THE DESIGN OF DISCRIMINATORS FOR

SUBMINIATURE FREQUENCY MODULATED TRANSCEIVERS

A THESIS

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DISCRIMINATORS FOR FREQUENCY MODULATED PORTABLE TRANSCEIVERS

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1.0 INTRODUCTION

1.1 PURPOSE OF THE THESIS

The purpose of this thesis is to investigate the design of discriminators suitable for use in portable frequency modulated transceivers.

1.2 FREQUENCY MODULATION IN COMMUNICATION ENGINEERING

During the past decade, frequency modulation, which was intended originally as a means of extending and improving broadcasting facilities has found extensive use in the field of mobile communications. The public has been generally indifferent to the high fidelity and freedom from inteference inherent in frequency modulation broadcasting, and indeed, many of the home receivers designed for this type of service do not make full use of its advantages. In the field of mobile communications on the other hand, although high fidelity is not generally a major requirement, high signal-to-noise ratios and freedom from interference are important factors in establishing a system of maximum reliability. Industrial, public utility and military communications have made good use of these advantages of frequency modulation and it may be that the science has found its most useful application in these fields.

Communication systems employing frequency modulation are usually of the "narrow band" or "low deviation" type. The maximum carrier deviation from the unmodulated value is restricted to about fifteen kilocycles and the voice band transmitted is generally limited to about three kilocycles. This is quite a different situation from that existing in frequency modulation broadcasting where carrier deviations of seventy-five kilocycles are used in transmitting an audio band of fifteen kilocycles. These operating conditions of the low deviation system are fixed mainly by bandwidth restrictions and by the requirements of a frequency modulation [illegible] system which is to exhibit a useful gain over an amplitude modulation system of the same audio band. It will be of interest to review these factors here.

1.3 <u>FREQUENCY BANDWIDTH</u>

The frequency spectrum of a frequency modulated wave has been investigated in detail by many authors (1, 2, 43) and for our purposes it will be sufficient to refer to some of the practical aspects of that work. In one of the earliest papers on the subject (1) Van der Pol showed that the bandwidth of a frequency modulated wave depended largely upon the ratio of the frequency deviation of the carrier from its unmodulated value to the modulation frequency. For practical purposes he showed that the bandwidth can be approximated by either twice the frequency deviation or twice the modulation frequency depending upon which of the two is the larger. More exact expressions have been developed for bandwidth and for the case of complex modulation but in general the bandwidth does not depart greatly from the value indicated by Van der Pol. It is evident that maximum spectrum economy demands the use of as low a value of frequency deviation as is consistent with satisfactory frequency modulation noise and interference gain. Mobile communication work, particularly in the military field, requires the use of a large number of closely spaced and well defined channels and this is one of the principal reasons for confining frequency deviations to approximately plus or minus fifteen kilocycles.

1.4 <u>THE RELATION OF MODULATION INDEX TO NOISE REDUCTION</u>

Van der Pol has defined the ratio of frequency deviation to modulation frequency as the index of frequency modulation. In his paper on frequency modulation noise characteristics (8), Crosby has shown that the signal-to-noise ratio gain of a frequency modulation system over an amplitude modulation system of the same audio band is directly proportional to the value of the modulation index. for fluctuation noise such as the thermal and tube noise, which is characterized by a random relation between the various frequencies in the spectrum, the improvement of frequency modulation over amplitude modulation is given by $\sqrt{3}$ times the modulation index. For impulse noise, characterized by an orderly phase and amplitude relation between the various noise frequencies, the gain in signal-to-noise ratio is given by twice the modulation index.

The gain of frequency modulation over amplitude modulation takes the place only when the carrier-to-noise level of the system exceeds a certain critical value which Crosby calls the threshold level. When the carrier-to-noise level is about three or four db above equality of peak carrier and peak noise, the peak improvement for either type of noise begins to fade and becomes zero when approximately equal carrier and noise levels exist.

It would appear from above that if frequency conservation were not a problem, the optimum condition of operation would be provided by using as high a deviation as possible, thus increasing the modulation index and the signal-to-noise ratio gain of the system. Crosby has shown however that the improvement threshold level discussed above occurs at a lower carrier-to-noise level in low deviation systems. This effect is due to the extra noise accepted by the wide receiver bandwidth required by high deviation systems. In mobile communication work therefore, where it is necessary to work at times with very low carrier to noise level it is evident that some advantage is gained by the use of smaller amounts of frequency deviation. This fact has been substantiated by further work in the field by Crosby and other investigators. (13, 15, 16)

In communication work, by limiting the audio band to three kilocycles and the frequency deviation to fifteen, it is possible to obtain the same modulation index, and therefore the same effective signal-to-noise ratio gain, as is obtained in frequency modulation broadcasting by using a deviation of seventy five kilocycles and an audio band of fifteen.

In summation, the narrow band system makes it possible to use many more channels and to work with lower carrier-to-noise ratios than would be possible were high deviations used. Greater frequency deviations would, of course, give a better signal-to-noise ratio for carrier-to-noise levels within the threshold. For communication purposes, however, the signal-to-noise ratio need not be greater than that necessary to ensure good intelligibility of the signal, and range is most important.

1.5 FREQUENCY MODULATION TRANSCEIVERS

The term transceiver as used here refers to a unit of a radio communication link which is capable of transmitting and receiving voice intelligence. As a general rule transceiver implies that the sending and receiving is done on the same frequency and this will be understood in this paper.

The transceivers which are used in police, military, forestry patrol and other communication systems may be intended for fixed station, mobile or portable use. The distinction made between mobile and portable is that which exists between units which are normally transmitted in vehicles(mobile) and those which may be conveniently carried and operated by a single operator. Generally the size, weight, power output and power consumption of the units is greatest for fixed station equipment and least for portable units. Since interoperation of the various units is frequently required however, we find that such requirements as frequency stability and deviation, receiver sensitivity and selectivity, do not vary greatly from one type of unit to the other. This has the effect of making the design of portable equipment somewhat more difficult since substantially the same performance, other than power output is required of a unit of much smaller size.

1.6 THE PROBLEMS OF PORTABLE TRANSCEIVERS

Transceivers which must be carried by the individual, occasionally for considerable periods of time, must of necessity be as light and compact as possible. They must be weatherproofed to a high degree and rugged enough to withstand a considerable amount of rough usage. They must be capable of normal operation over a wide temperature range. Power consumption must be low so that the lifeof the required power supply will be long and its weight andsize kept to a minimum.

To meet the weight and size requirements of portable transceivers, recourse has been made to the use of miniature, and more recently subminiature, tubes and components. Subminiature tubes are presently available in a sufficient variety of types for the complete design of an ulta high frequency receiver and transmitter and further improvements in these tubes may be expected. Resistors and condensers in a wide range of value and a fraction of their size are also available although tolerances are not as accurately maintained in the subminiature sizes. The use of printed circuits may be expected to contribute to further reductions in size when the techniques become better known and more readily applicable.

This reduction in size of transceiver components has naturally tended to increase the normal physical difficulties associated with the problems of production, alignment, and maintenance. To deal with these problems, it has been found useful to break down the construction of portable transceivers into a relatively large number of small sealed plug-in sub assemblies. By a construction of this nature production can be greatly facilitated. Repair and fault tracing can be reduced to a matter of plugging in successive test units until normal operation is obtained.

It is evident that a physical construction consisting largely of small plug-in units will place certain restrictions upon circuit design. The coupling between the plug-in units becomes a problem. All of the components in the plug-in unit will be in close proximity to the shielding walls and electrostatic coupling between the components will be difficult to control. The electromagnetic field of coils must be confined by toroidal or cupped coil construction or the shielding walls will contribute to coil losses. A physical and electrical design which seriously impairs the reliability and performance of a communications unit cannot be acceptable. Within the limits composed by the plug-in system, however it should be possible toeeffect a satisfactory circuit design.

It is the purpose of this paper to investigate the design of a subminiaturized discriminator suitable for use in a portable transceiver having the mechanical construction indicated above. It will be in order then to examine closely the function of a discriminator in a portable frequency modulated transceiver and the requirements of a satisfactory design.

1.7 THE DISCRIMINATOR FUNCTION IN PORTABLE TRANSCEIVERS

The discriminators used in portable transceivers mustusually perform a dual task. When the transceiver is receiving the discriminator must translate the frequency modulated wave to an amplitude modulated wave and extract the audio signal frequency. When the set is transmitting the discriminator forms part of a control circuit designed to maintain constant frequency of the unmodulated carrier.

One of the principal problems associated with portable frequency modulated transceivers is that of frequency control. The limited frequency deviation and receiver bandwidth requires a high degree of frequency stability. Transmitter designs, such as the Armstrong indirect system which provides crystal control of the unmodulated carrier, are not practical in these sets where the number of stages and power consumption must be kept to a minimum. A method of direct frequency modulation must be employed and some provision must be made for controlling the carrier centre frequency.

One of the simplest and most satisfactory means by which frequency control may be maintained is illustrated in the block diagram of Fig. I.

The receiver section of the set is crystal controlled at the carrier frequency. The transmitter is directly frequency modulated at the carrier frequency or a submultiple of carrier frequency by means of a reactance connected in parallel with the oscillator tuned circuit. The reactance tube functions as an equivalent inductance or capacitance of a value depending upon the instantaneous grid voltage, and this voltage may be varied by the application of audio signal voltages. By proper design, the variation of reactance so produced may be made cause deviations of the carrier frequency having a linear relationship to the intensity of the applied audio signal.

When the transceiver is sending, the receiver may be left in operation A portion of the transmitter signal is then picked up by the crystal controlled receiver. The discriminator produces a direct voltage component whose average value changes with slow variations in the transmitter centre frequency. This direct voltage component may then be applied, in series with the audio modulation, to the grid of the reactance tube. If the voltage ia applied in the correct sense, the mean carrier frequency may be kept at substantially the same frequency as the crystal controlled receiver.

Different variations of this scheme may be employed but the function of the discriminator is usually much the same. It must provide distortionless detection of the received frequency modulated wave. When the transceiver is sending, it must provide a direct voltage component in the proper sense and at the proper level for automatic frequency control of the transmitter.

With the role of the discriminator thus established it remains to consider the requirements of a discriminator which is to accomplish this task most efficiently.

1.8 THE DESIGN REQUIREMENTS OF DISCRIMINATORS FOR F.M. TRANSCEIVERS

Many different schemes have been employed in the detection of frequency modulated signals. Before proceeding to a study of these methods it will be in order to establish in a general way the requirements of a good F. M. discriminator, keeping in mind thepurpose for which it is intended. This will supply a basis for studying the merits of the various methods which might be employed to achieve the same ends and help to eliminate those methods which are obviously unsuitable.

The requirements of a discriminator for distortionless F.M. reception have been investigated by Roder (9) and Chaffee (4) among others. Let us consider the wave equation for a frequency modulated wave as developed by Carson and Van der Pol.

$$e = E_1 \sin \ (wt + m \sin ut) \qquad (1)$$

Where E_1 = amplitude of the unmodulated carrier

W = unmodulated carrier angular frequency

U = modulation angular frequency

$$m = \frac{\text{carrier frequency deviation}}{\text{modulation frequency}} = \text{frequency modulation index}$$

$$\doteq \frac{\Delta w}{u}$$

When the expression for e is differentiated with respect to time we have

$$\frac{de}{dt} = Elw \left(1 + \frac{\Delta w}{w} \cos ut \right) \cos (wt + m \sin ut) \qquad (2)$$

This is an amplitude modulated wave with a modulation factor $\frac{\Delta w}{w}$. It is evident, if the wave of equation 1 is applied to a linear time rate demodulator of the proper bandwidth, it should be possible to extract the audio signal from the resultant wave of equation 2 byconventional amplitude modulation detection methods.

1.8.1 DISCRIMINATOR BANDWIDTH

The question of the required bandwidth of the linear demodulator or differentiator is interesting. It is stated by Hund (23) that the maximum range of line arity of the discriminator need only be twice the maximum frequency deviation since in effect it merely reverses the process of modulation at the transmitter where frequency deviations are limited to a definite maximum value. The bandwidth of all bandpass networks preceding the demodulator must however pass the entire frequency spectrum of the wave.

The problem was attacked by Roder in a more formal manner (9) He reduced the wave of equation (1) to the familiar Bessels series expansion and applied this wave to a network having a phase and amplitude characteristic linearly variable with frequency. When the proper change in amplitude and phase is attached to each sideband and the resulting series is recombined an output wave of the following form results.

$$e_2 = E_2(1 + Sum \cos(\mu t - \nabla\Phi))\sin(\omega t + m\sin(\mu t - \nabla\Phi)) \quad (3)$$

This again is an amplitude modulated wave whose radio frequency envelope is

$$1 + Sum \cos(\mu t - \nabla\Phi)$$

Where S is the slope of the network frequency amplitude function and $\nabla\Phi$ is the incremental phase change. The factor of percentage modulation of the envelope is sum and since $m = \frac{\Delta w}{\mu}$ it is evident that the audio signal amplitude is proportional to the slope of the network frequency amplitude function and to the frequency deviation Δw. Since reconstruction of the output wave of equation 3 requires the combination of all the side bands of the frequency modulated wave, it is pointed out by Roder that, if the network is to serve as a perfect Frequency-amplitude converter, it must provide amplitude and phase characteristics which are both linear functions of frequency over the whole bandwidth occupied by the signal.

In practice Frequency modulation discriminators are usually designed to have a linear characteristic extending somewhat beyond the band pass of the receiver. The receiver selectivity tends to round off the characteristic at the extemities of the band pass and this design procedure ensures that the characteristic will have maximum possible range of linearity.

1.8.2 CONVERSION EFFICIENCY

The slope of the discriminator characteristic may be defined as the conversion efficiency of the discriminator. It has been pointed out that the modulation factor of the resultant amplitude modulated wave is directly proportional to this slope, so it is obviously desirable that the slope be as high as possible. Crosby has pointed out that low conversion efficiency is not serous provided there is no amplitude modulation at the input to the discriminator. (8) This is because the low conversion efficiency reduces the noise output in the same proportion as the signal, but only when no amplitude modulation is present. In general, therefore, with low conversion efficiency a greater degree of amplitude limiting is necessary to ensure that amplitude modulation noise does not become comparable with frequency modulation noise.

High conversion efficiency in frequency modulated transceivers is very important because of the control function of the discriminator. When a slow drift in the transmitter frequency takes place, the direct voltage component of discriminator output is applied as bias to the oscillator reactance valve and this has the effect of reducing the transmitter drift. The greater the bias which the discriminator can develop for a given off-frequency voltage, the smaller will be the net transmitter drift. The situation is similar to that which exists in negative feed-back applications and in general the same stability conditions apply. Too much frequency control cannot be used or the frequency variation of the transmitter will oscillate or " hunt ". In application to the transmitter control function, the problem is more likely to be one of too little control rather than too much and it is desirable that high conversion efficiency should be maintained.

1.8.3 DISCRIMINATOR BALANCE

In effecting maximum noise and amplitude modulation reduction, the symmetry and balance of the frequency modulation discriminator are very important. One of the original methods used in the detection of frequency modulated signals, was tuning the carrier to the centre of the slope of a resonant circuit characteristic. (3) This scheme had many disadvantages. Its linearity and conversion efficiency were poor. A signal above the resonant frequency would be received as well as one below. Further, if the signal was not completely limited, amplitude modulation would give rise to a variation in the detector output.

Discriminators of this nature, based simply on the slope of some filter characteristic, have the disadvantage that an output can exist when no frequency modulated signal is applied. This means that the output is not symmetrical with respect to noise voltages above and below the centre frequency, and if amplitude limiting is not complete, residual amplitude modulation will produce variations in the detector output.

For best results the discriminator should be arranged to have zero output at the centre frequency and equal outputs above and below the centre frequency for equal frequency departures. With this arrangement, the rejection of amplitude modulation at the centre frequency will be complete. Further equal noise voltages applied simultaneously at equal off-tune frequencies will tend to cancel each other.

1.8.4 STABILITY

It is important that the discriminator in portable transceivers be highly stable with respect to the centre frequency output voltage. It is obvious that a change in the centre frequency output voltage from the design value will not only reduce the distortionless working range of the discriminator but will also produce an undesirable shift in the transmitter frequency. Since the transceivers are normally required to work over a wide range of temperatures, consideration must be given to temperature compensation and this should be done with as few extra components as possible. This means that the temperature compensation problem will be simpler if the number of tuned circuits associated with the detection is kept to a minimum.

1.8.5 ALIGNMENT

The adjustment and alignment of the discriminator should be simple. As a general rule the number of adjustments required will vary as the complexity of the circuit used, which again argues in favor of maximum simplicity. It must be recognized the necessity of working with small components in confined spaces will tend to exaggerate the normal difficulties of adjustment and alignment.

1.8.6 POWER CONSUMPTION

Some consideration should be given to the power consumed by the discriminator. Although the total power so involved will be a small proportion of the total transceiver power, it is nevertheless of some consequence in battery operated sets. Methods of detection which employ several tubes are obviously undesirable.

1.8 [7] SUMMARIZATION OF REQUIREMENTS

Summarizing the above discussion it appears that the required discriminator features may be listed.

1. The discriminator should provide both an audio frequency output for the receiver and a direct voltage component for the transmitter control.
2. The discriminator should have maximum linearity over the band pass of the receiver.
3. The conversion efficiency or discriminator slope should be as great as possible.
4. The reduction of noise and amplitude modulation should be maximum.
5. The stability over a large temperature range should be high.
6. The components should be as few in number and as small in size as as possible.
7. The number of alignment adjustments required should be a minimum.

With the discriminator requirements established in this rather general fashion we are now in a somewhat more favorable position to consider the different methods by which detection can be carried out.

2.0 AN HISTORICAL SURVEY OF FREQUENCY MODULATION DETECTION

2.1 FIRST CONSIDERATIONS OF FREQUENCY MODULATION

Prior to Armstrong's historic paper on frequency modulation (6), the theoretical analyses of frequency modulation as a means of communication were either unfavorable or noncommittal. In 1922 Carson had developed the frequency spectrum for a signal modulated in frequency, and had condemned all frequency modulation on the basis of the large bandwidth required. In 1931 Roder, comparing phase, frequency and amplitude modulation, recognized that it would be possible to expect a transmitter gain in a frequency modulated system since the transmitter could be operated with full output during modulation, an impossible condition in amplitude modulation. He felt, however, that the problems of reception of a frequency modulated signal made such a system impractical quite apart from consideration of bandwidth. (2)

In 1932 Andrew analysed a system of frequency modulation reception which has been mentioned before. The method consisted of tuning the midpoint of the slope of one of the receiver tuned circuits to the carrier centre frequency. He concluded that frequency modulation reception in this fashion would require about eleven times the power of an equivalent amplitude modulated system.(3)

The same problem was considered from an experimental and theoretical point of view by Chaffee in 1935. (4) He reasoned that the loss in receiver output entailed by mistuning one of the receiver stages in the manner indicated above, could be entirely offset by the fact that the transmitter could be adjusted for maximum output during modulation. He developed formulae for calculating the distortion to be expected from such a detection system but came to no particular conclusions as to the possible merit of using frequency modulation.

This was the general state of the subject before Armstrong's work.

2.2 ARMSTRONG'S CONTRIBUTION

It was reasoned by Armstrong that it should be possible to introduce into a transmitted wave a characteristic not found in random noise such as atmospheric noise, tube shot noise or conductor thermal noise. If then a receiver could be employed which would not be responsive to the random noise effects but would be fully responsive to the special characteristics of the transmitted wave, a substantial improvement in signal to noise ratio should be effected. Since random noise is primarily an amplitude disturbance, with no orderly frequency variations, he argued that a system of frequency modulation, employing large frequency swings, would provide the desired results if the receiver could be made to respond to changes in frequency but not to changes in amplitude.

This was an entirely new concept of the use of frequency modulation. Before Carsons paper on the spectrum distribution, frequency modulation had been proposed as a possible means of narrowing the bandwidth occupied by the transmitted wave. Carson had demonstrated the fallacy in this argument by showing that the bandwidth could not be less than twice the highest modulation frequency. Armstrong now proposed to use much greater frequency swings than had originally been contemplated and to move the transmissions into the ultra high frequency band where the greater bandwidth could be tolerated. The scheme was greeted with considerable doubt at first but eventually rigorous theoretical analysis and extensive field trials proved the effectiveness of frequency modulation when used as Armstrong had proposed. (8,13, 15)

The requirements for distortionless frequency modulation reception were well known before Armstrong's work but he appears to have originated the use of amplitude limiting before the trnaslation of frequency to amplitude modulation.

Since some of the discriminators which we will study have inherent self limiting properties which can reduce by one or two the number of stages required in the receiver, We should consider the methods by which limiting must otherwise be accomplished and the reasons for it.

2.3 AMPLITUDE LIMITING

It is sometimes supposed that the limiter stage in frequency modulation receivers is the factor which reduces the interference through clipping of the noise peaks. It has been found however that noise peaks which exceed the signal will be just as noisy in frequency modulation receivers as in amplitude modulation receivers. In fact Landon has shown that in the case of impulse noise, limiting can make the noise worse. (20)

The true function of the limiter is to strip the signal of amplitude modulation and allow the frequency modulated signal to do its job. Goldman (18) has pointed out that the real reason for the effectiveness of frequency modulation in reducing interference lies in the peculiar distribution of the frequency modulation sidebands and in the selective circuit, which is the discriminator and its associated audio system. The combination of the carrier and its sidebands has the property that, when applied to a slope detector, it can produce extremely large frequency shifts at a low audio frequency rate as compared with what can be produced by a random distribution of noise carrier and sidebands of the same energy. The slope detector and its audio amplifier thus effectively selects only those signals with the proper distribution of carrier and sidebands. The limiter by removing the amplitude modulation from the signal makes this selection process possible without distortion. The most widely used method of effecting amplitude limiting has been through plate and grid limiting, either separately or in combination.

2.3.1 PLATE LIMITING

Plate limiting is carried out by operating a stage of intermediate frequency amplification with low plate and a screen voltages. The tuned circuit in the plate ensures that symmetrical limiting of the positive and negative peaks will be obtained when plate current saturation occurs.

Plate limiting has the disadvantage (20) that excessive grid current may flow during periods of high signal level. This causes detuning of the previous tuned circuit and results in variable receiver selectivity. A further disadvantage is the fact that the low plate impedence caused by plate current saturation lowers the effective Q of the plate tuned circuit. If the tuned circuit happems to be the primary of a discriminator circuit, this effect can be quite detrimental particularly in the case of narrow band discriminator.

2.3.2 GRID LIMITING

Grid limiting is accomplished by using a valve with normal voltages except for the grid bias. The bias is obtained through the use of a grid leak resistor, and is thus proportional to the signal strength. Landon has shown that the relatively high values of grid resistance necessary to obtain limiting with low signal levels has a detrimental effect on impulse noise. The extra bias built up by a noise pulse has the effect of blocking of the receiver for a short time unless the resistor-condenser time constant is fast enough to follow the pulsation. Grid limiting has however some important advantages when compared with plate limiting. There is little or no detuning of the grid tuned circuit with consequent loss of receiver selectivity. A further advantage of the system is the fact that the grid resistor is a convenient source of receiver automatic volume control voltage since the bias developed is directly proportional to the signal level.

For best limiter action Landon has recommended a combination of plate and grid limiting and this practice had been widely used in commercial receivers. The use of two stages of limiting has also been found desirable, and makes possible a gain of about five in the limiter stages. The degree of amplitude limiting should not in general be greater than about 20 db per stage, if amplitude distortion is to be avoided. The limiter must provide in its limiting condition sufficient output to drive the audio amplifier. If the degree of limiting is too great it is evident that for small frequency deviations, the output will be insufficient and amplitude distortion of the audio will result.

2.3[3] DYNAMIC LIMITING

Recently another form of limiting known as dynamic limiting has been introduced into frequency modulation receivers. (44) For dynamic limiting, a crystal rectifier in series with a parallel resistor-condenser combination is placed across one of the receiver tuned circuits. The rectifier conducts on peaks of radio frequency voltage and the resultant damping of the tuned circuit produces a constant voltage across the tuned circuit. The time constant of the rectifier load is made long with respect to the lowest expected amplitude modulation frequency and a variable threshold device thus results, which adjusts itself to the average signal amplitude level.

Tests on the dynamic limiter in comparison with a two stage grid bias limiter have indicated that the dynamic limiter has an amplitude modulation reduction factor about six to ten db better than the grid bias limiter for signal levels below the threshold of the grid bias limiter. Above its threshold level however the grid bias limiter is capable of considerably greater amplitude modulation reduction. The principal advantage of the dynamic limiter appears to be its very low threshold level since a significant amount of receiver quieting can be produced at a signal level about half that required by the grid limiter.

2.4 THE ARMSTRONG AMPLITUDE DISCRIMINATORS

The first useful systems of converting frequency modulation to amplitude modulation designed for that purpose, were used by Armstrong. While no longer in use these systems embody principles which are being used in other frequency modulation detector systems and are at least of some historic importance.

The first scheme employed by Armstrong is illustrated in Fig. 2. It may be seen that the input circuit consists of two parallel series tuned circuits. The circuits are tuned to resonance at the extremes of the frequency band over which linear detection is desired. The series resistances are chosen sufficiently high to make the current in each branch constant over the frequency band and the voltages developed across the two combinations are then proportional to the branch reactances. The variation of reactance in these circuits is not linear with frequency but by the proper choice of components and the use of shunts of high resistance and reactance (not shown in the schematic) a characteristic of good linearity can be obtained. The voltages across the two reactances are applied to equal linear amplifiers and rectifiers. The rectifiers are in series with equal output transformers whose secondaries are so arranged that frequency changes produce additive secondary voltages.

The method has of course been superseded by systems of greater simplicity, conversion efficiency and linearity but the principle of combining the frequency amplitude characteristics of two tuned circuits to obtain a linear balanced frequency amplitude characteristic has been widely used in frequency modulation detection.

2.5 THE ARMSTRONG PHASE DISCRIMINATOR

Another type of discriminator which was used by Armstrong in his tests on frequency modulation is shown in Fig. 3. In this circuit the input circuit is again series tuned in series with a resistance. The series circuit in this case is tuned to the centre frequency and the phase of the voltage across the combination undergoes 180 degrees of phase shift with respect to the current through it as the frequency passes through the resonant point.

Part of the input signal is impressed on the resiston R_1 and this is shifted in phase by ninety degrees, amplified and applied to the screen grids of the tetrode tubes in opposite phase. The signalling and heterodyne voltages are then in phase in one mixing tube and 180 degrees out of phase in the other. For variable signal frequency this results in a rectifier current characteristic which is linear with respect to frequency over the working range. The outputs are again cumulatively combined for frequency changes.

The discriminator requires careful balancing of the heterodyne voltages and is in general too complex to have found widespread use in frequency modulation detection. The principle of mixing two voltages 180 degrees out of phase has however been used in other discriminator circuits.

2.6 THE TRAVIS DISCRIMINATOR (5, 23, 27)

In 1935 Travis had described a system of frequency control designed to keep the carrier of amplitude modulated signals centred in the receiver intermediate frequency band. The system used an auxiliary tube across the receiver local oscillator tuned circuit which acted as an effective reactance of a value dependent on the grid bias of the tube. To control this action Travis used a device to produce bias voltages dependent on receiver intermediate frequency which he called a frequency discriminator. The Travis discriminator is shown in Fig. 4 appendix.

The composite primary circuit of the Travis discriminator is tuned to the centre of the receiver intermediate frequency band, and is heavily damped to prevent coupling between the secondaries. The two secondary circuits are loosely coupled to the primary and tuned to different frequencies, one above and one below the receiver mid-frequency. The secondaries act very much like isolated tuned circuits and the voltages impressed on the rectifier circuits thus maximize at equal departures from the mid frequency. Since the rectifier outputs are differentially connected an S shaped curve of D.C. voltages versus frequency results.

The Travis circuit was intended for use as a automatic frequency control auxiliary and as such was designed to have peaks about five kilocycles apart and to have maximum possible conversion efficiency. Following Armstrong's work it became apparent that if the circuit values and tuning frequencies could be adjusted for good characteristic linearity over the frequency modulation receiver band, the devide should also function satisfactorily as a frequency modulation detector. The Travis circuit or modificationsof it have been widely used in frequency modulation for controlling the transmitter centre frequency in a manner previously indicated and in frequency modulated receivers.

The circuit as proposed by Travis has a number of disadvantages. It requires careful tuning of three tuned circuits if good linearity and balance is to be achieved. Further, stray capacitive coupling between the primaries or the secondaries can cause considerable unbalance of the peaks in relation to the base frequency.

A modification of the Travis circuit which has been used in a number of cases is shown in Fig. 5. This is sometimes known as the Crosby Discriminator. The primary circuitsare in series. To offset the capacity coupling between the secondary windings a small amount of mutual inductance coupling is introduced between the windings in the proper sense to balance out the capacitive coupling. It is evident that the problems of alignment and coupling adjustment are by no means diminished by this arrangement. (27)

Another variation of the Travis circuit uses an untuned primary winding. This leads to greater alignment simplicity but the conversion efficiency is lower than for the Travis or Crosby circuits.

2.7 THE FOSTER - SEELEY DISCRIMINATOR (7,10,19,24,28,35)

In 1937 Foster and Seeley described a discriminator for use in connection with automatic frequency control systems. As was the case with the Travis circuit, the Foster-Seeley discriminator has found more widespread use in connection with frequency modulation transmitter control and reception than in the amplitude modulation receivers for which it was originally intended.

The Foster-Seeley circuit for automatic frequency control is illustrated in Fig. 6.

The action of the discriminator depends on the difference in phase of the primary and secondary potentials of a tuned transformer. When the primary and secondary circuits are tuned to the same frequency, a 90 degree phase difference exists between the primary and secondary voltages when the resonant frequency is applied. The voltages between the centre tap of the secondary and each end of the secondary coil are 180 degrees out of phase. The vector sum of the primary voltage and half the secondary voltage is applied to each diode circuit and at the resonant frequency since vector sums are equal no output is obtained from the differentially connected rectifiers. The vector condition is shown in Fig. 6a.

When the frequency of the signal departs from the resonant frequency, the phase relation between the primary and the secondary voltages changes. The Vector sum of the primary voltage with one of the secondary half voltages may then become greater or less than for the resonant condition depending upon the direction of the phase shift which in turn is dependent on whether the frequency is higher or lower than the resonant frequency. The vector relationships are shown in Figs. 6b and 6c for the two off resonant conditions.

The outputs from the rectifier circuits will thus maximize, one a bove and one below, the resonant frequency and since they are differentially connected an output frequency voltage characteristic is obtained as shown in Fig. 6d .

The exact shape of the discriminator characteristic is dependent upon the effective circuit Q's, the coupling between the tuned circuits, and the ratio of secondary and primary voltages. For automatic frequency control purposes, where linearity is not essential, loose coupling is generally used and the circuit Q's are kept very high. For this reason the choke coil L_3 is generally required to prevent excessive loading of the primary tuned circuit. For wide band frequency modulation reception the degree of coupling between the circuits is considerable higher than in the automatic frequency control case, and the circuit Q's are lower. The choke coil L_3 may be omitted to permit more reduction of the effective primary Q and a single load condenser is used, as shown by the dotted line, to prevent effective grounding of the radio frequency primary voltage .

The Foster-Seeley circuit has the advantage that only two alignment adjustments are necessary once the degree of coupling has been fixed. Since both the tuned circuits operate at the same frequency the alignment is further simplified.

The Foster-Seeley circuit requires rather elaborate precautions at times to eliminate electrostatic coupling between the primary and secondary windings which can adversely affect the linearity and balance of the circuit. (35)

When the output circuit is grounded on one side, the capacitive unbalance of the diodes to ground may also result in some unbalance of the characteristic.

A modification of the Foster-Seeley circuit which has been used in mobile F. M. equipment (21) is shown in Fig. 7. The coupling between the tuned circuits ishere high impedance capacitive coupling rather than inductive coupling. The capacity C_1 is made variable to effect a good balance. Coupling systems of this nature will be referred to in more detail in a later section.

2.8 THE COUNTER-DISCRIMINATOR (4, 27, 41)

Frequency modulation detection can be accomplished by means of a counter circuit which extracts the modulation frequency directly from the frequency modulation wave. Such a circuit is described in a paper by Seeley, Kimball and Barso (4) and is shown in Fig. 8.

Referring to Fig. 8 it may be seen that the last intermediate frequency tube has no tuned circuit and is subject to grid and plate limiting. During negative grid swings the plate potential of the tube rises to a fixed level equal to the supply voltage. On positive grid swing, plate current saturation takes place fixing a lower limit to the change in plate potential. The output wave is thus effectively squared between two definite fixed potentials. The "counter" condenser C. thus charges to plate supply level during the negative grid swings and discharges through the second diode and its load resistor on positive grid swings until its voltage is equal to the minimum plate potential. Since a single pulse of current is obtained through the load resistor of the second diode for each cycle, the average output current must be directly proportional to the frequency and therefore will reproduce the modulation content of the F. M. wave. The counter circuit is very simple in operation and requires no tuned circuits or critical alignment procedure. The distortion due to non linearity is said to be less than 0.02 percent over a bandwidth of 200 kilocycles

The disadvantage of the counter circuit is the necessity for double conversion in the receiver before the counting can be carried out. The frequency of the F. M. signal is limited by the time of discharge of the counter resistor- condenser combination which must be considerably less than the half period of the maximum frequency modulated signal frequency. This means that conversion of the signal frequency into the region of 100 to 300 kilocycles per second is required before distortionless counting can be carried out. To avoid spurious receiver responses it is then necessary to use double frequency conversion with considerable application of the design.

The conversion efficiency of the counter discriminator is quite low compared with the Foster Seeley and Travis circuits and it is necessary to employ considerable audio gain in the receiver. It follows also that the direct current voltage available for frequency control purposes will be low.

It should be recognized that the counter discriminator will produce a direct voltage output when an unmodulated signal is applied and in automatic frequency control systems this would require balancing out with some form of bias.

2.9 THE RATIO DETECTOR

During 1945 a balanced discriminator known as the ratio detector was developed in the RCA laboratories. The ratio detector was designed to be insensitive to amplitude modulation without the necessity for preliminary limiters. The circuit for the Seeley ratio detector is shown in Fig. 9

The ratio detector may be seen to resemble the Foster-Seeley discriminator in many respects. In particular the phase relationships between the voltages applied to the rectifier circuits are the same. It may be seen hpwever that a tertiary winding,having the same phase as the primary, is used to couple the primary voltage to the secondary centre tap. Further the diodes in the ratio detector are in series, as opposed to the differential arrangement in the Foster-Seeley, and a large condenser is shunted across the load resistances. The change in the signal take off point should also be noted.

The action of the ratio detector in removing amplitude variations can be explained simply. Because the diodes in the circuit are in series, the same rectified current flows through the load resistors and the ratio between [illegible] the rectified voltages at any particular frequency will be constant regardless of variations in the signal amplitude. If the sum of the rectified voltages is also held constant, with respect to rapid changes in amplitude, it follows that the rectified voltages must themselves remain constant. The audio output of the discriminator, which is proportional to the difference in the rectified voltages, must also remain constant and the discriminator is thus insensitive to changes in amplitude of the input signal.

It is the function of the large condenser across the diode load resistors to hold the sum of the rectified voltages constant with respect to rapid changes in signal amplitude. This condenser, with the diode load resistor has a time constant of the order of a 0.2 seconds and the voltage across it will thus vary slowly in accordance with changes in the average signal level. The condenser could be replaced by a battery but this would have a fixed threshold level below which the signal would not be strong enough to operate the discriminator. The condenser permits the voltage across the load to vary in proportion to the average signal level and thus automatically adjust to the optimum level.

A limiting factor in the ratio detector is the amount of downward amplitude modulation which may be tolerated. If the signal level falls to too low a value the condenser voltage will bias off the rectifiers and the detector will become inoperative. To ensure that this condition occurs only at very low input signals, it is necessary to use high values of secondary unloaded Q and low values of load resistance which will lower the operating Q to about one quarter of the unloaded Q. During the periods of low signal level, then, when the diodex conduction current is low, the secondary Q can rise, effectively increasing the discriminator sensitivity.

The low values of load resistance required as explained above is the factor which makes the tertiary winding necessary. If no impedence matching of this nature were used, the effect of diode conduction current damping would be to seriously reduce the effective primary Q and therefore the discriminator sensitivity.

The ratio detector is said to be capable of handling without distortion signals with as much as fifty percent amplitude modulation. To do this however the diode load resistors must be kept very low and conversion efficiency suffers. It should also be noted that the diode series arrangement results in an inherent six db loss over the Foster-Seeley arrangement since the audio output is only half the difference between the rectified voltages.

It has been found that the ratio detector requires considerably more careful attention to tolerances than the Foster-Seeley discriminator and grid limiter combination. The effective coupling between the primary and the secondary must be closely controlled since it may be shown to affect the degree of amplitude modulation rejection. Further any unbalance in the diodes or load resistances will result in an unbalanced amplitude modulation component in the output. These considerations are quite apart from the linear frequency-amplitude problems.

The discriminator alignment is no more complicated than for the Foster-Seeley but there is the added coupling adjustment required through the tertiary winding.

It should be noted that the audio output level of the ratio detector will vary with the average signal input and the ratio detector requires that automatic volume control be applied to the receiver intermediate frequency section. Such A. V. C. voltages can conveniently be taken off the voltage in stabilizing condenser.

2.10 THE BRADLEY F. M. DETECTOR (31)

The Bradley F. M. detector appeared in 1946 as a further effort to provide a satisfactory self limiting discriminator. The circuit of the Bradley discriminator is shown in Fig. 10.

The Bradley detector operates around a special heptode tube designed for high transconductance, sharp cutoff and maximum sheilding of the input signal grid from other tube elements.

The first three elements of the tube with their associated tuned circuits operate as a class C oscillator, with space current consisting of very short pulses. The proportion of the space current which reaches the plate circuit is dependent upon the potential of the input signal grid. The fundamental of the resulting plate current is fed back to the oscillator through a phasing network which is so adjusted that variations in the plate current cause pure frequency modulation of the oscillator.

The phase of the input signal with respect to the oscillator, by affecting the pulse current magnitude varies the oscillator frequency causing it to look to a fixed phase relation with the input signal. Since the oscillator frequency is directly proportional to the fundamental component of the plate current it follows that the mean plate current must vary linearly with the frequency deviation of the input signal.

It is then possible to extract the audio signal component from the plate circuit by a simple low pass filter.

The Bradley detector is stated to have a response to amplitude variations about fifty db less than to frequency change. This of course holds only for those signals which are sufficiently strong to cause the oscillator to lock in, but it is claimed that the required voltage is no more than half a volt rms

The adjustment of the circuit appears to be somewhat critical requiring careful adjustment of the phasing network. The degree of amplitude modulation rejection is affected by stray coupling capacity between the input signal and the oscillator and this may require a neutralizing circuit particularly if heptode valves of less specialized design must be used.

It may be noted that the detector output is taken off at a high direct voltage level and could not be used directly for automatic frequency control purposes.

2; 11 INTEGRATOR DETECTORS

The Bradley detector mentioned above appears to be a special refinement of a class of detectors which Sturley has referred to as " integrator detectors" (27)

The integrator detector which is self limiting may be a regenerative amplifier, a multivibrator or a blocking oscillator which is triggered by the intermediate frequency.

All of these circuits produce short pulses of plate current whose repetition rate is controlled by the input signal, but whose amplitude is dependent only on the circuit constants. The mean plate current in each case is then proportional to the signal frequency, varies with the frequency deviation, and in conjunction with a low pass filter will reproduce the audio signal.

Such circuits have not found widespread use either because of excessive complexity or because of the difficulties of adjustment. The problem of extracting a direct voltage component at the low levels required for automatic frequency control has also made them unsuitable in most cases for that purpose.

2.12 THE CBC DETECTOR (42)

An interesting circuit employing the phase principle in combination with a heptode tube has recently been developed in the CBC Laboratories. The circuit for this detector is shown in Fig. 11.

Signal voltages, derived from the primary and secondary of an intermediate frequency transformer, are fed to what normally constitute the signal and oscillator grids of a heptode valve. The phase angle between these voltages is 90 degrees at the resonant frequency and changes rapidly in the neighborhood of resonance.

Multiplication of the voltages in the mixer tube results in a product plate current term proportional to the cosine of the phase angle. For phase shifts in the neighborhood of ninety degrees, then, the plate current component will vary almost linearly with frequency. The audio signal may then be obtained from the output of a low pass filter in the plate circuit.

The discriminator is said to provide 50 volts of audio signal for 75 kilocycles deviation. Limiting is required prior to detection.

The required phase shift at resonance may also be obtained by tuning the primary of the transformer to parallel resonance at the centre frequency and tuning the secondary, in series with a small capacity, to series resonance. This is said to minimize the unbalancing effect of stray capacity between the mixer grids.

2.13 RECAPITULATION

It is believed that this summary covers most of the methods which have been used in the detection of frequency modulated signals. It appears that a wide choice is available to the design engineer and probably most of the discriminators reviewed here could be udes in a portable transceiver if modifications were made, where necessary, to obtain an automatic frequency control voltage. Practical consideration, however, will serve to rule out those methods which are least suitable for our purpose.

The Bradley detector requires a special heptode tube a large number of components, and is somewhat critical of adjustment. More important, there is no simple means of obtaining a suitable low level automatic frequency control voltage. The same considerations apply to the CBC detector. The integrator detectors may be ruled out on the basis of excessive complexity or critical adjustment.

The ratio detector offers attractive self limiting properties but requires two coupling adjustments and a high degree of electrical balance which may not be obtainable with subminiature components. Its performance in the presence of severe multi-path fading is also suspect.

The counter discriminator has great simplicity and requires no tuned circuits. To obtain a constant output at the centre frequency, however, requires a constant supply voltage since the averaged load current is dependent upon the supply. Further double frequency conversion is necessary in the receiver which offsets much of the simplicity of the counter circuit.

The choice narrows down to the Foster-Seeley and Travis discriminators, and the greater simplicity of adjustment of the Foster-Seeley appears to swing the balance in its favor.

2.14 FURTHER COUPLING CONSIDERATIONS OF THE FOSTER-SEELEY CIRCUIT

It has been pointed out that the Foster-Seeley circuit is subject to some unbalance due to electrostatic coupling between the primary and secondary circuits. Capacitive unbalance of the rectifier circuits can also adversely affect the performance.

The elimination of primary-secondary electrostatic coupling has been accomplished in practice by the interposition of a grounded electrostatic shield between the windings. (35) In a small plug-in unit with all the components in close proximity to a grounded shield, the elimination of this coupling, while adjusting for correct inductive coupling is apt to be more difficult. Since a certain amount of capacitive coupling appears to be inevitable, it would seem that the use of capacitive coupling without inductive coupling might have some advantages.

One advantage of capacitive coupling is the possibility of using permeability tuning. While permeability tuning may be used with inductively coupled circuits, the circuits cannot be tuned without changing the effective coupling and the problem of obtaining the correct coupling at the tuned frequency complicates the design considerably. Further, if a centre tapped secondary is used, a special "bifilar" construction is required so that the coupling of each half secondary to the primary will be the same for any given position of the tuning slug. (29)

It is generally more difficult to produce a small variable capacity with a lower temperature coefficient than a variable inductance. Pressure type mica dielectric capacitors show appreciable increase in capacity with temperature, and air dielectric capacitors, with ceramic insulators tend to be considerably larger than variable inductance providing the same reactance change. Permeability tuning is thus seen to have some important advantages.

Capacitive coupling has been used in intermediate frequency amplifiers (23, 34) and in discriminator circuits (21) but the literature does not contain a great deal of information on its application.

The capacitive coupling between two circuits may be series capacitance (high impedence) or shunt capacitance (low impedance). It has been shown that the transmission characteristic for two simple coupled circuits tuned to the same frequency may maximize at two frequencies when the coupling exceeds a certain critical value. When the coupling is simple mutual inductive coupling the two peaks are symmetrically disposed with respect to the tuned frequency and move outward symmetrically as the coupling is increased. For band pass circuit design this is a very desirable characteristic.

When capacity coupling is used, as coupling is increased beyond the critical value, one of the peaks remains on the tuned frequency while the other shifts higher or lower depending upon the nature of the coupling. With series coupling, it has been shown (45) that the shifted peak is lower in frequency and the peak separation increases with increase in the coupling component. With shunt capacity coupling, the shifted peak is higher and peak separation increase with a decrease in the coupling component.

It is evident that if we try to apply capacitive coupling to the Foster-Seeley circuit we can expect some asymmetry of the discriminator characteristic with respect to frequency. This unbalance must be evaluated and made negligible over the discriminator working range if satisfactory operation is to be obtained

Series capacitive coupling has been used in connection with phase discrominators (21) but very little design information is given. The writer has not found any indication that low impedence capacity coupling has ever been used in this connection. The coupling condenser required for low impedence systems is much larger in capacity than that required for high impedence coupling but need not be much greater in a physical sense. Further, the use of a large coupling capacity should make it possible to minimize the effects of stray capacity on the effective coupling coefficient of the circuit. Any changes in the distributed capacity associated with a small series coupling condenser should result in a much greater change in effective coupling coefficient than would occur if a large coupling capacity were used.

It was decided to experiment with a low impedence coupled circuit arranged to provide the Foster-Seeley voltage combinations. Referring to the Foster-Seeley circuit, it will be recalled that the radio frequency voltage applied to each rectifier circuit was the vector sum of the primary voltage and half the secondary voltage of a tuned coupled circuit. Such a voltage combination can be approximated very closely by the circuit shown in the following section which employs permeability tuning.

3.0 A LOW IMPEDENCE CAPACITY-COUPLED PHASE DISCRIMINATOR

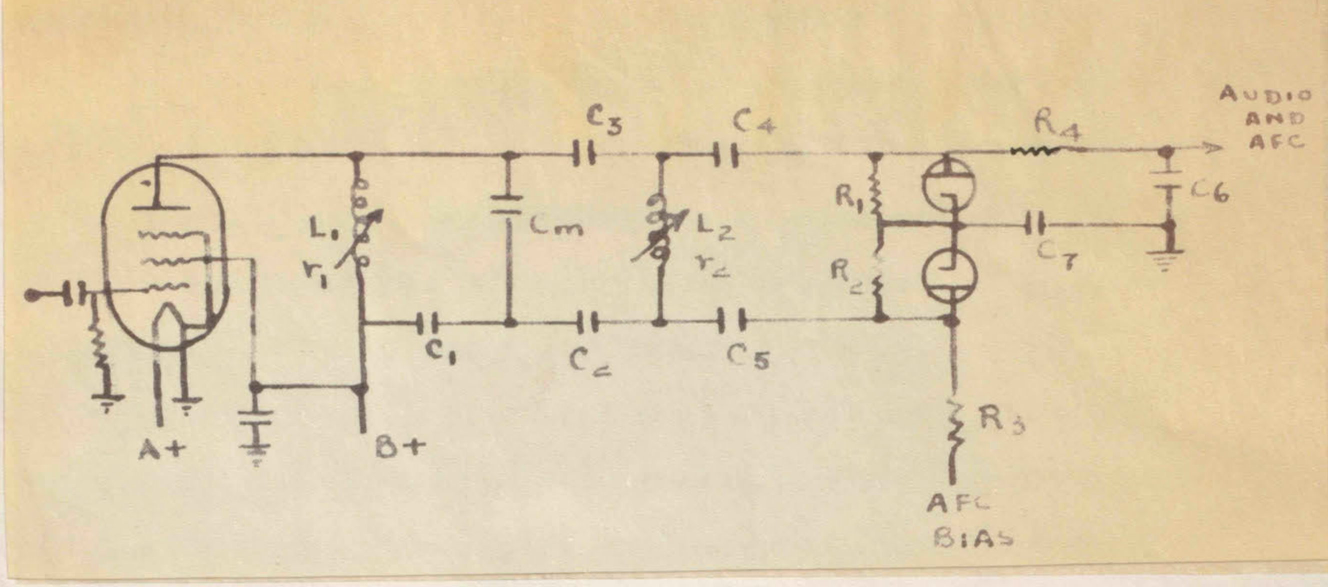


Fig. 12

3.1 DESCRIPTION OF THE CIRCUIT

In this circuit the capacity c_m is much greater than C_1, C_2, or C_3. The primary voltage across the coil L_1 is then effectively the same at the points A and B. If C_2 and C_3 are approximately equal the voltage appearing across each of them will then be substantially equal to half the secondary voltage. The voltage applied to each rectifier circuit will then be effectively primary voltage in series with half a secondary voltage, the capacity C_7 providing a low impedance return for the high frequency voltage.

In considering phase relations it may be seen that the voltage drop across C_m due to primary will be in phase with the voltage across C_1. This voltage is effectively coupled into the secondary circuit and at resonance the secondary current will be in phase with it. The voltages across C_2 and C_3, lagging secondary current, will thus be 90 degrees out of phase with the primary voltage at resonance. As the frequency departs from resonance, the secondary current is no longer in phase with the coupled primary voltage and the voltage across C_2 and C_3 will thus depart from the 90 degree phase relation with the primary voltage.

The voltage combinations and phase relationships are thus seen to approximate closely those found in the Foster-Seeley discriminator.

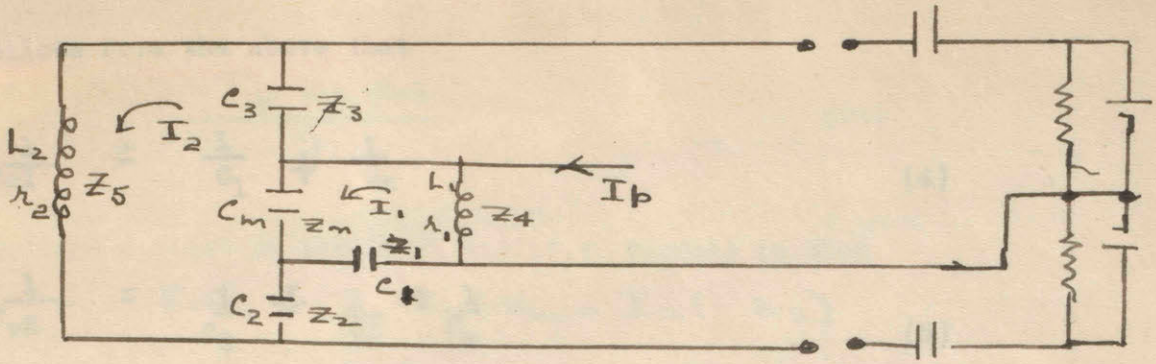
It will be noticed that shunt loading of the discriminator rectifiers is employed. Series loading as used in the Foster-Seeley circuit requires the provision of a secondary direct voltage centre tap on the coil. That makes the use of permeability tuning difficult, as has been explained. Shunt loading raises other difficulties which will be mentioned later.

The condensers C_4 and C_5 are the load condensers andas in the more conventional second detector circuits operate to maintain high detection efficiency discharging through the load resistors on negative peaks of the radio frequency voltage when the rectifiers are not conducting.

Bias for the reactance tube of the automatic frequency control circuit may be applied through R_3 and the load resistors as shown and the direct voltage component of the discriminator circuit will be effectively added in series with it. R_4 and C_6 may be used as a de-emphasis circuit if desired. This will be discussed in more detail later.

3.2 DETERMINATION OF THE OUTPUT EQUATIONS

To analyze the circuit it is convenient to assume that the loading effects of the driver tube and the rectifier circuits are taken into account by modification of the effective coil Q factors. The load and plate circuit impedances may then be assumed to be infinite for purposes of simplifying the circuit. In the final analysis of course the relation between the unloaded and loaded circuit Q's must be determined.



3.2.1 Nomenclature

The nomenclature used is as follows

C_1 Primary condenser

C_m Coupling Condenser

C_2 ,C_3 Seocondary condenser

L_1 Primary Tuning inductance

L_2 Seocondary tuning inductance

Q_1 Effective Primary Circuit Q

Q_2 Effective Secondary circuit Q

r_1 Effective series resistance of the primary circuit including the effects of driver tube and output cif-cuit loading

r_2 Effective secondary series resistance including the effects of output circqit loading

F Frequency of the applied signal

ω Angular Frequency of the applied signal 2nF

ω_0 Resonant angular frequency of the primary and secondary circuits

C_{e1} Effective primary tuning capacity

C_{e2} Effective secondary tuning capacity

It follows from the above that

$$\frac{1}{C_{e1}} = \frac{1}{C_1} + \frac{1}{C_m} \qquad (4)$$

and

$$\frac{1}{C_{e2}} = \frac{1}{C_2} + \frac{1}{C_3} + \frac{1}{C_m} \qquad (5)$$

$$w_o^2 = \frac{1}{L_1 C_{e1}} = \frac{1}{L_2 C_{e2}} \qquad (6)$$

$$Q_1 = \frac{W_o L_1}{r_1} = \frac{1}{W_o C_{e1} r_1} \qquad (7)$$

$$Q_2 = \frac{W_o L_2}{r_2} = \frac{1}{W_o C_{e2} r_2} \qquad (8)$$

3.22 DERIUATION OF THE RECTIFIER EQUATIONS

The mesh equations for the system may be written down, the impedances and meshes being numbered as in the simplified circuit.

From mesh 1 we have

$$I_1, (z_1 + z_m + z_4) + I_2 (-z_m) = I_p z_4 \qquad (9)$$

From mesh 2

$$I_1 (-z_m) + I_2 (z_2 + z_3 + z_5 + Z_m) = 0 \qquad (10)$$

The determinant of the system is given by

$$D = (z_1 + z_4 + z_m)(z_2 + Z_3 + z_5 + z_m) - Z_m^2 \qquad (11)$$

And the currents may be written

$$I_1 = \frac{I_p z_4 (z_2 + z_3 + z_5 + z_m)}{D} \qquad (12$$

$$I_2 = \frac{I_p Z_4 Z_m}{D} \tag{13}$$

The voltage applied to the first rectifier circuit is then

$$E_{d_1} = I_1 Z_1 + (I_1 - I_2) Z_m + I_2(-Z_3)$$

$$= I_1(Z_1 + Z_m) - I_2(Z_m + Z_3) \tag{14}$$

and to the second rectifier circuit is

$$E_{d_2} = I_1 Z_1 + I_2 Z_2 \tag{15}$$

Substituting for the currents I_1 and I_2 from (12 and (13) leads to

$$E_{d_1} = \frac{I_p Z_4}{D}\left[(Z_1 + Z_m)(Z_2 + Z_3 + Z_5 + Z_m) - Z_m(Z_m + Z_3)\right] \tag{16}$$

and

$$E_{d_2} = \frac{I_p Z_4}{D}\left[Z_1(Z_2 + Z_3 + Z_5 + Z_m) + Z_2 Z_m\right] \tag{17}$$

Let us now set $z_m + z_3 = z_2$ in equation 16 which means in effect that we make

$$C_2 = \frac{C_3 C_m}{C_3 + C_m} \tag{18}$$

If we further assume that $Z_1 = Z_1 + Z_m$ in equation (17) which in practice will be nearly true since $C_m >> C_1$ then equations (16 and (17) become identical except for the sign between the bracketed terms

$$E_{d_1} = \frac{I_p Z_4}{D}\left[(Z_1 + Z_m)(Z_2 + Z_3 + Z_5 + Z_m) - Z_m Z_2\right] \tag{19}$$

$$E_{d_2} = \frac{I_p Z_4}{D}\left[(Z_1 + Z_m)(Z_2 + Z_3 + Z_5 + Z_m) + Z_m Z_2\right] \tag{20}$$

The rectifier voltages must now be evaluated in tems of the circuit components. Before doing this it will be useful to introduce an expression for the coefficient of coupling between the two circuits. The coupling coefficient for two simple low impedance coupled circuits is given in a general way by the following (47)

$$k = \frac{X_m}{\sqrt{X_1 X_2}}$$

Where X_m is equal to the reactance of the coupling, and X_1 and X_2 are the reactances, of the same nature as the coupling, by which the circuits are tuned to resonance. Application of this formula to the case we have here leads to

$$k = \frac{\sqrt{C_{e1} C_{e2}}}{C_m} \qquad (22)$$

The series impedance of the primary is given by

$$Z_1 + Z_4 + Z_m = r_1 + j\left(\omega L_1 - \frac{1}{\omega C_{e1}}\right)$$
$$= r_1 + j\,\omega_0 L\left(\frac{\omega}{\omega_0} - \frac{1}{\omega\,.\,\omega_0 L_1 C_{e1}}\right)$$

Using equation (6) this reduces to

$$Z_1 + Z_4 + Z_m = r_1 + j\,\omega_0 L_1\left[\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right]$$

It is convient to introduce a notation which has been widely used in circuit theory. We may write for the frequency term in the preceding equation.

$$\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} = \frac{(\omega - \omega_0)(\omega + \omega_0)}{\omega_0\omega} = \frac{\Delta\omega}{\omega_0}\frac{(\Delta\omega + 2\omega_0)}{\omega_0 + \Delta\omega} \cong \frac{2\Delta\omega}{\omega_0} = V \qquad (23)$$

The percentage error introduced by use of this notation is given by $-\dfrac{(f - f_0)}{f + f_0} \times 100$

and for $\Delta f =$ 100 kilocycles and

f_o = 4300 kilocycles the percentage error is about 1.15

We may then write for the primary impedance

$$Z_1 + Z_4 + Z_m = r_1 + j\omega_o L_1 V = r_1(1 + jQ_1 V) \qquad (24)$$

Similarly for the secondary impedence

$$Z_2 + Z_3 + Z_5 + Z_m = r_2(1 + jQ_2 V) \qquad (25)$$

For the impedance of the primary coil we may write

$$Z_4 = r_1 + j\omega L_1 = r_1\left(1 + j\frac{\omega_o L_1}{r_1} \cdot \frac{\omega}{\omega_o}\right)$$

Since the ratio of $\frac{w}{wo}$ is very nearly equal to one over a small range of frequencies we may then write

$$Z_4 \cong r_4(1 + jQ_1) \cong jr_1 Q_1 \cong jX_1 \qquad (26)$$

Where $x_1 = \frac{\omega_o L_1}{r_1} = \frac{1}{\omega_o C_{e1}}$

For the impedance term $Z_1 + Z_m$ we may write

$$Z_1 + Z_m = -j\frac{1}{\omega C_{e1}} = -j\frac{1}{\omega_o C_{e1}} \cdot \frac{\omega_o}{\omega} \cong -j\frac{1}{\omega_o C_{e1}} \qquad (27)$$

Another identity of some use is the following

$$\frac{Z_m^2}{r_1 r_2} = \frac{\left(-j\frac{1}{\omega C_m}\right)^2}{r_1 r_2} = -\frac{k^2}{\omega_o C_{e1} \cdot \omega_o C_{e2} \cdot r_1 r_2}\left(\frac{\omega_o}{\omega}\right)^2$$

Assuming as before that the ratio $\frac{wo}{w}$ is nearly equal to one

$$\frac{Z_m^2}{r_1 r_2} \cong -k^2 Q_1 Q_2 \qquad (28)$$

For the determinant D we have from equation (11)

$$D = (Z_1 + Z_4 + Z_m)(Z_2 + Z_3 + Z_5 + Z_m) - Z_m^2$$

Substituting equations (24),(25) and (28) in (11) we have

$$D = r_1 r_2 \left[(1 + jQ_1 v)(1 + jQ_2 v) + k^2 Q_1 Q_2 \right] \qquad (29)$$

Substituting (24),(27),(28) and (29) into equation (19) we obtain an expression for the voltage applied to the first rectifier circuit as follows.

$$E_{d_1} = I_p\, j x_1 \frac{\left[-j \frac{r_2}{\omega_0 C_1} (1 + jQ_2 v) - Z_m^2 \cdot \frac{Z_2}{Z_m} \right]}{r_1 r_2 \left[(1 + jQ_1 v)(1 + jQ_2 v) + k^2 Q_1 Q_2 \right]}$$

$$= I_p\, j x_1 \frac{\left[-jQ_1 (1 + jQ_2 v) + k^2 Q_1 Q_2 \cdot \frac{C_m}{C_2} \right.}{(1 + jQ_1 v)(1 + jQ_2 v) + k^2 Q_1 Q_2}$$

$$= I_p \frac{Q_1 x_1 \left[1 + j \left(Q_2 v + k^2 Q_2 \frac{C_m}{C_2} \right) \right.}{(1 + jQ_1 v)(1 + jQ_2 v) + k^2 Q_1 Q_2} \qquad (30)$$

Similarly substitution in equation (20) leads to

$$E_{d_2} = I_p \frac{Q_1 x_1 \left[1 + j \left(Q_2 v - k^2 Q_2 \frac{C_m}{C_2} \right) \right.}{(1 + jQ_1 v)(1 + jQ_2 v) + k^2 Q_1 Q_2} \qquad (31)$$

3.2.3 THE OUTPUT EQUATION

The voltage across each of the load resistors will be proportional to the detection efficiency of the rectifier circuit and the absolute value of the peak voltage applied. If the discriminator output be designated by the symbol Y we have

$$Y = \eta \left[|\overline{E_{d_1}}| - |\overline{E_{d_2}}| \right] \qquad (32)$$

Where $\overline{E_{d_1}}$ = peak value of E_{d_1} given by equation (30)

$\overline{E_{d_2}}$ = peak value of E_{d_2} given by equation (31)

η = rectifier detection efficiency.

η = rectifier detection efficiency.

It is evident that direct substitution of (30) and (31) into (32) will lead to a very complicated expression. It is possible to simplify (30) and (31) considerably by assuming equal primary and secondary Q factors.

Let $Q_1 = Q_2 = Q$ and define $Q_1 v = Q_2 v = Qv = x$ (33)

$$Qk = a \quad (34)$$

$$\frac{k C_m}{C_2} = p \quad (35)$$

Then we have the following expressions for the peak values of rectifier voltages

$$\overline{E_{d_1}} = \overline{I_p} \frac{Q X_1 (1 + j(x + ap))}{1 + a^2 - x^2 + j2x} \quad (30a)$$

$$\overline{E_{d_2}} = \overline{I_p} \frac{Q X_1 (1 + j(x - ap))}{1 + a^2 - x^2 + j2x} \quad (31a)$$

and the discriminator output becomes

$$Y = n \overline{I_p} Q X_1 \left[\frac{(1 + (x+ap)^2)^{\frac{1}{2}} - (1 + (x-ap)^2)^{\frac{1}{2}}}{x^4 + 2x^2(1-a^2) + (1+a^2)^2} \right]^{\frac{1}{2}} \quad (32a)$$

3.2.4 A MORE EXACT EXPRESSION FOR THE OUTPUT EQUATION

In deriving the above expression for discriminator output from equations (16) and (17) for E_{d_1} and E_{d_2}, two assumptions were made to derive the similar equations (19) and (20). In determining equation (19) it was assumed that $C_2 = \frac{C_3 C_m}{C_3 + C_m}$

This condition can be readily compiled with by the proper choice of components. In the case where $C_m >> C_3$ and C_2, little unbalance would be created by making $C_2 = C_3$.

In obtaining equation (20) from equation (17) it was assumed that $Z_1 = Z_1 + Z_m$ This is nearly true since $C_m >> C_1$ but the expression is nevertheless exact and the error is the basis for the inherent asymmetry of capacitive coupling. The error to be expected by using equation (20) in place of (17) can be evaluated in terms of frequency and the circuit Q. Writing equation (17) again we have.

$$E_{d_2} = \frac{I_p Z_4}{D}\left[Z_1(Z_2 + Z_3 + Z_5 + Z_m) + Z_2 Z_m\right]$$

This may be written

$$E_{d_2} = \frac{I_p Z_4}{D}\left[(Z_1 + Z_m)(Z_2 + Z_3 + Z_5 + Z_m) + Z_2 Z_m\right]$$

$$- \frac{I_p Z_4}{D} Z_m(Z_2 + Z_3 + Z_5 + Z_m) \quad (17a)$$

The first term in this expression is the value of E_{d_2} as given by equation (20). The second term is

$$- \frac{I_p Z_4 Z_m (Z_2 + Z_3 + Z_5 + Z_m)}{D} \quad (36)$$

$$\cong - \frac{I_p (j x_1)(-j \frac{1}{\omega C_m})(r_2(1 + jx))}{r_1 r_2 (1 + a^2 - x^2 + j2x)}$$

If we set $\frac{C_1}{C_1 + C_m} = b$ then $\frac{C_1 C_m}{C_1 + C_m} = C_{e_1} = C_m b \quad (37)$

Substituting $\frac{C_{e_1}}{b}$ for C_m in equation (36) gives

$$\frac{-I_p Q x_1 b (1 + jx)}{1 + a^2 - x^2 + j2x} \quad (36a)$$

A better approximation to the value of $\overline{E_{d_2}}$ than is given by (31a) is therefore

$$\overline{E_{d_2}} = \frac{\overline{I_p} Q x_1 (1 - b)\left[1 + j\left(x - \frac{ab}{1-b}\right)\right]}{1 + a^2 - x^2 + j2x} \quad (38)$$

A better approximation to the true discriminator output is therefore given

$$Y = n\overline{I_p}QX_1\left[\frac{(1+(x+ap)^2)^{\frac{1}{2}} - (1-b)(1+(x-\frac{ap}{1-b})^2)^{\frac{1}{2}}}{x^4+2x^2(1-a^2)+(1+a^2)^2}\right] \quad (39)$$

When $b = \frac{c_1}{c_1+c_m}$ is very much less than one then (39) reduces to (32a)

3.2.5 RELATIONS BETWEEN THE EQUATION PARAMETERS

It is possible to derive some useful relationships between p, k and b.

We have defined $k = \frac{\sqrt{C_{e1}C_{e2}}}{C_m}$ (22)

and $p = \frac{k\,C_m}{C_2}$ (35)

Substituting for k in the expression for p

$$p = \frac{\sqrt{C_{e1}C_{e2}}}{C_2}$$

Combining equations (5) and (18) yields the result

$$C_2 = 2C_{e2}$$

Hence substituting for C_2

$$p = \frac{1}{2}\sqrt{\frac{C_{e1}}{C_{e2}}} \quad (40)$$

If we take the product pk we have

$$pk = \frac{1}{2}\sqrt{\frac{C_{e1}}{C_{e2}}}\cdot\frac{\sqrt{C_{e1}C_{e2}}}{C_m} = \frac{1}{2}\frac{C_{e1}}{C_m} = \frac{b}{2}$$

and therefore

$$b = 2pk \quad (41)$$

3.25 COMPARISON OF THE EQUATIONS FOR OUTPUT

We have seen that the approximate output equation(32a) differs somewhat from that given by euqation (39). Since it is obviously simpler to analyze than equation (39) we should like to know how much the characteristic it represents departs from the more accurate characteristic opf equation (39) A comparison of the discriminator cross over points should indicate what error might be expected.

From equation (32a) we see that at the resonant frequency when $x = 0$ the output of the discriminator is zero. The more exact expression of equation (34) shows that the output is not zero when $x = 0$

For zero output the numerator of equation (34) must obviously be zero. This then fixes the condition

$$1 + (x + ap)^2 = (1 - b)^2 + (x(1-b) - ap)^2$$

from which

$$x^2 + \frac{2ap}{b}x + 1 = 0$$

and for zero output

$$x = x_o = -\frac{ap}{b} \pm \left[\left(\frac{ap}{b}\right)^2 - 1\right]^{\frac{1}{2}}$$

$$= -\frac{ap}{b} \pm \frac{ap}{b}\left(1 - \left(\frac{b}{ap}\right)^2\right)^{\frac{1}{2}} \qquad (42)$$

Expanding the square root into a binomial series gives

$$x_o = -\frac{ap}{b} \pm \frac{ap}{b}\left[1 - \frac{1}{2}\left(\frac{b}{ap}\right)^2 - \frac{1}{8}\left(\frac{b}{ap}\right)^4 - \cdots\right]$$

Since $\frac{b}{ap}$ will be much less than one in any practical case we can terminate the series after the second term. The negative root may be neglected since it gives a value of x well out of the range for which x is a good approximation for $Q\left[\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega}\right]$

As a result we have

$$x_o = -\frac{1}{2}\frac{b}{ap} \qquad (42a)$$

Since a = Qk and b = 2 pk from equations (34 and 41)

$$x_0 \cong -\frac{1}{Q} \qquad (42b)$$

In terms of frequency since $x = Q\nu = \frac{2Q\Delta f}{f_0}$

$$\Delta f = \frac{-f_0}{2Q^2} \qquad (42c)$$

Thus if Q = 46 and f_0 = 4300 kilocycles the discriminator cross over point is shifted about one kilocycle below the resonant frequency.

In general the output of the discriminator as given by the more accurate expression of equation 39 will be slightly greater for positive x and slightly less for negative than that given by equation (32a). Figure 14 show the curves for characteristics calculated from the two output equations for the condition where Q = 46, p = 0.707 and a = Qk = 1.5. The resonant frequency is taken as 4300 kilocycles.

It can be seen that if the more accurate curve were shifted slightly to the right the two curves would be almost coincident over the useful working range of the characteristic. The more accurate curve has a slightly larger useful working range babove the crossover point than below but the effect is not serious.

In analyzing the characteristic we may then confine the analyis to the approximation of equation (32a) recognizing that some asymmetry is unavoidable but can be made to have a negligible effect upon the discriminator working range.

3.3 CONVERSION EFFICIENCY OF THE DISCRIMINATOR

The conversion efficiency of the discriminator may be defined as the slope of the frequency-voltage characteristic at the centre frequency. Mathematically this can be determined by differentiation of (39a) so that we have

$$\text{Conversion Efficiency} = \left[\frac{dY}{df}\right]_{\Delta f=0} = S_1 \qquad (43)$$

Let us set the frequency sensitive portion of equation (39a)

equal to

$$y = M.N. \qquad (44)$$

$$\text{where } M = \left[1+(x+ap)^2\right]^{\frac{1}{2}} - \left[1+(x-ap)^2\right]^{\frac{1}{2}} \qquad (45)$$

$$\text{and } N. = \left[x^4 + 2x^2(1-a^2) + (1+a^2)^2\right]^{-\frac{1}{2}} \qquad (46)$$

We have then $\dfrac{dy}{dx} = M\dfrac{dN}{dx} + N\dfrac{dM}{dx}$

when $x = 0$, $M = 0$, $N = (1+a^2)^{-1}$, $\dfrac{dm}{dx} = 2ap(1+a^2p^2)^{-\frac{1}{2}}$

$$\text{therefore } \left(\frac{dy}{dx}\right)_{x=0} = S = \frac{2ap}{(1+a^2)(1+a^2p^2)^{\frac{1}{2}}} \qquad (47)$$

$$\left(\frac{dY}{dx}\right)_{x=0} = S = n\bar{I}_p Q X_1 \left[\frac{2ap}{(1+a^2)(1+a^2p^2)^{\frac{1}{2}}}\right]$$

since $\dfrac{dY}{df} = \dfrac{dY}{dx}\cdot\dfrac{dx}{df} = \dfrac{2Q}{f_o}\dfrac{dY}{dX}$

we may write for conversion efficiency

$$S_1 = \left(\frac{dY}{df}\right)_{\Delta f=0} = \frac{4n\bar{I}_p Q^2 X_1 \quad ap}{f_o \quad (1+a^2)(1+a^2p^2)^{\frac{1}{2}}} \qquad (43a)$$

Equation (43a) turns out to be identical with that given by Foster- Seeley for the inductively coupled case, the notation being somewhat different. This is to be expected since the approximations made in the capacity coupled output equation reduced it to a mathematically symmetrical equation. The coupling coefficients in the two cases are different of course but they indicate equivalent relationships between stored and coupled energy.

3.4 COUPLING FACTOR FOR MAXIMUM CONVERSION EFFICIENCY

The coupling factor required for maximum conversion efficiency may be obtained by equating the first derivative of 43a to zero, holding the factor p constant.

This leads to

$$\frac{dS_1}{da} = \frac{4 n I_p Q^2 X_1 p}{F_0} \frac{d}{da}\left[\frac{a}{(1+a^2)(1+a^2p^2)^{\frac{1}{2}}}\right] = 0$$

from which $$2a^4p^2 + a^2 - 1 = 0$$

and $$a = \frac{1}{2p}\left[(1+8p^2)^{\frac{1}{2}} - 1\right]^{\frac{1}{2}} \qquad (48)$$

for maximum conversion efficiency with fixed primary to secondary capacity ratio.

Fig. 16 shows a plot of the function of equation 48. for values of p from zero to two. It may be seen that the coupling factor is always less than one, irrespective of the value of p.

3.5 VARIATION OF CONVERSION EFFICIENCY WITH CAPACITY RATIO - p

For the manner in which conversion efficiency varies with p when the coupling factor a is fixed we note that equation 43a may be written

$$S_1 = \frac{4 n \overline{I_p} Q^2 X_1}{f_0} \frac{a}{(1+a^2)(1+a^2p^2)^{\frac{1}{2}}}$$

If $X_1 = \frac{1}{\omega_0 C_{e1}}$ is fixed, therefore the conversion efficiency must increase, with increase in p, to a limiting value

$$S_1 = \frac{4 n \overline{I_p} Q^2 X_1}{f_0 (1+a^2)} \tag{49}$$

Since $X_1 = \frac{1}{\omega_0 C_{e1}}$ it is important that C_{e1} be small and this places a practical limitation on the value of p. The primary stray capacity is generally considerably greater than the secondary strays so that as a practical matter C_{e2} can be made considerably smaller than C_{e1}. If C_{e1} is made great enough to minimize the effects of stray primary variations then the minimum permissible value of C_{e2} will limit the maximum value of p which can be employed.

3.6 DISCRIMINATOR PEAK SEPARATION

The location of the peaks of a discriminator curve was of much interest to the designers of automatic tuning circuits. The peaks were generally designed to be about five kilocycles apart, thus falling within the receiver band pass. The location of the discriminator peaks thus determined to a large extent the range of frequency control, or the"pull-in" and "throw-out" frequencies.

Foster-Seeley gave nom solution to the problem for the phase discriminator but stated that the peaks would fall in the range corresponding to $x = 0.5.$ to $x = 1.0$ for small coupling coefficients. Roder has given an approximate solution for the inductive coupled case froma a circle diagram based on the assumption of constant primary voltage. For coupling factors less than Q.75, Roder shows that the peaks occur at approximately $x = 1.0$ (10)

It is not difficult to see why no simple splution to the problem exists. The points of zero slope can be determined by taking the derivative with respect to x of equation 44, equating the result to zero and solving for x in terms of a and p. A complicated expression with x raised to the fifth power results and the solution of this equation for particular value s of a and p would be as tedious as calculating the curve for values of x and determining the peak in that manner.

The solution given by Roder can be supplemented by an approximate solution for coupling factors greater than one. Considering equation 44 again we have

$$y = M.N.$$

where $M = \left(1+(x+ap)^2\right)^{+\frac{1}{2}} - \left(1+(x-ap)^2\right)^{+\frac{1}{2}}$

and $N = \left(x^4 + 2x^2(1-a^2) + (1+a^2)^2\right)^{-\frac{1}{2}}$

The function of M is the difference in two hyperbolic functions and does not change very rapidly with x for values of x beyond $x = ap$.

Differentiation of the function N with respect to x shows that two maxima occur at

$$x = \pm\left(a^2 - 1\right)^{\frac{1}{2}} \qquad (50)$$

The function Y will then maximize at two values of x less than a and if $a^2 >> 1$, the peaks will occur at approximately

$$x \cong \pm a \qquad (50a)$$

Discriminators for frequency modulation reception are generally designed for linearity over a range somewhat greater than the receiver band pass and the peaks are thus fixed to a greater extent by the actual band pass of the receiver.

3.7 TYPICAL DISCRIMINATOR CURVES

Fig. 16a shows a set of relative discriminator output curves calculated from the function of equation 44. The value of p chosen corresponds to a primary to secondary capacity ratio of 1.5 to 1. Arbitrary coupling factors were used as indicated in the figure and the functions was plotted over a range of x from zero to three. Slide rule accuracy only is obtained.

It may be seen from the figure that the conversion efficiency or slope of the characteristic for a = 0.75 is about the same as that for a = 1. The maximum slope would be given by a curve for a = 0.81 as determined from Fig.15.

For greater or less coupling factors it may be seen that the slope of the characteristic falls off.

The peaks occur at greater values of x as the coupling factor increases. It is necessary to interpret this fact properly in terms of frequency deviations.

Since $x = \frac{2Q\,\Delta f}{f_0}$, the abscissae are directly proportional to Q. If Q is constant then, increasing the coupling factor through increase of k will result in the peaks occuring at greater values of Δf. If, on the other hand, k is constant, increasing the coupling factor through increase of Q, will reduce the value of at which the peaks occur providing $\frac{x_2 Q_1}{x_1 Q_2} < 1$ where x_1 and x_2 are the peaks abscissae at which the peaks occur. Equation 50 shows that

$$\frac{x_2}{x_1} \cong \frac{\left((Q_2 k)^2 - 1\right)^{\frac{1}{2}}}{\left((Q_1 k)^2 - 1\right)^{\frac{1}{2}}}$$

For coupling factors greater than one since $Q_2 k > Q_1 k$

$$\frac{x_2}{x_1} < \frac{Q_2}{Q_1}$$

and the value of Δf at which the peaks occur is therefore less for the case of greater Qk.

The curves show that increasing the value of Qk beyond one results in an increase in the working range of the discriminator curve but do not show the linearity of the curves very accurately. It is necessary to determine by how much the characteristic departs from the straight line which is necessary for distortionless frequency modulation reception and the manner in which this may be done is indicated in the following section.

3.8 LINEARITY CONSIDERATIONS

3.8.1 DEFINITION OF LINEARITY

The conditions governing the linearity of phase discriminators have been treated from two rather different points of view by Sturley (24) and Arguimbau (28)

Sturley defines the departure from linearity of the curve in terms of the ratio of the output, at a particular value of x, to the output which would be obtained if the characteristic were a straight line having a slope equal to the slope of the characteristic at $x = 0.1$. Arguimbau defined the departure from linearity at any given value of x as the difference between the output and the tangent to the characteristic at $x = 0$. This appears to be a more fundamental definition than that given by Sturley although the results should be virtually identical since the slope at $x=0.1$ will not depart from that at $x = 0$.

In this paper, expanding Arguimbaus lead, the non linearity is defined as the percentage departure of the characteristic from the straight line tangent to the characteristic at $x = 0$. Expressed mathematically the percentage non-linearity becomes

$$\frac{x\left[\frac{dY}{dx}\right]_{x=0} - Y}{x\left[\frac{dY}{dx}\right]_{x=0}} \times 100$$

$$= \left(1 - \frac{y}{sx}\right) \times 100 \qquad (51)$$

Non-linearity as defined in this fashion may be positive or negative depending upon whether $\frac{y}{sx}$ is greater or less than one. Geometrically the characteristic is concave downward from the tangent for positive nonlinearity, and is concave upward from the tangent for negative non-linearity. The two conditions are illustrated in fig. 17 a and b.

3.8.2 STURLEY'S TREATMENT OF LINEARITY

Sturley treats the problem of linearity in a rather indirect fashion. Assuming that the primary voltage is constant with x, it has been shown by Roder (10) that the output may be determined from a circle diagram. This is done by Sturley for ratios of secondary to primary voltage of 1,2,3,4,5,and 6. The ratio decrease in linearity of the resulting output curves is then plotted against x. For coupling factors greater than one, the primary voltage increases as the frequency departs from the centre frequency and the familiar double humps appear in the voltage characteristic. The ratio increase of primary voltage with x is calculated for coupling factors of 1.0,11, 1.2, ... 2.0 and the resulting curves are superimposed on the ratio-decrease-in-linearity curves calculated on the assumption of constant primary voltage. By this means the coupling factor which gives maximum linearity for the secondary to primary voltage ratios considered, can be found.

While the method is seemingly accurate enough for practical purposes, it amounts to compensating by means of two separate characteristics, both of which are determined by assuming arbitary values having discrete differences. The method of superposition can also be expected to lead to some error.

3.8. ARGUIMBAU'S TREATMENT OF LINEARITY

The problem is essentially a mathematical one and a more fundamental approach seems to be suggested by Arguimbau. He considers the problem as applied to a simplified discriminator having a centre tapped primary as well as secondary and having equal primary and secondary circuit constants. Such a discriminator has a frequency sensitive equation given by

$$y = \frac{1}{(1+(x-a)^2)^{\frac{1}{2}}} - \frac{1}{(1+(x+a)^2)^{\frac{1}{2}}}$$

where the symbols have the same meaning used previously.

When the expression for y is expanded in a power series in the neighborhood of $x = 0$ we have

$$y = a_0 + a_1 x + a_2 x^2 + a_3 x^3 + \cdots a_n x^n$$

where the general coefficient is an

$$a_n = \frac{1}{\lfloor n} \frac{d^n y}{dx^n}$$

Because of symmetry a_0 and all even order terms will vanish at the origin.

For this simplified discriminator Arguimbau shows that

$$\frac{d^3 y}{dx^3} = 6a(2a^2-3)(1+a^2)^{-\frac{7}{2}}$$

and the coefficient of x^3 therefore vanishes when

$$a = \sqrt{\frac{3}{2}}$$

The first term contributing to curvature is then the term $a_5 x^5$ and for small frequency deviations the curve for $a = \sqrt{\frac{3}{2}}$ is the closest approximation to a straight line. If greater range of frequency is desired and more distortion can be tolerated for a small value of x a greater value of a can be used.

3.8.5. THE APPLICATION OF ARGUIMBAU'S METHOD TO THE GENERAL EQUATION

The general expression of equation 44 is more complex than thatmgiven by Arguimbau since it involves the primary to secondary capacity ratio and takes into account the fact that full primary voltage, rather than half primary voltage, is applied in series with the half secondary voltage to the rectifiers. The method should nevertheless be applicable to the problem.

We have from equation 44

$$y = m.N$$

where
$$m = \left(\left(1+(x+ap)^2\right)^{\frac{1}{2}} - \left(1+(x-ap)^2\right)^{\frac{1}{2}}\right)$$

and
$$N = \left(x^4 + 2x^2(1-a^2) + (1+a^2)^2\right)^{-\frac{1}{2}}$$

taking derivatives of y with respect to x we then have

$$\frac{dy}{dx} = m\frac{dN}{dx} + N\frac{dm}{dx} \qquad (52)$$

$$\frac{d^2y}{dx^2} = m\frac{d^2N}{dx^2} + 2\frac{dN}{dx}.\frac{dm}{dx} + N\frac{d^2m}{dx^2} \qquad (53)$$

$$\frac{d^3y}{dx^2} = m\frac{d^3N}{dx^3} + 3\frac{dm}{dx}.\frac{d^2N}{dx^2} + 3\frac{d^2m}{dx^2}.\frac{dN}{dx} + N\frac{d^3m}{dx^3} \qquad (54)$$

When x is equal to zero we then have

$$N = (1+a^2)^{-1}, \qquad m = 0$$

$$\frac{dN}{dx} = 0, \qquad \frac{dm}{dx} = 2ap(1+a^2p^2)^{-\frac{1}{2}}$$

$$\frac{d^2N}{dx^2} = \frac{-2(1-a^2)}{(1+a^2)^3}, \qquad \frac{d^2m}{dx^2} = 0$$

$$\frac{d^3N}{dx^3} = 0, \qquad \frac{d^3m}{dx^3} = -6ap(1+a^2p^2)^{-\frac{5}{2}}$$

substituting these quantities into equations 52, 53, 54 leads to

$$\left[\frac{dy}{dx}\right]_{x=0} = \frac{2ap}{(1+a^2)(1+a^2p^2)^{\frac{1}{2}}} = A$$

$$\left[\frac{d^2y}{dx^2}\right]_{x=0} = 0$$

$$\left[\frac{d^3y}{dx^3}\right]_{x=0} = \frac{6ap}{(1+a^2)(1+a^2p^2)^{\frac{1}{2}}}\left[\frac{2(a^2-1)}{(1+a^2)^2} - \frac{1}{(1+a^2p^2)^2}\right]$$

$$= 3A^3\left[\frac{2(a^2-1)(1+a^2p^2)^2-(1+a^2)^2}{4a^2p^2(1+a^2p^2)}\right] \qquad (55)$$

For small values of x, the coefficients of the series become

$$a_0 = 0$$

$$a_1 = A$$

$$a_2 = 0$$

$$a_3 = \frac{1}{\lfloor 3}\frac{d^3y}{dx^3} = A^3\left[\frac{2(a^2-1)(1+a^2p^2)^2-(1+a^2)^2}{8a^2p^2(1+a^2p^2)}\right]$$

The series expansion for y is then

$$y = a_0 + a_1x + a_2x^2 + a_3x^3 + \cdots \quad a_nx^n$$

For small x

$$y = Ax + A^3x^3\left[\frac{2(a^2-1)(1+a^2p^2)^2-(1+a^2)^2}{8a^2p^2(1+a^2p^2)}\right] \qquad (56)$$

The percentage non-linearity becomes

$$\left[1 - \frac{y}{Ax}\right]100$$

$$= -A^2x^2\left[\frac{2(a^2-1)(1+a^2p^2)^2-(1+a^2)^2}{8a^2p^2(1+a^2p^2)}\right]100 \qquad (57)$$

For small frequency deviations the percentage departure from linearity may be made substantially equal to zero, when the following condition is obtained from equation 57

$$2(a^2-1)(1+a^2p^2)^2 = (1+a^2)^2 \qquad (58)$$

Expressed in explicit form, this becomes

$$p = \frac{1}{a}\left[\left[1+a^2\right]\left[2(a^2-1)\right]^{-\frac{1}{2}} - 1\right]^{\frac{1}{2}} \qquad (58a)$$

It may be seen from this equation that a must be greater than one for real values of p. The condition on a for maximum conversion efficiency therefore cannot be met if a straight line characteristic about the origin is desired.

If p isless than the value indicated by equation (58a) it is evident that positive nonlinearity will be obtained. If p is greater, negative non linearity will be obtained.

Fig 18 shows equation (58a) with p plotted as a function of Qk , for values of Qk from 1.2 to 3.4. From this curve it is possible to select a value of coupling factor which will give maximum linearity in the vicinity of x-0 for a given value of p. As x departs farther from zero, of course, the other terms in the series will assume greater importance and the approximation of equation 56 no longer holds.

Fig.16 b shows how the percentage departure from linearity varies with x for the coupling factors used in Fig.16a and for the critical coupling factor Qk = 1.65 obtained from Fig.18 for the particular value of p used. The curves are based on slide rule calculations but they indicate the general trends. For small coupling factors the non-linearity is positive and increases rapidly with x. For large coupling factors, the non-linearity is initially negative andoveventually becomes positive after the tangent intersects the characteristic as indicated in Fig.17 b. The critical value Qk = 1.65 is seen to give the maximum range of undistorted characteristic. If more distortion can be tolerated at small x it can be seen that some value of coupling factor between 1.65 and 2 would give a greater range of useful characteristic.

3.85 COMPARISON OF RESULTS WITH STURLEY'S WORK

It is interesting to compare the value of p obtained from equation (58a) with that given by Sturley as an optimum design value. Sturley gives a suitable compromise design condition a = 1.5 and secondary-to-primary voltage ratio of 2. The conditions may be shown to be equivalent to a value of p = 2/3 = 0.667.

From equation (58a) a value of Qk = 1.5 demands that p be 0.685 corresponding to a voltage ratio of 2.055 or a difference of about 2.5 percent from that indicated by Sturley.

The conditions are not thus very different but it is interesting to compare the range of x in each case for similar conditions of non-linearity. To do this accurately five place logarithms were employed and the results are indicated in Fig. 19.

7 The curves plotted here are the function of equation (51) for arbitrary coupling factors and values of p as given by equation (58a).

It may be noted that for Qk = 1.5 the value of p indicated by Sturley gives a two percent departure from linearity at x = 0.75. The value of p given by equation (58a) gives this departure from linearity at x = 0.79, a difference of about five percent in the working range. Sturley gives the range of x as 0.8 but the method of calculation is based on the slope at x = 0.1 and is rather less refined since it involves the superposition of two curves calculated with arbitrary values. As a practical matter it is probable that the difference between the two conditions is negligible.

If p is increased to 0.707 for Qk = 1.5, corresponding to a voltage ratio of 2.12, the range of two percent nonlinearity is increased to x = 0.84 butit may be noted that negative non-linearity is then present at lower values of x, where the critical p value gives an essential linear characteristic.

It may be seen from the curves of Fig. 19 for different values of Qk that the range of distortionless characteristic increases as Qk increases and the value of p required is less. Further it may be noted from Fig. 18 that with high values of "a" the critical value of p becomes less sensitive to changes in "a" and the coupling and Q need not be so accurately controlled. A value of Qk = 2.0 and p = 0.5 is commonly used in wide band frequency modulation phase discriminators. The curves of Fig. 19 indicate that a value of p = 0.51 would be somewhat better.

Fig.20 shows the relative discriminator outputs for the critical p and Qk relationships of Fig 19 assuming thatm Q and X_1 are constant. It may be seen that the output is reduced considerably as the coupling factor increases. The output for Qk = 1.4 is about 1.6 times that obtained for Qk = 2.0. It appears that Qk should be kept as low as possible consistent with obtaining the required linearity.

3.9 LOADING EFFECTS

The Q which has been referred to thus far is the effective Q of the primary and secondary circuits withall the loading effects assumed taken into consideration by modification of the unloaded coil Q factors. It is necessary to relate the effective Q of the circuit to the unloaded coil Q's for design purposes.

The shunt loading used in this discriminator has a worse loading effect than the more conventional series loading. Sturley has given a method for calculating the effective Q of the series loaded phase discriminator and the method is applicable with some modifications to the shunt loaded case.

Consider first an unloaded parallel tuned circuit with a Q of Q_0, inductance L and capacity C. The resonant parallel impedance is a pure resistance $R = Q_0 w_0 L$ where $w_0{}^2 = \frac{1}{LC}$ and the circuit may be represented by the equivalent circuit shown below.

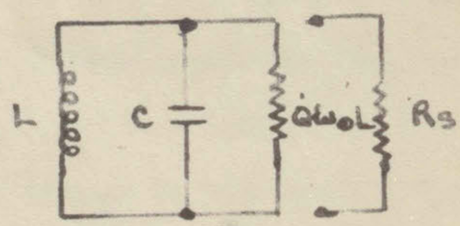


Fig. 21

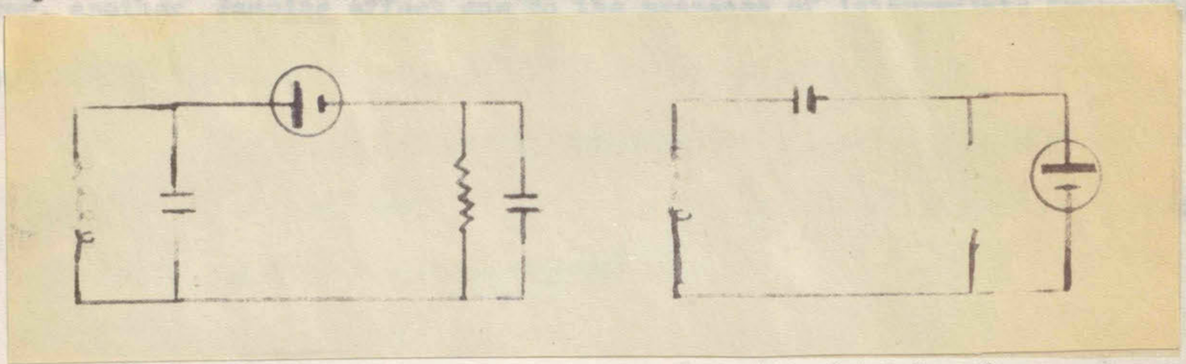
When the circuit is loaded with a shunt resistance R_s, the resonant impedance becomes $\frac{R\,R_s}{R + R_s}$ and therefore the effective Q is given by

$$Q_{eff}\, w_0 L = \frac{RR_s}{R + R_s} = \frac{Q_0\, w_0\, L\, R_s}{Q_0\, w_0\, L + R_s} \qquad (59)$$

The effective circuit Q is then given by

$$Q_{eff} = \frac{Q_o}{1 \pm \frac{Q_o w_o L}{R_s}} \qquad (59a)$$

Consider now the conditions of shunt and series loading of ordinary amplitude modulation second detectors shown below.



Series Loading (a) Shunt loading (b)

Fig. 22

The load condenser in each case fills in the gaps in the output voltage pulses when the rectifier is cut off on negative voltage peaks of the signal. The mean D.C. voltage across the load resistor then approaches the peak value of the tuned circuit voltage. The detection efficiency is defined as the ratio of D.C. load voltage to peak radio frequency input voltage and when the forward resistance of the rectifier is negligible with respect to the load resistance, the detection efficiency approaches unity.

If the detection efficiency is unity we then have

$$E_{dc} = E_p = \sqrt{2}E$$

Where E_{dc} = output D. C. voltage

E_p = Peak frequency signal voltage

E = Root mean square radio frequency signal voltage

The power absorbed by the load is then

$$\frac{(E_{dc})^2}{R_1} = \frac{2E^2}{R1} = \frac{E^2}{R1/2} \qquad (60)$$

This is equivalent to the power absorbed by a resistance $R_1/2$ connected across the tuned circuit. When the detection efficiency is less than unity the equivalent damping resistance is greater and approaches R_1 when the detection efficiency = 0.55. (45)

This conduction current damping is substantially the same in both the series and shunt loading cases. In the case of the shunt loaded circuit there is, however, another damping effect due to the presence of intermediate frequency voltage across the load resistance. (48)

The A. C. resistance of fixed composition resistors has been shown to be less than D.C. resistance. (48) If the ratio of A.C.toD.C. resistance is denoted by e we have for the A.C. resistance of the load.

$$(R_1)_{ac} = eR_1$$

The damping of the tuned circuit due to the A.C.load resistance is then equivalent to a resistance eR_1 shunting the tuned circuit.

For the shunt loaded case the two damping effects are then equivalent to a damping resistance of value given by

$$\frac{eR_1 \cdot \frac{R_1}{2}}{eR_1 + \frac{R_1}{2}} = \frac{eR_1}{1 + 2e} = R_1/3 \text{ if e approaches 1} \qquad (61)$$

The value of e depends upon the type and nominal resistance of the resistor and also upon the frequency. In general, e decreases with increasing frequency and nominal resistance. For a resistance in the order of 1 megohm at four megacycles, e may have a value in the order of 0.6 so the effect may be rather serious.

To apply these results to the shunt loaded phase discriminator of Fig. 12, it may be seen that the rectifier circuits are effectively in parallel across theprimary circuit. Additional damping is also contributed by the tube plate impedance R_p. The total effective primary damping resistance is then given by

$$R_{s1} = \frac{\frac{R_p}{2} \frac{e R_1}{(1+2e)}}{R_p + \frac{eR_1}{2(1+2e)}} = \frac{eR_p R_1}{2R_p (1+2e) + eR_1} \qquad (62)$$

If $R_p >> R_1$ and e approacheS 1, $R_{s1} \cong R_{1/6}$ (62a)

In this case the effective primary Q is given by

$$Q_1 = \frac{Q_{p1}}{1 + \frac{Q_{o1} \; w_o L_1}{\frac{R_1}{6}}} \qquad (63)$$

Q_{o1} is the unloaded Q of the primary coil.

As far as the secondary circuit is concerned the rectifier circuits are in series and the damping resistance is then given by

$$R_{s2} = \frac{2 eR_1}{1 + 2e} = \frac{2R_1}{3} \quad \text{if e approaches 1} \qquad (64)$$

The effective secondary Q is then given by

$$Q_2 = \frac{Q_{o2}}{1 + \frac{Q_{o2} \, w_o L_2}{\frac{2}{3} R_1}} \qquad (65)$$

Where Q_{o2} is the unloaded Q of the secondary coil.

It is evident that the primary loading effects are much more serious than the secondary loading effects. The ratio of the damoing resistances is in fact

$$\frac{R_{2s}}{R_{1s}} = 4 + \frac{2e}{(2e+1)R_p} R_1 \qquad (66)$$

The effective primary damoing resistance is less than one quarter of the secondary damping resistance and a limiting design factor in any particular case may then be the effective primary Q.

In maintaining a high primary Q, it may be seen from equation 63 that if $\omega_o L1 = \frac{1}{\omega_o C_e1}$ is kept small, the reduction of primary unloaded Q will be less. If equal primary and secondary effective Q's are designed for, the secondary resonant reactance may be considerably higher than the primary since the effective damping resistance is more than four times as high. Since $p = \frac{1}{2}\sqrt{\frac{C_{e1}}{Ce_2}}$, values of p greater than 0.5 will therefore help to some extent in equalizing primary and secondary Q values.

3.10 THE EFFECTS OF STRAY CAPACITY ON THE SHUNT COUPLED CIRCUIT

In experimental application of the circuit of Fig.12 some difficulty was experienced in obtaining agreement between the experimental and calculated values of coupling coefficient, $k = \frac{\sqrt{C_{e1} \quad Ce_2}}{Cm}$. It was recognized by the writer that stray capacities, particularly in the primary circuit, would modify the coupling coefficient through changes in C_e1 and C_e2 and calculations were made on that basis. The writer is much indebted to Mr. J. R. G. Bennett for pointing out that the effective value of C_m would also change appreciably and suggesting that an analysis of the following nature be carried out.[3]

The coupling network of Fig.12 is modified considerably by the effects of stray capacity. When stray capacity effects are considered, the coupling network of Fig.12 may be redrawn as shown below.

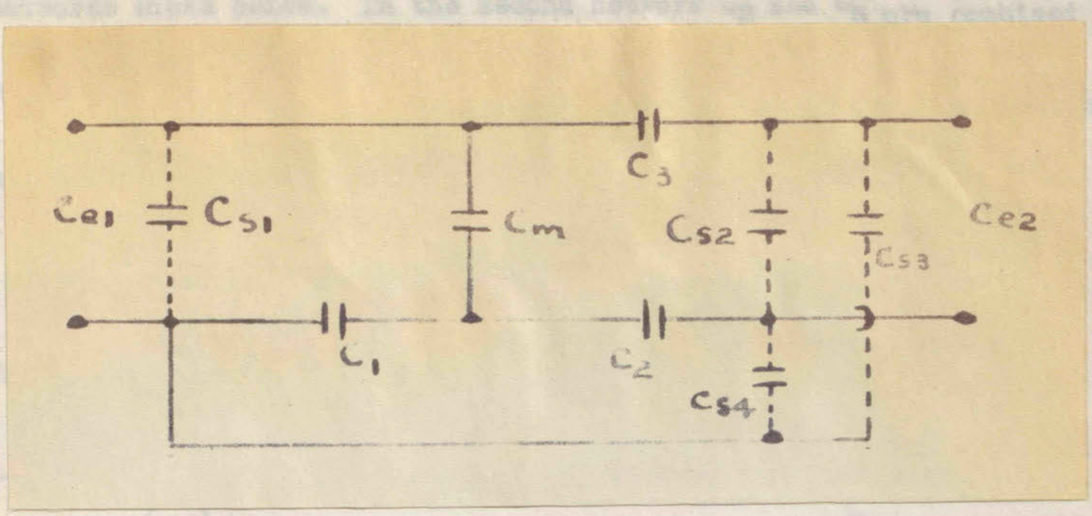


Fig.23

Here C_1 , C_2, C_3 and C_m have the same meaning as previously used.

C_s1 = Primary stray capacity including the valve output capacity coil distributed capacity, and wiring capacity.

C_s2 = Secondary coil distributed capacity + wiring capacity.

C_s3 = Capacity to ground of first rectifier circuit.

C_s4 = Capacity to ground of second rectifier circuit.

Representative measured values for these quantities have been found to be

C_s1 = 12 uuf.

C_s2 = 5 uuf.

C_s3 = and C_s4 = 3 uuf.

The network of Fig.23 can be simplified considerably if C_s4 and C_s3 are lumped in with other capacities. It may be seen that the effect of C_s3 and C_s4 on the secondary tuning is approximately equivalent to a capacity $\frac{C_s3\ C_s4}{C_s3 + C_s4}$ in parallel with C_s2. As far as the primary circuit is concerned, since C_2 is much greater than C_s4, the effect of C_s4 is to add to C_1. Similarly since C_3 is much greater than C_s3, the effect of C_{s3} is to add to C_{s1}.

A good approximation to the network is then given by either of the two networks shown below. In the second network C_2 and C_3 are combined into $C_4 = \frac{C_2 C_3}{C_2 + C_3}$ for convenience of analysis. As far as coupling or tuning of the circuits is concerned no generality is lost by this latter step.

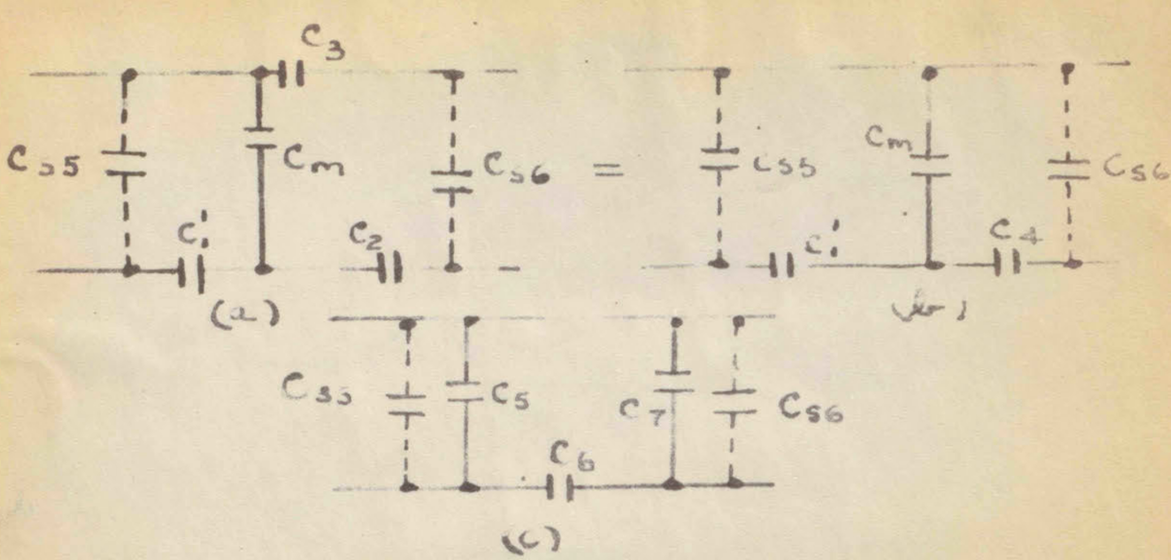


FIG 24

here we have

$$C_{s5} \cong C_{s1} + C_{s3} \tag{67}$$

$$C_{s6} \cong C_{s2} + \frac{C_{s3} . C_{s4}}{C_{s3} + C_{s4}} \tag{68}$$

$$C_1' = C_1 + C_{s4} \tag{69}$$

$$C_4 = \frac{C_2 C_3}{C_2 + C_3} \tag{70}$$

To transform the network of Fig. 24 b into that of Fig.12 we note that the star network C_1', C_4, C_m can be transformed into a delta C_5, C_6, C_7

where
$$C_5 = \frac{C_1' C_m}{\Sigma C} \tag{71a}$$

$$C_6 = \frac{C_1' C_4}{\Sigma C} \tag{71b}$$

$$C_7 = \frac{C_4 C_m}{\Sigma C} \tag{71c}$$

$$\Sigma C = C_1' + C_4 + C_m \qquad (71d)$$

Adding the strays C_s5 and C_s6 results in a new delta network in which the branches are $C_5 + C_s5$, C_6, $C_7 + C_s6$.
This delta may be transformed into a star network.

C_{1e}, C_{me}, C_{4e}

Where
$$C_{1e} = \frac{\Sigma CC}{C_7 + C_{s6}} \qquad (72a)$$

$$C_{me} = \frac{\Sigma CC}{C_6} \qquad (72b)$$

$$C_{4e} = \frac{\Sigma CC}{C_5 + C_{s5}} \qquad (72c)$$

and
$$\Sigma CC = (C_5 + C_{s5})C_6 + (C_7 + C_{s6})C_6 + (C_7 + C_{s6})(C_5 + C_{s5}) \qquad (72d)$$

The resulting star network is shown below in Fig. 25a and it may be seen that the only modification required to make it identical with the network of Fig. 12 is the division of C_{4e} into the two components C_{2e} and C_{3e} as shown in Fig. 25b.

Then
$$C_{4e} = \frac{C_{2e} C_{3e}}{C_{2e} + C_{3e}} \qquad (73)$$

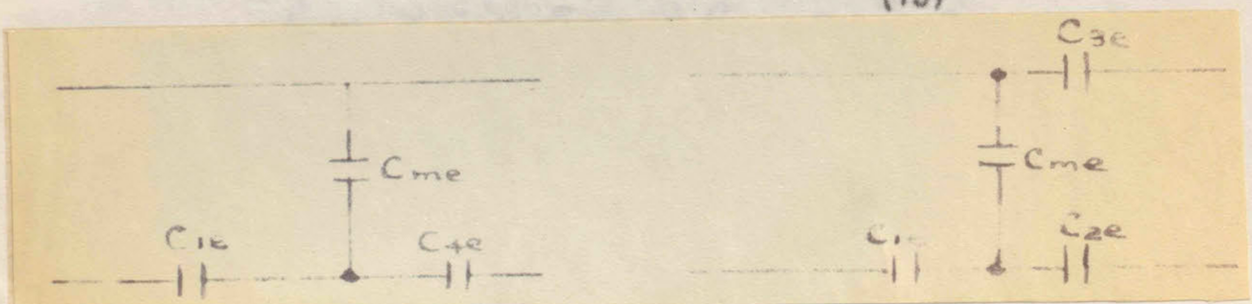


Fig. 25

From equation (18) it follows that $C_{2e} = \dfrac{C_{3e} C_{me}}{C_{3e} + C_{me}}$

Therefore $$C_{2e} = \frac{2 C_{4e} C_{me}}{C_{me} + C_{4e}} \qquad (74)$$

and $$C_{3e} = \frac{2 C_{4e} C_{me}}{C_{me} - C_{4e}} \qquad (75)$$

To simplify the calculations of C_{1e}, C_{me}, C_{4e}

assume first that $C_m >> C_1' + C_4$

then $$\sum C \cong C_m, \quad C_5 \cong C_1', \quad C_6 \cong \frac{C_1' C_4}{C_m}, \quad C_7 \cong C_4$$

For $\sum$ C C we have then

$$\frac{C_1' C_4}{C_m}(C_1' + C_{s5} + C_4 + C_{s6}) + (C_1' + C_{s5})(C_4' + C_{s6})$$

The first term here will be much less than the second because of C_m in the denominator.

Therefore $$\sum CC \cong (C_1' + C_{s5})(C_4 + C_{s6})$$

And from equations (72a)(72b) (72c) and (72d) we may write

$$C_{1e} \cong C_1' + C_{s5} \cong C_1 + C_{s4} + C_{s5} \qquad (75a)$$

$$C_{4e} \cong C_4 + C_{s6} \qquad (75b)$$

$$C_{me} \cong \frac{(C_4 + C_{s6})(C_1' + C_{s5})\, C_m}{C_1' C_4} \qquad (76a)$$

The two approximations made for $\sum C$ and $\sum CC$ tend to balance out in calculations of C_{1e} and C_{4e} but not in calculations of C_{me}. The result is that the error in calculation of C_{1e} and C_{4e} is negligible. C_{me} as calculated from equation will be slightly high, the error increasing with the degree of x coupling or inversely as C_m.

A sample calculation will serve to indicatethe extent of the error involved in the approximations.

Suppose $C_1 = 50$ uuf., $C_4 = 25$ uuf, $C_m = 1000$ uuf.

$C_{s5} = 15$ uuf $\quad C_{s6} = 5$ uuf.

For exact values of C_{1e}, C_{me}, C_{4e}

$$\sum C = 50 + 25 + 1000 = 1075 \text{ uuf}$$

$$C_5 = 46.5, \quad C_6 = 1.16 \quad C_7 = 23.3 \text{ uuf}$$

$$C_5 + C_{s5} = 61.5 \quad C_7 + C_{s6} = 28.3$$

$$\sum CC = 1847 \text{ uuf}$$

$$C_{1e} = \frac{1847}{28.3} = 65.2, \quad C_{me} = \frac{1847}{1.16} = 1580, \quad C_{4e} = \frac{1847}{61.5} = 30 \text{ uuf}$$

FROm the approxiamate equations we have

$$C_{1e} = 50 + 15 = 65 \text{ uuf}$$

$$C_{4e} = 25 + 5 = 30 \text{ uuf}$$

$$C_{me} = \frac{65}{50} \cdot \frac{30}{25} \cdot 1000 = 1560 \text{ uuf}$$

The error in C_{1e} is negligible and C_{me} is about 1% low.

If we take $C_m = 500$ uuf and calculate the values again.

We get for the exact values $C_{1e} = 65.3$, $C_{4e} = 29.8$, $C_{me} = 805$ uuf.

For the approximate values $C_{1e} = 65$ $C_{4e} = 30$ $C_{me} = 780$ uuf.

The error in C_{me} is about 3%

3.11 DETERMINATION OF THE CIRCUIT COMPONENTS

In a practical case the actual circuit components must be calculated on the basis of selected values of p and k and on measured or estimated values of the stray capacities.

Using equation 37 we may write for the equivalent circuit of Fig. 25.

$$\frac{C_{1e}}{C_{me} + C_{1e}} = b = 2kp$$

therefore $$\frac{C_{1e}}{C_{me}} = \frac{b}{1 - b} = \frac{2kp}{1 - 2kp} \qquad (77)$$

From equation 35 $$\frac{C_{2e}}{C_{me}} = \frac{k}{p}$$

Combining this with equation 74 leads to

$$\frac{C_{4e}}{C_{me}} = \frac{k}{2p - k} \qquad (78)$$

Combining equations (77) and (78) leads to

$$C_{1e}\frac{1}{C_{4e}} = \frac{4p^2 - 2kp}{1 - 2kp} = d \qquad (79)$$

From (76a) and (76b) we then have

$$\frac{C_{1e}}{C_{4e}} = d = \frac{C_1' + C_{s5}}{C_4 + C_{s6}}$$

Therefore $$C_4 = \frac{C_1' + C_{s5} - dC_{s6}}{d} = \frac{C_1 + C_{s4} + C_{s5} - dC_{s6}}{d} \qquad (80)$$

Combining (76a) and (76c) leads to

$$\frac{C_{1e}}{C_{me}} = \frac{C_1 C_4}{(C_4 + C_{s6})\, C_m} = \frac{b}{1 - b}$$

Separating C_m and using equation 80 then leads to

$$C_m = \frac{1-b}{b}\left[C_1'\left[1 - \frac{d\,C_{s6}}{C_1' + C_{s5}}\right]\right]$$

$$= \frac{1-b}{b}(C_1 + C_{s4})\left[1 - \frac{d\,C_{s6}}{C_1 + C_{s4} + C_{s5}}\right] \qquad (81)$$

By hypothesis $C_4 = \dfrac{C_2 C_3}{C_2 + C_3}$ and $C_2 = \dfrac{C_3 C_m}{C_3 + C_m}$

and it follows that

$$C_2 = \frac{2 C_4 C_m}{C_m + C_4} \qquad (82)$$

$$C_3 = \frac{2 C_4 C_m}{C_m - C_4} \qquad (83)$$

The parameters b and d are functions of the design factors k and p. The stray capacities may be measured or estimated as a first approximation. A value may then be assigned to C_1 in equations 80 and 81 and C_4 and C_m calculated. C_2 and C_3 will then follow from 82 and 83.

Alternatively a value may be assigned to C_4, and C_1 and C_m, calculated from 80 and 81.

For the coil values we note that

$$C_{e1} = \frac{C_{1e} C_{me}}{C_{1e} + C_{me}} = (1-b)\,C_{1e} = (1-b)(C_1 + C_{s4} + C_{s5}) \qquad (84)$$

and L_1 follows from $L_1 = \dfrac{1}{\omega_o^2 C_{e1}}$ (85)

then $L_2 = \dfrac{1}{\omega_o^2 C_{e2}} = \dfrac{1}{\omega_o^2 \dfrac{C_{e1}}{4p^2}} = 4p^2 L_1$ (86

3.12 DESIGN PROCEDURE

As the first step in a particular design, it is necessary to select one of the curves of Fig.19, from which the required Q, k and p factors may be determined. The optimum curve is a compromise between several conflicting factors.

Let us consider a design for a discriminator with a centre frequency of 4300 kilocycles and having a linearity requirement of 40 kilocycles. The bandwidth factor in this case $f_o / 2\Delta f = 53.8$. If a two percent departure from linearity at the extremes of the discriminator curve is tolerable, the critical curve from Fig. 19 for the condition $Qk = 2.0$, $p = 0.510$ has a range of $x = 1.02$. For the curve $Qk = 1.4$, $p = 0.761$ the range of x for a similar departure from linearity is then 0.75. Since $x = 2Q\Delta f / f_o$ the Q required for the condition of $Qk = 2.0$ is then $1.02 (53.8) = 55$. The Q required for the condition $Qk = 1.5$ is then $0.75(53.8) = 40$. Provided the Q can be obtained, both of these curves should then have the same range of linearity.

Let us consider that the conversion efficiency for the two cases. From equation 43 a the conversion efficiency may be written

$$S_1 = \frac{\frac{2\eta I_p}{f_o}(X_1 \; Q^2 s)}{} \qquad 43b$$

$$= \frac{2\eta I_p}{f_o} \frac{(X_2 \; Q^2 s)}{4_p^2} \qquad 43c$$

Where s, as given by equation 47, is equal to $\dfrac{2 a_p}{(1 + a^2) \quad (1 + a^2 p^2)^{\frac{1}{2}}}$

Let the conversion efficiency for the case where $Qk = 2.0$ be S_1' and for the the case where $Qk = 1.4$ be S_1''. Let us assume first that X_1 is designed for its maximum value as permitted by the minimum value of C_{e1} required to minimize variations in primary stray capacity. From equation 43b it then follows that the ratio of the conversion efficiencies is

$$\frac{S_1'}{S_1''} = \frac{(Q')^2 \ s'}{(Q'')^2 \ s''} = \frac{(55)^2 \ 0.286}{(40)^2 \ 0.493} = 1.1$$

If, on the other hand, X_2 is designed for its maximum value as permitted by minimum C_{e_2} required to minimize secondary strays, the ratio of conversion efficiencies as given by 43_c becomes.

$$\frac{S_1'}{S_1''} = \frac{(Q')^2 \ s' \ (p'')^2}{(Q'')^2 \ s'' \ (p')^2} = 1.1 \frac{(0.510)^2}{(0.761)^2} = 0.495$$

In the first case the assumption is made that when C_{e1} has its minimum permissible value the, value of C_{e1} as demanded by p, will be sufficient to minimize the effects of secondary stray capacity. In the second case, the assumption is that when C_{e2} has its minimum permissible value, the value of C_{e1} demanded by the p factor, will be sufficient to swamp out variations in primary stray capacity.

It appears from this that the optimum design conditions will be obtained when p is so chosen that the ratios of stray capacity to total tuning capacity are the same for primary and secondary circuits and when both equivalent capacities are minimum. Sturley gives as an optimum value of p as 0.667 corresponding to a capacity ratio of 1.77. Since primary strays are considerably greater than secondary strays, this is probably a good approximation to the best operating conditions.

As a practical matter, since the secondary tuning is the major factor in determining the point of zero D.C. output, the use of smaller p factors than that required for equal primary and secondary tuning stability may be desirable. This will result in less conversion efficiency but increased secondary tuning stability and therefore greater stability of the zero output point.

Another important consideration is the effective Q demanded by the linear range of x. In the case we are discussing an effective Q of 55 may be impossible to obtain using subminiature coils and a shunt loaded discriminator. Qk can still be adjusted to 2.0 through increase of k but the linear range will be extended beyond that required and the conversion efficiency will be lower .

To continue with the design procedure let us suppose that the curve for $Qk = 1.4$, $p = 0.761$ is chosen as the design curve.

Since $Q = 40$ then $k = 1.4/40 = 0.35$

The design parameters b and d may now be calculated

Here $b = 2k_p = 0.0533$

$$d = \frac{4_p^2 - 2k_p}{1 - 2k_p} = 2.40$$

The estimated or measured stray capacities are

C_{s1} = Coil distributed capacity (estimated) plus valve output capacity (measured or estimated) plus wiring capacity estimated = 12 uuf say

C_{s2} = Secondary coil distributed capacity and wiring capacity = 5 uuf

C_{s3} ≠ Earth capacity of top rectifier circuit 3 uuf.

C_{s4} = Earth capacity of bottom rectifier circuit 3 uuf.

Therefore from equations 68 and 67

$C_{s5} = 12 + 3 = 15$
$C_{s6} = 5 + 1.5 = 6.5$

Suppose as first step C_1 be chosen = 47 uuf.

Then from equation 80

$$C_4 = \frac{C_1 + C_{s4} + C_{s5} - d\,C_{s6}}{d} = 20.6 \text{ uuf}$$

From 81

$$C_m = \frac{1 - b}{b}\,(C_1 + C_{s4})\left(1 - \frac{dC_{s6}}{C_1 + C_{s4} + C_{s5}}\right) = 675 \text{ uuf}$$

From 62 and 63

$$C_2 = \frac{2C_4\, C_m}{C_4 + C_m} = 39.9 \text{ uuf.}$$

$$C_3 = \frac{2C_4\, C_m}{C_m - C_4} = 43.2 \text{ uuf.}$$

From equation 64

$$C_{e1} = 1 - b\,(C_1 + C_{s4} + C_{s5}) = 61.5 \text{ uuf}$$

From equation 65

$$L_1 = 1/w_o^2\, C_{e1} = 22.2 \text{ uh}$$

$$L_2 = 4_p^2\, L_1 = 51.6 \text{ uh}$$

To consider the loading effects let us suppose the initial coil Q of the primary and secondary coils is 100. If we assume, to simplify the calculation, that plate resistance is infinite, and that the A.C. and D.C. resistance which will give the required primary Q of 40 is given by equation 63.

$$R_1/6 = \frac{Q_{o1}\, Q_1\, w_o\, L_1}{Q_{o1} - Q_1} \quad \text{and } R_1 = 6\,\frac{(100)(40)(6.28)(4.3)(51.6)}{60}$$

$$= 240{,}000 \text{ ohms.}$$

With this load resistance the damping effect of the secondary circuit becomes

$2\, R_1/3 = 160{,}000$ ohms.

From equation 65 required damping resistance for a secondary Q of 40 is

$$R_{2e} = \frac{(100)(40)(6.28)(4.3)(51.6)}{60} = 93{,}000 \text{ ohms}$$

An additional secondary damping resistance is then required of value

$$\frac{(160)(93)K}{67} = 220{,}000 \text{ ohms.}$$

This is an oversimplified solution to the loading problem, neglecting the true detection efficiency, primary plate circuits loading and the fact that there may be other restrictions on the load resistance. These effects will be discussed more fully under the section on experimental work.

4.0 EXPERIMENTAL WORK

4.1 SPECIFICATIONS

The experimental work was mainly directed towards testing the shunt coupled discriminator described in section 3.0. No definite specification was given for the discriminator performance and a rather arbitrary one was set up by the writer from what was known of the system requirements.

4.1.1 BANDWIDTH

The intermediate frequency, nominal maximum frequency deviation and nominal audio band were given as 4300 kilocycles, 15 kilocycles and three kilocycles respectively. Taking the modulation index as five, the bandwidth for sidebands of amplitude in excess of one percent of the unmodulated carrier is given by Corrington (43) as $2f_d\ (1 + 0.7) = 30\ (1.7) = 51$ kc/s.

Some tolerance must be provided in bandwidth for the mistuning of tuned circuits due to temperature and humidity effects. It may be expected that frequency deviations in excess of fifteen kilocycles per second will be obtained in some cases. Further, the operation of a number of sets on the same frequency, with their transmitters controlled to a degree determined by the receiver tuned circuit stability will require additiotolerance inn bandwidth. It was estimated that a bandwidth of about 70 kc/s at 3db down would be a fair figure for these portable transceivers. Since the shunt coupled discriminator has a certain amount of asymmetry, it was desided to design for ± 40 kc/s linearity to ensure that linearity would be obtained over the receiver bandwidth. The peaks of the discriminator would then be effectively fixed by the actual receiver selectivity.

41.2 DISTORTION

The distortion limit was arbitrarily taken as two percent, with a modulation frequency of 400 cycles per second and a deviation plus or minus 35 kilocycles per second. With the high modulation index given by this combination, the bandwidth is effectively 70 kilocycles. The limit of two percent distortion on the discriminator characteristic was considered reasonable since distortion will also be introduced by the receiver audio system and by the transmitter. The system distortion should not exceed about ten percent and a two percent limitation on discriminator distortion will allow reasonable receiver and transmitter distortion limits to be set.

41.3 LIMITING

It was decided to use grid limiting with the limiting voltage level taken as one to two volts on the limiter grid. This was based on an arbitrary limiting signal voltage level of one to two microvolts and a receiver gain, from antenna to limiter grid, of 120 db. In the final analysis if more limiting than this is required, additional grid limiting may be used in other intermediate frequency amplifier stages and plate or dynamic limiting may be included to reduce the effects of impulse noise. Grid limiting was felt to be preferable for ease of maintaining high discriminator primary Q which may otherwise be made difficult in view of the shunt loading effects.

4.1.4 CONVERSION EFFICIENCY

The conversion efficiency was not specified. The degree of transmitter control has been seen to depend upon the slope of the discriminator characteristic but it is impossible to fix a discriminator requirement based on this consideration without involving the complete design of the transmitter, which is felt to be beyond the scope of this paper. With the limiting drid voltage and distortion limit of the discriminator fixed, the maximum conversion efficiency will, in any event, be fixed by the minimum permissible discriminator tuning capacities, as seen from section 3.12. The sensitivity of the transmitter reactance valve might then be adjusted to give the required degree of frequency control.

4.2 COMPONENTS

The components used in the experimental work by the writer were, so far as possible of a subminiaturized nature.

4.2.1 TUBES

The limiter and intermediate frequency amplifier tube used in the tests by the writer was the CK5678. This is the most satisfactory sharp cut off subminiaturized frequency pentode thus far developed in the 1.3 volt filament series. The tube closest to the CK5678 in performance in the 1.3 volt series is the 2E31 with about half the transconductance and twice the grid-plate capacity of the CK 5678.

4.2.2 COILS

The coils used in the test were supplied by CSRDE and the construction of the coils is illustrated in Fig. 25. Two types of powdered iron core and shell were experimented with, Crowley Type 5HF and Type 7HF, the Carbonyl equivalents being E and TH respectively. The Carbonyl TH was found to give somewhat better results at 4.3 megacycles than the Carbonyl E, with the particular coils used.

The coils were initially intended for use at two megacycles and the stationary universal winding was used with 3/41 Litz wire. The advantage of Litz in reduction skin effect at 4.3 megacycles is questionable but it was hoped by the writer that the coils could be adapted for use at that frequency. The Q values obtained were quite reasonable but the use of these coils was largely a matter of convenience and further experimentation with different wire sizes and form factors would be necessary to determine the optimum coil dimensions.

Fig. 27 shows how the apparent inductance and Q of the coil vary with the number of turns and slug position for the Carbonyl E core and shell. These curves were obtained using a Bonton Q-Meter Type 160-A. Resonance was obtained by varying the frequency, with a fixed tuning capacity of 50 uuf, and the frequency range so covered extended from about 1.4 to 23 megacycles. The true Q and L may therefore be considerably different from these figures due to change in the effective iron permeability and due to distributed capacity affects. The curves were found useful however in predicting the design of other coils.

The position of maximum inductance was obtained by positioning the core for minimum resonant frequency for any given number of turns, The position of minimum inductance was fixed by the tuning the core in as far as permitted by the physical construction of the coil.

It is evident from these curves that the tuning range of the coils is much greater than necessary for trimming an intermediate frequency circuit at 4.3 megacycles. The required tuning range is unlikely to be more than 20% in inductance or about ten percent in frequency. It was found possible to shorten the coil forms and therefore the travel of the tuning slugs by half inch and still maintain adequate tuning range.

Using the Carbonyl TH iron the apparent Q and inductance were again measured for various turns. (Fig.24) The measurements in this case were made only for the case of maximum inductance and Q.

Comparing the curves of Figs. 27 and 28 for the apparent maximum inductance and Q it may be seen that the Carbonyl TH gave a better Q and inductance for the same number of turns.

Fig. 28 shows also a further set of Q and inductance curves based upon correction of the apparent Q and inductance for distributed capacity effects. The distributed capacity was measured by adjusting the resonant frequency of a standard Q meter test coil until no change in the resonant frequency was observed when the unknown coil was connected across the Q-meter condenser terminals. (49)

Under this condition the unknown coil is resonating with its distributed capacity C_d and the true inductance L is then given by

$$L = 1 / 4\pi^2 f_2^2 C_d$$

From measurements of the apparent coil inductance we have

$$L = 1 / 4\pi^2 f_1^2 (C_1 + C_d)$$

Where C_1 is the Q meter tuning capacity. The true inductance, assuming that L is constant, is then given by

$$L = \left[1 - \left(\frac{f_1}{f_2}\right)^2\right] / 4\pi^2 f_1^2 C_1$$

The method assumes that the true inductance remains constant with frequency over the range $f_2 = f_1$. For iron core coils this may not be entirely accurate if the frequency differential is very great. It was observed that the distributed capacity as measured in this manner did not change very radically with the number of turns. This suggests that the true inductance may increase at the higher frequencies and / or that a large percentage of the distributed capacity lies in the core and shell material. It has been found that that the distributed capacity of carbonyl iron cores may be appreciable since the cores are in effect small particles of highly conductive material separated by infinitely thin layers of insulation. (38) The assumption that the inductance does not change with frequency leads to values for the distribute capacity of from four to six uuf, which appears to be reasonable for this type of multi layer coil.

The true Q of the coil is then given by

$$Q_t = Q_a((C_1 + C_d)/C_1) = Q_a f_2^2/(f_2^2 - f_1^2)$$

4.2.3. RECTIFIERS

For rectification purposes, Germanium crystal diodes type 1N34 were used. (40) Seperate cathode dual diode vacuum tubes have not yet been made avaible in the subminiature 1.3 volt filament series although single diodes of a size comparable with the germanium diode have been produced in the 6.3 volt series.

Germanium crystals have some advantages over vacuum tube diodes in that no filament voltage is required and the forward conductance is generally higher then in the vacuum tube diode. Contact potential effects in the diode elements are also eliminated, which removes a source of static unbalance between the two halves of the circuit.

Offsetting these advantages is the fact that crystal diodes exhibit a finite back resistance and this resistance is subject to variations with changes in the ambient temperature and impressed voltage. This effect may be very detrimental to maintaining a good balance over the wide range of temperatures through which the discriminator must have normal operation.

Fig. 29 shows reverse current voltage characteristics for 6 crystals of random selection at temperature of 23, 40 and 60 degrees centigrade. The method of measurement is indicated in Fig. 30. The temperature of the oven was maintained at the operating temperatures for a period of three hours so that thermal stability could be assured.

The results are not intended to be conclusive but they di indicate that considerable unbalance of the D.C. back resistance may exist between units selected at random. The difference in the current scales should be noted It appears that the six crystals fall roughly into three divisions of reverse resistance. Crystals 1 and 2 have regular low reverse resistance even at room temperature. The back resistances at 23 °C with an impressed voltage of twenty volts are 670 and 400 kilohms respectively, falling off to 330 and 220 kilohmsa at 60 °C. The corresponding reverse resistance for cyrstals No's 3 and 4 are 5.7 and 2.7 megohms at 23°C falling off to 720 and 930 kilohms at 60°C, and for crystals 5 and 6 are 1 and 1.3 at 23°C falling off to 280 and 400 kilohms at 600°C.

It appears that crystals 1 and 2 are considerably worse than average even at room temperature and would probably be discarded on the basis of normal testing. The variations in back resistance of the other units with temperature changes could scarcely be anticipated however on the basis of measurement at room temperature.

The forward resistance of the crystals is said to be little affected by increases in temperature less than one hundred degrees centigrade but may double with a temperature drop of about 70°C. This effect however is not nearly so important as the decrease in back resistance.

Balance of the crystal characteristics must also be obtained, of course at the frequency of operation so that the problem of getting balanced crystals over the temperature range is more complex than indicated by the direct voltage measurements.

In the absence of a suitable vacuum tube rectifier it was necessary to proceed on the assumption that balanced units could be produced with maintenance of high back resistance under temperature variations.

.2.4 RESISTORS AND CONDENSERS

The writer was not able to obtain subminiaturized resistors and only a very few subminiaturized condensers. The condenser problem may create some added difficulties since the maintenance of the correct coupling coefficient depends upon stable condensers of rather critical values if trimming of the condenser is not to be used.

4.3 MEASUREMENT OF THE DISCRIMINATOR CHARACTERISTICS

The test setup which was used to measure the discriminator characteristics is illustrated in Fig. 31. Two signal generators were used. Static characteristics were obtained with a General Radio 605B Signal Generator Serial 1968 and a Boonton FM Signal Generator Type 150-A Serial No. 870 was used in measuring the dynamic characteristics. The FM Signal Generator could not be conveniently used for the static measurements because of the contracted frequency scale.

The vernier dial of the 605B Signal Generator was calibrated against a Ferris Crystal Calibrator Model III at 10 kilocycle points over a frequency range of 320 kilocycles by beating the two signals together in a Hallicrafter Communication receiver. This calibration is shown in Fig. 32. The calibration was checked several times during the period of experimental work and the variation from the curve shown here was negligible. The vernier dial spacings are approximately equivalent to 4 kilocycles but interpolation to the nearest half division was carried out throughout the measurements and the error in frequency is estimated at less than 2 kilocycles.

It may be seen from Fig. 31 that a stage of intermediate frequency amplification was included in the test setup. The maximum voltage obtainable from either of the signal generators is approximately one volt and a stage of amplification was included to ensure that limiting voltage could be obtained. The amplifier was purposely made broad band so that its effects on the range of discriminator linearity would be negligible.

The distortion characteristics were obtained by measuring the relative amplitudes of the fundamental and harmonic components of the audio output with a general Radio Wave Analyzer Type 736A. The total distortion was taken as the percentage ratio of the square root of the sum of the squares of the harmonic amplitudes to the fundamental amplitude. In most cases it was not necessary to consider harmonics beyond the third in these measurements.

A cathode ray oscilloscope connected across the output was found useful for studying the output wave form and for making preliminary alignment.

The static characteristics were measured with a Barber V.T.V.M. Model 27 Serial 549, which has a DC input impedance of about seven megohms.

The input voltage to the limiter was measured with a General Radio V.T.V.M. Type 726A connected across the amplifier tuned circuit. The capacity of the instruments had some effect upon the amplifier selectvity and it was left in position throughout the measurements. This served also as tuning indicator for the amplifier circuit.

The plate and screen currents and the filament current of the tubes were metered to ensure that the tubes were being operated within the rated value

4.4 Test Circuit No. 1

The first circuit which was tested is illustrated in Fig. 33. The amplifier design will be considered first since this was the procedure followed in the experimental work.

It was found useful to establish arithmetic values for the reactance of unit inductance and capacitance at 4.3 mc/s since these were used continually in the design work. These values are indicated here.

1 microhenry at 4.3 mc/s has 27 ohms reactance.

1 micromicrofarad at 4.3 mc/s has 37 kilohms reactance.

For resonance at 4.3 mc/s it follows that, if C is in uufs and L in uhs, $L = 1370/C$.

4.4.1 Amplifier Design

The amplifier was designed to have a 3 db bandwidth of 400 kilocycles. The gain for a single stage tuned amplifier is given by Sturley (45) as

Gain $= \dfrac{g_m Q X}{1 + jx}$ where g_m is the valve transconductance and the other symbols have the meaning previously assigned.

The effective Q for a bandwidth of 400 kc/s at 3 db down is then obtained when $x = 1$ or

$Q = 4300/400 = 10.7$.

A fixed tuning capacity of 39 uuf was chosen. The coil used was a 40 turn, 3/41 Litz universal wound coil with the tuning core and shell of Type 7HF iron. Tuning was accomplished at 4.3 mc/s with the core detuned considerably from the centre of the coil. The coil was then checked for this tuning condition on the Boonton Q meter and tuned to 4.3 mc/s with a Q meter tuning capacity of 51 uuf and an apparent Q of 40. This indicated a stray capacity effect for the tube, vacuum tube voltmeter, and wiring, of 12 uuf.

To obtain an effective Q of 11 with a total effective capacity of 51 uuf then requires an additional damping resistance of value given by equation

$$R = \frac{Q_0 Q_1}{Q_0 - Q_1} \frac{1}{\omega_0 C} = \frac{40 \cdot 11}{29} \left[\frac{37000}{51}\right] = 11{,}000 \text{ ohms}$$

A damping resistance of 10,000 ohms was used across the tuned circuit and the bandwidth of the amplifier as given by the frequencies at which the voltage across the tuned circuit is 3 db down was 380 kc/s.

The voltage gain of the amplifier was only about six and might have been increased by reduction of the grid bias resistor and/or increase in the plate voltage, which was set at 45 volts. The gain was more than ample to provide limiting voltages from either of the signal generators and it was decided to let it stand.

4.4.2 Approximate Discriminator Design

The first discriminator circuit shown in Fig. 38 was set up, before making an extensive analysis of the circuit, to check the alignment procedure and operating characteristics. The primary to secondary capacity ratio was chosen rather arbitrarily as two to one and from Sturleys paper (24) it was estimated that the required Qk factor should be between 1.4 and 1.5 for a linear range of x of approximately 0.8.

The condensers C_1, C_2 and C_3 were surplus service stock of unknown vintage which were tagged as nominal 27 uuf. Measurements showed them to be approximately 25, 25 and 26 uuf respectively. The coupling condenser C_m was a nominal 470 uuf mica condenser of 20% tolerance.

The apparent coupling coefficient, neglecting stray capacities, was then anticipated as approximately

$$k \cong \frac{\sqrt{C_1 C_4}}{C_m} \quad \text{where } C_4 = \frac{C_2 . C_3}{C_2 + C_3}$$

$$K \cong \frac{\sqrt{25.(12.7)}}{470} \cong 0.0380$$

For a Qk factor of 1.4, the required circuit Q was then $\frac{1.4}{0.0380} = 37$ and then linear range of frequency under these conditions might then be expected to be, from $\Delta f = \frac{x F_0}{2Q}$, approximately . 45kc.

The data on the coils used is indicated in Fig. 38. The coils were purposely designed for rather large inductance with maximum core insertion in order to provide a wide tuning range about the resonant frequency of 4.3 mc/s.

The discriminator was aligned by static measurements by tuning of the primary and secondary coils. The primary coil was first tuned for maximum voltage across one of the diode load resistors with 4.3 mc/s applied to the input. The secondary was then tuned for zero output across the two load resistors. The primary was retuned for equal off tune peaks so far as could be obtained and the secondary was finally readjusted for zero output at 4.3 mc/s. This final readjustment to the secondary circuit was found to be of a very minor nature, the tuning of the primary circuit having little effect upon the cross over point.

The limiter voltage was fixed at 6.5 volts and its signal input at the limiter grid of 2 v rms was used in determining the discriminator characteristic. To obtain a limiter characteristic, the signal generator output was varied and the output across one of the load resistors was observed for various recorded values of limiter grid voltage.

The characteristics so obtained are shown in Fig. 34. The Boonton FM Signal Generator was not yet available and a distortion characteristic could not be made. It is evident from the discriminator static characteristic, however, that considerable nonlinearity and unbalance is present even at small frequency deviations.

The limiter characteristic shows that limiting is effective at about 1.5 v limiter grid voltage, but the drooping voltage characteristic at higher grid voltage levels indicates that amplitude modulation would not be entirely suppressed at higher input voltages.

4.4.3 Effect of Increasing Load Resistors

As a first step in obtaining greater conversion efficiency through maintenance of a higher primary effective Q, the load resistors were increased to one megohm. The resultant discriminator characteristics for a limiter grid voltage of 2 volts and various plate voltages is shown in Fig. 35.

It is evident that considerable improvement in conversion efficiency is effected by the use of higher load resistors. Some improvement was also effected in the symmetry and balance of the circuit. It was possible to obtain as much output with a plate voltage of 45 v as was previously obtained with 65 v.

Fig. 36 shows a set of limiting characteristics obtained for the various plate voltages with this load condition. It may be seen that lowering the plate voltage and screen voltage has the effect of lowering the input level at which limiting occurs. The curve for plate and screen voltages of 25 v shows maximum flatness above the limiting level indicating that grid limiting is being accompanied by plate current saturation. The discriminator output for the lower values of plate voltage is also lowered of course.

4.4.4 <u>Effect of Increasing the Coil Q's</u>

The coils used in the circuit had purposely been designed with a higher value of inductance for maximum core insertion than had been anticipated since it was easier to lower the inductance by tuning the cores than to increase it by adding more turns.

This results, however, in a considerable lowering of the coil Q if the cores are very far out of the centre region of the coils. The primary and secondary coils were removed from the circuit and the inductances measured, and the input and output capacities measured on the Q meter.

The primary coil when measured on the Q meter tuned to

4.3 mc/s with a tuning capacity of 33.7 uuf and an apparent Q of 40. This indicated that the required tuning inductance was $1370/33.7 = 40.6$ uh. The secondary coil could not be resonated at 4.3 mc/s due to the minimum Q Meter capacity, but tuned to 3.25 mc/s with 29.8 uuf, indicating an apparent inductance of 80 uh and Q of 48. The measured distributed capacity was 4.8 uuf, which indicates that the true inductance of the coil is $25300/3.25^2(34.8)$ or 68.7 uh.

The secondary coil was trimmed to 56 turns and with the core approximately centred in the shell tuned to 3.12 mc/s with a Q meter tuning capacity of 30 uuf. The measured distributed capacity was 4.5 uuf, so that the true inductance of the coil for this maximum inductance condition was 75.2 uh. This was about 8% higher than the required true inductance and was close enough to give adequate tuning range without too much decentering of the core. The apparent Q at 3.12 mc/s was 73.

The primary coil was also trimmed to give an apparent inductance of 45 at 4.3 mc/s with the core centred, but the apparent Q obtained was only 55 so it was decided to use a new coil for the primary. A new primary coil using the Type 7HF iron core and shell with 38 turns of 3/41 Litz was used. This coil had an apparent inductance of 45 uh and apparent Q of 85 at 4.3 mc/s. This apparent inductance was about 12% higher than required for tuning the primary circuit as indicated by previous measurements but close enough to give trimming range without excessive core decentring.

These primary and secondary coils were used in the circuit of Fig. 33, and the circuit modified slightly as indicated in Fig. 37 for measurement of distortion characteristics. The Boonton FM Signal Generator was obtained on loan from RCA Victor at this time through the good offices of Dr. F. S. Howes, and it was possible to make some distortion measurements. The network R_4C_7 was used to introduce a de-emphasis of 100 microseconds to correspond with the pre-emphasis factor used in the Signal Generator.

The static and distortion characteristics so obtained are indicated in Fig. 38.

The static curve is seen to be a definite improvement over that of Fig. 36 showing somewhat more conversion efficiency, more symmetry and better linearity. The distortion measurements indicate, however, that the characteristic does not meet the requirements for permissible distortion. The total distortion curve shown is compounded of almost equal parts of second and third harmonic indicating that the curve is both asymmetrical and non linear.

The methods used thus far in the experimental work showed that the circuit had some possibilities, but that some refinement of the design procedure would be necessary in order to obtain the required coupling components and loading effects. As a first step in this direction it was decided to make some measurements of the stray capacity effects.

4.4.5 Measurement of Stray Capacities

Previous measurements indicated that some consideration would have to be given to the effects of stray capacity on the circuit if the proper component design was to be carried out. Stray capacities modify the coupling coefficient to some extent, but a more important consideration is the necessity for maintaining high circuit Q with permeability tuning. High Q requires that the core be fairly close to the centre of the coil at the resonant frequency and yet not so close to the condition of maximum inductance that adequate tuning range cannot be obtained on the low frequency side of resonance. It is necessary, therefore, that the total tuning capacity be known so that the correct tuning inductance can be designed to have high Q at the resonant frequency.

For the measurements of stray capacity the test set up indicated in Fig. 39 was used. Capacities were measured by a substitution method, the measured capacity being the difference in the Q meter capacity required to resonate a standard Q meter test coil to the same frequency, with the unknown connected and disconnected.

The Q meter was arranged in close proximity to the test chassis and connections made to the required test points by means of spring clips soldered to two-inch leads of No. 18 magnet wire. In making measurements the grounded terminal of the Q Meter and chassis was left connected and the high terminal disconnected at the test point. The capacity effects of the test leads were thus included in both measurements and could be ignored for all practical purposes.

Measurements of this nature were made on the circuit of Fig. 37. The equivalent primary tuning capacities was measured, with the primary and secondary coils open, for the following conditions and results:

	Equivalent Cel
(1) Circuit as in Fig. 39	34.6 uuf
(2) Tube removed	29.0
(3) Circuit opened at A	27.0
(4) Circuit opened at B	25.7
(5) Circuit opened at C	23.9

The tube and its associated wiring is thus seen to contribute a primary stray capacity of 7.6 uuf, the top diode load circuit a stray capacity of 1.3 uuf and the bottom diode load a stray capacity of 1.8 uuf.

A similar set of measurements was made on the secondary circuit with the Q Meter connected to test points A and B with the coils again disconnected. A common ground potential between the Q Meter and the chassis could not be established in this case since the secondary is above the chassis potential. To obtain stable measurements it was found necessary to ground the Q Meter and allow the relatively much smaller test chassis to take up an arbitrary potential.

The measured values of C_{e2} with the coils disconnected was observed to be 15.0 uuf. With the circuit opened at A, no effect on the secondary tuning was observed. With the circuit opened at B or C the apparent tuning capacity was observed to be 13.2 uuf, indicating that the shunting capacity of the load circuits is about 1.8 uuf so far as the secondary is concerned. This is less than the effect on the primary because the load distributed capacities are effectively in series as far as the secondary tuning is concerned.

4.5.1 <u>Test Circuit No. 2</u> -- <u>Capacity Tuning</u>

For experimental purposes permeability tuning is rather inconvenient since any appreciable variation in the coupling components requires redesign of the coils to obtain high Q with the new required inductance values. Since it was desired to provide maximum flexibility to the experimental work, the circuit of Fig. 40 was set up to permit capacity tuning of the circuit.

With capacity tuning, the coils may be designed for high Q with any given values of C_{e1} and C_{e2}. The coupling condenser may then be replaced and the condensers C_1, C_2 and C_3 varied as demanded by the design equations of Section 3.11. These exact equations demand at times rather odd values for the circuit condensers and while these values may be closely approached by commercially available values, the stock available to the writer did not provide much leeway in this respect.

To ensure that the alignment procedure would be no more complicated than with permeability tuning, the procedure followed was to set the capacity of C_2 to the value required by the design equations and to carry out the alignment by means of variation of C_1 and C_3.

4.5.2 Variation of Coupling Condenser

To examine the effects of variation of the coupling condenser C_m, the circuit of Fig. 40, was aligned and measured with four different values of coupling condenser C_m. The coils and the condenser C_2 were left unchanged and alignment was carried out in each case by variation of the condensers C_1 and C_3. With the same inductance in each case it follows that C_{e1} and C_{e2} remain constant. Leaving C_2 fixed for the various coupling condensers results in some unbalance since the correct value of C_2 depends upon C_m, but the results show the effect on the discriminator characteristic of variation of the coupling condenser.

The discriminator static and distortion characteristics are shown in Figs. 41, 42, 43 and 44 for different values of the coupling condenser C_m. Examination of these curves shows that the circuit passes from a condition of undercoupling to one of much overcoupling. A table has been drawn up below to indicate the bandwidth between the peaks and the approximate conversion efficiency expressed in volts per kilocycle for each of the four coupling condensers used.

Coupling Nominal	Condenser Measured	Bandwidth Between Peaks	Conversion Efficiency
1000 uuf	1070 uuf	85 kc/s	0.2 v/kc
700	685	105	0.2
500	490	145	0.1
200	195	400	0.025

4.5.3 Q Meter Measurements

Measurements of the primary and secondary equivalent tuning capacity were made in a manner which has been previously indicated. The measurements were made using a Q Meter standard coil of approximately 25 uh inductance and 6 uuf capacity. With the coils disconnected, it was also possible to get some indication of the loading effects through measurement of the damping of the Q Meter.

The measurements for the circuit of Fig. 41 gave results as indicated below.

Measurements of C_{e1}

Q_1 and Q_2 are the unloaded and loaded Q Meter readings, C_1 and C_2 the Q Meter tuning capacity for the unloaded and loaded condition.

Frequency	Q_1	Q_2	C_1	C_2	C_{e1}
3.06 mc/s	218	100	100 uuf	52.1 uuf	47.9 uuf

For measurement of C_{e2}, the following results were obtained:

Frequency	Q_1	Q_2	C_1	C_2	C_{e2}
3.06 mc/s	218	150	100	78.5	21.5

If we add in the distributed coil capacities as indicated from Fig. 40, the true values of C_{e1} and C_{e2} become:

$$C_{e1} = 47.9 + 5.3 = 53.2 \text{ uuf}$$

$$C_{e2} = 21.5 + 5.2 = 26.7 \text{ uuf}$$

Measurements on the components of the coupling network resulted as follows:

$C_1 = 37.6$, $C_2 = 36$, $C_3 = 37.5$, $C_m = 1070$ uuf

It is desirable to interpret these measurements in terms of the coupling coefficient k, the effective circuit Q and the capacity ratio factor p. This may be done in an approximate manner as indicated below.

4.5.4 Calculation of Coupling Coefficient

The true coupling coefficient is given by

$$k = \sqrt{\frac{C_{e1}\, C_{e2}}{(C_{me})^2}} = \sqrt{\frac{C_{1e}\, C_{4e}}{(C_{1e} + C_{me})(C_{4e} + C_{me})}}$$

where C_{1e}, C_{4e} and C_{me} are the values indicated in the figure 25 b of section 3.10. Substituting for C_{1e}, C_{4e} and C_{me} from equations a, b and c leads to the following expression for k:

$$k = \sqrt{\frac{(C_1' C_4)^2}{(C_1'C_4 + (C_1' + C_{s5})C_m)(C_1'C_4 + (C_4 + C_{s6})C_m}}$$

$$= \frac{C_1' C_4}{C_m}\sqrt{\frac{1}{\left(\frac{C_1'C_4}{C_m} + C_1' + C_{s5}\right)\left(\frac{C_1'C_4}{C_m} + C_4 + C_{s6}\right)}}$$

The diode stray capacities were not itemized in the measurement, but a good approximation for k may nevertheless be obtained.

$$C_{e1} \cong C_{1e} \cong C_1 + C_{s5}$$

$$\text{and } C_{e2} \cong C_{4e} \cong C_4 + C_{s6}$$

Then k is approximately

$$k \cong \frac{C_1 C_4}{C_m}\sqrt{\frac{1}{\left(\frac{C_1C_4}{C_m} + C_{e1}\right)\left(\frac{C_1C_4}{C_m} + C_{e2}\right)}}$$

From the measurements $C_4 = \frac{C_2 C_3}{C_2 + C_3} = \frac{36(37.5)}{73.5} = 18.35$ μμF

$$\frac{C_1 C_4}{C_m} = 0.645 \text{ μμF}$$

$C_{e1} = 53.2$, $C_{e2} = 26.7$ uuf

Then $k = 0.0168 = 0.017$

4.5.5 Calculation of Effective Q

The measurements on C_{e1} and C_{e2} provide some indication of the loading on the primary and secondary circuits. The equivalent damping resistance may be calculated from the apparent damping of the Q Meter.

From the measurements on C_{e1} and noting that the standard test coil has a distributed capacity of 6 uuf, the damping effect on the primary is given by

$$R_{1S} = \frac{Q_1 Q_2}{Q_1 - Q_2} \frac{1}{\omega_0 C} = \frac{(218)(150)(490)}{118} = 91 \text{ Kohms}$$

Similarly from measurements on C_{e2} the damping on the secondary

$$R_{2S} = \frac{218(150)(490)}{68} = 236 \text{ Kohms}$$

The unloaded primary Q is 72 so the effective primary Q is at 4.3 mc/s with tuning capacity of 53.2 uuf is from equation 59a

$$Q_1 = \frac{72}{1 + \frac{(72)(37000)}{(53.2)(236000)}} = 46.5$$

and the effective secondary Q is similarly

$$Q_2 = \frac{81}{1 + \frac{(37000)\,81}{(26.7)(236000}} = 47.6$$

The apparent Q of the circuit may then be taken as approximately equal to 47.

In the circuits of Figs. 41, 42, 43 and 44, the value of C_{e1} and C_{e2} are substantially constant. The coupling will therefore vary in almost inverse ratio to the value of the coupling condenser used. The effective Q of the circuits will not change since the circuit reactance is the same in each case. For the different coupling condensers used then we have approximately the conditions shown in the table below:

C_m	Approximate k	Approximate Qk
1070 uuf	0.017	0.780
685	0.027	1.25
490	0.037	1.70
195	0.094	4.38

From the measurements on C_{e1} and C_{e2} the approximate value of the p factor is $\frac{1}{2}\sqrt{\frac{53.2}{26.7}} = 0.705$. Examination of the curve of Fig. 18 indicates that this value of p requires a value of Qk in the neighborhood of 1.45 for a linear characteristic, so it is not too surprising that poor distortion characteristics were obtained. It appears that the optimum coupling coefficient for the particular value of p used in these circuits lies somewhere between 490 and 685 uuf.

The method used in determining the effective Q here is questionable since the voltages applied to the rectifier circuits by the Q Meter are rather small. The maximum injected Q Meter voltage is about 0.02 volts, a current of 500 milliamps through a 0.04 ohm resistor. This voltage even when magnified by the Q of the Q Meter circuit is not more than one to two volts. In operation, the voltages applied to the rectifier circuits are much higher than this.

The forward conductance of the 1N34 crystals is rather low for small voltages less than about three volts so the normal rectification efficiency may be expected to be higher than indicated here with somewhat greater damping effects. Further the tube loading is not taken into account by the Q Meter measurement and under limiting conditions this may amount to a substantial proportion of the primary loading. This surprising thing in the light of these facts is the relatively sma ll value of the damping resistance measured.

4.6 Q Meter Measurement of A C Resistance

Considering equation for the effective primary damping resistance, and neglecting the tube damping, the damping resistance is given by

$R_{1s} = \frac{1}{2}\frac{a R_1}{(2e+1)}$ where R_1 is the DC resistance of one of the load resistors and e is the ratio of AC to DC resistance of R_1. Since R_1 is 2.2 megohms and $f_o = 4.3$ mc/s, it was anticipated that e would be in the neighborhood of 0.5 from curves given by Zepler and Pavlasek. Measurements were made on some composition resistors of different values to determine what value of e might be expected. The table given below indicates the results of determining the value of e through Q Meter damping effects. The measurements were made by connecting the resistors directly across the Q Meter condenser terminals with a standard test coil connected across the coil terminals.

The Q Meter test coil was a 25 uh, 6 uuf distributed capacity coil. The resistors were quarter watt, 10% tolerance carbon composition resistors except for the 1 megohm resistors which were 20 % tolerance. The DC resistance of the resistors was not measured apart from a check on a voltohymst.

Nominal R Megohms	Frequency mc/s	Q_1	Q_2	C_1 uuf	C_2 uuf	Rac Megohms	e R_{ac}/R_{dc}
2.2	4.3	233	177	46.45	46.05	.521	.236
2.2		234	174	46.45	46.0	.485	.220
5.6		235	192	46.3	45.90	.735	.131
5.6		235	192	48.3	46.00	.735	.131
1.0		243	155	48.2	47.8	.291	.291
1.0		243	152	48.2	42.8	. 276	.276
1.0		237	153	48.2	47.7	.294	.294
1.0		236	155	48.2	47.7	.305	.305
1.0		237	150	48.2	47.7	.282	.282

As indicated by these measurements the apparent value of AC to DC resistance is somewhat less than expected.

Terman, however, indicates that carbons resistors of the order of 5 megohms may be reduced to about one megohm at five megacycles so that the values of e shown here are probably fairly accurate (49). Subminiaturized resistors are generally of a metallized type and could be expected to show somewhat better characteristics at high frequencies but unfortunately none were available at this time.

Assuming a value of $e = 0.23$ for the 2.2 megohms load resistors, the damping resistance with a perfect diode with rectification efficiency of one is given by

$$R_{1s} = \frac{1}{2}\,\frac{e\,R_1}{(2e+1)} = \frac{1}{2}\,\frac{0.23\,(2.2)}{1.46} = 0.165 \text{ megohms}$$

The effect of lower detection efficiencies will be to increase this apparent damping resistance and the damping resistance should therefore be expected to be in excess of 165,000 ohms.

The actual measured damping resistance was 91 K ohms, and the only apparent source of this extra loading is the impedance of the crystals themselves, since the de-emphasis network and output measuring instruments were not included in the measurement of primary and secondary damping.

4.7 Q Meter Measurement of Crystal Impedance

In order to investigate the impedance of the germanium crystals, damping measurements were made on a number of crystals using the test set up shown in Fig. 45. The rectifiers were connected in series with a high Q condenser of nominal value 1000 uuf. This condenser Q was checked by first placing it in series with a 3 uuf condenser across the condenser terminals of the Q Meter and the reduction of the effective Q of the Q Meter standard test coil was negligible.

When the crystals were placed in series with the condenser, however, there was a considerable reduction of the effective Q of the test coil in all cases.

If the crystal were a perfect rectifier with infinite impedance in the reverse direction, the condenser would charge, after a few cycles of the radio frequency voltage, to the peak voltage available across the Q Meter. There would be no diode current flowing, therefore negligible damping of the Q Meter circuit. With a finite crystal back resistance, however, the condenser will discharge through the source and the crystal back resistance and power will be dissipated in the crystal circuit, resulting in damping of the Q Meter. The impedance of the crystal may then be measured in terms of the damping of the Q Meter at 4.3 mc/s at least for the low voltage ranges levels provided by the Q Meter.

The manner in which the effective damping resistance of the germanium crystals varies for different crystals, is indicated in the table below.

Note: Q Meter test coil 25 uh 6 uuf distributed capacity.

Crystal No	Frequency mc/s	Q_1	Q_2	C_1 uuf	C_2 uuf	Effective Crystal Impedance
1	4.3 mc/s	238	113	46.6	45.9	151 Kohms
2			113		46.1	151 Kohms
4			187		46.1	615 Kohms
5			130		46.2	190 Kohms
7			120		45.9	170 Kohms
8			144		46.1	256 Kohms
9			165		46.2	380 Kohms
10			198		46.2	826 Kohms
11			128		46.2	195 Kohms
12			160		46.2	342 Kohms

Some of the losses indicated here are of course in the condenser circuit, but these are of a secondary nature. In any case it is evident that there is considerable variation in the effective impedance of the crystals tested here. It is interesting to compare the equivalent damping resistances of Crystals 1, 2, 4 and 5, with their DC back current characteristics shown by Fig. 25. Crystals No. 3 and 6 had unfortunately been lost in the shuffle, but comparison of the other four with their DC characteristics indicates that the equivalent impedance at 4.3 mc/s is definitely related to the DC impedance measured previously, those crystals having the lowest DC back resistance having as well the lowest High Frequency impedance.

The radio frequency impedance measured in this manner is rather low even under the low voltage conditions of the Q Meter circuit. Voltages of the order of 20 volts or more may be expected in operation, and this will have the effect of making the impedance look even lower.

The Q Meter measurements of stray capacity effects indicated that rather close control of the condenser components would be necessary if the required coupling coefficient was to be obtained. At this stage in the experimental work, the theory of section 3.11 had been developed relating the components to the design parameters and including the stray capacity effects. It was felt that the required k and p factors could be designed for quite closely.

The measurements of AC resistance and crystal damping showed that the method of shunt loading and the relatively low back resistance of the crystals would make it difficult to obtain a very high effective circuit Q.

Because of these factors, it appeared that it would be necessary to measure the effective Q under conditions of voltage and loading similar to those to be used in the actual circuit.

This made it necessary to establish the design up to the point of obtaining the required Q and then to adjust the load resistors to obtain the necessary Q.

The manner in which a complete design was carried out is shown in the following section.

4.8 A Working Design Based on Estimated Stray Capacities

4.8.1 Calculation of Components

From measurements already made on the stray capacity effects it appeared that it should be possible to calculate fairly closely the required circuit components from the equations of Section 3.11.

Referring to Fig. 23c, the following estimated values of C_{s5} and C_{s4} were chosen:

$C_{s5} = 5$ (coil) + 5 (tube) + 3 (top load capacity and primary wiring strays) = 13 uuf

$C_{s4} = 5$ (coil) + 2 (wiring and load capacity) = 7 uuf

The ratio of estimated stray capacities is $\frac{13}{7}$ and for equal primary and secondary tuning stability the ratio of C_{e1} to C_{e2} should be of the same order. An approximate value for p is therefore $p = \frac{1}{2}\sqrt{\frac{13}{7}} = 0.690.$

Referring to Fig. 19, it may be seen that the curves for $Qk = 1.5$ require p to be in the order of 0.7. The first condition selected was $Qk = 1.5$, $p = 0.707$, which, as shown by Fig. 19, is slightly over the critical value of p for maximum low deviation linearity but within 2 per cent of linearity to $x = 0.84$.

It was decided to use capacity tuning, partly for ease of obtaining high Q coils with the proper inductance but more for ease of meeting any requirements for odd values of capacity in the coupling network.

For $Qk = 1.5$, $p = 0.707$ the value of x for a two per cent positive departure from linearity is $x = 0.84$. For a frequency departure of 40 kc/s at 4300 kc/s the required value of Q is

$$Q = \frac{0.84\,(4300)}{2\,(40)} = 45$$

then $K = \dfrac{1.5}{45} = 0.0333$

Therefore $k = 0.0333$

$b = 2kp = 0.0471$

$$\frac{b}{1-b} = 0.0494$$

$$d = \frac{4p^2 - 2kp}{1 - 2kp} = \frac{1.953}{0.953} = 2.05$$

With C_{s5} 13, a value of C 50 uuf appears to be a reasonable compromise between maximum reactance and stability. $C_1' = C_1 + C_{s4}$ and the value of C_{s4} is not likely to be more than 2 uuf.

From equation 81

$$C_m = \frac{1-b}{b} C_1' \left[1 - \frac{C_{s6} d}{C_1' + C_{s5}}\right]$$

$$= \frac{50}{0.0494} \left[1 - \frac{7(2.05)}{50+13}\right] = 780 \text{ uuf}$$

The closest values to C_m in the preferred value series are 750 and 820 uuf unless parallel condensers are used. The writer had a tubular ceramic condenser of 820 uuf, and it was decided to use this.

This made it necessary to recalculate C_1 from the above equation. Solving for C_1 in terms of C_m leads to

$$(C_1')^2 + C_1' \left[C_{s5} - C_{s6} d - \frac{C_m b}{1-b}\right] - \frac{C_m b}{1-b} C_{s5} = 0$$

or $$C_1' = q + (q^2 + h)^{\frac{1}{2}}$$

where $$q = \frac{C_{s6}d + C_m b}{1-b} - C_{s5}$$

$$h = \frac{C_m b}{1-b} C_{s5}$$

This leads to $$C_1' = 21 + (21^2 + 526)^{\frac{1}{2}}$$
$$= 52 \ \mu\mu F$$

From the equations of Section 3.11

$$C_4 = \frac{C_1' + C_{s5} - C_{s6}d}{d} = \frac{52 + 13 - 14.4}{2.05} = 24.9 \ \mu\mu F \cong 25 \ \mu\mu F$$

$$C_3 = \frac{2C_4C_m}{C_m - C_4} = \frac{2(25)(820)}{795} \cong 51 \ \mu\mu F$$

$$C_2 = \frac{2C_4C_m}{C_m + C_4} = \frac{2(25)(820)}{845} \cong 47 \ \mu\mu F$$

$$C_{e1} = (1-b)(C_1' + C_{s5})$$
$$= 0.953(52 + 13) = 62 \ \mu\mu F$$

$$L_1 = \frac{1370}{62} = 22.1 \ \mu h$$

$$L_2 = 4p^2 L_1 = 44.2 \ \mu h$$

With the component values thus established, it remains to consider the loading problem. It was decided to use crystals Nos. 4 and 10, which Q Meter measurements showed to have high back resistance, at least for the low voltages obtainable by the Q Meter method.

4.8.2 Detection Efficiency

To determine the rectification efficiency and loading effects of those crystals with various load resistors, the circuit of Fig. 46 was used. The load condenser C was chosen as 50 uuf, which was considered large enough for adequate by passing of the intermediate frequency and small enough to offer high impedence to the audio frequencies. It has been shown that providing the ratio $\frac{R}{X_c}$ is greater than about 100, the condenser will have negligible effect on detection efficiency (45) X_c is the reactance of the load condenser at the intermediate frequency and R is the load resistance. With C 50 uuf, $X_c = 740$ ohms and the minimum value of R is therefore about 100,000 ohms if detection efficiency is not to suffer. The coil used was an inductance equal to the primary design value of 22.1 uh.

Detection efficiency measurements were made with load resistances of 0.120, 0.330, 0.470, 1.0 and 2.2 megohms. Measurements were also made with the back resistance of the crystal providing the only load. To determine the rectification efficiency, the DC voltage across the load resistance was measured for various root mean square values of tuned circuit voltage. The level of the tuned circuit voltage was set by adjusting the signal generator output. Detection efficiency was then given by $\sqrt{2}\,\frac{E_{Rms}}{E_{dc}}$ and the results are shown in Figs. 47 and 48 for the two crystals. It may be seen that the detection efficiency is from 60 to 65 per cent and little improvement is obtained by using high load resistance because of the low crystal back resistance.

4.8.3 Measurement of Effective Q

The Q of the circuit varied considerably with the load resistance. To determine the effective Q of the circuit under conditions of limiting, the plate voltage was set at 45 volts and a limiting signal was fed in. The Q of the circuit was then measured by detuning the signal generator and measuring the 3 db bandwidth. The effective Q measured in this way includes the effect of the amplifier selectivity but since this was very broad band, it is of secondary importance. In any case the ultimate discriminator characteristic depends upon the amplifier selectivity so the effective Q would be the same. The effective Q of the primary tuned circuit for the two crystals is indicated below.

Load Resistor	Effective Q Crystal 4	Effective Q Crystal 10	Average	Average Damping Resistance
120k	35	31	33	32k
330	49	46	47.5	65k
470	54	52	53	86k
1000	60	56	58	115k
2200	62	60	61	139k
Infinite	66	62	64	170k

With tube damping only, the average circuit Q measured by the 3 db bandwidth, was 82.

The average damping resistance was calculated from

$$R_{15} = \frac{Q_o Q_{eff}}{Q_o - Q_{eff}} X_1$$

Thus for an effective Q of 33, since $X_1 = \omega_o L = (2\pi)(22.1) = 577$ ohms

$$R_{15} = \frac{(82)(33)}{49}(577) = 32{,}000 \text{ ohms.}$$

For the discriminator circuit the primary damping resistance will be approximately half the value indicated in this table, since the detector circuits are paralleled as far primary damping is concerned. The effective primary Q can then be calculated on the basis of an unloaded circuit Q of 82, X = 577, and damping resistance half the average value indicated. This is shown below.

R	Primary Damping	Effective Q
120k	16k	27
330k	33k	34
470k	43k	39
1000k	58k	45
2200k	70k	49
Infinite	85k	53

These measurements indicated that a minimum load resistance of 1 megohm would be required when using this coil. The secondary damping is much less severe and the secondary Q can be adjusted by additional damping resistors if required.

4.8.4. Test Circuit No. 3

The calculated components were arranged as shown in Fig. 49.

A secondary coil was made up with a true inductance of 44.2 uh, a distributed capacity of 5.0 uuf and an apparent Q of 85. The true Q, corrected for distributed capacity, was 97.

The condensers C_1, C_2 and C_3 were each made up of a 39 uuf ceramic condenser in parallel with a 2-20 air dielectric variable condenser. The condenser C_2 was set at 47 uuf.

The load condenser and the loaded crystals were interchanged as shown to eliminate one condenser. This does not change the action of the circuit. The condenser discharges as in previous circuits through one load resistor in parallel with a crystal back resistance, through the coil and the forward resistance of the other crystal. Because of the finite back resistance of the load crystal, the load resistance even at audio frequencies is considerably less than one megohm.

The de-emphasis network used in other circuits was not included and distortion measurements were made without signal generator pre-emphasis. The pre-emphasis used in narrow band transmitters must be of the order of about three hundred microseconds to be effective and the signal generator could not provide this. To avoid unnecessary "weighting" of the output harmonics by a de-emphasis circuit of 300 microsecond time constant it was decided to leave the de-emphasis out.

The maximum imput impedance of the Wave Analyzer Type 726 A is one megohm. To avoid excessive loading of the discriminator by this instrument, the audio signal was effectively tapped down in approximately a three to one ratio as shown. The measured audio was reduced somewhat from that across the load condenser, but could be measured quite easily with the Wave Analyzer.

The discriminator was aligned by varying C_1 and C_3 as previously indicated.

Fig. 50 shows static and distortion characteristics obtained with secondary damping resistors of 150k and 330k. The more heavily loaded circuit may be seen to provide a somewhat greater working range as indicated by the distortion characteristic but has lower conversion efficiency, about 0.15 volts per kilocycle as against 0.18 volts per kilocycle.

The distortion was less than two per cent with 40 kilocycles deviation in both cases, indicating that secondary damping might be further reduced and still provide the required working range.

Fig. 51 shows the characteristics obtained with no secondary damping resistors. The two per cent distortion limit extends to 40 kc/s deviation and a conversion efficiency of approximately 0.28 volts per kilocycle is obtained. An additional distortion curve was made by using a detuned carrier with deviation fixed at $\pm$ 15 kc/s and a modulation frequency of 400 cps. It may be seen that a drift of 10 kilocycles may be tolerated in either direction.

The distortion consists of nearly equal parts of 2nd and 3rd harmonic distortion out to 30 kilocycles deviation at which point 3rd harmonic distortion causes a sharp break in the distortion characteristic.

4.8.5. Checking Stray Capacity Effects

To check the actual values of C_{s5} and C_{s6} against the assumed values of 13 and 7 uuf, the capacity effects were measured as indicated before. In addition the crystal rectifier circuit capacities to ground C_{s3} and C_{s4} were measured by connecting the Q Meter across each rectifier circuit in series with its associated secondary condenser C_3 or C_2, with the rest of the coupling network disconnected.

The results are itemized below.

Tube & wiring capacity = 7.5 uuf

Diode Capacity C_{s3} = 2.8

Coil = 5.3

Total C_{s5} = 15.6 uuf

Secondary coil = 5.0 uuf

Diode series capacity = 1.3

Total C_{s6} = 6.3

C_1 = 49.3 uuf

Diode capacity C_{s4} = 2.3

C_1' = 51.6 uuf

C_3 = 57.7 uuf

C_2 = 47 uuf

Using $C_{s5} = 15.6$ and $C_{s6} = 6.3$ uuf the required component values were recalculated, and the new components C L and L_2 were, $C_2 = 50$uuf

$L_1 = 21.8$ uuf

$L_2 = 43.6$ uuf

The coils were not changed but C_2 was changed to 50 uuf and the circuit re-aligned.

The characteristics obtained under the changed balance of secondary tuning condensers are shown in Fig. 52. The conversion efficiency was increased slightly to approximately 0.3 volts per kilocycle, the peaks are about 10 kc farther apart and the range of 2 per cent distortion is increased from 40 to 45 kc/s.

It may appear surprising that a change of only 3 uuf in the value of C_2 can produce such a change in the characteristic. This amounts to about a six per cent change, however, and since C_3 must be changed in the opposite direction to maintain resonance, there is a total change of 12 per cent in the level of secondary voltages applied to the rectifier circuits.

It points to the necessity for rather close control of the secondary voltage dividing condensers.

Fig. 53 shows a limiter characteristic obtained for the circuit of Fig. 52. The fundamental of the audio output signal is plotted against the root mean square limiter grid voltage. The characteristic is within one db of limiting for a limiter voltage of less than one volt, but falls off slightly for higher grid voltages.

4.8.6 Design For Optimum Linearity

From the curves of Fig. 19, the condition for optimum linearity with $Qk = 1.5$ calls for a value of $p = 0.685$. It was decided to design for this condition using the same primary coil as in Section 4.8. This means the maximum effective primary Q will be the same. We may expect some reduction in the conversion efficiency because of the lower p.

With $Qk = 1.5$, $Q = 45$, $p = 0.685$, and assuming $C_{s5} = 15.6$, $C_{s6} = 6$. the required values of C_1', C_m, C_2 and C_3 were calculated and the results are shown below.

C_1'	C_m	C_2	C_3
49 uuf	845 uuf	54 uuf	58 uuf

The coil $L_1 = 22.1$ uh and since $L_2 = 4p^2 L_1$, $L_2 = 41.5$ uh.

The same components as in section 4.8 were used. The coil L_2 was used and the inductance value obtained by detuning the slug slightly. There was no loss in apparent Q of the coil for this small change.

A capacitance of 27 uuf was added in parallel with the 820 used in section 4.8 to obtain the required C_m. C_2 was set to 54 uuf and the circuit aligned as before with C_1 and C_3.

The results of distortion and static measurements are shown in Fig. 54. The measured distortion was well within the two per cent distortion limit at 40 kc/s. The conversion efficiency is down to 0.25 volts per kilocycle, a drop of approximately fifteen per cent from that obtained with $p = 0.707$.

The peaks are approximately in the same position which is to be expected since the Qk factor is the same, 1.5 in both cases.

4.8.7 <u>Design for Higher Secondary Stability</u>

It was pointed out in Section 3.12 that the use of higher secondary capacity than necessary to achieve equal primary and secondary tuning stability would provide better stability of the cross over frequency.

The secondary equivalent tuning capacity for the condition $Qk = 1.5$, $p = 0.685$ above is only about 33 uuf.

To increase this capacity and therefore the circuit stability, the circuit was adjusted for the condition $Qk = 1.8$, $p = 0.556$, as taken from Fig 19. The maximum Q was again taken as 45, using the same primary coil and this calls for a value of $k = \frac{1.8}{45} = 0.0465$. The coupling components were calculated assuming $C_{s5} = 16$ and $C_{s6} = 5$ uuf and gave results as below.

C_1'	C_m	C_2	C_3	
49	952	91	99	uuf

C_2 and C_3 were made up with parallel condensers of $47 + 39 + 2 - 20$ uuf. C_2 was set to 91 uuf. The same coil was used in the primary having a value 22.1 uh. The secondary coil required was

$$L_2 = 4p^2 L_1 = 4(0.556)^2(22.1) = 27.3 \mu h .$$

The coupling condenser was obtained by adding a nominal 100 uuf condenser to the 847 uuf used in the previous circuit.

The static and distortion characteristics are shown in Fig. 55 for this circuit. It may be seen that the low value p used in this circuit has had the effect of further lowering the conversion efficiency to approximately 0.2 volts per kilocycle. The distortion meets the requirement for 2 per cent distortion at 40 kc/s.

The peaks of the discriminator are considerably farther apart, the bandwidth between the peaks being approximately 150 kilocycles. This is due, as shown in section 3.7, to the higher value of coupling coefficient.

9.0 Discussion of Results

9.1 Comparison of Figures 52, 54 and 55

A comparison of the distortion characteristics of Figures 52, 54 and 55 shows that in each case the 3rd harmonic distortion rises at small frequency deviations of 5 to 10 kilocycles, falls off to a low value in the intermediate range and rises again at deviations approaching the peak separation.

As shown in Figures 16b and 19, this is indicative of a condition of negative non-linearity and occurs for values of Qk in excess of that demanded by the particular capacity ratio factor p. In the circuit of Figure 52 this was anticipated since the design figures were $Qk = 1.5$, $p = 0.707$ and this value of p as shown by Fig. 18 requires a Qk value of about 1.46 for a linear low deviation characteristic.

It was expected that for the circuits of Figures 54 and 55, the 3rd harmonic distortion would be negligible for small frequency deviations and would increase sharply above a frequency deviation of about 40 kilocycles. While the design for $Qk = 1.5$, $p = 0.685$ gave the best distortion measurements, the low deviation 3rd harmonic present, indicates that Qk is slightly higher than required by the capacity ratio.

When the Qk factor was reduced by damping of the secondary circuit, it was possible to extend the useful range of the characteristic but the 3rd harmonic distortion still peaked at low deviation frequencies indicating that the coefficient of coupling was probably high.

The effect is not serious, however, providing the low deviation distortion is within the distortion limit and, in fact, is advantageous in that it extends the useful working range.

The effect of residual magnetic coupling was thought to be a possible cause of this slight overcoupling. When the coils were shielded with small aluminum cans, however, the principal effect was a reduction in effective Q which lowered the conversion efficiency about 10 per cent and extended the working range slightly.

A further possible source of the low deviation distortion observed here may lie in non-linearity of the crystal voltage current characteristic. The voltage applied to each crystal increases with frequency deviation and a departure of the crystal characteristic from linearity, could result in some frequency distortion of this nature.

The approximations made in the caluclation of the components in Section 3.11 will, as has been shown, lead to some small error, and it may be that the magnitude of this error is of the order of the departure from theoretical results obtained here.

The characteristic of Fig. 55 for $Qk = 1.8$, $p = 0.556$, was expected to give the maximum range of linearity, but the curves show it to have about the same range of 2 per cent distortion as the condition for $Qk = 1.5$, $p = 0.707$. It has a greater range of 3 per cent distortion, however. Examination of Fig. 55 shows that the curve has more 2nd harmonic distortion than found in the other two cases, and this sharply breaking 2nd harmonic distortion is the main reason for the smaller range of two per cent distortion.

In most cases the second harmonic distortion was found to be fairly constant over the working range and to lie between 0.5 and 1 per cent. In the capacity coupled discriminator, it has been shown that some asymmetry of the two halves of the characteristic is inherent and some second harmonic distortion is therefore inevitable.

Second harmonic distortion has been kept low by aligning the circuits for equal positive and negative voltage peaks. This must result in distorting the characteristic in some other manner and probably accounts to some extent for the difficulty in obtaining more exact coincidence with the curves of Fig. 19. It must be remembered, of course, that the curves of Fig. 19 are based on the assumption of a symmetrical discriminator and when applied to the capacity coupled discriminator, the approximation must lead to some error.

9.2 Effective Circuit Q

The principal difficulty which has been experienced with this circuit is the difficulty in maintaining high circuit Q. With shunt loading as used here, it has been seen that the A C resistance of the load resistors can have a severe loading effect upon the primary circuit. In order to obtain the required circuit Q, it has been necessary to use load resistors of a value approaching the D C back resistance of the rectifying crystals.

The unloaded coil Q values obtained could probably be improved with smaller shell construction to provide a more complete magnetic circuit. The clearance between shell and coil used was of the order of 1/16 of an inch, so that there appears to be room for considerable improvement in this respect.

With higher Q coils, it should be possible to use shunt loading and maintain adequate Q with lower values of load resistance. There is a possibility of improving the stability of the circuit balance with lower values of shunt loading resistors since variations in crystal back resistance with temperature can be effectively swamped by low load resistors.

9.3 <u>Tuning of the Circuit</u>

It has been seen to be necessary to estimate or measure the stray circuit capacity effects in order to obtain the required coupling coefficient and circuit balance. With permeability tuning, it is then necessary to employ capacities of rather close tolerance and of value as close as possible to the design value. The writer has found that it is generally possible to calculate a design for components which can be obtained in the preferred value series, with perhaps one of the condensers being obtained through a parallel combination of two preferred values.

Capacity tuning has the advantage that coils may be designed for the required inductance with maximum iron in the magnetic circuit and therefore with maximum Q. The use of toroidal cores could be expected to give a further improvement in Q and better confinement of the magnetic field. Capacity tuning also requires that only two of the network condensers be exact design value, and when one of the secondary condensers is variable, an exact balance of the rectifier voltages is possible. Series loading could also be used with a fixed secondary coil.

Capacity tuning then appears to be more satisfactory for ease of obtaining exact component requirements. From the space factor and temperature stability view point, permeability tuning was more satisfactory.

9.4 The Rectifiers

The germanium crystal temperature and impedance measurements indicated that considerable variation can exist between units selected at random. The finite back resistance and low high frequency impedance made it difficult to maintain high circuit Q with the shunt loaded circuit. The variation in back resistance with temperature can be expected to introduce considerable unbalance in the characteristic unless crystal balance over a wide temperature range can be guaranteed.

A dual diode vacuum tube in the subminiaturized 1.3v series would be very desirable for this application and would probably compare favorably in size with two of the germanium diodes.

9.5 Application of The Circuit To Unitized Construction

The bread board measurements made on the circuit pointed to no reason why it should not be adaptable to the plug-in unit construction described in section 1.6. The physical design and electrical layout of the unit is most important, however, and the writer did not have time to make any extensive investigation of an optimum test unit.

A small test unit (permeability tuned) was built up with a complete limiter and discriminator in a unit of dimensions 1½x2x1 inches. The procedure followed by the writer was to wire in the tube and load circuits, measure the stray capacities to the test chassis, and design the coupling network on the basis of these capacities and selected p and Qk factors. Unfortunately, time ran out before it could be determined whether this design procedure would prove fruitful or not. The shielding walls will alter the stray capacities somewhat and may require consideration in the design calculations, depending upon the electrical layout.

9.6 Limiting

The limiting obtained here was not generally satisfactory, giving a drooping voltage characteristic at high grid voltage levels. Adjustment of grid resistance and acreen voltage is necessary to improve this characteristic, but this must be done for the complete range of probable limiter grid voltage, and may be done more readily with a complete I. F. strip.

10. Conclusions

10.1

The shunt capacity coupled phase discriminator shows good possibilities for use in frequency modulation portable transceivers or broadcast receivers. The application of the circuit to plug-in unit construction seem to be possible but will require further experimentation to evolve a satisfactory layout.

10.2

The design equations developed permit satisfactory calculation of the required components. The equations seem more complex than those normally associated with discriminator design, but this is due to the attempt to compensate for stray capacity effects which are usually ignored from the analytical point of view.

10.3

It is possible to obtain a theoretical relation between the design parameters for a phase discriminator, which will give a characteristic slightly better than indicated by Sturley. In practical application, however, the inherent unbalance of the capacity coupled circuit and the difficulty of obtaining precise Q value appears to wipe out any effective difference in the theoretical optimum conditions.

10.4

The required coupling components circuit balance and Q may be more easily obtained with capacity tuning. Permeability tuning, at least with those components with which the writer is acquainted, provides a better space factor and wider tuning range. Although no temperature measurements were made, it is probable that better temperature stability can be achieved with permeability tuning.

10.5

The Germanium rectifier is unlikely to give satisfactory results in a balanced discriminator operated over a wide temperature range. Shunt loading with much smaller resistances than used here may provide some additional stability with temperature by making the effective back resistance much lower than that likely to be produced by temperature variations. This will result in a much lower conversion efficiency, however.

10.6

Higher values of coil Q than were obtained in these tests would be desirable to provide better conversion efficiencies and production tolerance. The coils used provided little tolerance in effective Q even with large loading resistors. In application to a complete transceiver circuit, the permissible D C resistance in the grid circuit of the transmitter reactance tube may limit the load resistance so that higher Q coils may be even more essential to maintain a reasonable conversion efficiency.

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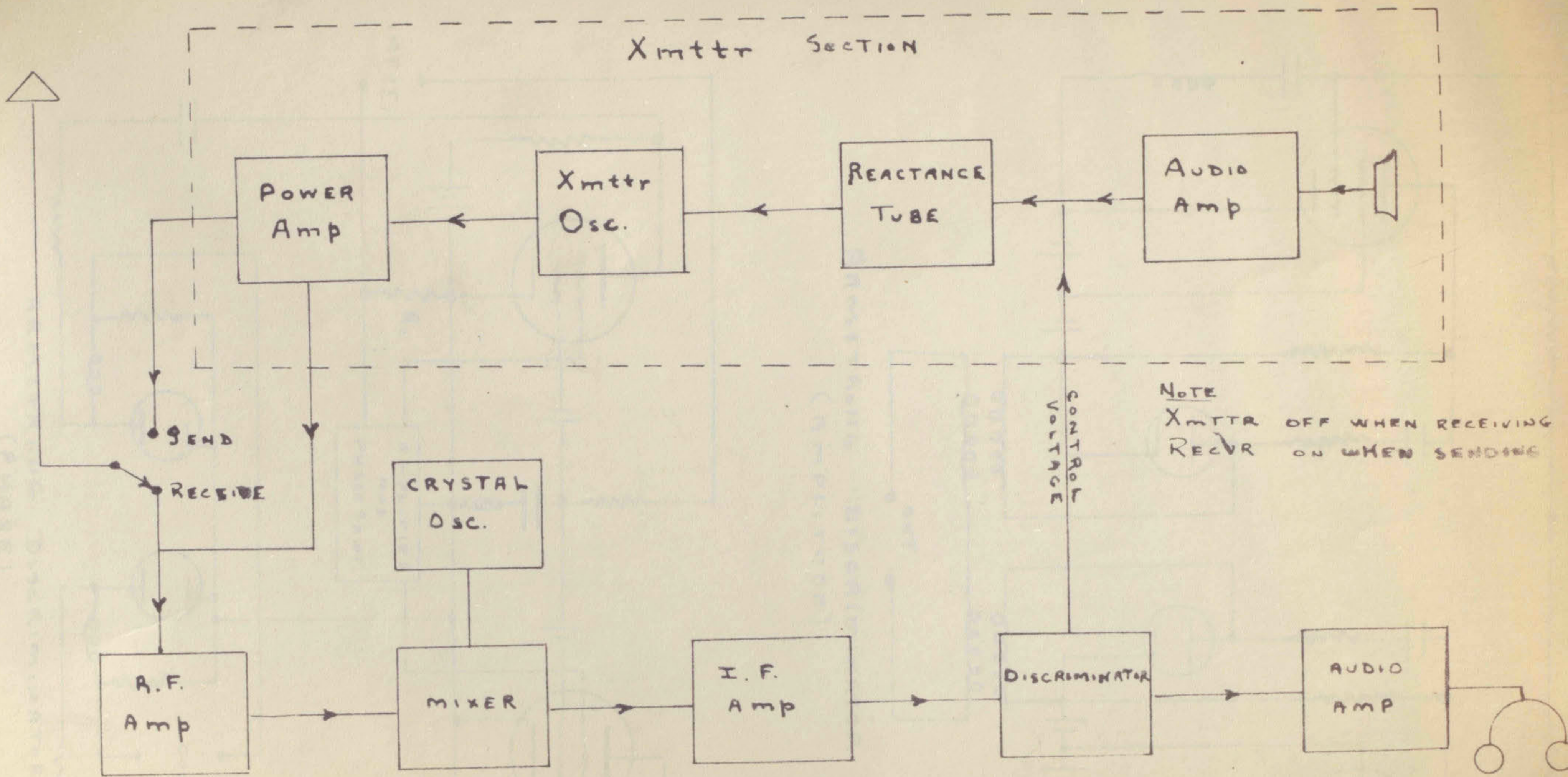


BLOCK DIAGRAM OF SIMPLE TRANSCEIVER

FIG. 1

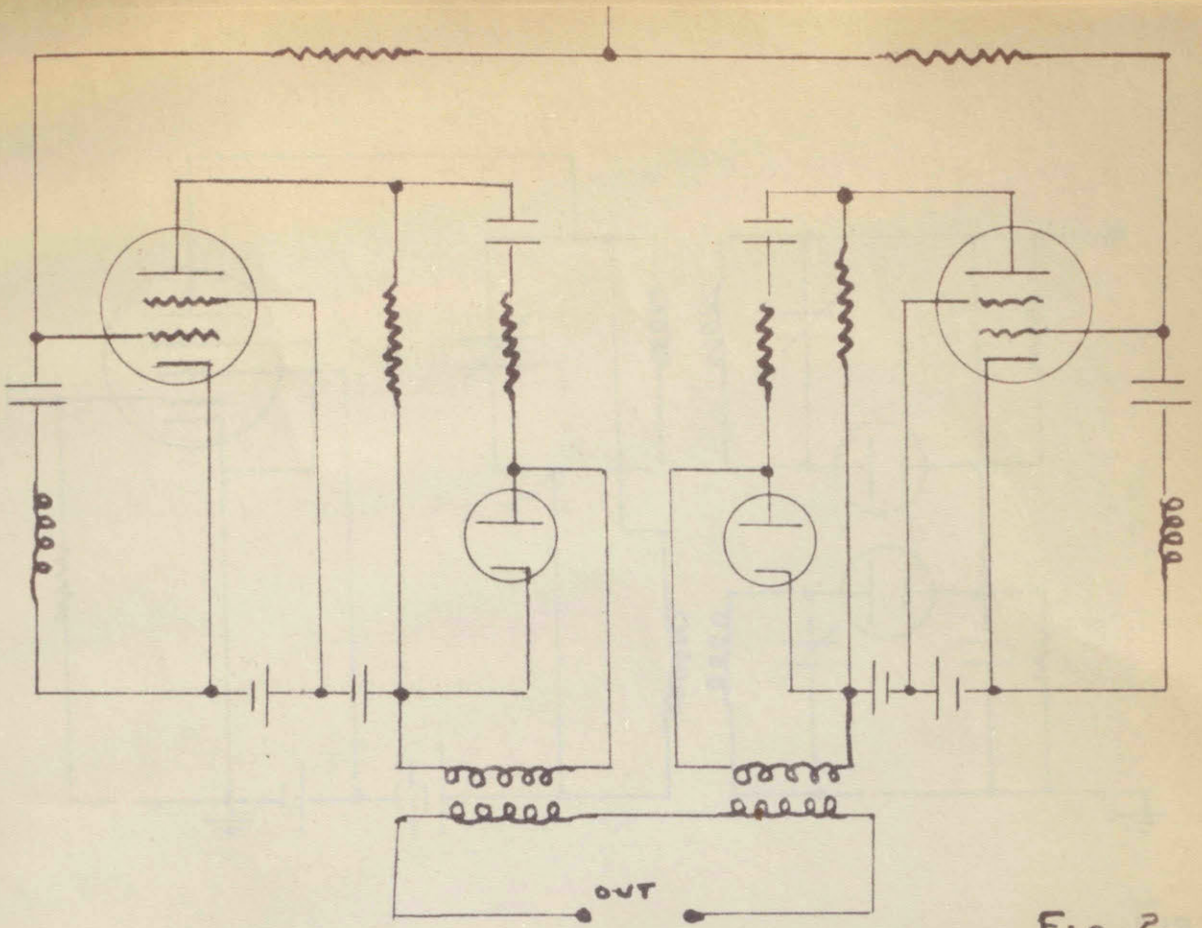
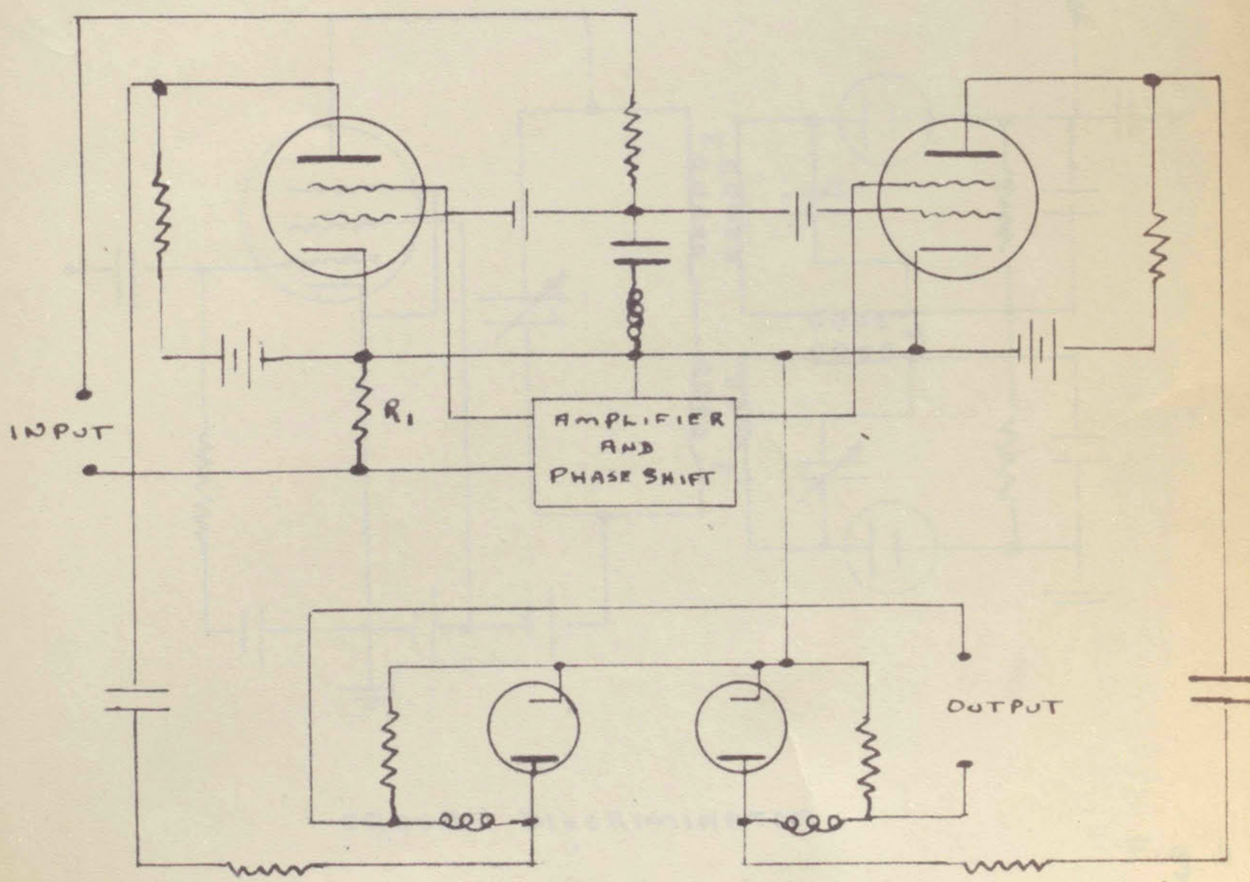


Fig. 2.

ARMSTRONG DISCRIMINATOR
(AMPLITUDE)



ARMSTRONG DISCRIMINATOR
(PHASE)

FIG. 3

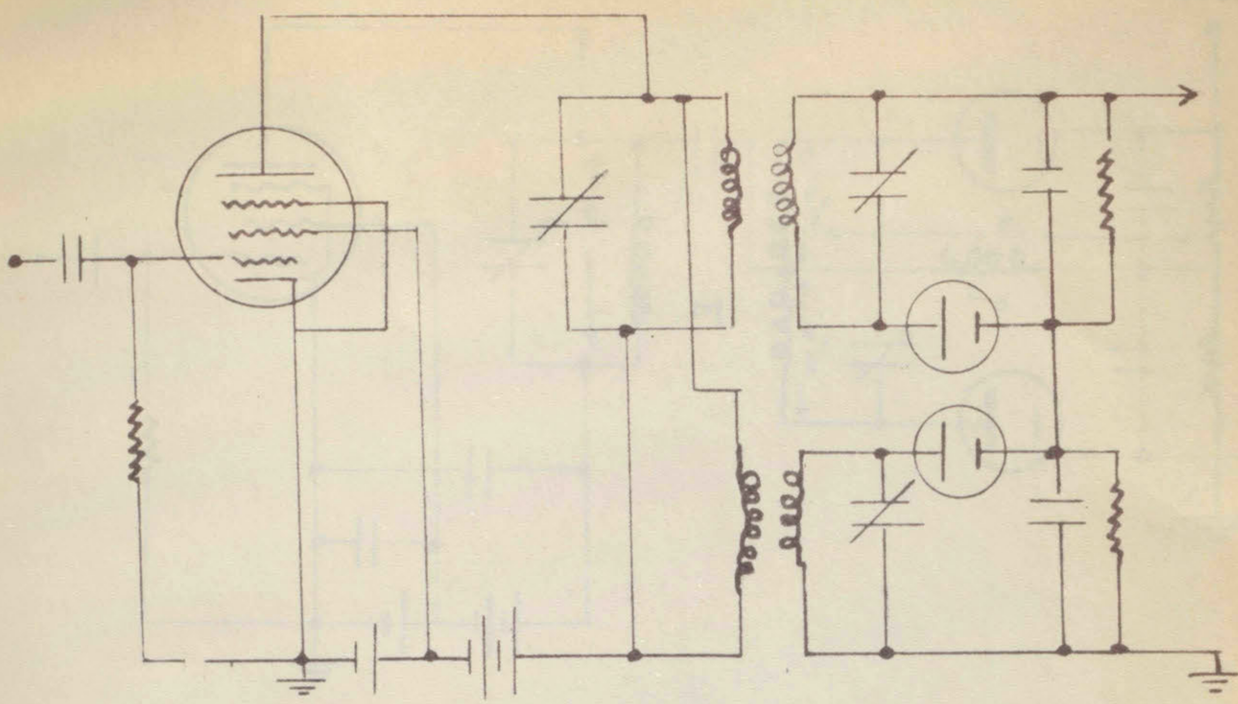
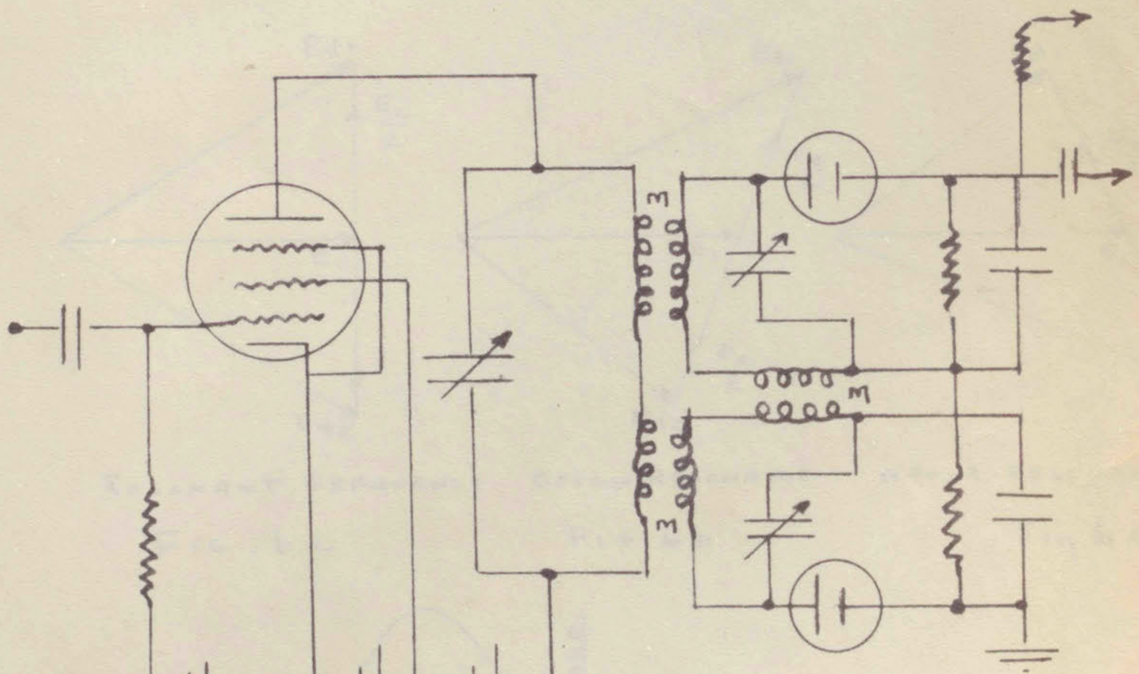


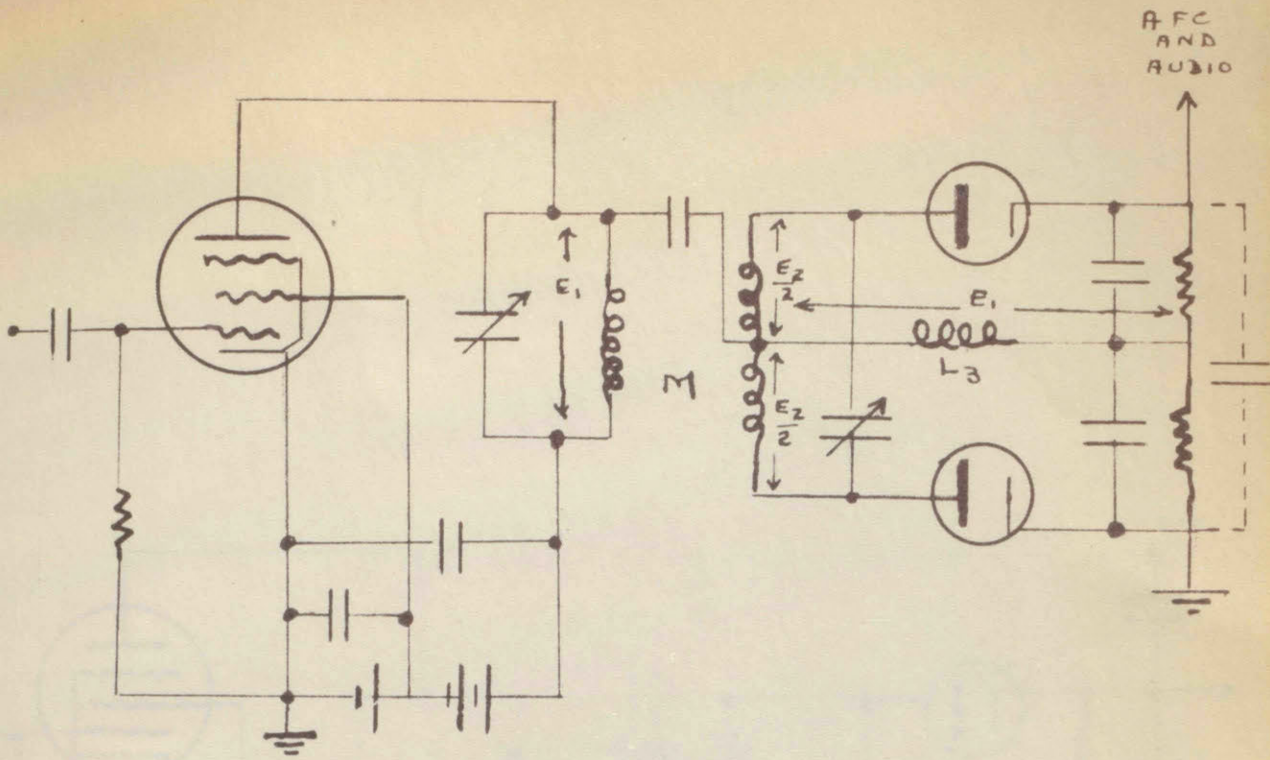
Fig.4

TRAVIS DISCRIMINATOR



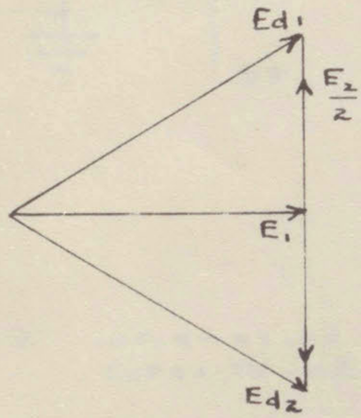
CROSBY DISCRIMINATOR

Fig 5



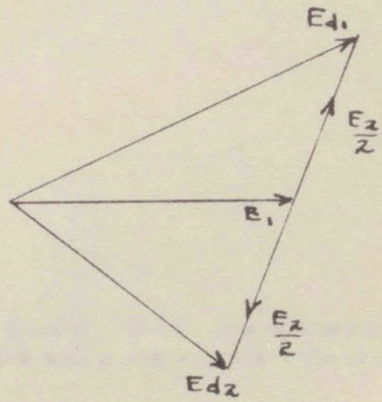
FOSTER SEELEY DISCRIMINATOR

Fig 6



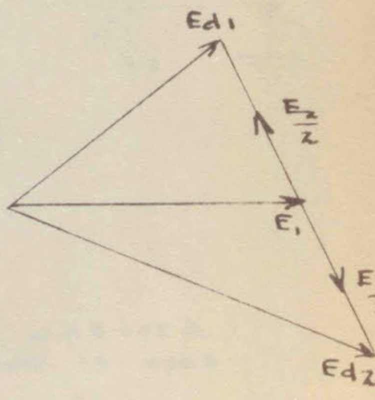
RESONANT FREQUENCY

FIG. 6a



BELOW RESONANCE

FIG 6b



ABOVE RESONANCE

Fig 6c

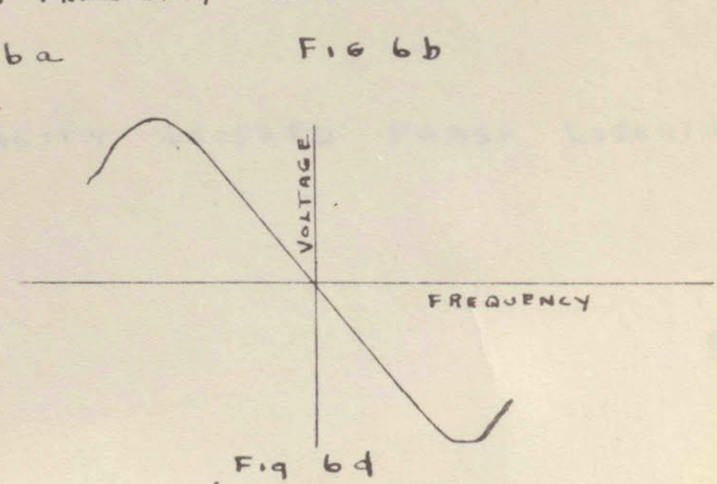
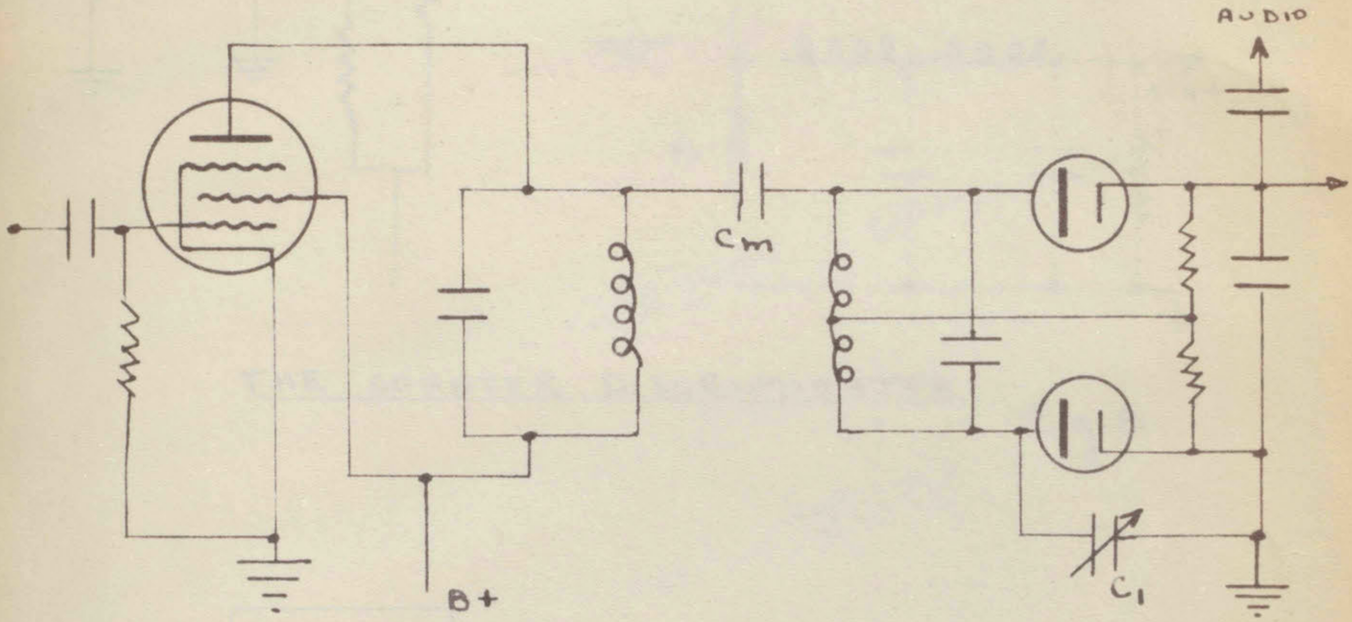


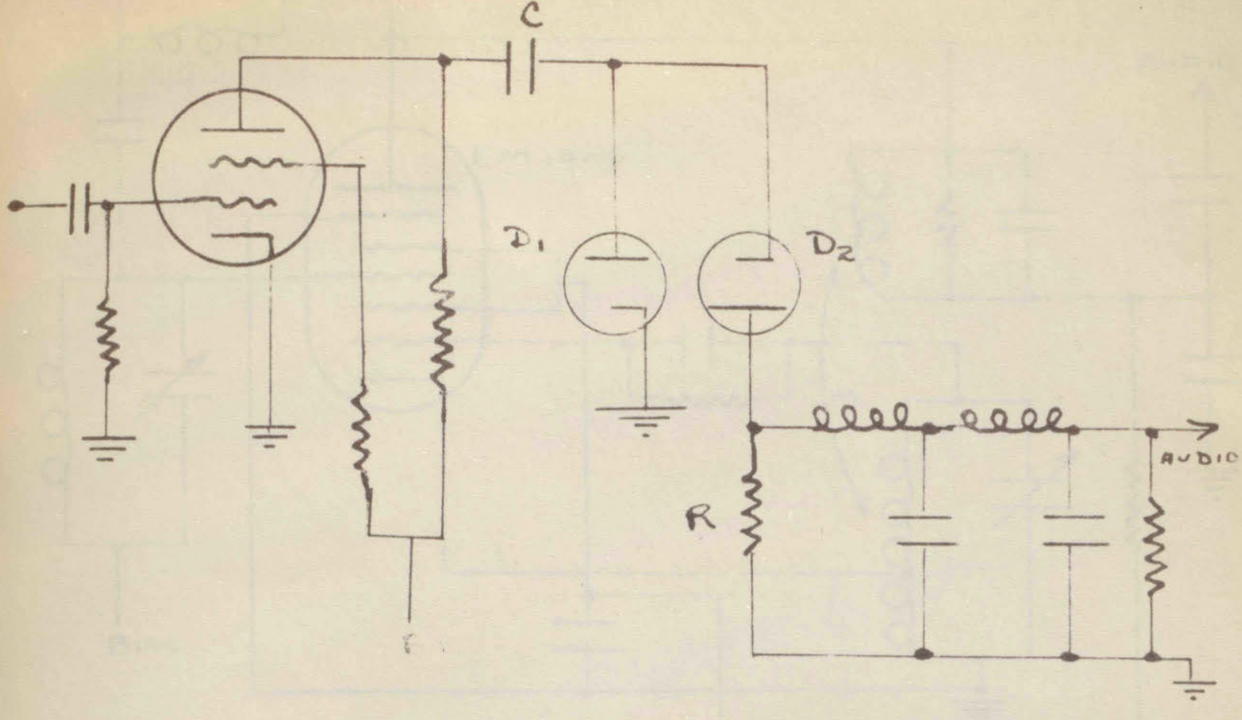
Fig 6d



NOTE INFORMATION DOES NOT INDICATE WHETHER CAPACITY OR PERMEABILITY TUNING IS USED

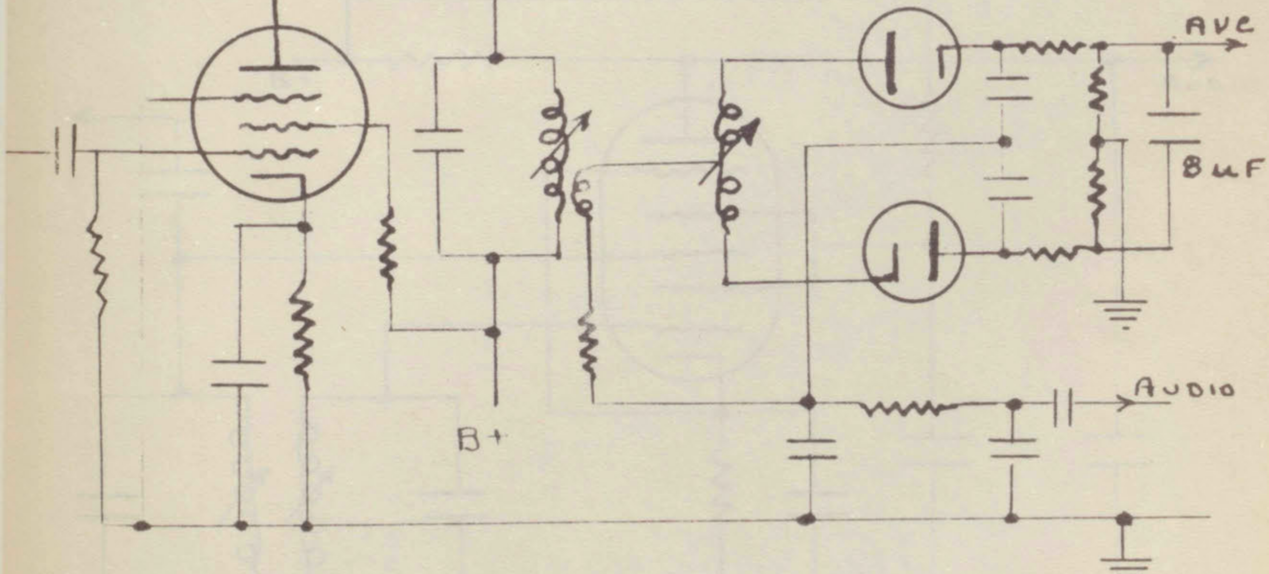
SERIES CAPACITY COUPLED PHASE DISCRIMINATOR

Fig 7.



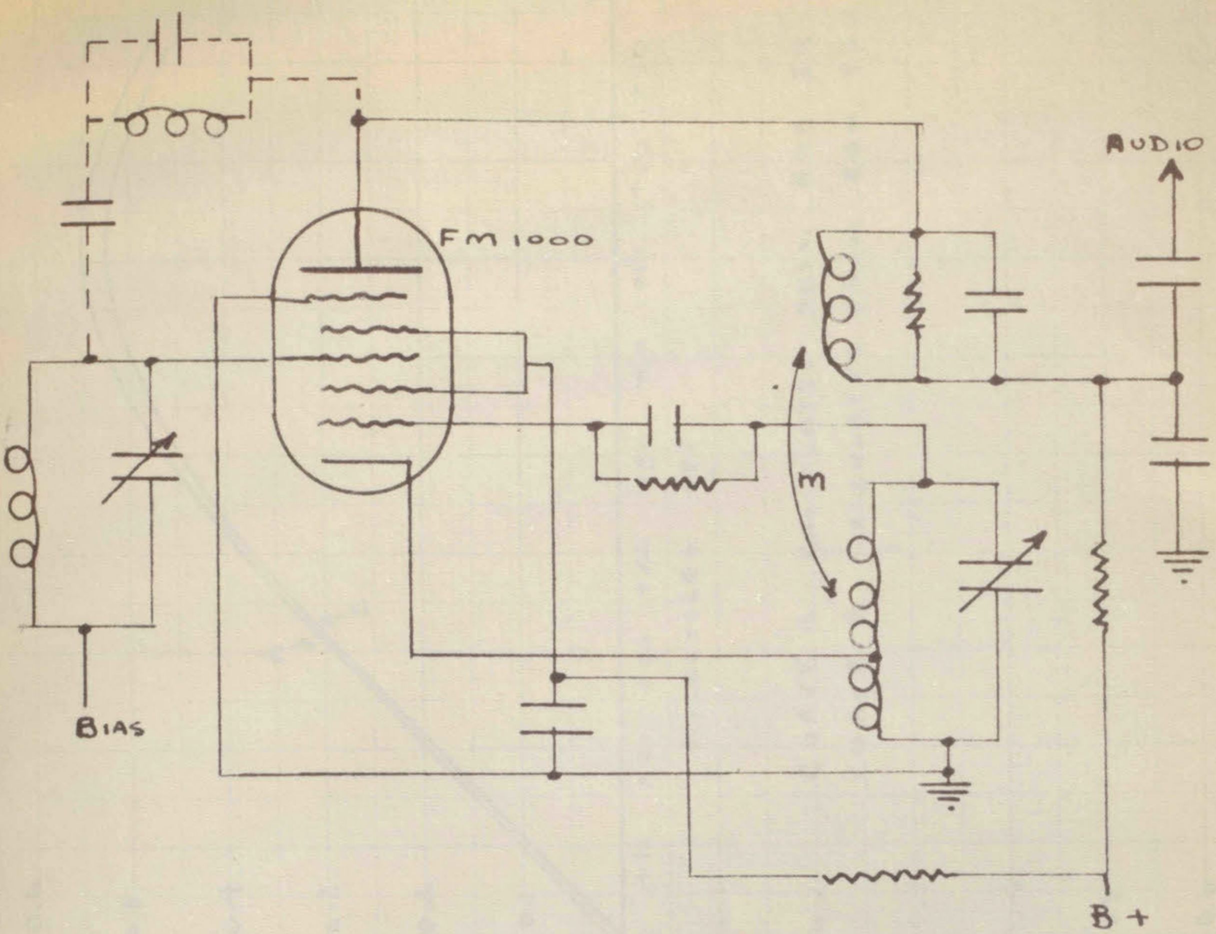
THE COUNTER DISCRIMINATOR

Fig 8



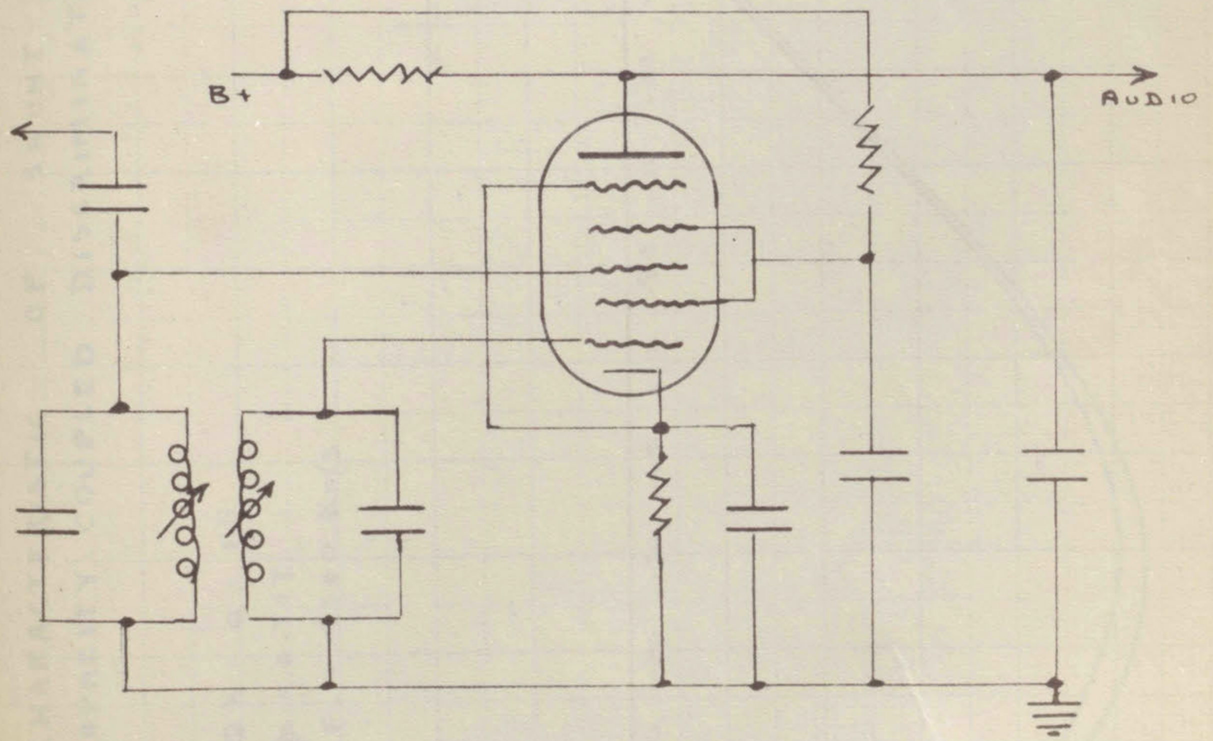
THE RATIO DETECTOR

Fig 9



THE BRADLEY F.M. DETECTOR

Fig 10



THE CBC DISCRIMINATOR

Fig 11.

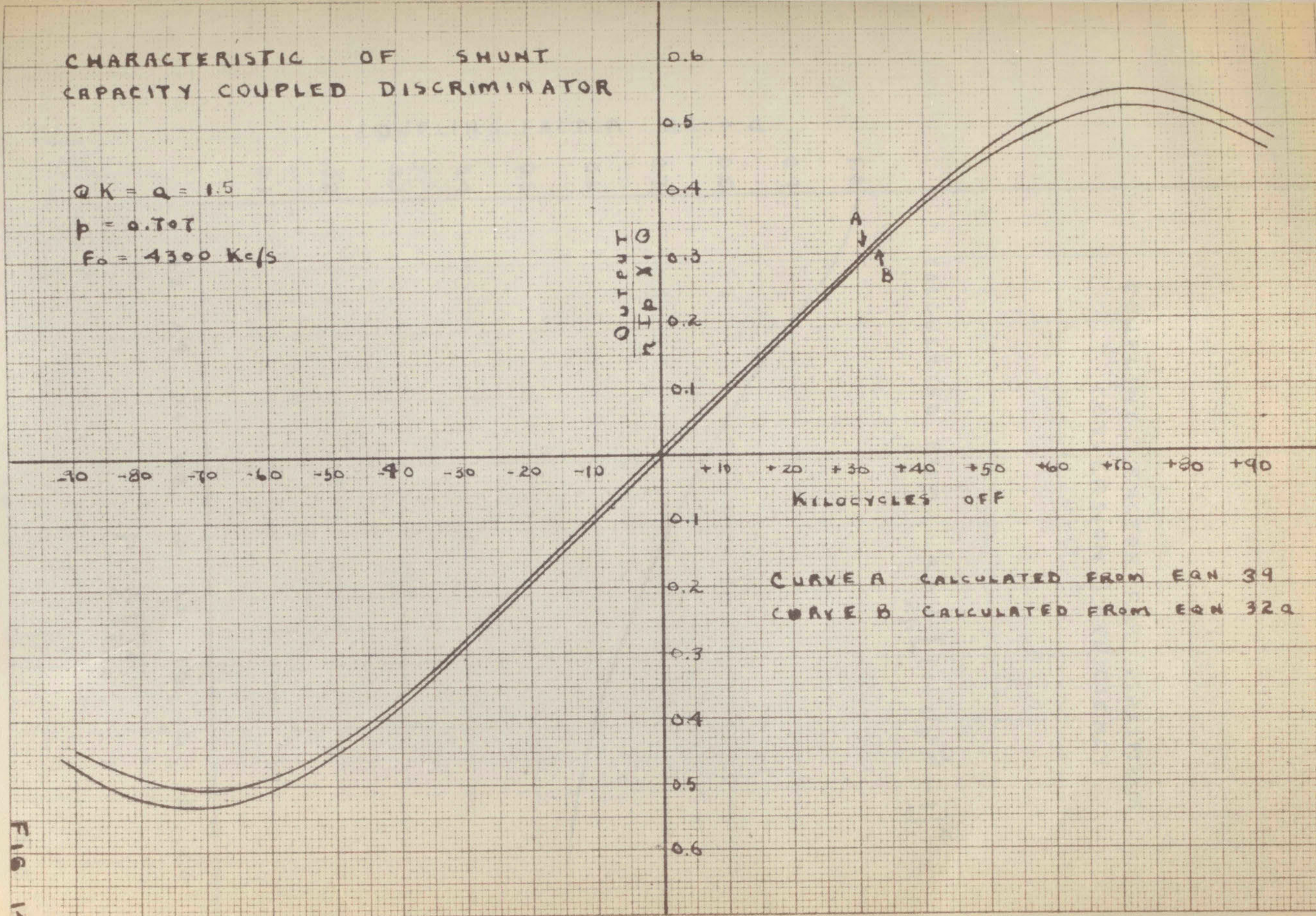

CHARACTERISTIC OF SHUNT
CAPACITY COUPLED DISCRIMINATOR
QK = a = 1.5
p = 0.707
Fo = 4300 Kc/s
OUTPUT
KILOCYCLES OFF
A
B
CURVE A CALCULATED FROM EQN 39
CURVE B CALCULATED FROM EQN 32a

FIG 14

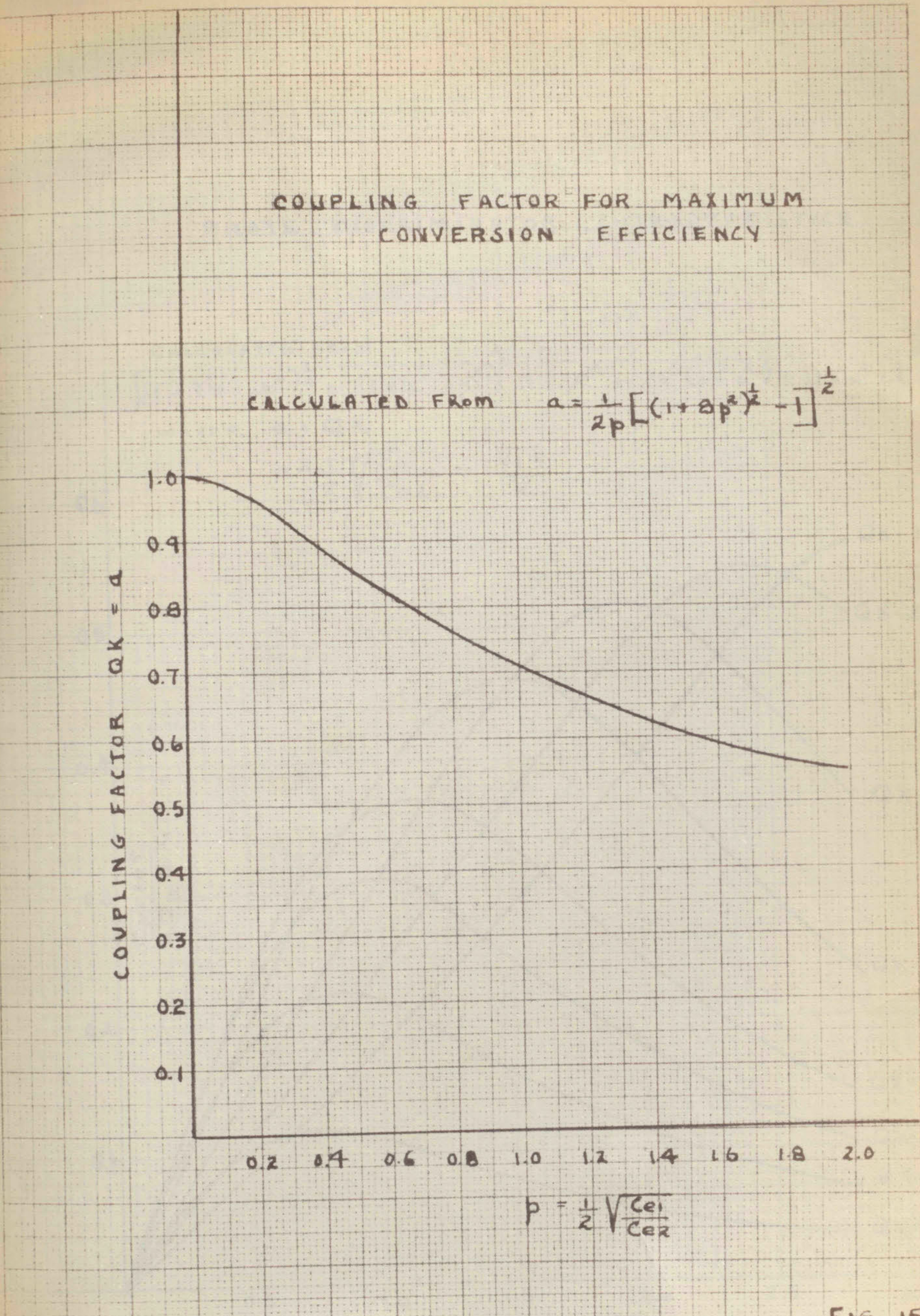

COUPLING FACTOR FOR MAXIMUM CONVERSION EFFICIENCY
CALCULATED FROM $a = \frac{1}{2p}\left[(1+8p^2)^{\frac{1}{2}} - 1\right]^{\frac{1}{2}}$
COUPLING FACTOR QK = a
1.0
0.9
0.8
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0.2 0.4 0.6 0.8 1.0 1.2 1.4 1.6 1.8 2.0
$p = \frac{1}{2}\sqrt{\frac{C_{e1}}{C_{e2}}}$

FIG 15

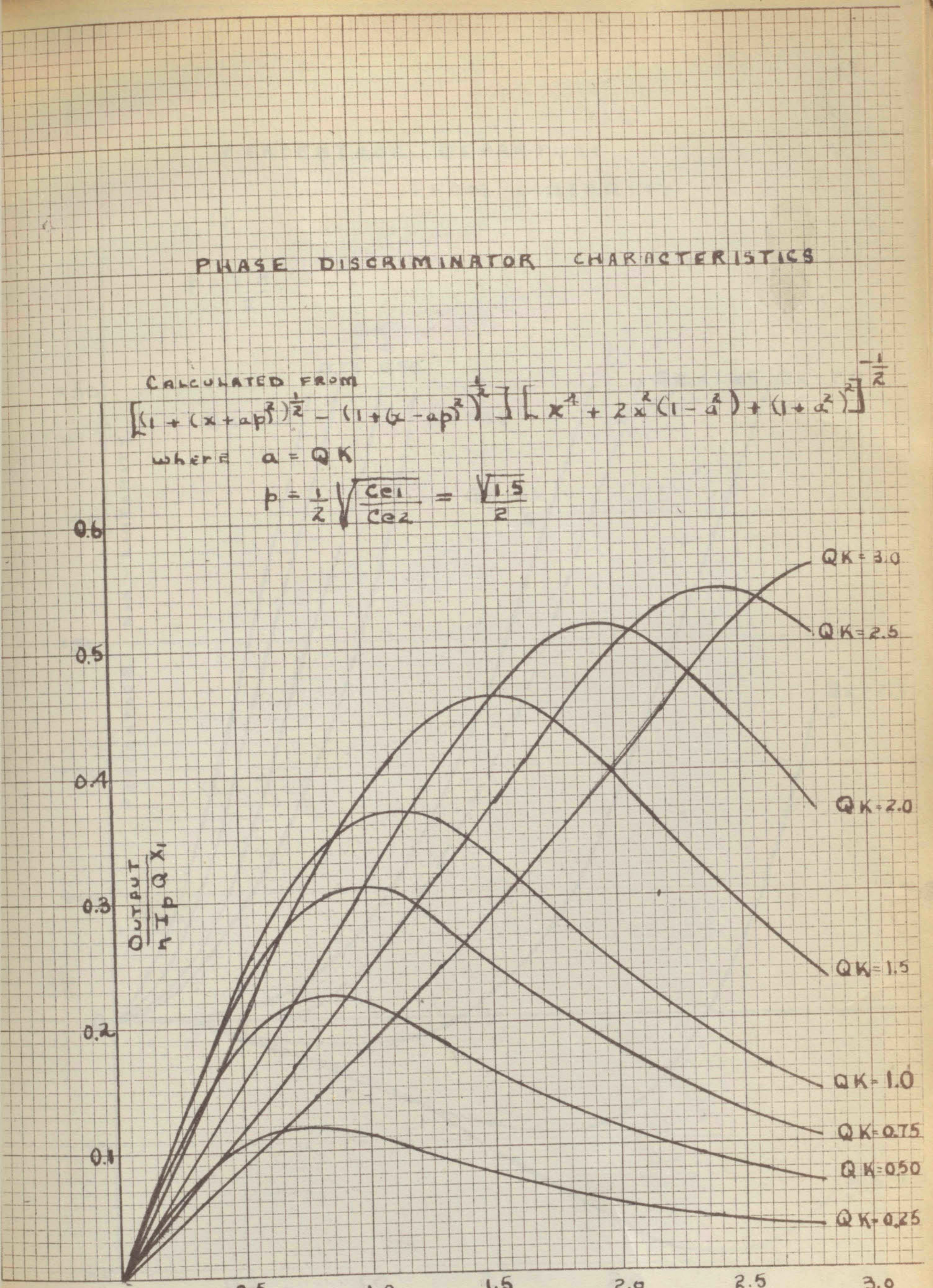

PHASE DISCRIMINATOR CHARACTERISTICS
CALCULATED FROM
$\left[(1+(x+ap)^2)^{\frac{1}{2}} - (1+(x-ap)^2)^{\frac{1}{2}}\right]\left[x^4+2x^2(1-a^2)+(1+a^2)^2\right]^{-\frac{1}{2}}$
where $a = QK$
$p = \frac{1}{2}\sqrt{\frac{Ce1}{Ce2}} = \frac{\sqrt{1.5}}{2}$
$\frac{\text{OUTPUT}}{n I_p Q X_1}$
0.1
0.2
0.3
0.4
0.5
0.6
0.5
1.0
1.5
2.0
2.5
3.0
$x = \frac{2Q\Delta F}{F_0}$
QK = 3.0
QK = 2.5
QK = 2.0
QK = 1.5
QK = 1.0
QK = 0.75
QK = 0.50
QK = 0.25

Fig 16a

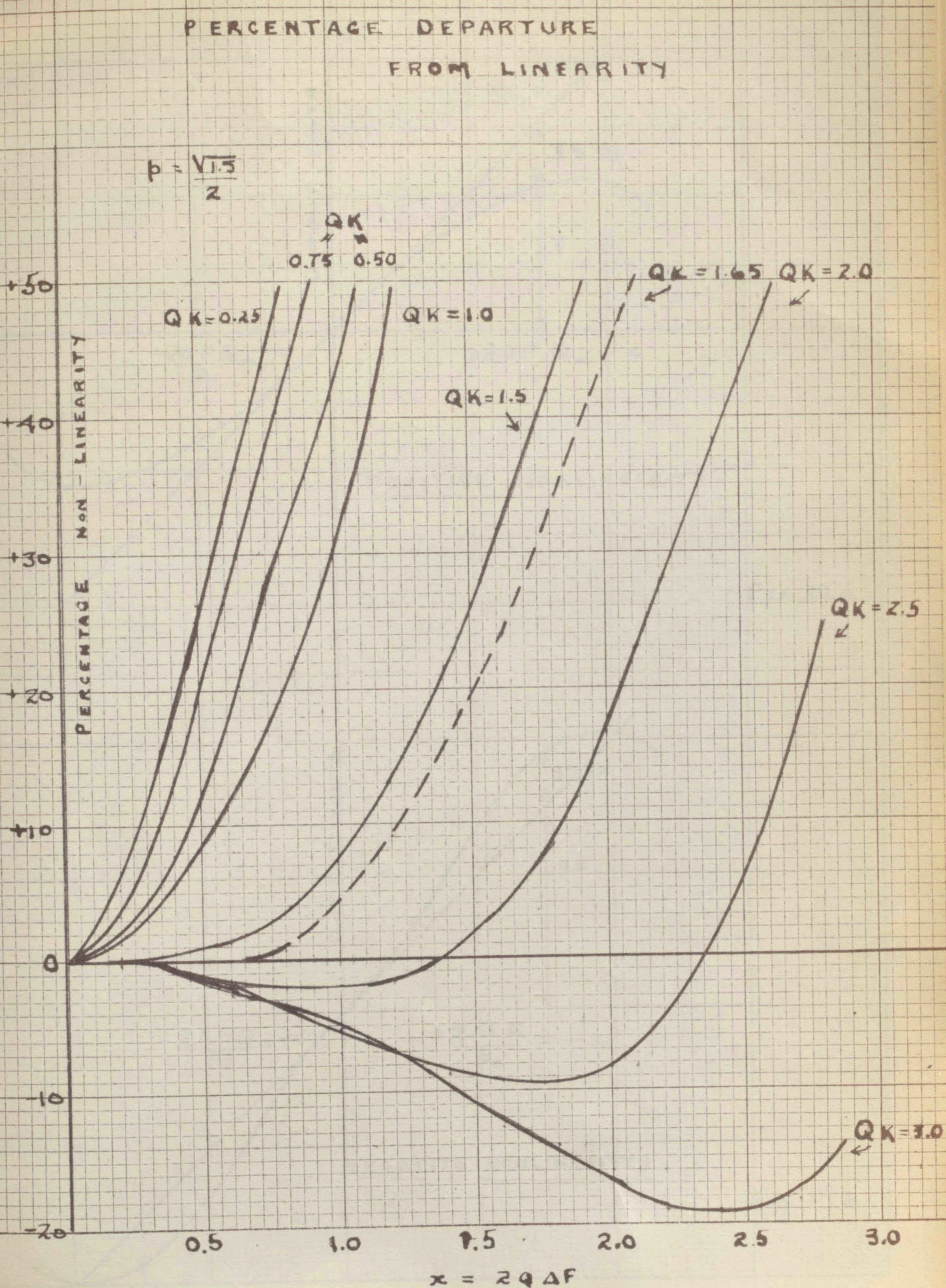

PERCENTAGE DEPARTURE FROM LINEARITY
p = √1.5 / 2
QK
0.75
0.50
QK = 0.25
QK = 1.0
QK = 1.65
QK = 2.0
QK = 1.5
QK = 2.5
QK = 3.0
PERCENTAGE NON - LINEARITY
+50
+40
+30
+20
+10
0
-10
-20
0.5
1.0
1.5
2.0
2.5
3.0
x = 2Q ΔF / Fo

Fig.(16 b)

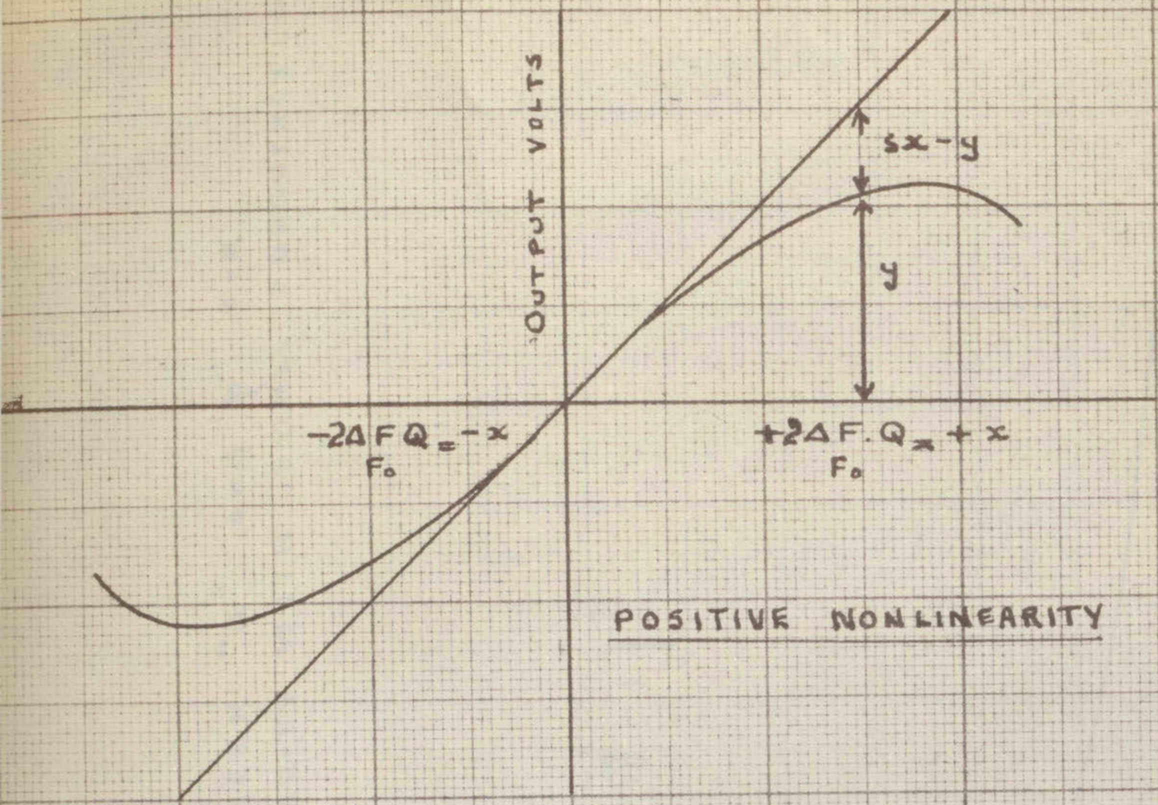


Fig 17a

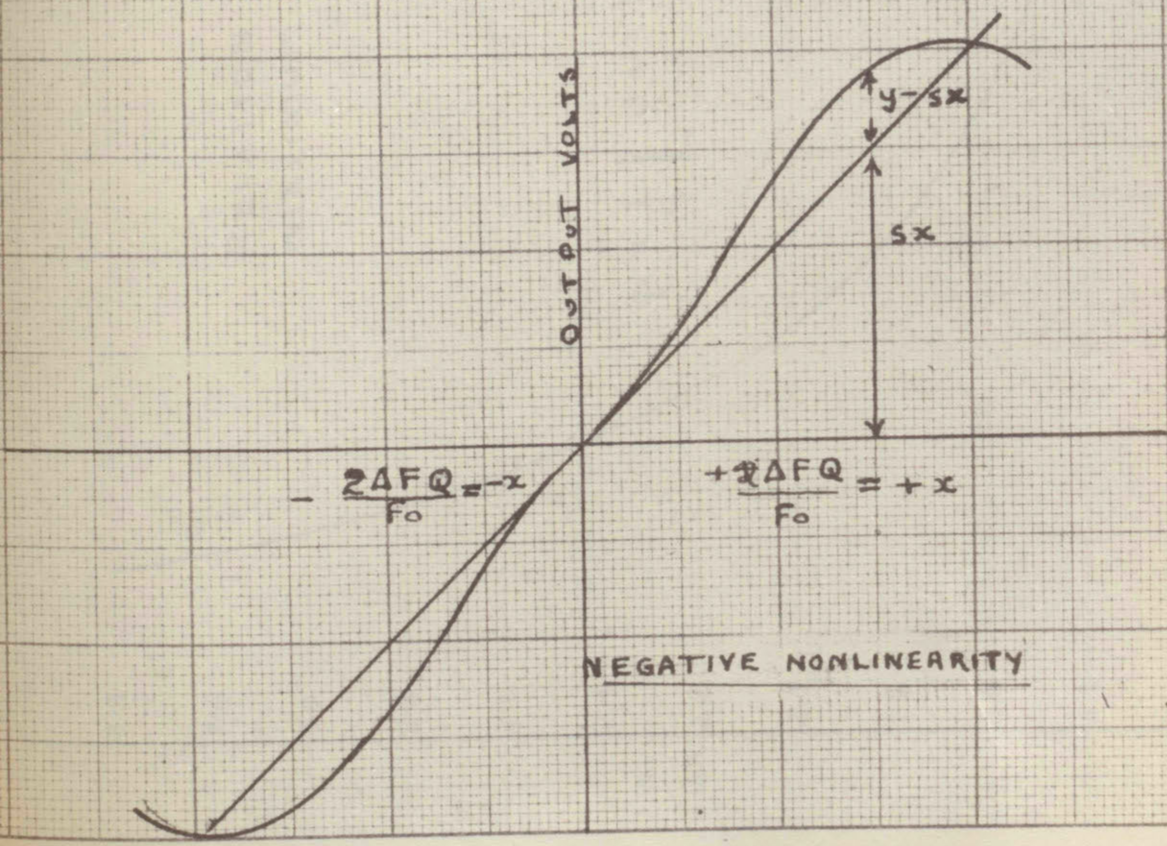


Fig 17b

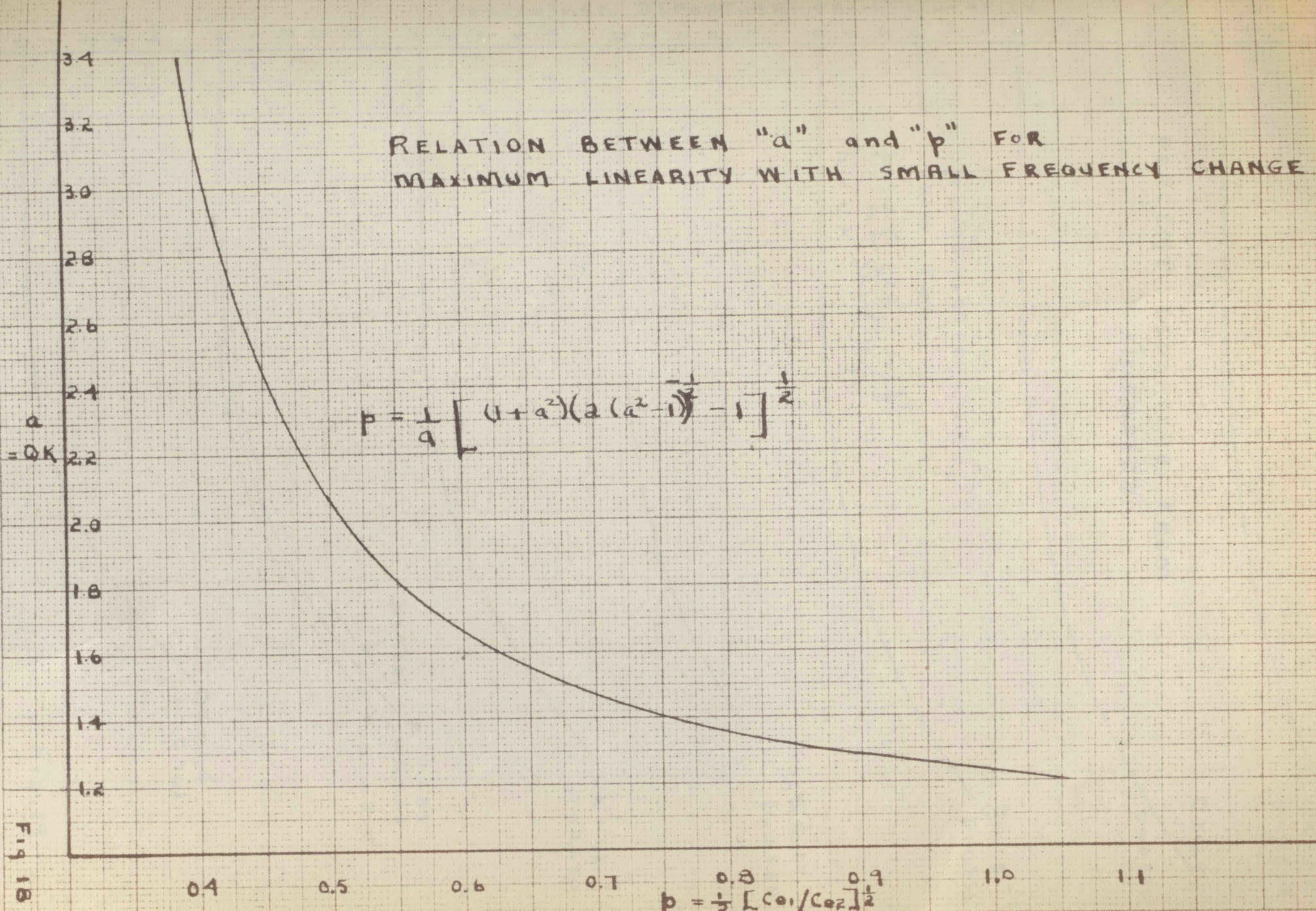
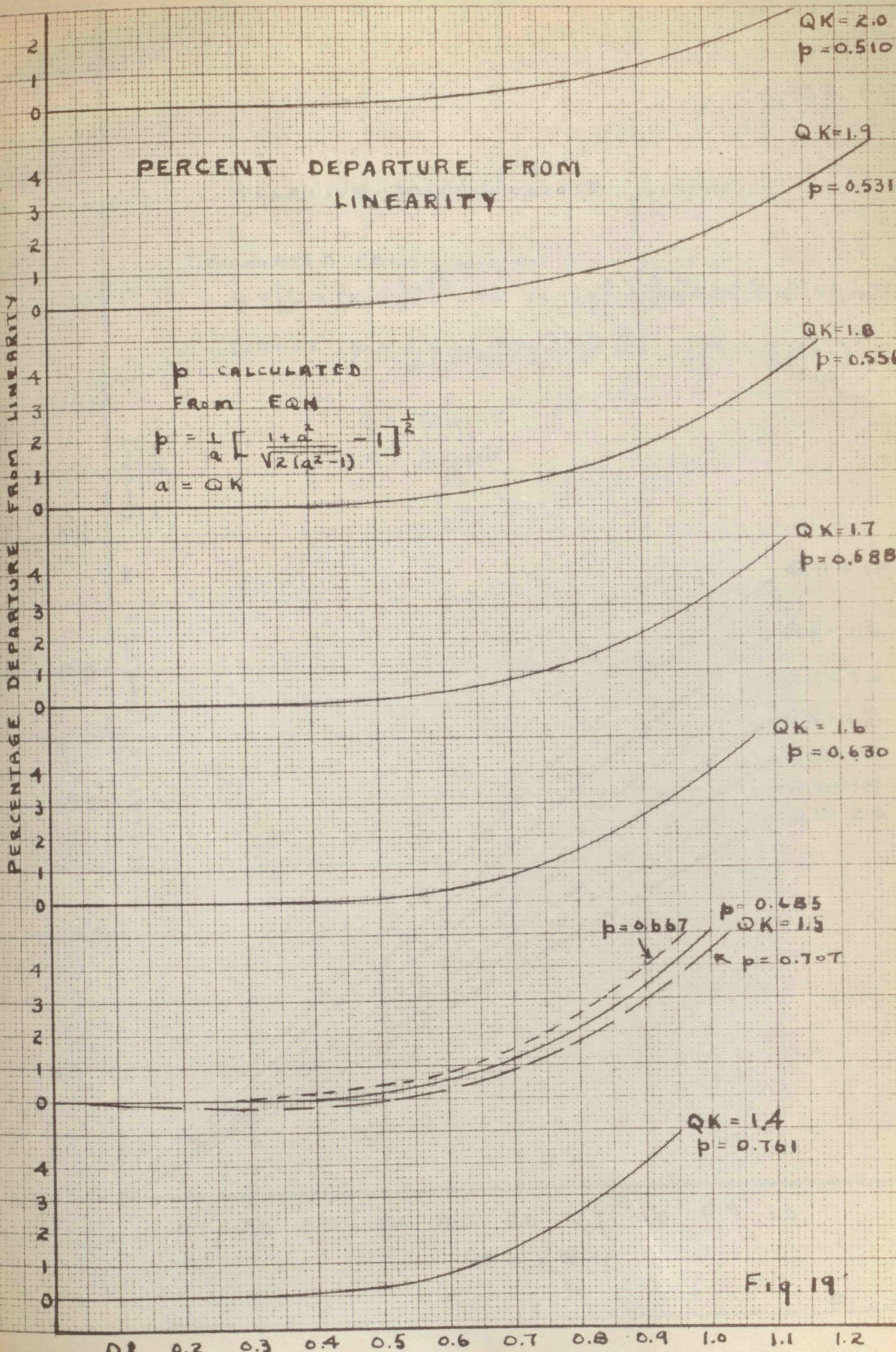

RELATION BETWEEN "a" and "p" FOR
MAXIMUM LINEARITY WITH SMALL FREQUENCY CHANGE
$p = \frac{1}{a}\left[(1+a^2)(2(a^2-1)^{-\frac{1}{2}}-1\right]^{\frac{1}{2}}$
a
= QK
3.4
3.2
3.0
2.8
2.6
2.4
2.2
2.0
1.8
1.6
1.4
1.2
0.4
0.5
0.6
0.7
0.8
0.9
1.0
1.1
$p = \frac{1}{2}\left[C_{01}/C_{02}\right]^{\frac{1}{2}}$

Fig 18



PERCENT DEPARTURE FROM LINEARITY
p CALCULATED FROM EQN
$p = \frac{1}{a}\left[\frac{1+a^2}{\sqrt{2(a^2-1)}} - 1\right]^{\frac{1}{2}}$
a = QK
QK = 2.0
p = 0.510
QK = 1.9
p = 0.531
QK = 1.8
p = 0.556
QK = 1.7
p = 0.588
QK = 1.6
p = 0.630
p = 0.685
QK = 1.5
p = 0.667
p = 0.707
QK = 1.4
p = 0.761
PERCENTAGE DEPARTURE FROM LINEARITY
x
0.1
0.2
0.3
0.4
0.5
0.6
0.7
0.8
0.9
1.0
1.1
1.2

Fig. 19

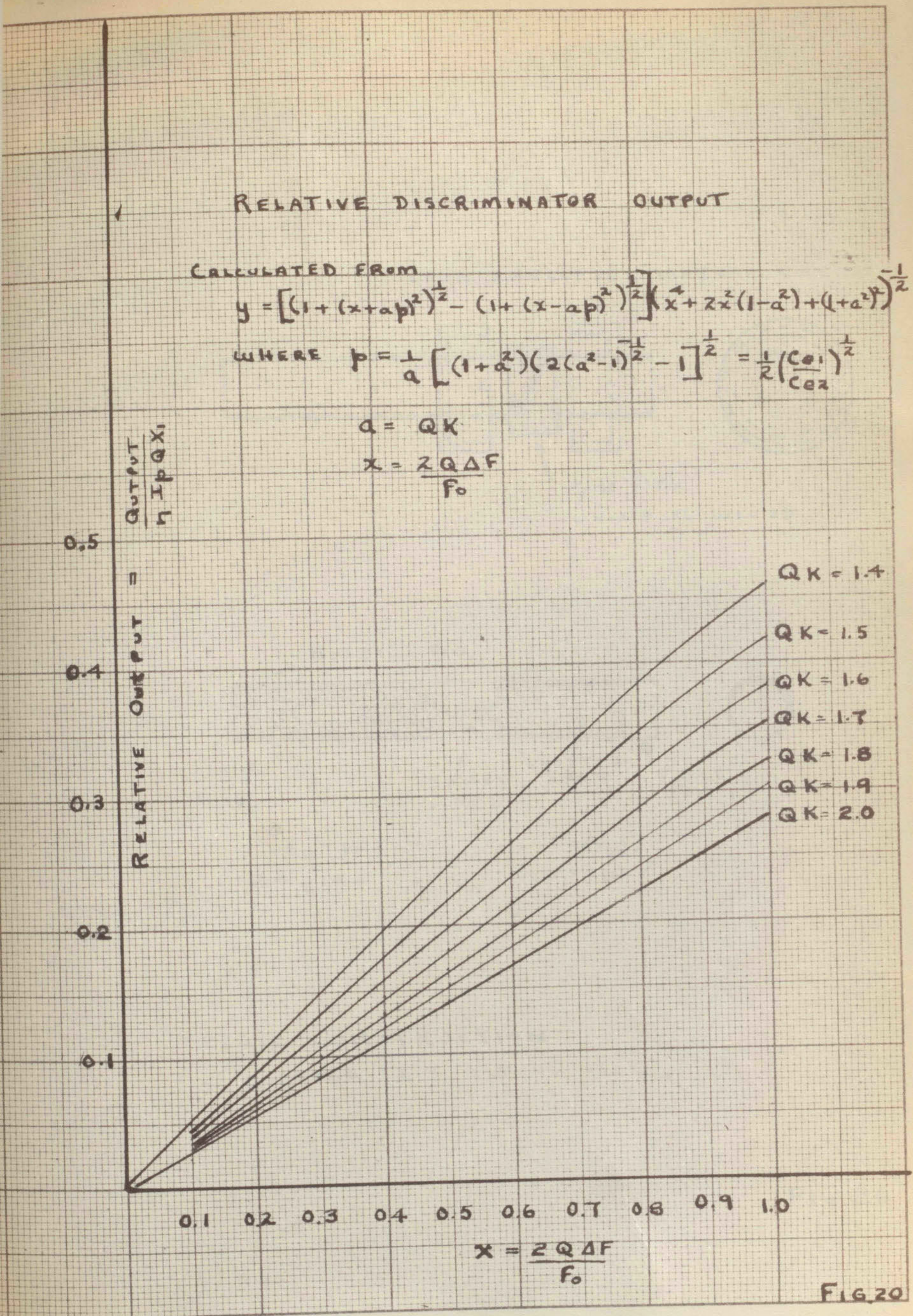
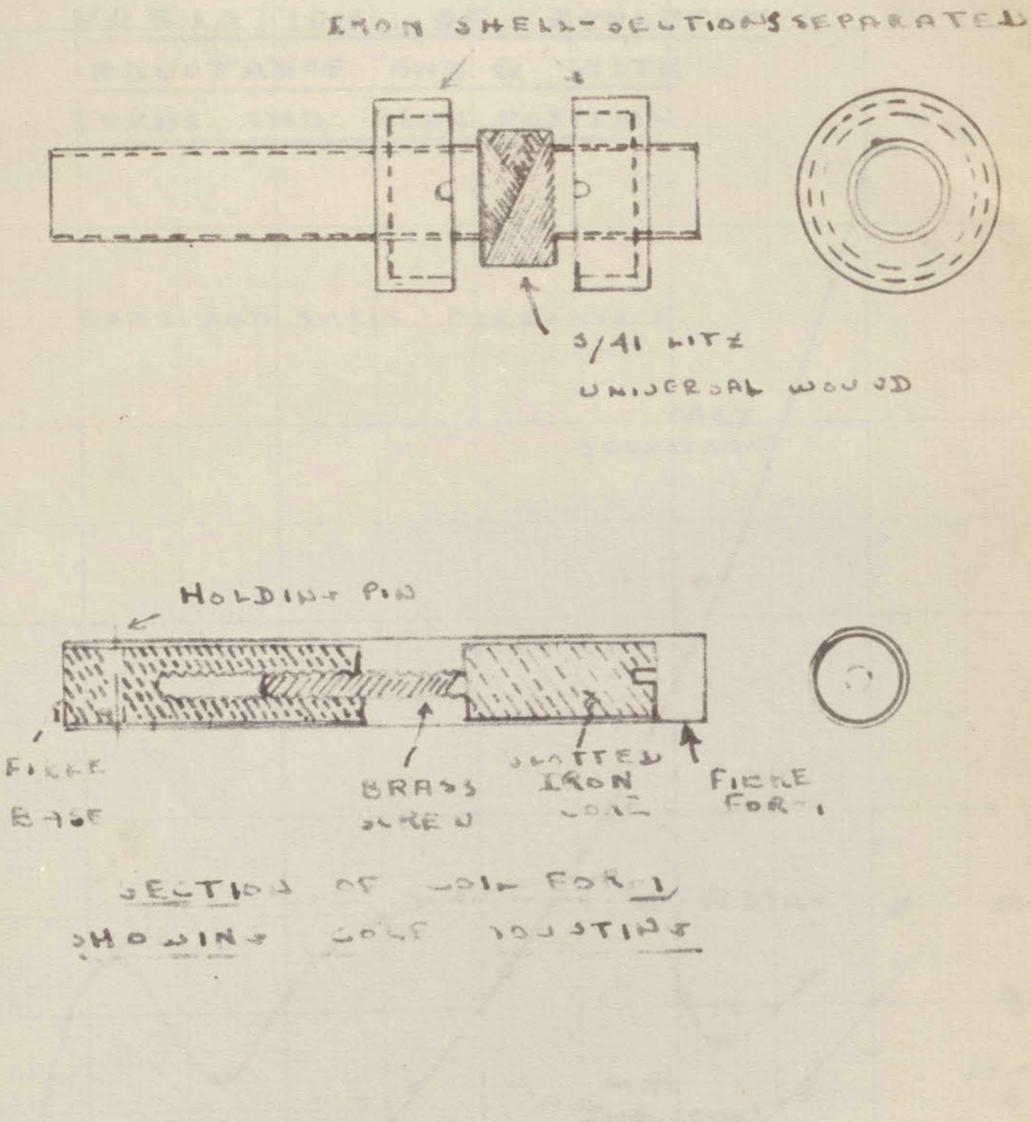

RELATIVE DISCRIMINATOR OUTPUT
CALCULATED FROM
$y = \left[(1+(x+ap)^2)^{\frac{1}{2}} - (1+(x-ap)^2)^{\frac{1}{2}}\right]\left(x^4+2x^2(1-a^2)+(1+a^2)^2\right)^{-\frac{1}{2}}$
WHERE $p = \frac{1}{a}\left[(1+a^2)(2(a^2-1))^{-\frac{1}{2}} - 1\right]^{\frac{1}{2}} = \frac{1}{2}\left(\frac{C_{e1}}{C_{e2}}\right)^{\frac{1}{2}}$
$a = QK$
$x = \frac{2Q\Delta F}{F_0}$
RELATIVE OUTPUT = $\frac{\text{OUTPUT}}{\eta I_p Q X_1}$
0.1
0.2
0.3
0.4
0.5
QK = 1.4
QK = 1.5
QK = 1.6
QK = 1.7
QK = 1.8
QK = 1.9
QK = 2.0
0.1 0.2 0.3 0.4 0.5 0.6 0.7 0.8 0.9 1.0
$x = \frac{2Q\Delta F}{F_0}$

FIG. 20



SECTION OF COIL FORM
SHOWING CORE ADJUSTING

COIL CONSTRUCTION
[SCALE 2X ACTUAL SIZE]

FIG 26

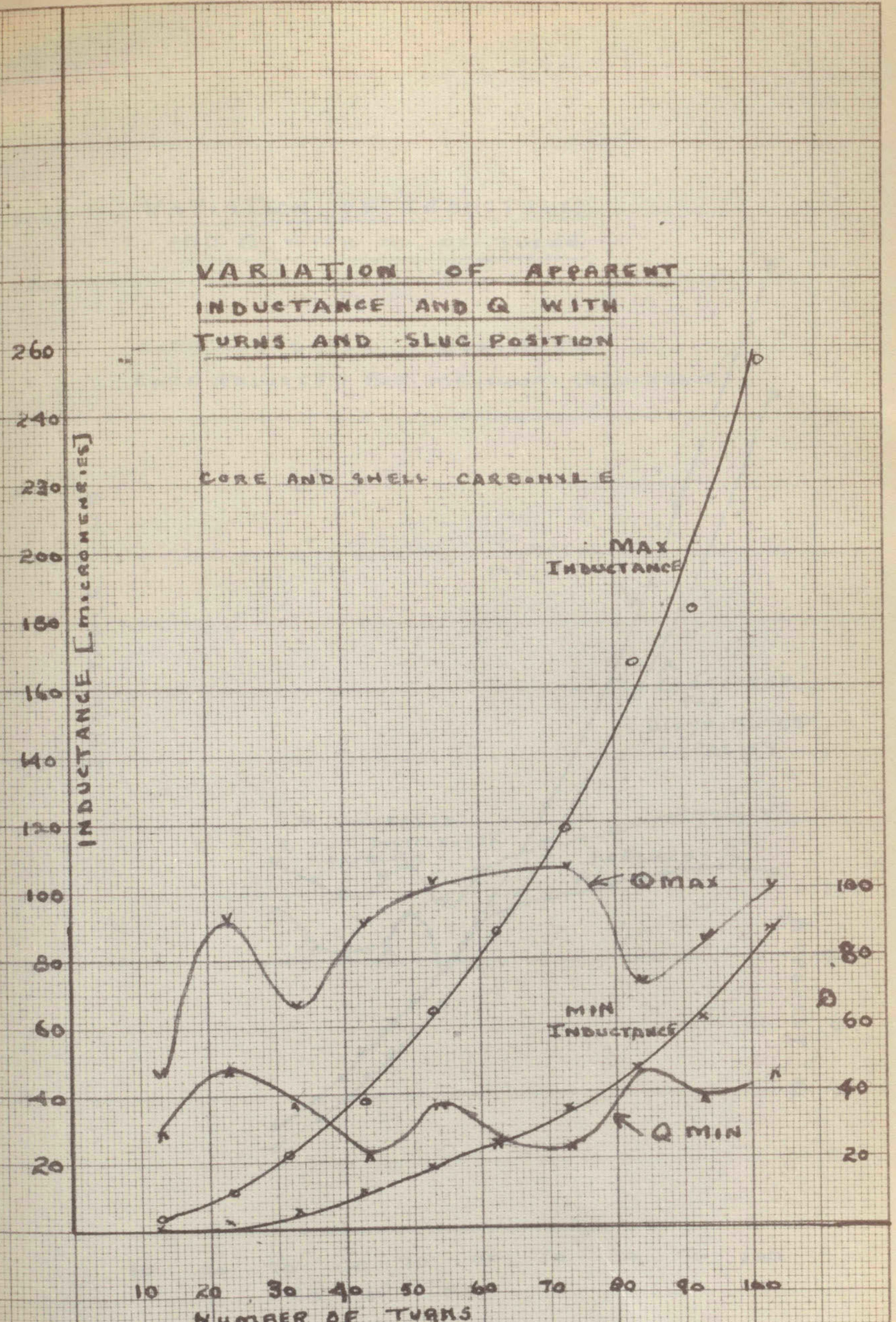
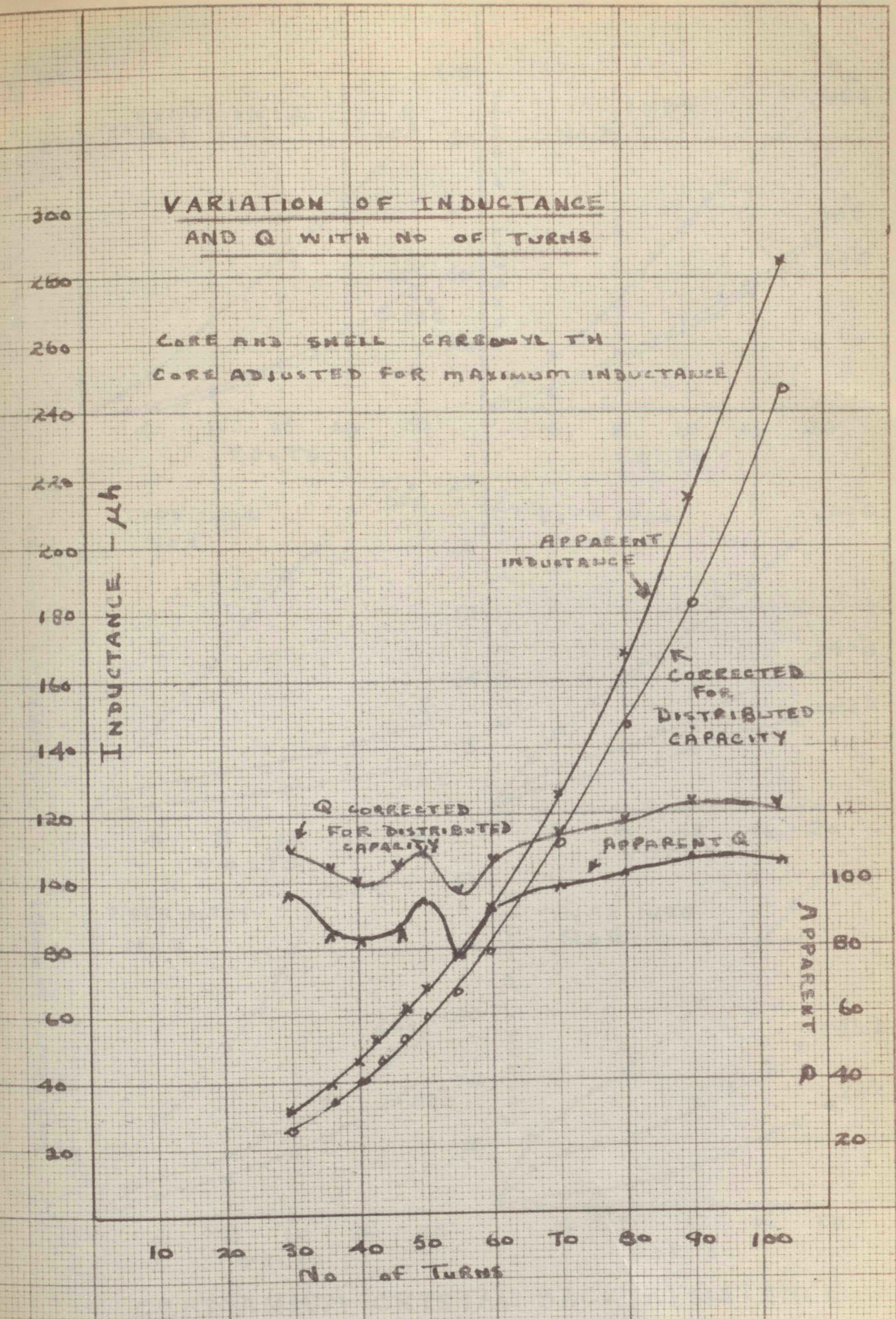
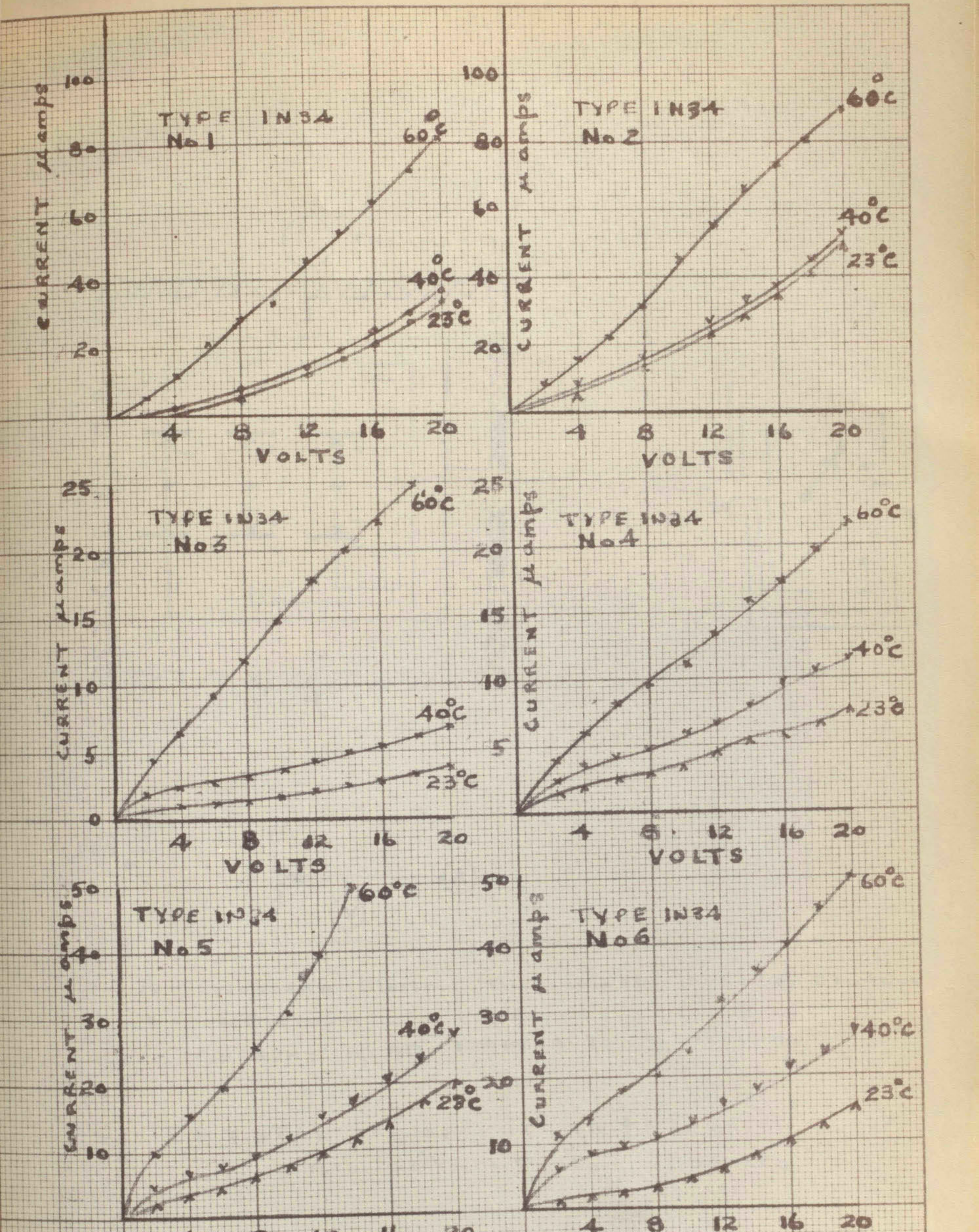

VARIATION OF APPARENT INDUCTANCE AND Q WITH TURNS AND SLUG POSITION
CORE AND SHELL CARBONYLE
MAX INDUCTANCE
Q MAX
MIN INDUCTANCE
Q MIN
INDUCTANCE [MICROHENRIES]
Q
NUMBER OF TURNS
260
240
220
200
180
160
140
120
100
80
60
40
20
10 20 30 40 50 60 70 80 90 100

FIG 28



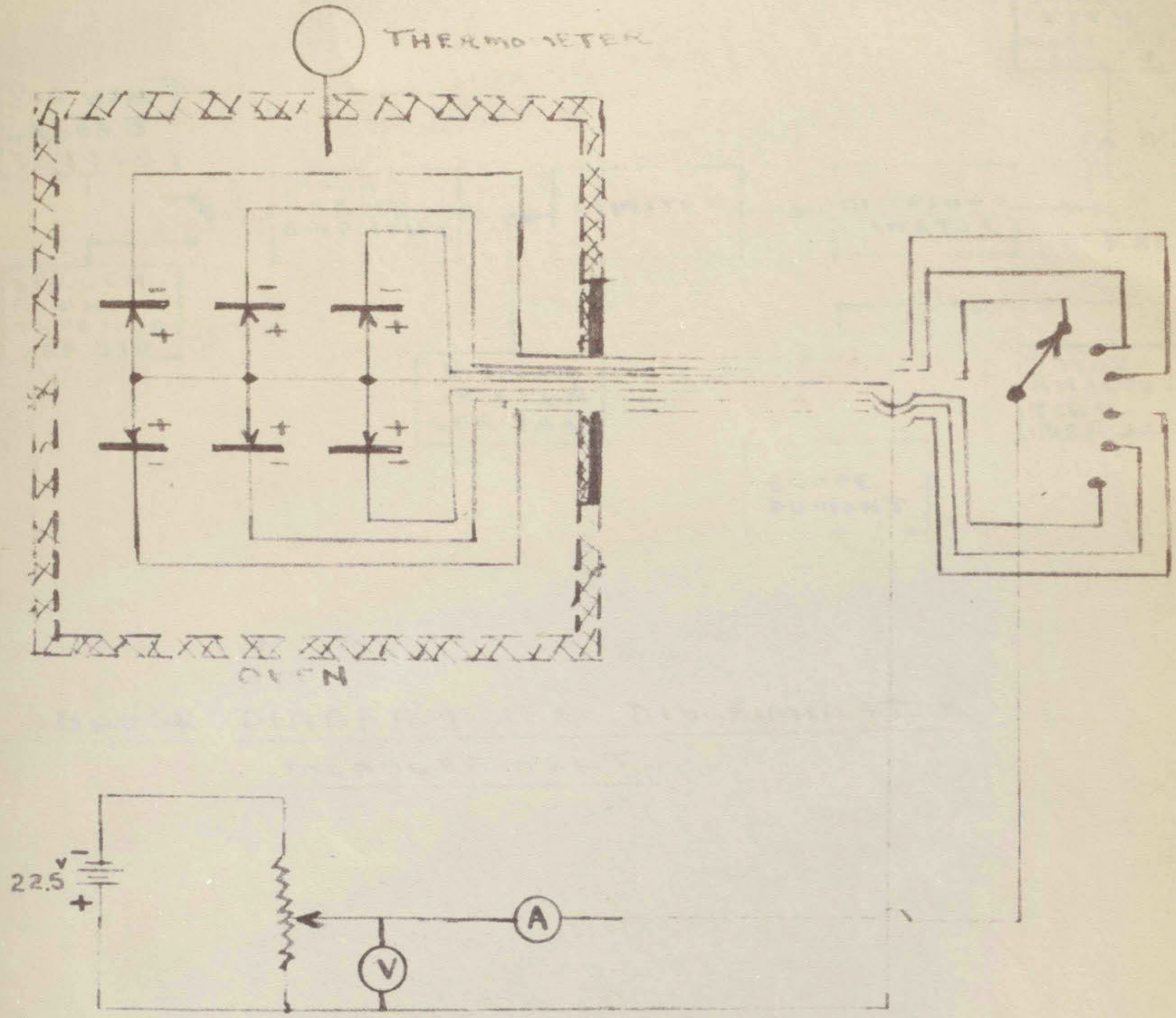
VARIATION OF INDUCTANCE
AND Q WITH NO OF TURNS
CORE AND SHELL CARBONYL TH
CORE ADJUSTED FOR MAXIMUM INDUCTANCE
APPARENT INDUCTANCE
CORRECTED FOR DISTRIBUTED CAPACITY
Q CORRECTED FOR DISTRIBUTED CAPACITY
APPARENT Q
INDUCTANCE - μh
APPARENT Q
No of TURNS
10 20 30 40 50 60 70 80 90 100
20 40 60 80 100 120 140 160 180 200 220 240 260 280 300

Fig 28



GERMANIUM CRYSTAL BACK RESISTANCE
VARIATION WITH TEMPERATURE

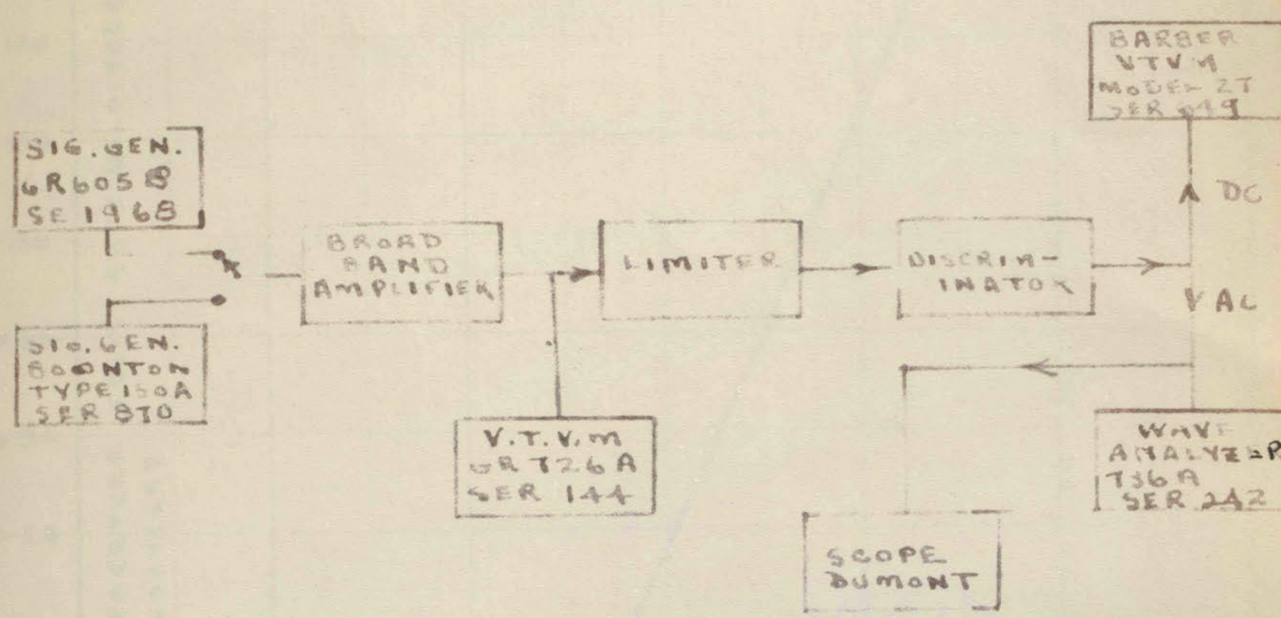
FIG 29



A - WESTERN Electric Microammeter 0-50 Ser D164601
V WESTINGHOUSE PX 4
CURRENT READ 1 MIN AFTER SWITCHING

CRYSTAL TEMPERATURE & BACK
BACK CURRENT MEASUREMENTS

FIG 30



BLOCK DIAGRAM FOR DISCRIMINATOR MEASUREMENTS

FIG 31

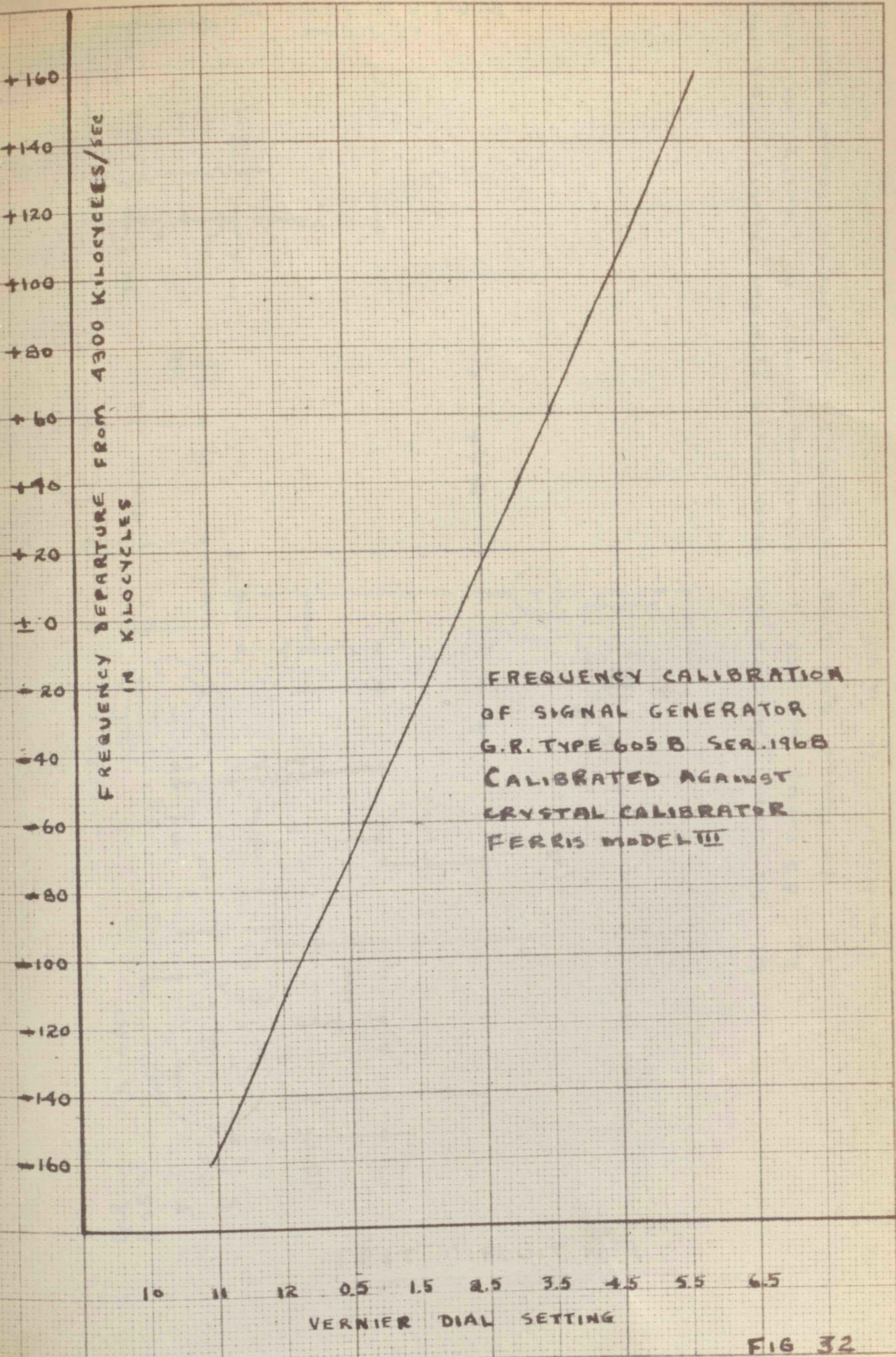
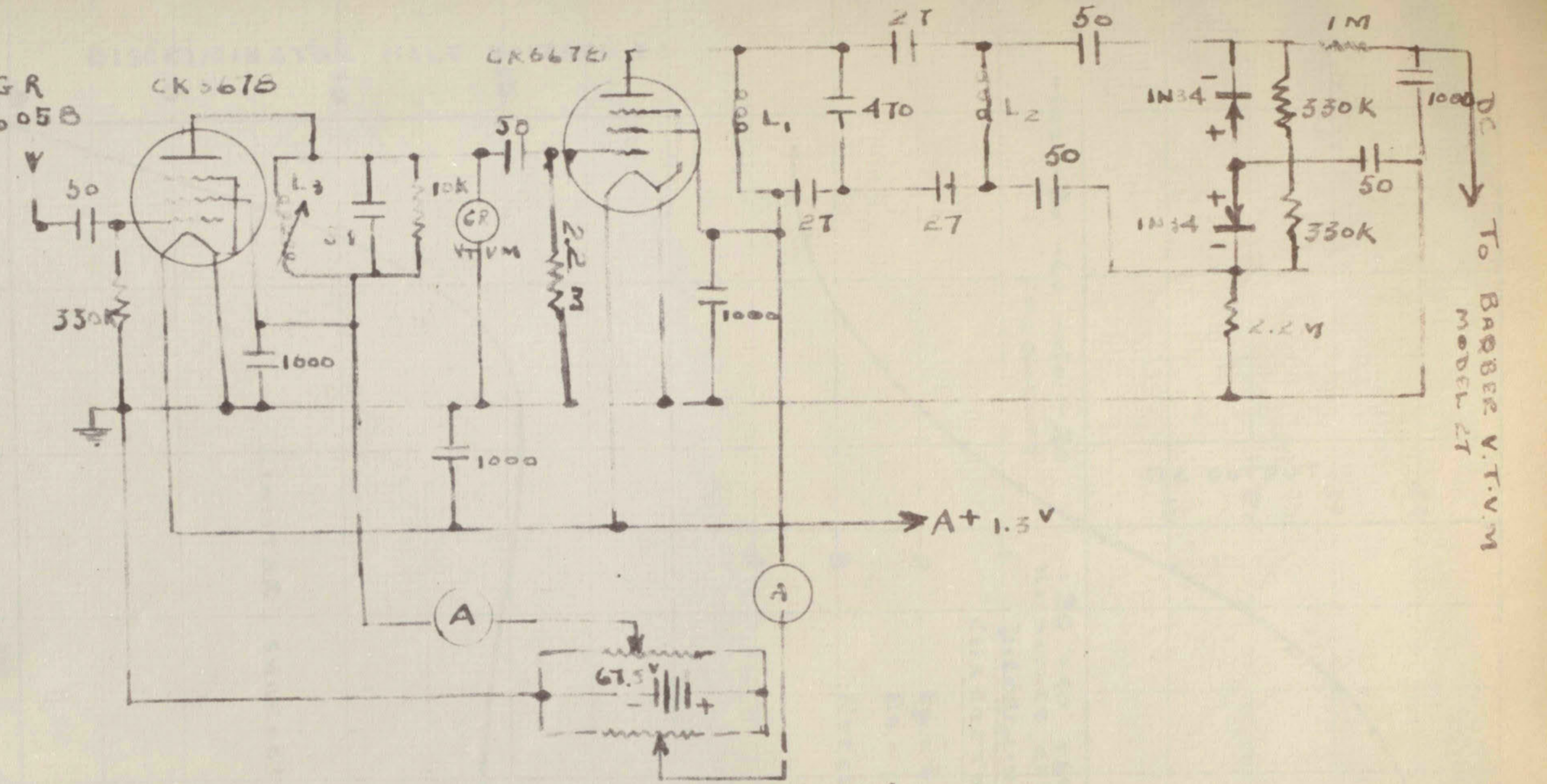

FREQUENCY DEPARTURE FROM 4300 KILOCYCLES/SEC
IN KILOCYCLES
+160
+140
+120
+100
+80
+60
+40
+20
± 0
-20
-40
-60
-80
-100
-120
-140
-160
FREQUENCY CALIBRATION
OF SIGNAL GENERATOR
G.R. TYPE 605 B SER. 1968
CALIBRATED AGAINST
CRYSTAL CALIBRATOR
FERRIS MODEL III
10
11
12
0.5
1.5
2.5
3.5
4.5
5.5
6.5
VERNIER DIAL SETTING

FIG 32

TEST CIRCUIT No 1



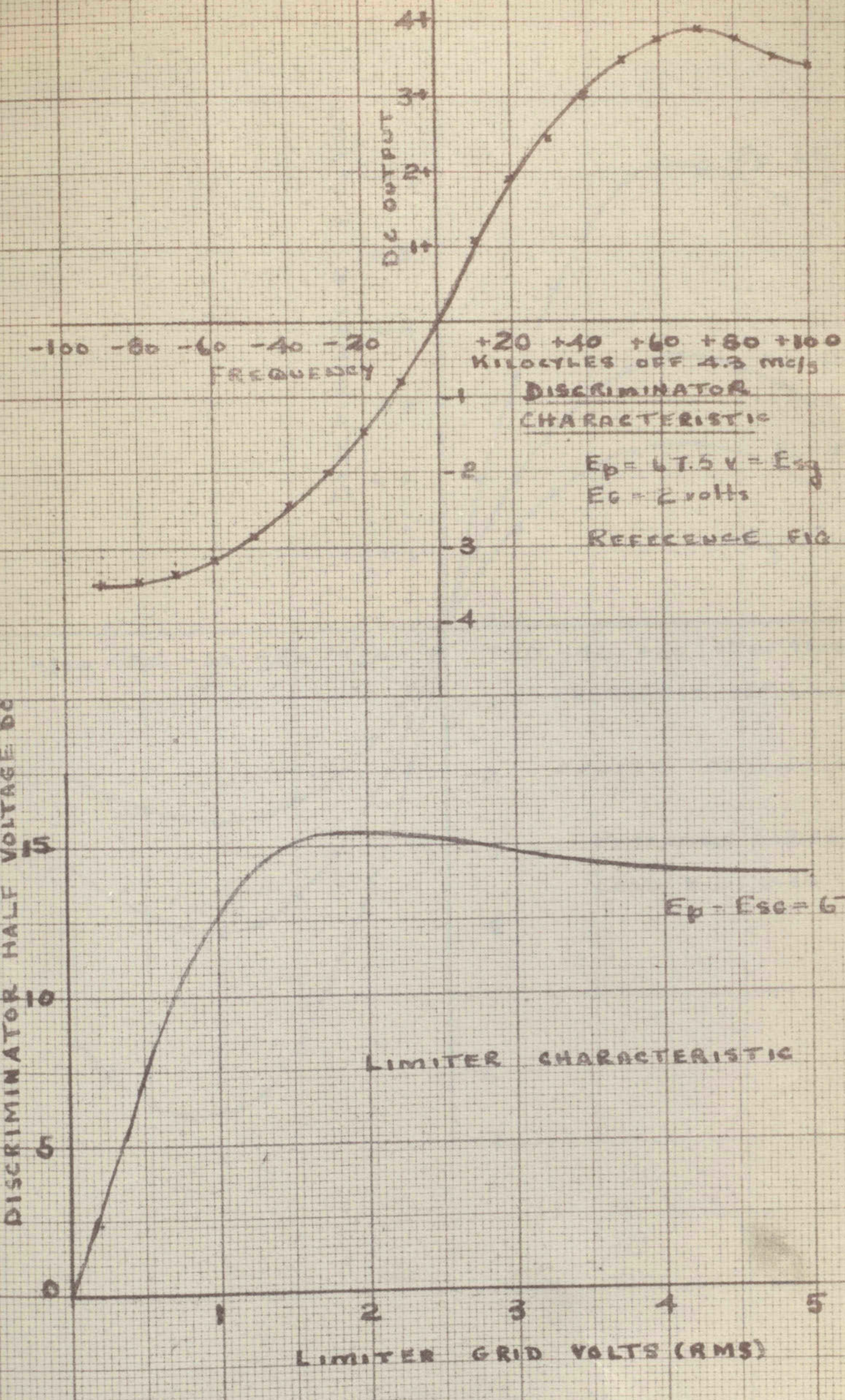
L1 53 TURNS 3/41 LITZ UNIVERSAL WOUND TYPE 5MF CROWLEY CORE & SHELL

L2 65 TURNS 3/41 LITZ UNIVERSAL WOUND TYPE 5MF CROWLEY CORE & SHELL

L3 40 TURNS 3/41 LITE UNIVERSAL WOUND TYPE 1MF CROWLEY CORE & SHELL

Condensers in μμf

FIG 33



DC OUTPUT
4
3
2
1
-1
-2
-3
-4
-100 -80 -60 -40 -20
+20 +40 +60 +80 +100
FREQUENCY
KILOCYCLES OFF 4.3 mc/s
DISCRIMINATOR CHARACTERISTIC
Ep = 67.5 v = Esg
Ec = 2 volts
REFERENCE FIG 33
DISCRIMINATOR HALF VOLTAGE DC
0
5
10
15
Ep = Esc = 67.5 v
LIMITER CHARACTERISTIC
1
2
3
4
5
LIMITER GRID VOLTS (RMS)

Fig 34

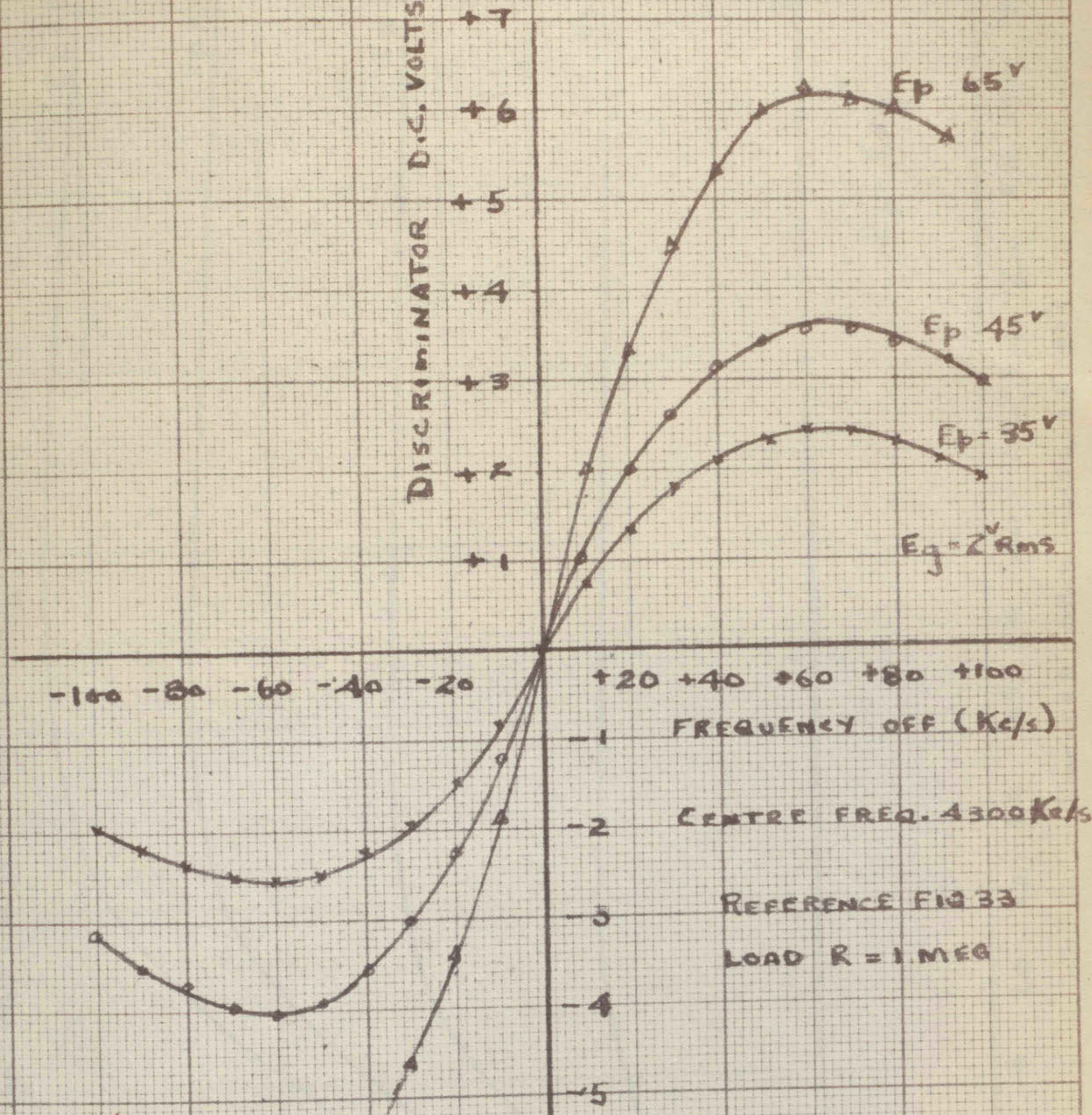

DISCRIMINATOR D.C. VOLTS
+7
+6
+5
+4
+3
+2
+1
-1
-2
-3
-4
-5
-6
-7
-100 -80 -60 -40 -20
+20 +40 +60 +80 +100
FREQUENCY OFF (Kc/s)
Ep 65v
Ep 45v
Ep 35v
Eg = 2v Rms
CENTRE FREQ. 4300 Kc/s
REFERENCE FIG 33
LOAD R = 1 MEG

FIG 35

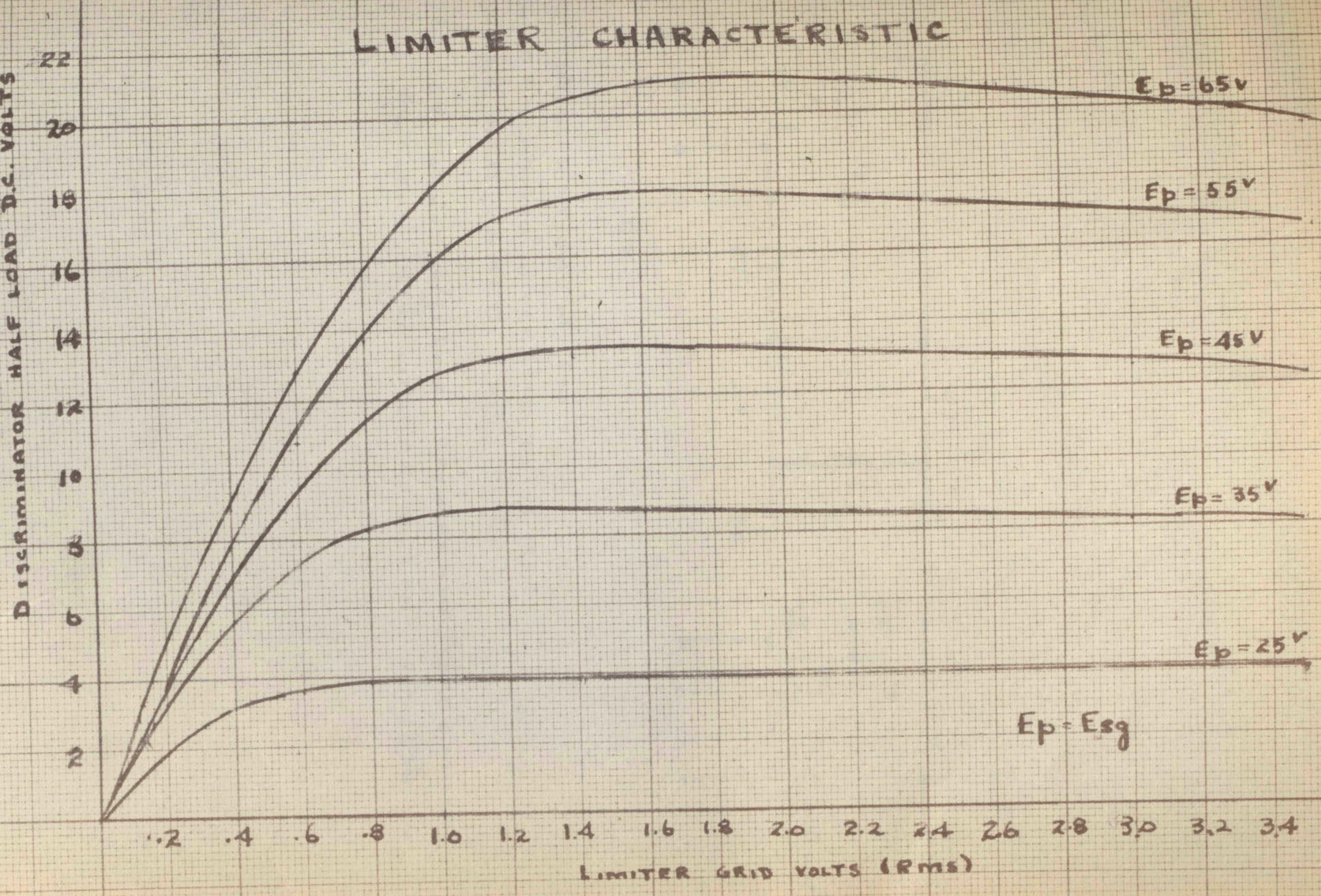
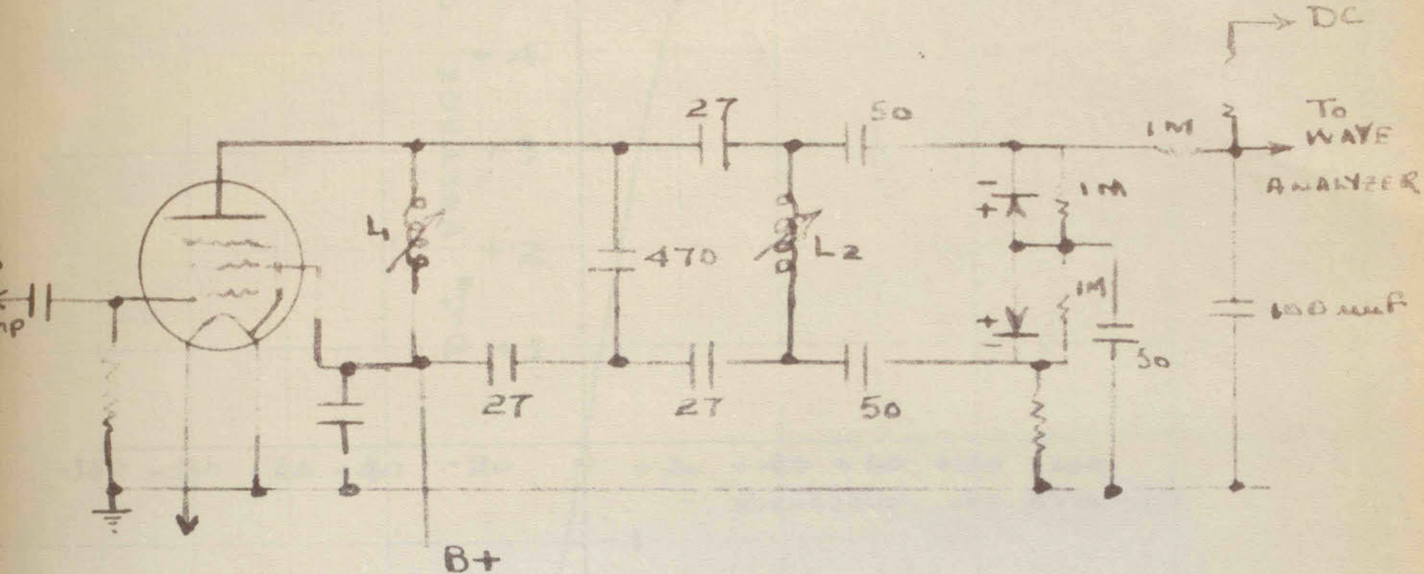

LIMITER CHARACTERISTIC
DISCRIMINATOR HALF LOAD D.C. VOLTS
22
20
18
16
14
12
10
8
6
4
2
.2 .4 .6 .8 1.0 1.2 1.4 1.6 1.8 2.0 2.2 2.4 2.6 2.8 3.0 3.2 3.4
LIMITER GRID VOLTS (RMS)
Ep = 65v
Ep = 55v
Ep = 45v
Ep = 35v
Ep = 25v
Ep = Esg

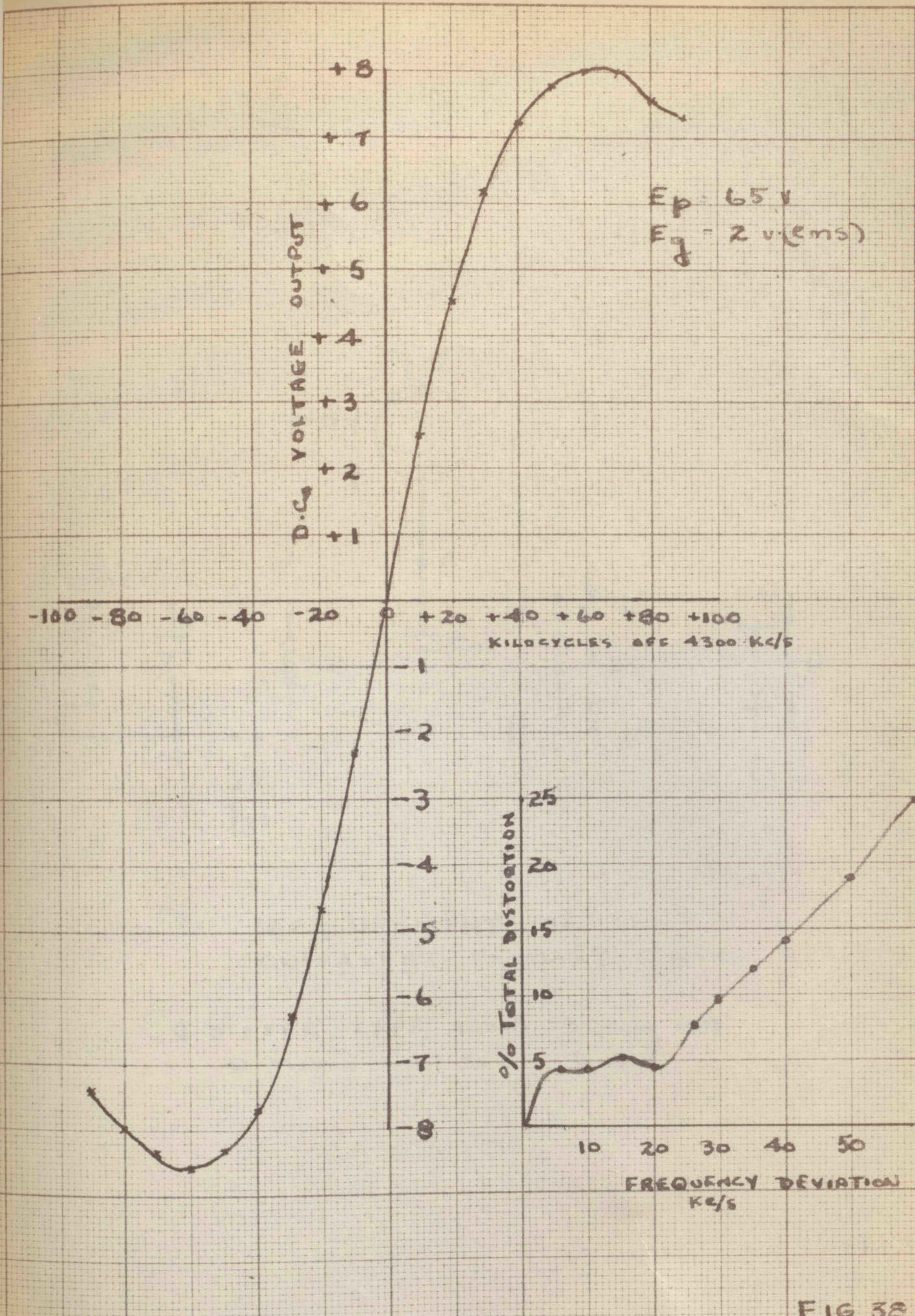
Fig 36



MODIFICATION OF FIG 55
LOAD RESISTORS INCREASED TO 1 MEG
COILS REDESIGNED

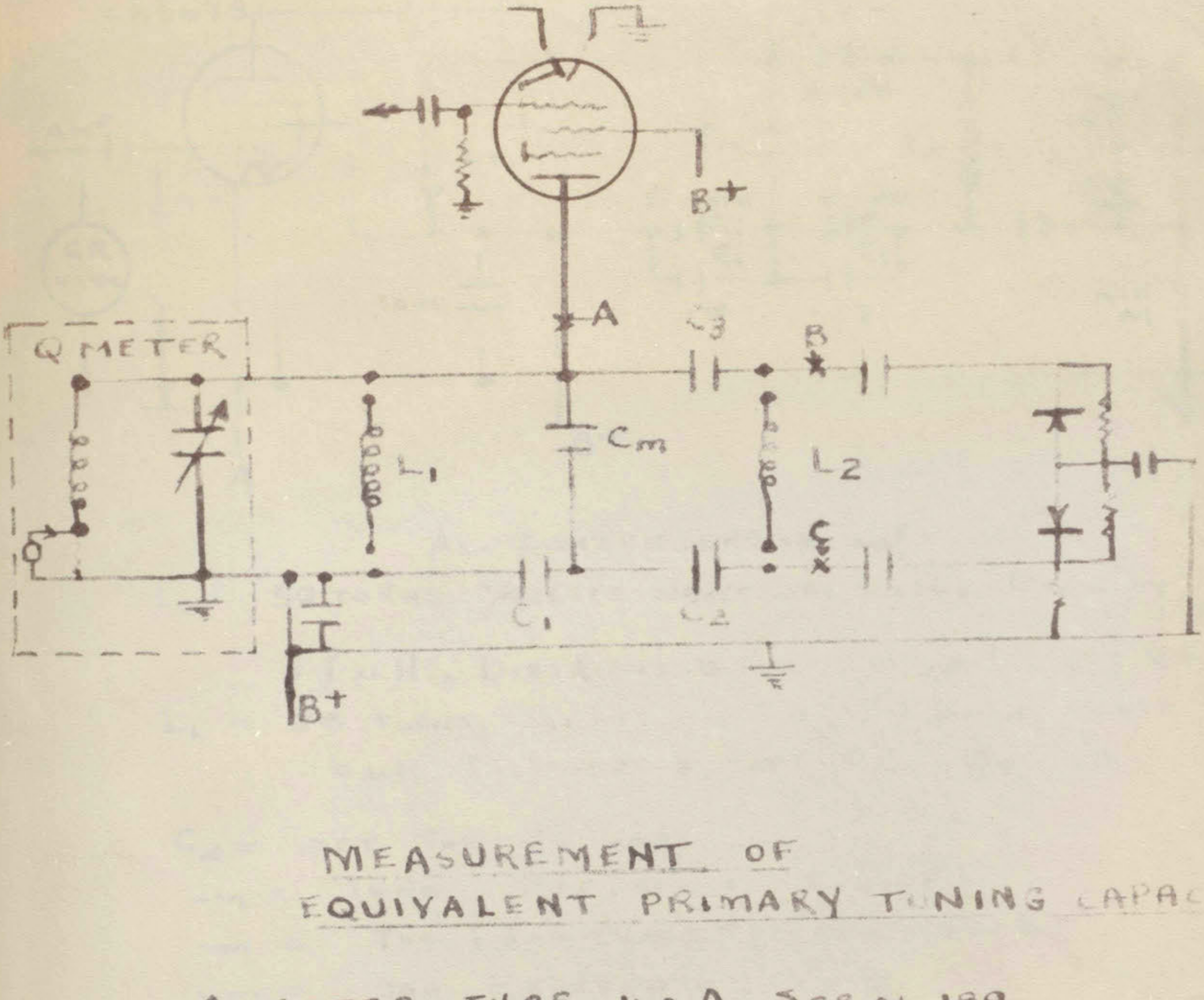
L1 58 TURNS - 3/41 LITZ CROWLEY THF CORE & SHELL
L2 56 TURNS - 3/41 LITZ CROWLEY SHF - CORE & SHELL

FIG 57



+8
+7
+6
+5
+4
+3
+2
+1
D.C. VOLTAGE OUTPUT
Ep 65 v
Eg 2 v (rms)
-100 -80 -60 -40 -20 0 +20 +40 +60 +80 +100
KILOCYCLES OFF 4300 Kc/s
-1
-2
-3
-4
-5
-6
-7
-8
25
20
15
10
5
% TOTAL DISTORTION
10 20 30 40 50
FREQUENCY DEVIATION Kc/s

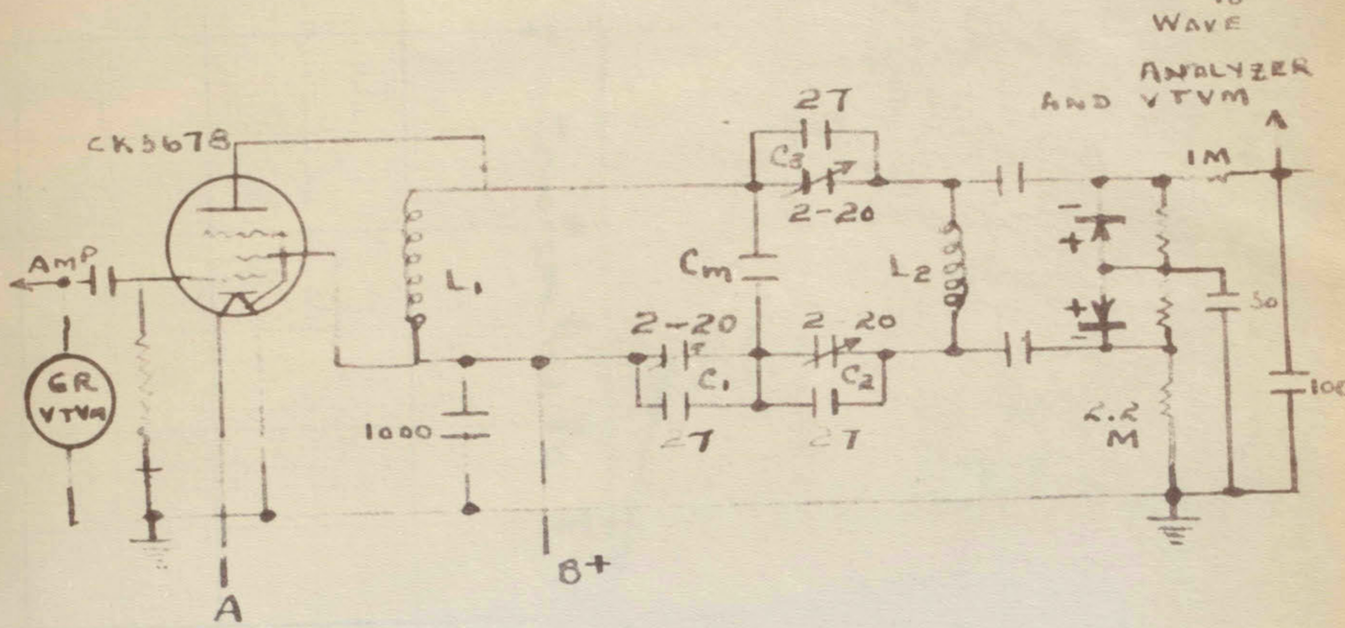
FIG 38



MEASUREMENT OF
EQUIVALENT PRIMARY TUNING CAPACITY

Q METER TYPE 160 A SERIAL 189
Q METER COIL 25 uh, 6 uuF

FIG 37



ALL CONDENSERS IN $\mu\mu$F

L_2 = 50 TURNS 3/41 LITZ UNIVERSAL WOUND – CROWLEY CORE AND SHELL THF

57 μH, DISTRIBUTED CAP. – 5.3 $\mu\mu$F Q_T 81

L_1 = 38 TURNS, 3/41 LITZ, UNIVERSAL WOUND, TYPE THF

26 μh. DISTRIBUTED CAP – 5.2, Q_T = 72

C_2 = SET FOR 36 $\mu\mu$F

C_m = 1000 REFERENCE FIG 41

C_m = 700 [500 + 200] REFERENCE FIG 42

C_m = 500 REFERENCE FIG 43

C_m = 200 REFERENCE FIG 44

VARIABLE TUNING CONDENSERS – WINGROVE & ROGERS AIR DIELECTRIC

CAPACITY TUNED

SHUNT CAPACITY COUPLED

FIG 40

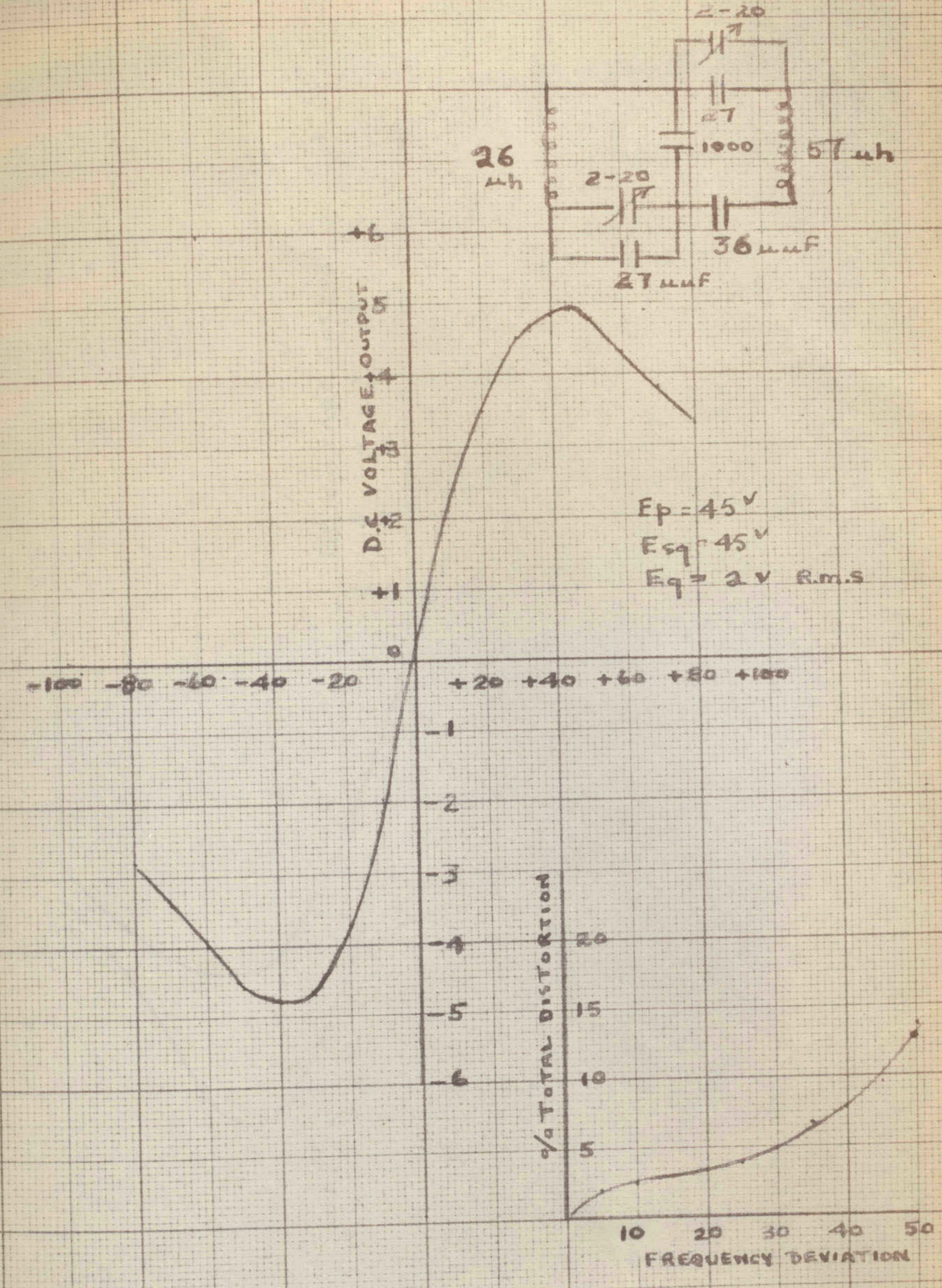

2-20
27
1000
26 μh
57 μh
2-20
36 μμF
27 μμF
D.C. VOLTAGE OUTPUT
Ep = 45V
Esg = 45V
Eg = 2V R.m.s
% TOTAL DISTORTION
FREQUENCY DEVIATION

FIG 41

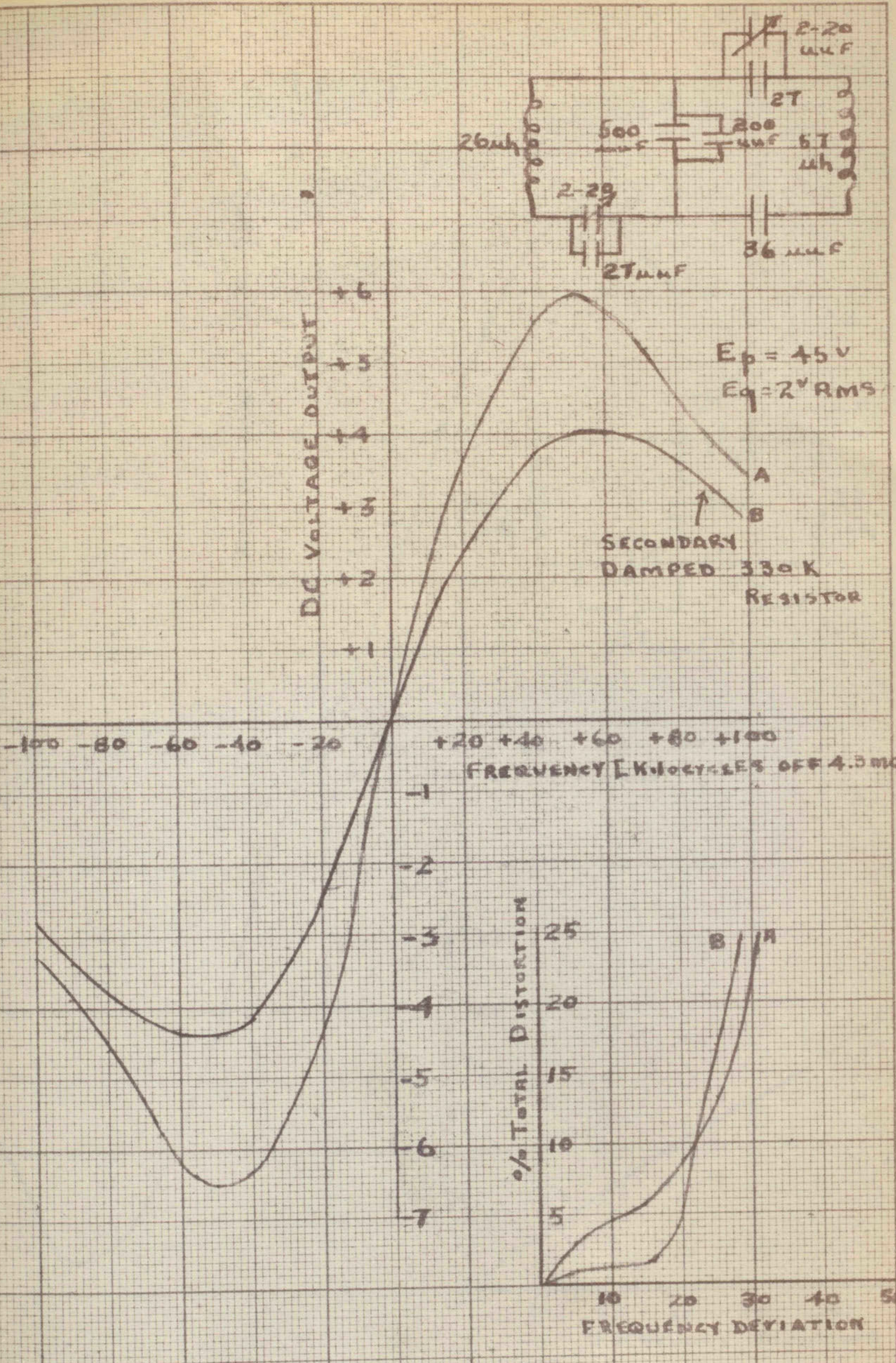

2-20 μμF
2T
26μh
500 μμF
200 μμF
67 μh
2-29
27μμF
36 μμF
DC VOLTAGE OUTPUT
+6
+5
+4
+3
+2
+1
-1
-2
-3
-4
-5
-6
-7
Ep = 45V
Eg = 2V RMS
A
B
SECONDARY DAMPED 330K RESISTOR
-100 -80 -60 -40 -20
+20 +40 +60 +80 +100
FREQUENCY [KILOCYCLES OFF 4.5 mc/s]
% TOTAL DISTORTION
25
20
15
10
5
10 20 30 40 50
FREQUENCY DEVIATION

FIG 42

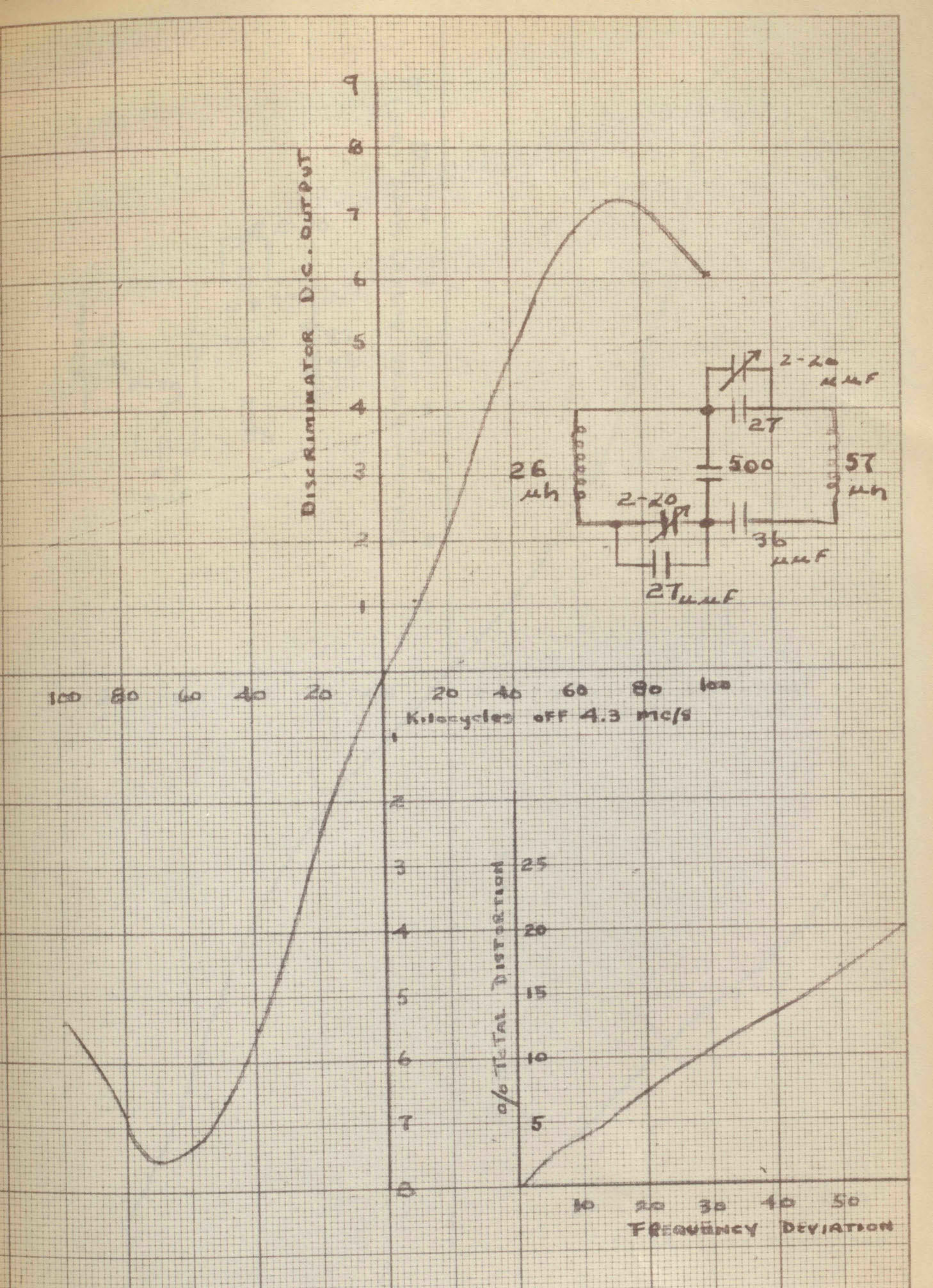

DISCRIMINATOR D.C. OUTPUT
Kilocycles off 4.3 mc/s
% TOTAL DISTORTION
FREQUENCY DEVIATION
2-20 μμF
27
500
26 μh
57 μh
2-20
36 μμF
27 μμF

FIG 43

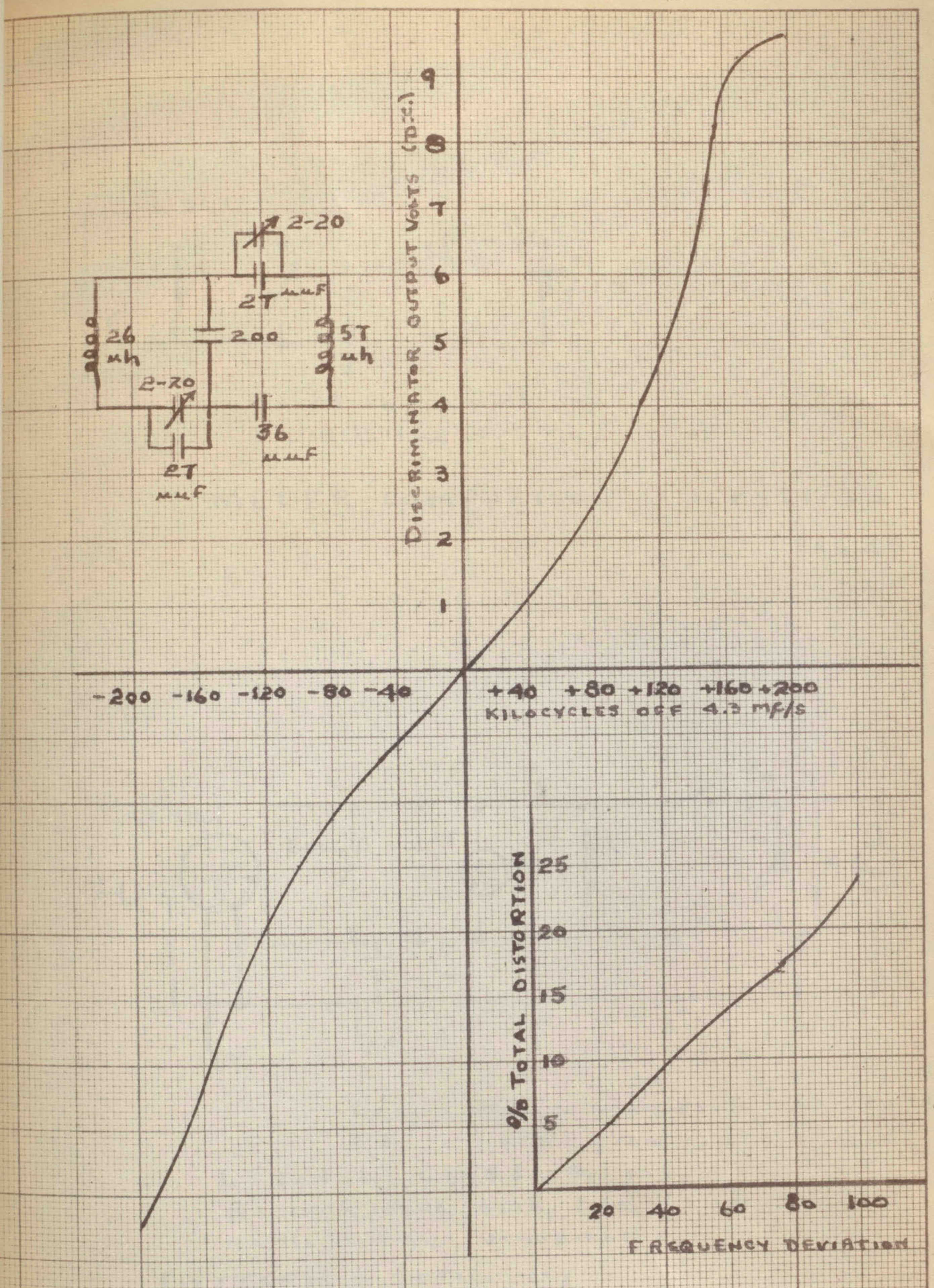
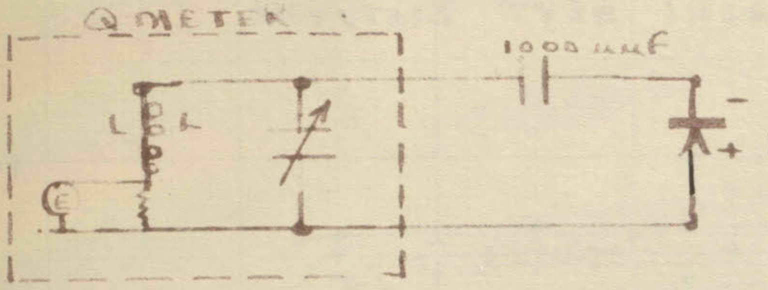

2-20
2T μμF
26 μh
200
25T μh
2-20
36 μμF
2T μμF
DISCRIMINATOR OUTPUT VOLTS (D.C.)
9
8
7
6
5
4
3
2
1
-200 -160 -120 -80 -40
+40 +80 +120 +160 +200
KILOCYCLES OFF 4.3 mc/s
% TOTAL DISTORTION
25
20
15
10
5
20 40 60 80 100
FREQUENCY DEVIATION

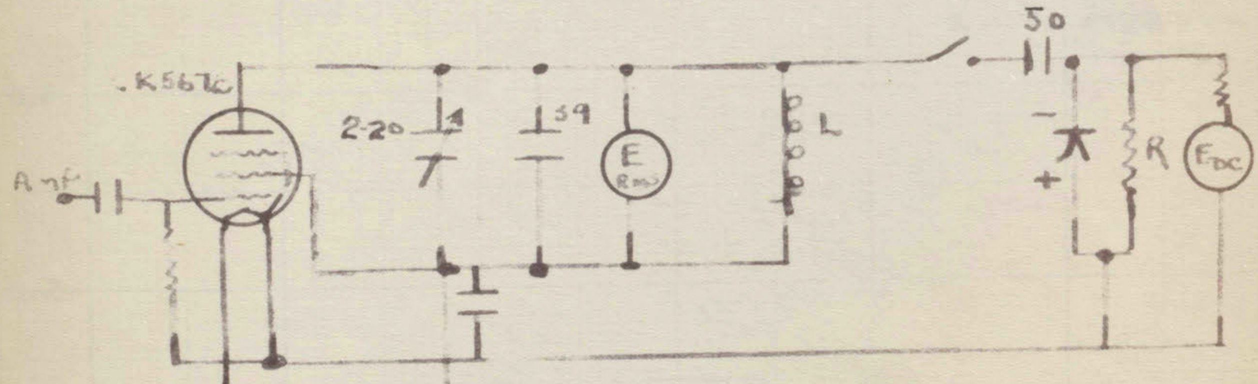
FIG 44



L - 25 uh, 6 uuf
FREQUENCY 4.3 Mcs

Q METER COMPARISON OF CRYSTALS

FIG 45



ERMS - GENERAL RADIO VTVM 726A SN 184
EDC - BARBER VTVM MODEL 27 SN 647

L = 22.1 uh, Cd = 5.3, QT - 104
R - 120K, 330K, 470K, 1.0 M., 2.2 M, INFINITE
SIGNAL 4.3 Mc/s UNMODULATED

FIG 46

DETECTION EFFICIENCY

DETECTION EFFICIENCY OF GERMANIUM CRYSTALS TYPE 1N34 AT 4.3 Mc/s

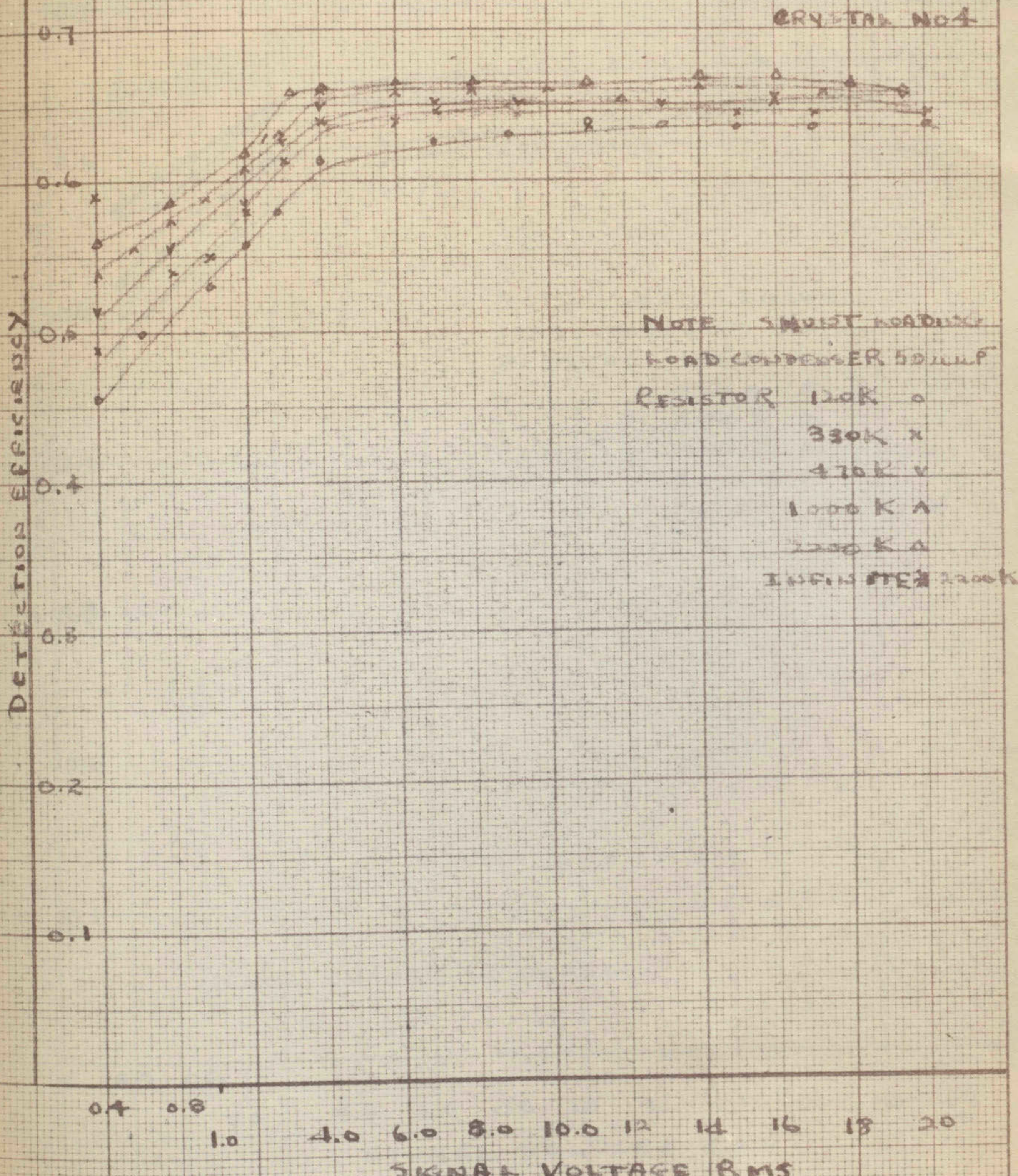


FIG 47

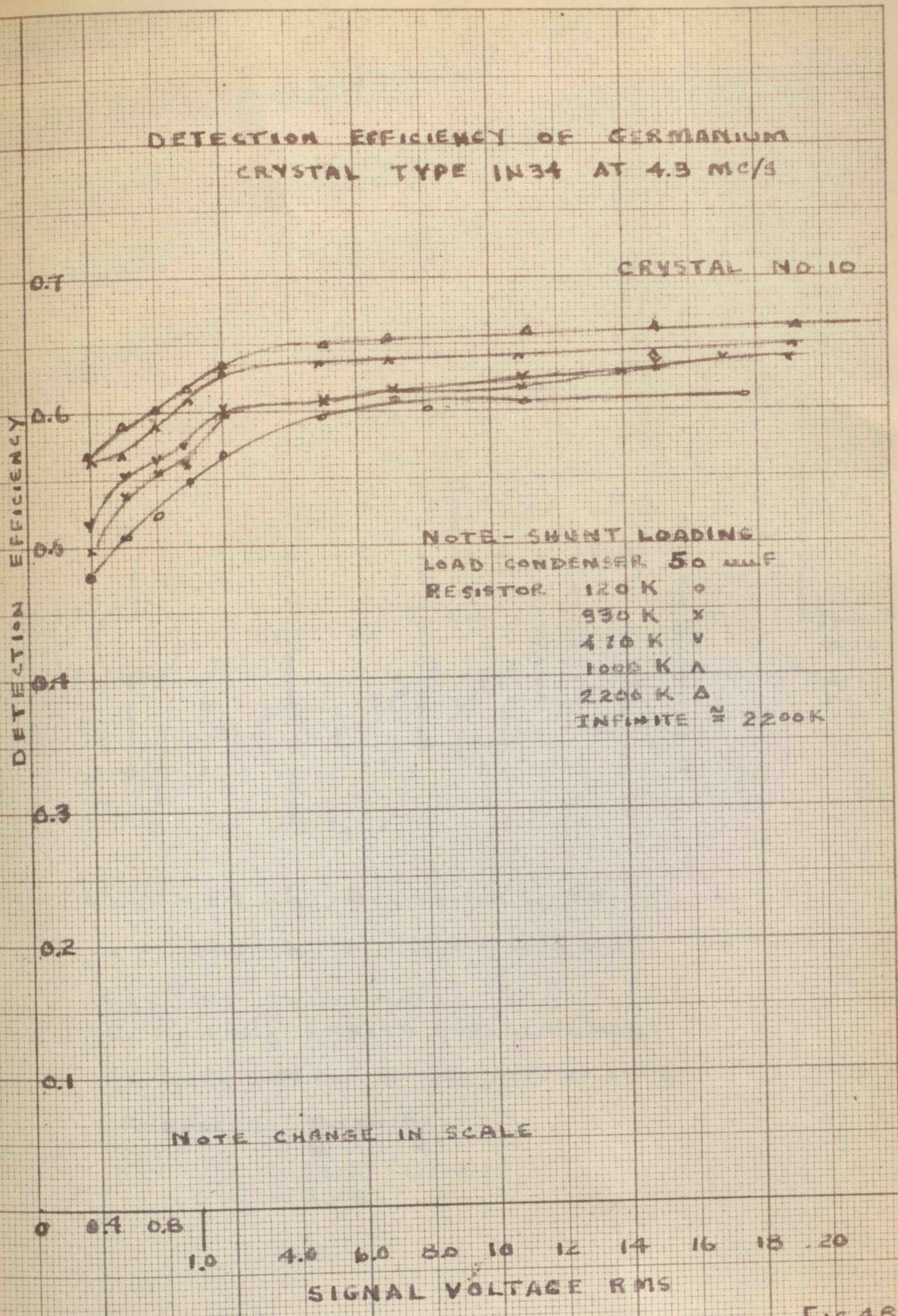
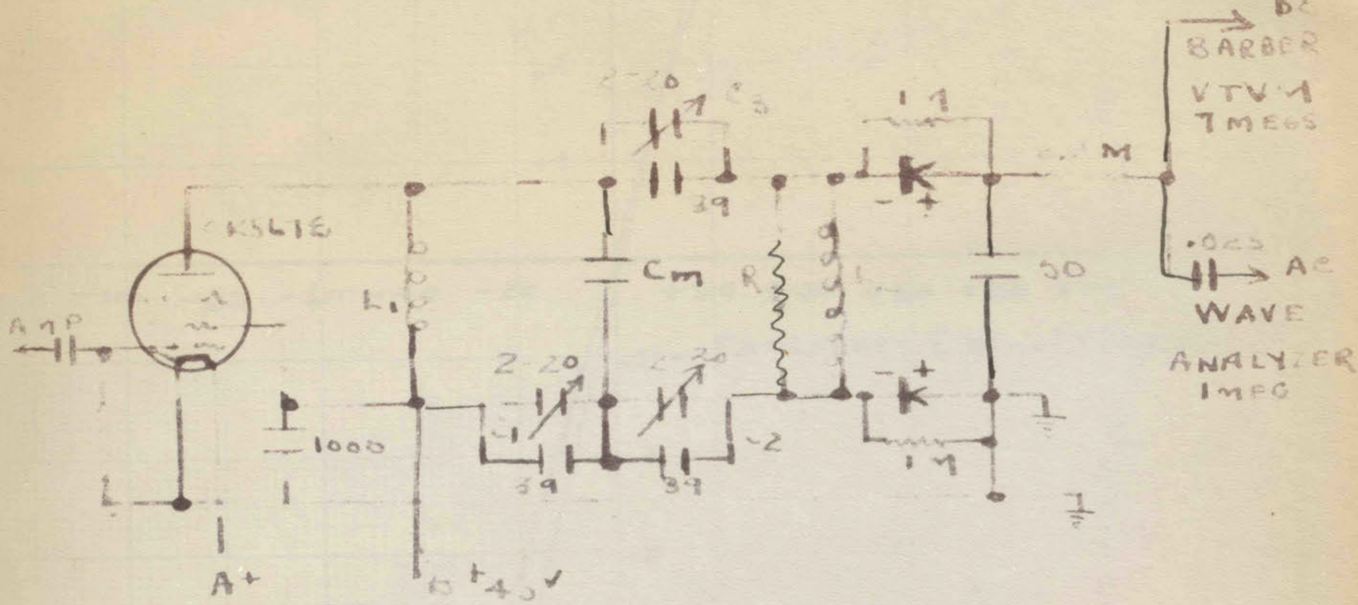

DETECTION EFFICIENCY OF GERMANIUM
CRYSTAL TYPE IN34 AT 4.3 MC/S
CRYSTAL NO 10
0.7
0.6
0.5
0.4
0.3
0.2
0.1
DETECTION EFFICIENCY
NOTE - SHUNT LOADING
LOAD CONDENSER 50 μμF
RESISTOR 120 K o
330 K x
470 K v
1000 K ^
2200 K Δ
INFINITE ≅ 2200K
NOTE CHANGE IN SCALE
0 0.4 0.8
1.0 4.0 6.0 8.0 10 12 14 16 18 20
SIGNAL VOLTAGE RMS

FIG 46



L_2 = 44.2 uh, Distributed C 5.0 uuf, Q_T = 97

L_1 = 22.1 uh, Distributed C 5.3 " Q_T = 104

C_2 SET TO 41 uuf

C_1 and C_3 VARIED FOR ALIGNMENT

C_1, C_2, C_3 = 39 uuf ceramic + 2-20 AIR DIELECTRIC VARIABLE

REFERENCE FIG 50, 51

R , R

150K INFINITE

330K

FIG 49

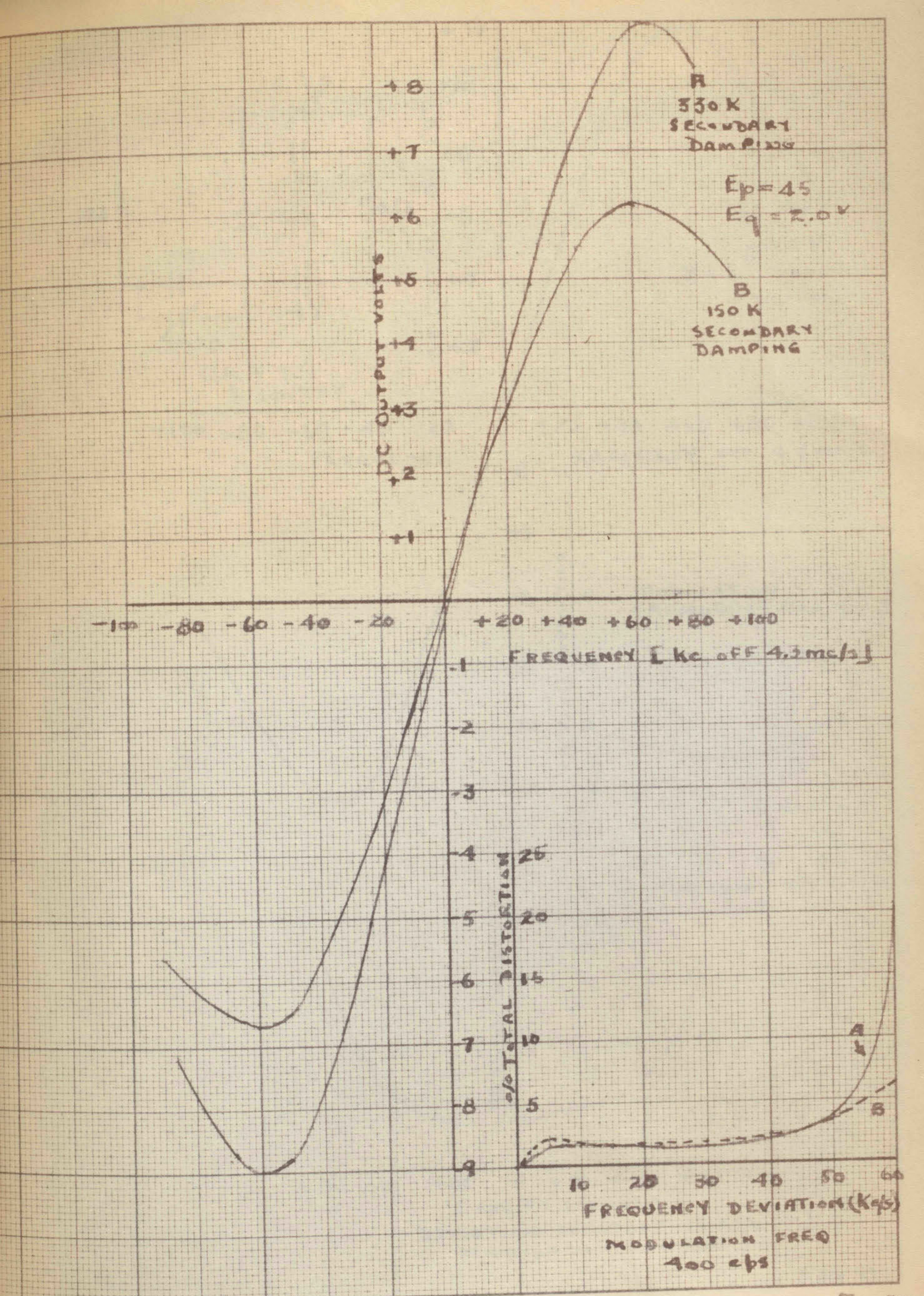

A
330 K
SECONDARY
DAMPING
Ep = 45
Eg = 2.0V
B
150 K
SECONDARY
DAMPING
DC OUTPUT VOLTS
+8
+7
+6
+5
+4
+3
+2
+1
-100 -80 -60 -40 -20
+20 +40 +60 +80 +100
FREQUENCY [Kc off 4.5 mc/s]
-1
-2
-3
-4
-5
-6
-7
-8
-9
% TOTAL DISTORTION
25
20
15
10
5
10 20 30 40 50 60
FREQUENCY DEVIATION (Kc/s)
MODULATION FREQ
400 c/s
A
B

FIG 50

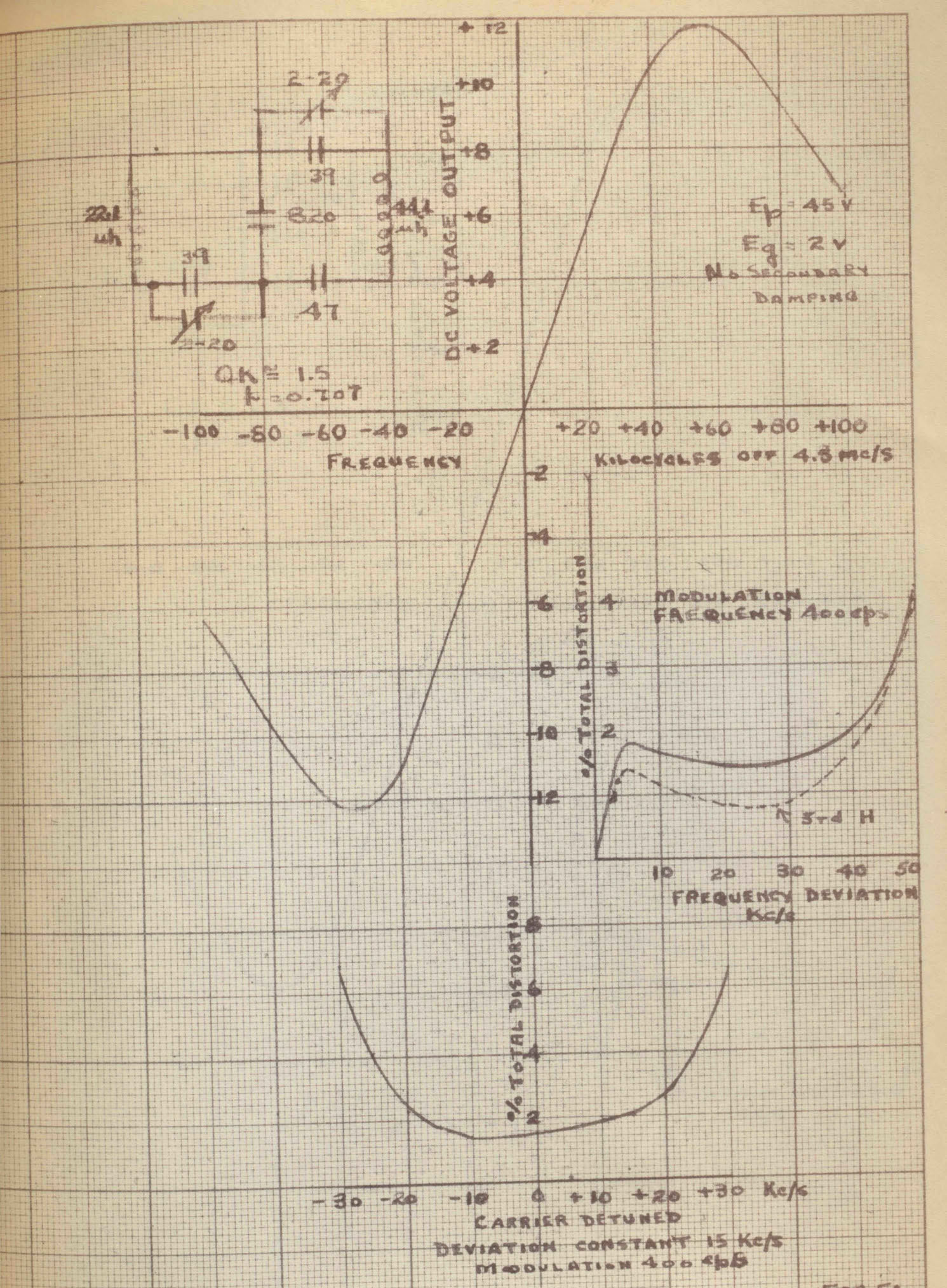
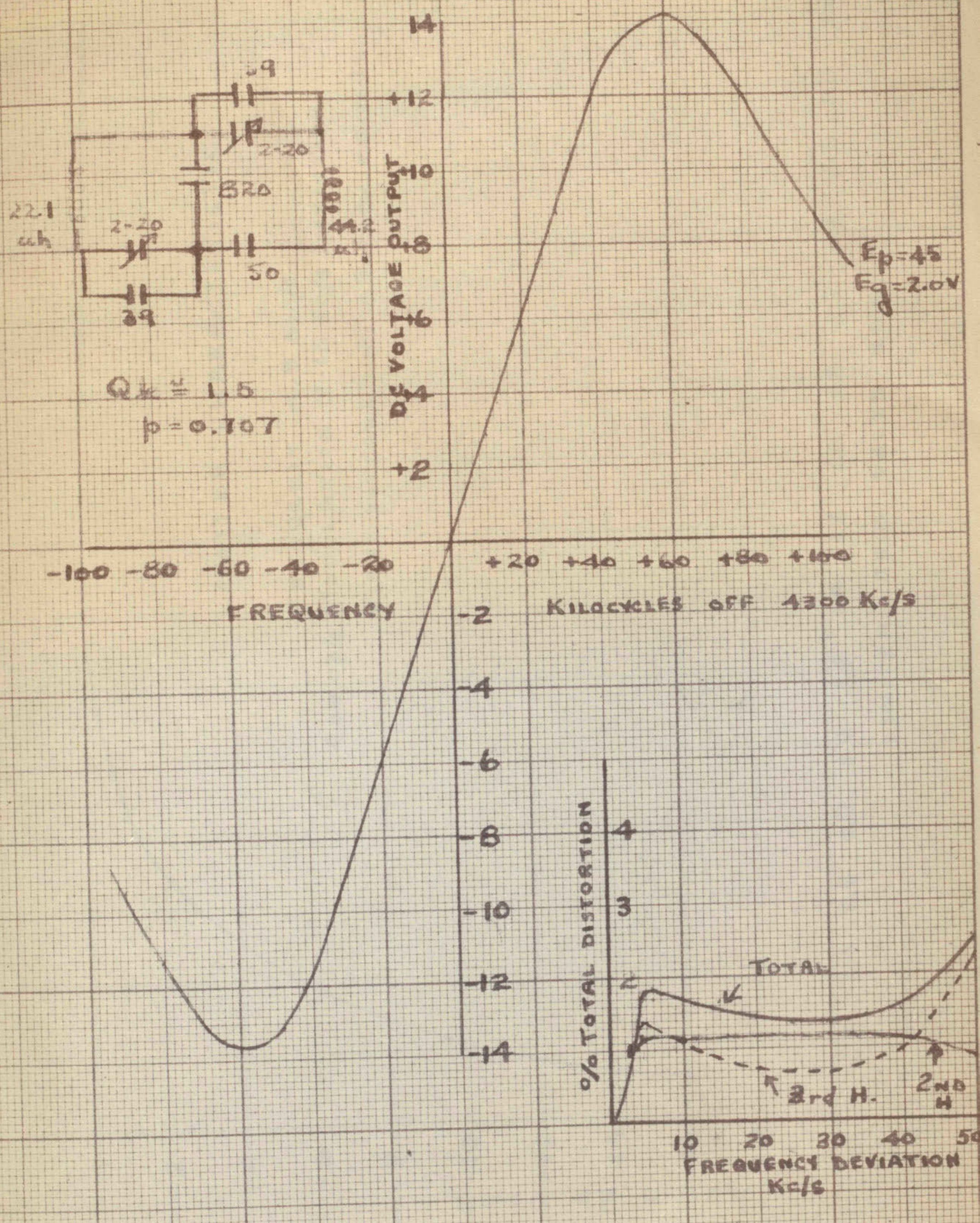

2-20
39
820
221 uh
441 uh
39
47
2-20
QK = 1.5
k = 0.707
DC VOLTAGE OUTPUT
+12
+10
+8
+6
+4
+2
-100 -80 -60 -40 -20
+20 +40 +60 +80 +100
FREQUENCY
KILOCYCLES OFF 4.8 mc/s
-2
-4
-6
-8
-10
-12
Ep = 45 V
Eg = 2 V
No SECONDARY DAMPING
% TOTAL DISTORTION
4
3
2
MODULATION FREQUENCY 400 cps
3rd H
10 20 30 40 50
FREQUENCY DEVIATION Kc/s
% TOTAL DISTORTION
8
6
4
2
-30 -20 -10 0 +10 +20 +30 Kc/s
CARRIER DETUNED
DEVIATION CONSTANT 15 Kc/s
MODULATION 400 cps

FIG 51



49
2-20
820
22.1 uh
2-20
44.2 uh
50
39
Qk ≅ 1.5
p = 0.707
DC VOLTAGE OUTPUT
14
+12
+10
+8
+6
+4
+2
-2
-4
-6
-8
-10
-12
-14
-100 -80 -60 -40 -20
+20 +40 +60 +80 +100
FREQUENCY KILOCYCLES OFF 4300 Kc/s
Ep=45
Eg=2.0V
% TOTAL DISTORTION
4
3
2
TOTAL
2rd H.
2ND H
10 20 30 40 50
FREQUENCY DEVIATION Kc/s

FIG 52

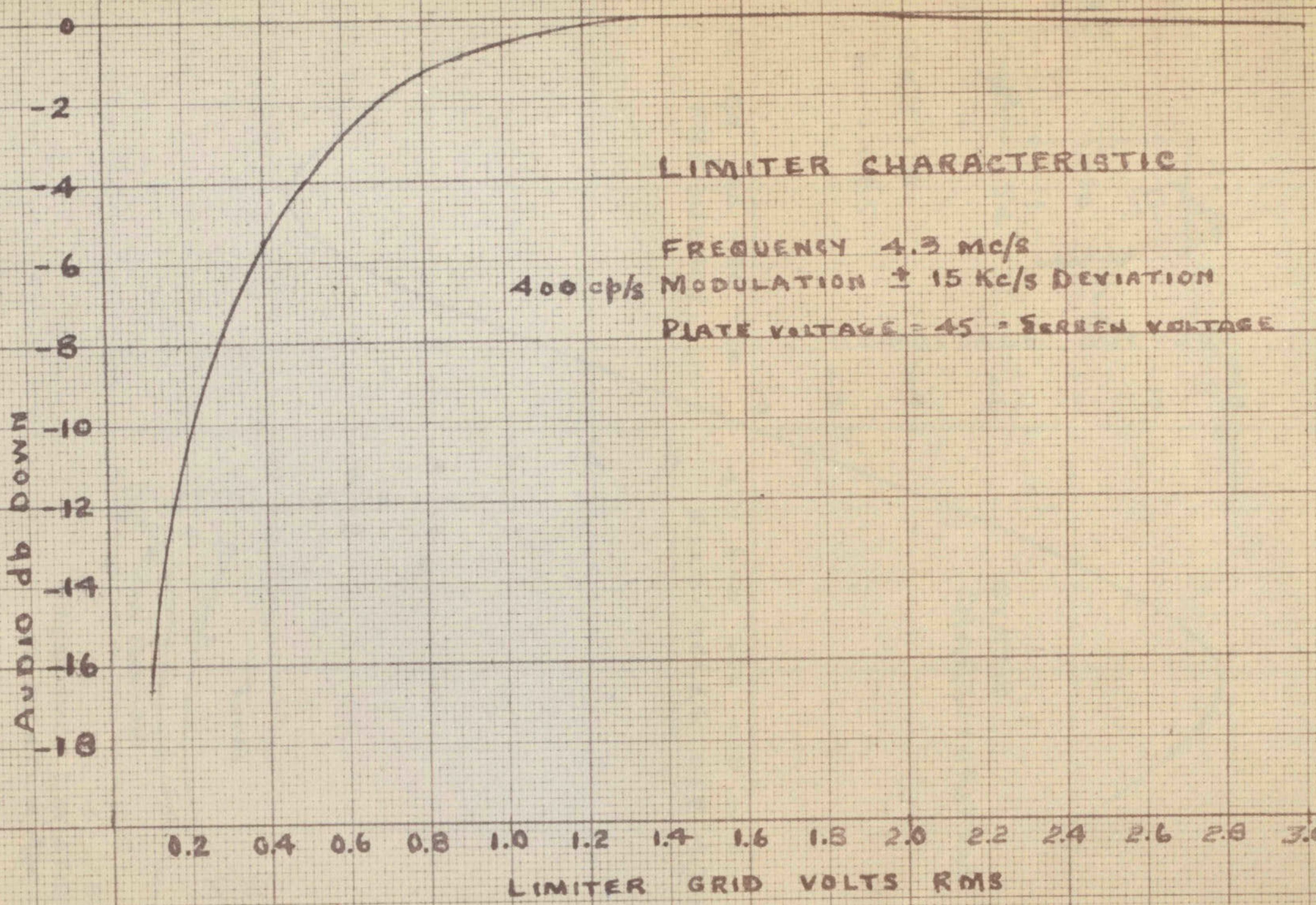
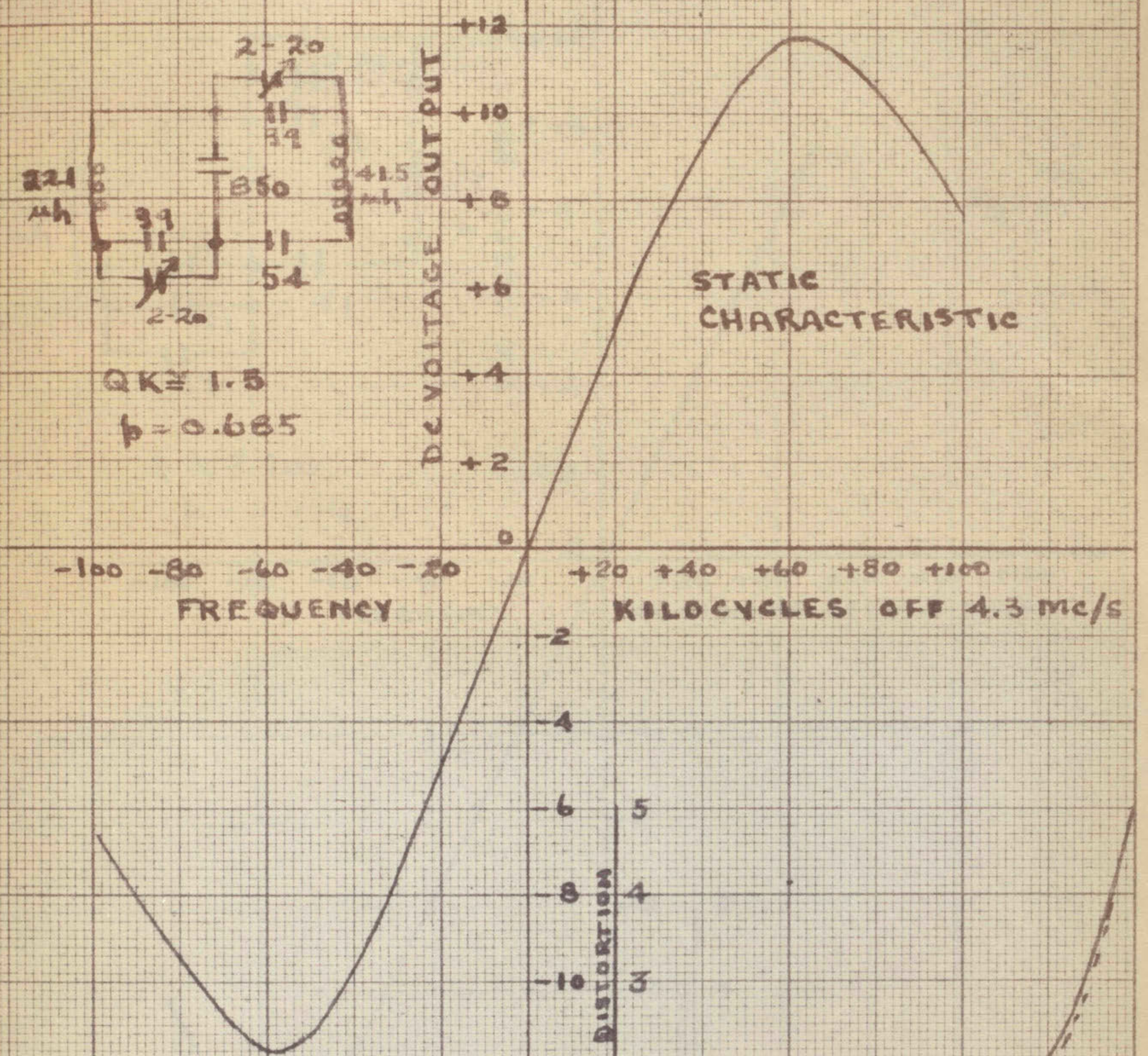

LIMITER CHARACTERISTIC
FREQUENCY 4.3 Mc/s
400 cp/s MODULATION ± 15 Kc/s DEVIATION
PLATE VOLTAGE = 45 = SCREEN VOLTAGE
0
-2
-4
-6
-8
-10
-12
-14
-16
-18
AUDIO db DOWN
0.2
0.4
0.6
0.8
1.0
1.2
1.4
1.6
1.8
2.0
2.2
2.4
2.6
2.8
3.0
LIMITER GRID VOLTS RMS

Fig 53



2-20
39
221 μh
850
41.5 μh
39
54
2-20
QK≅ 1.5
p = 0.685
DC VOLTAGE OUTPUT
+12
+10
+8
+6
+4
+2
0
-2
-4
-6
-8
-10
-12
STATIC CHARACTERISTIC
-100 -80 -60 -40 -20
+20 +40 +60 +80 +100
FREQUENCY
KILOCYCLES OFF 4.3 mc/s
% TOTAL DISTORTION
5
4
3
2
1
TOTAL
3rd H
2nd Harmonic
10 20 30 40 50 60
FREQUENCY DEVIATION Kc/s
MODULATION 400 cps

FIG 54

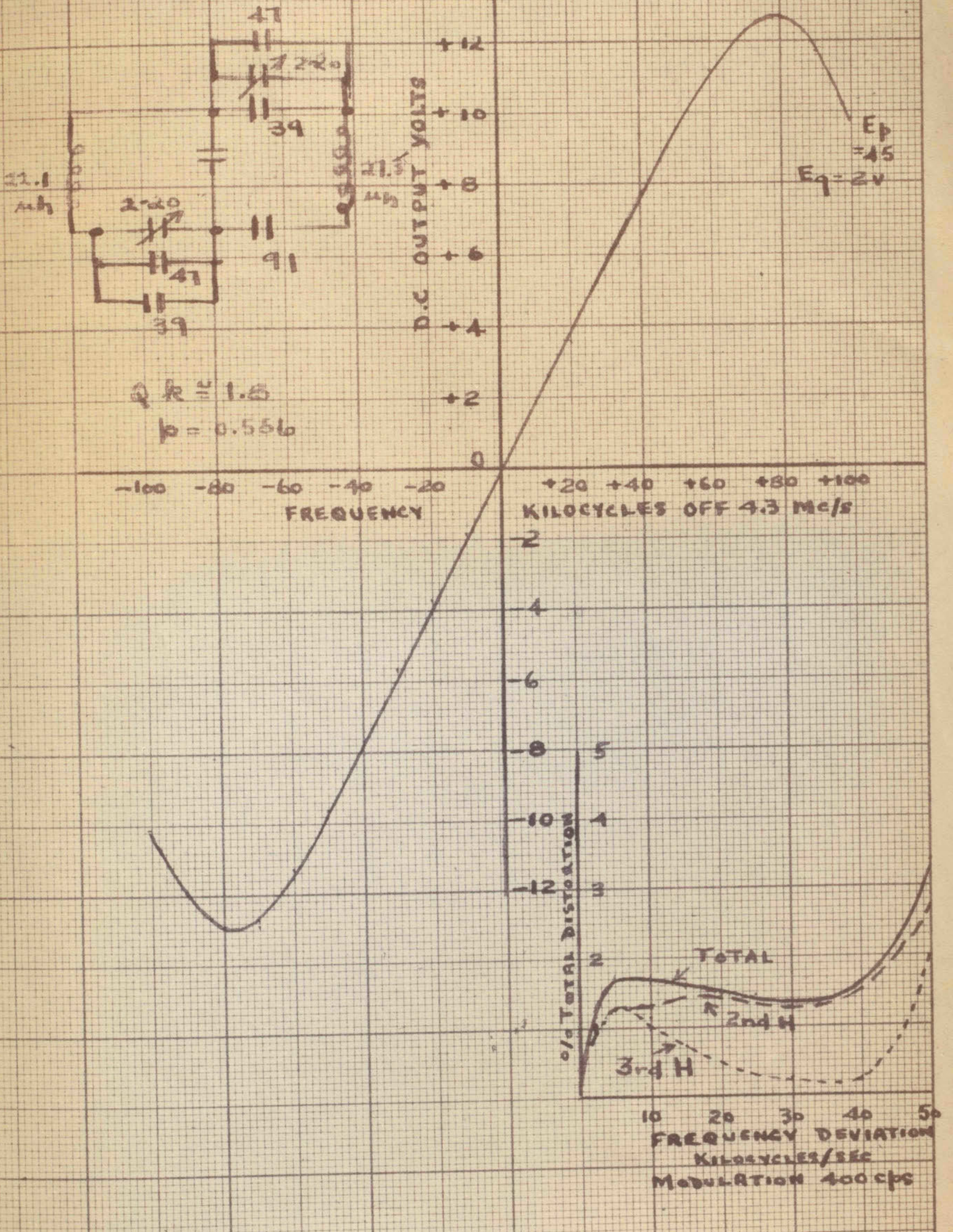

47
220
39
21.1 mh
2-20
21.5 mh
91
47
39
Q k ≅ 1.6
p = 0.556
D.C OUTPUT VOLTS
+12
+10
+8
+6
+4
+2
0
-2
-4
-6
-8
-10
-12
-100 -80 -60 -40 -20
+20 +40 +60 +80 +100
FREQUENCY KILOCYCLES OFF 4.3 Mc/s
Ep =45
Eg=2v
% TOTAL DISTORTION
5
4
3
2
TOTAL
2nd H
3rd H
10 20 30 40 50
FREQUENCY DEVIATION
KILOCYCLES/SEC
MODULATION 400 cps

FIG 55