The experimental receiver has reasonably good response to pulsed signals and has a large linear dynamic range of amplification such as is required in a broad-bandwidth amplifier.

Narrow-bandwidth reception could be provided by adding a sixth LC tank circuit and narrow-bandwidth i-f amplifiers in a superheterodyne arrangement. The five r-f amplifiers, when operated at an over-all gain of no more than 20 db, have sufficient amplitude linearity to allow narrow-bandwidth reception of a small amplitude signal which is in the receiver r-f passband with a signal 80 db greater in amplitude.

It is recommended that the concepts involved in the experimental receiver be considered in the design of future countermeasures receivers intended to intercept, and present for analysis, signals occupying fractional bandwidths greater than those encountered in standard communications use.

ACKNOWLEDGMENTS

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APPENDIX

Properties of an Inductance-Capacitance-Tuned Parallel Circuit

When in a parallel tuned LC circuit the values of inductance are made proportional to the capacitance values, i.e., $L = kC$ the frequency of resonance is inversely proportional to either the inductance or the capacitance:

$$f = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{kC^2}} = \frac{k_1}{C},$$

where

f = resonant frequency,

L = inductance,

C = capacitance, and

k, k_1 = constants.

In the receiver described, the inductance, capacitance, and frequency all have maximum and minimum values differing by a factor of five.

At resonance the reactances are equal in magnitude:

$$\overline{X}_L = 2\pi f L = \overline{X}_C = \frac{1}{2\pi f C}.$$

But since the variable capacitor and variable inductor have inversely proportional relationships with the resonant frequency, it is seen by substituting k_2/f for L and k_1/f for C that the reactances are independent of the frequency and are constant:

$$\overline{X}_L = 2\pi f\left(\frac{k_2}{f}\right) = K$$

and

$$\overline{X}_C = \frac{1}{2\pi f\left(\frac{k_1}{f}\right)} = K$$

where k_2 and K are constants.

When the parallel-tuned circuit has inherent losses much smaller than are required to obtain the desired bandwidth, an external loss which is large enough to obtain the desired bandwidth can be added and thereby fix the impedance at resonance. The external loss can be in the form of a resistor placed in parallel with the circuit.

The parallel resistance, composed of the damping resistance and the equivalent parallel LC circuit resistance, divided by the reactance of L or C is the Q factor of the circuit. Since the parallel resistance and the reactances are constant, the Q factor is constant.

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The bandwidth of a parallel-tuned circuit of $Q \geqq 10$ is approximately equal to the frequency of resonance divided by the Q factor, and since the Q factor is constant in this circuit, the bandwidth is proportional to the resonant frequency.

The impedance at resonance is equal to the product of two constants, the reactance and Q factor, and is therefore also constant.

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