Linearity of some modulation circuits.

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L I N E A R I T Y o f S O M E M O D U L A T I O N C I R C U I T S

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" Linearity of Some Modulation Circuits "

(A Thesis submitted as part of the requirements for the Degree of Master of Science of the University of Aberdeen.)

by Maurice Clifford Hately

I hereby declare that the thesis has been composed by myself, that it has not been accepted in any previous application for a higher degree, that the work of which it is a record has been done by myself, and that all quotations have been distinguished by quotation marks and sources of information specifically acknowledged.

April 1975

SUMMARY

of Thesis

LINEARITY OF SOME MODULATION CIRCUITS

by M C Hately

The work consists of a detailed study of the methods of analogue modulation and demodulation of electrical carrier waves by amplitude and frequency with the object of presenting improved techniques where necessary.

The thesis is divided into five parts of which the last is devoted to the conclusions.

The first and second parts deal with amplitude modulation and demodulation. A novel form of mismatched diode modulator circuit is proposed to improve the linearity of such processes. Developed forms are tested with sinusoidal carrier drive. It is shown that the mismatched demodulator will eliminate distortion due to diagonal clipping.

By Fourier analysis it is shown that chopped sine wave modulators, of which there are many types presently used, cannot be deeply modulated and remain linear. The mismatched modulators are re-tested with square wave drive and improved performance is obtained.

The third and fourth parts are given to an examination of frequency modulation and demodulation. For applications of low cost, two simple linear frequency modulators are recommended. For receivers, a new theoretical approach to the spectrum of a frequency modulated wave is presented, in which the Bessel series are derived using a straightforward expansion technique. The sensitivity and linearity of the Foster-Seeley type discriminators are investigated in detail.

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MAIN CONTENTS

DETAILED CONTENTS

Part C FREQUENCY MODULATION

Part D FREQUENCY DEMODULATION

OBJECTIVE

The objective of the work has been to consider the linearity of presently used modulation circuits both in theory and practice with a view to devising improved techniques where necessary.

Owing to the early application of new electronic inventions to communications, many of the established circuits were introduced practically before the theory of their operation had been fully analysed.

In this project the general procedure will be to:-

<u>EXAMINE</u> — Previous Techniques
Previous Theories

in order to:-

<u>DEVISE</u> — New Theories
New Techniques

verified by tests leading to:-

<u>New Applications</u>

Part A

A M P L I T U D E   M O D U L A T I O N

Part A AMPLITUDE MODULATION

Section A1 Previous Techniques and Theories

A1.1 Previous Techniques

A1.1.1 Resistive Modulation

The most elementary technique for modulation has been the control of a current flowing from source to load by means of a variable resistance under the command of the signal to be modulated upon the carrier. Such a device is exampled by the carbon microphone usually attributed to Edison though earlier invented by Henning in Britain. The electromagnetic microphone (identical to the receiver headphone) invented by Bell had provided such a feeble current that as soon as the signals passed through a few switch contacts they became inaudible. Edison's use of the carbon microphone to modulate a comparatively large direct current from a battery enabled the previous difficulties to be overcome.

The basic system of Resistive Modulation is shown in Figure A1.1.1 . The resistor R_c is tightly controlled by the signal voltage in some way, and therefore changes the current flow to the load resistor R_L across which changes of voltage occur, v_o .

For the crude d.c. telephone mentioned above, the source is the battery and the control resistor is varied by air pressure. For an a.c. modulation circuit, v_c

the source, will be an a.c. generator (here taken to be constant voltage and zero impedance) and the control resistor

will be either a field effect transistor below pinch-off

or a bipolar transistor biased to a region of linear transconductance

or a tetrode, pentode, or heptode valve on some variable transconductance region

suppose $R_c = R_1/v_s$

then $$v_o = \frac{v_c R_L}{R_L + R_1/v_s}$$

$$= \frac{v_c}{1 + R_1/R_L v_s}$$

so $$\frac{v_o}{v_c} = \left(1 + R_1/R_L v_s\right)^{-1}$$

$$= 1 - \left(\frac{R_1}{R_L}\right)\frac{1}{v_s} + \left(\frac{R_1}{R_L}\right)^2\left(\frac{1}{v_s}\right)^2 - \ldots\ldots\ldots$$

by Binomial Theorem

which is linear only for small fractional depths of modulation, since R_L must exceed R_1 greatly, to neglect $1/v_s^2$.

Similar deductions result if output current, or output power is calculated.

Al.1.2 Non-linearity Modulation

The application of carrier wave and signal wave added together across a non-linear device has been the most usual form of a.c. modulator. It is the basis of the following devices

i) diode modulator Fig Al.1.2.i

ii) switch modulator Fig Al.1.2.ii

iii) Class C modulators such as

Grid modulated valve Fig Al.1.2.iii

Screen modulated valve

Anode modulated valve

Base bias modulated transistor

It may be said of all these circuits the load element receives from the non-linear device a chopped carrier wave in which the fraction of the original sinusoid is changed by the instantaneous magnitude of the signal. The phenomenon is illustrated in Figure Al.1.2.iv .

Considering the long history of such devices and the number installed throughout every field of radio and telecommunication engineering, it may seem churlish to say that such circuits are inherently distorting. (The term non-linear must unfortunately be retained here for the waveform chopping device). Such distortion is hinted at in some distinguished radio engineering circles. Schwartz (26) page 102 speaking of the Class C plate modulated amplifier says " the magnitude of the plate current varies very nearly linearly with the plate voltage ". BBC television transmitters contain a video pre-distortion circuit prior to the modulator. However the full investigation of linearity necessitates a degree of academic ruthlessness, and so careful analysis of a chopped-up sinusoidal wave train is made below, in paragraph A2.2 and Appendix A2.2

A1.2 Previous Theories

There have been three approaches to amplitude modulation theory. Each has produced some facts, but none alone has been found to give the explanation under all conditions. The approaches may be designated:-

i) polynomials (and multiplication)

ii) switching functions (and Fourier analysis)

iii) graphics and waveform sketching

These will be looked at in succession. The best of each will be extracted in order to synthesise an improved viewpoint for the future.

A1.2.i) Polynomials and Multiplication

Assuming that modulation is multiplication of sinusoidal functions, and assuming that a fair approximation to a non-linear characteristic may be represented by a polynomial expression some useful general deductions have been made. For instance Everitt (11) page 398 uses a polynomial to show that first order terms are harmless, second order are the necessary function, third order are harmful and that higher order terms are negligible provided signal magnitudes are small.

A more recent authority, Betts (8) introduces the second chapter "Modulation" by attempting a pure multiplication method. Unfortunately there results immediately, double sideband suppressed carrier modulation and he must therefore abandon multitiplication and re-introduce another method leading to " full AM ".

Brown & Glazier (9) page 60 say that modulation is "theoretically very simple" but thence proceed to remark that the multiplier devices quoted, such as pentode, Hall effect device, the analogue computer multiplier, are either insufficiently linear or low in output or complex. A normal polynomial approach is given for low level modulation and then finally a " carrier frequency valve" is cited as the ultimate in high quality, though why is not explained.

A1.2.ii) Switching Functions (and Fourier Analysis)

For simplicity the switching function approach is based upon piecewise linear representations of the non-linear devices and large valued sinusoidal functions. Thus initially pure sinusoids become broken up into harmonic series reducible to algebraic functions only by Fourier analysis. The list of references to this mathematical technique is not large apparently commencing with Bennett (7) hypergeometric series, Belevich (2), (3),(4) studying the effect of changing restance values for source , load and diode, followed very closely by Tucker (30) involving much mathematics and later following a thesis by Belevich of 1959 Tucker published a renewed attack on the problem in a book (31) in which the Fourier series are taken away from the voltages and assigned to the resistors, reactors and other circuit components. Unfortunately the mathematics remains as intractable as ever. "We have an infinite number of equations with an infinite number of terms" ibid p 21. But the subject was kept alive at this time by Howson and Tucker (22) embarking on the more significant problem of intermodulation products. Latterly there have been papers by Gardiner & Surana (13), Gardiner (14), Yousif & Gardiner(32), and Gardiner(15), most being concerned with the piecewise linear and switched current functions but with more extensive application of these ideas. In ref(14) Gardiner actually reaches the point of saying that "benefit is obtained from the offset" (meaning diode offset voltage) which he has added into the Tucker theory to make it more realistic. A crude experimental technique in which shunt connected capacitor-resistor pairs are connected in series with the diodes to increase the offset voltage artificially is said to produce results which verify the theory. Possibly some unsuspected resonance phenomenon between the large capacitor used (10 000 pF) and the reverse current pulse due to diode charge storage (not considered by Gardiner) may have given rise to the apparent improvement. It can hardly be true that the larger the diode forward voltage ("offset"of Gardiner) the better the diode is as a low distortion modulator. A more meaningful method of artificially increasing forward voltage would have been to use two or three diodes in series.

To sum up, it is interesting to note that the last reference brings Gardiner right to the position in which the present author found himself in regard to mismatch as the best way to reduce intermodulation; Patent Application 23 April 1971, (20).

Al. 2. iii) Graphics and Waveform sketching

In most of the elementary textbooks there has been a graphical first account to the techniques of amplitude modulation. Some presentations have given a piecewise linear representation of the non-linear device. Some have given a smoothly changed conductance picture. But in either case the result has been a very good introduction to waveform reproduction from signal frequency up onto the carrier wave. Unfortunately such drawings have looked rather too convincing. It has been assumed that linear modulation has been achieved when a train of part sinusoids of linearly changing peak amplitude has been supplied to the output element. It will be shown below a more careful analysis should be performed as a check.

Al. 3 Conclusions on Previous Techniques & Theories

The three basic theoretical approaches to amplitude modulation which apply to the two mainstream techniques have been reviewed and found to have contributed a partial understanding to any of the research workers concerned. It may therefore be deduced that a better understanding (in the absence of a better non-linear mathematical calculus) must be based upon a selective application of theories as seems best in the circumstances.

Section A 2 AM Linearity by Mismatch and Square Wave Carrier

A2.1 AM Linearity by Mismatch

A2.1.1 Single-ended Mismatched Modulator

A well established axiom of measurement technique is to construct each instrument so as to cause the least disturbance to the physical situation in which it is operating. Thus an ideal voltmeter would be one which took zero current from the electro-motive source it sought to observe. An ideal ammeter would be one which could indicate an accurate current value without developing any voltage drop by resistance.

Secondly, it is a further axiom of good practice in physical measurement to see that any non-linear tendency is overwhelmed by exterior major phenomena whose linearity can be relied upon.

Since the above measurement principles apply to the modulation process (a signal amplitude measured in the base band is to be accurately reproduced in the carrier band) and since the chief vague feature of the previously known modulators has been their characteristic curvature, it would seem wise to try to devise a modulator whose behaviour in regard to the above points can be properly designed.

Such proper designation will be most nearly achieved if the diode is employed in its most appropriate role Viz. A CURRENT SWITCH. In other words the diode will be acting as the current control element between a current source and a current sink. To ensure that the desired conditions can never alter, it is arranged that the diode lies between a high impedance (or high impedances) and a low impedance. In other words a MISMATCHED condition.

Presently used modulator circuits do not follow such principles. Taking for example the diode detector (a modulator in strictly theoretical terms by virtue of its ability to take a carrier plus two side frequencies and generate the new base-band frequency from the triple). A diode works between a low impedance source into a medium or high impedance load. Presumably an arrangement which is a legacy from the days of power rectification when efficiency was the prime objective.

Under these circumstances:

i) the conduction interval is a very short part of the cycle

ii) the diode operates near its bend for most of the conduction interval

iii) the diode peak current is very large so that carrier storage effects become pronounced.

For such reasons the behaviour is analytically intractable, and also there are few evident possibilities for practical improvement.

If a re-appraisal of the situation is made in the light of the above measurement axioms, then a new family of mismatched diode modulator or demodulators is created of which the Single-ended Mismatched Modulator is the first example.

The basic Single-ended Mismatched Modulator is shown in Fig.A2.1.1.i. The circuit consists of two current sources feeding through a single junction to a diode to a current sink. One current source is that of the carrier the other is that of the signal.

It is well known that modulation cannot be achieved by simple addition of two trains of sinusoidal currents (or voltages). Such an attempt viewed on an oscilloscope would be seen as a serpentine wave of the higher frequency lifted up and down by the lower.† Subsequent application to a band pass filter near carrier frequency would remove the audio frequency component of the sum, and restore an unmodulated carrier wave.

If however the serpentine wave is sliced in half by a diode in series (or shunt), there is formed a train of part sine waves of carrier frequency undulating in magnitude according to the signal wave (provided it is not greater than the carrier wave amplitude). Filtration of this train in a band pass filter centred around carrier frequency will give an approximation to a modulated wave. The chopped nature of the part sinusoids are restored to sinusoids by the Q of the filter.

Practical realisation of the current sources is usually approximated by connecting high impedances in series with input sources of a more or less constant voltage characteristic from which come carrier and signal in commonly found arrangements. Resistors of between 2 and 40

† See Photograph Page A2.1.1.R.1 photo 16

thousand ohms are useful for this requirement, see Fig A2.1.1.ii . Provided the source e.m.f. of the carrier is more than 2 volts peak, the curvilinear region of the diode is of minimal significance because during the wave the only curved portion traversed by the voltage is between the high impedance reverse biased diode resistance and the forwards biased region having less resistance than the current limiting resistor deliberately included above. The action is explained graphically in Figure A2.1.1.iii .

The signal source may be similarly connected. So that "over modulation" may not be caused, the current peak value from the signal current source arrangement must not exceed the peak current value of the carrier; Fig A2.1.1.iv .

Moving to consideration of the current sink, it is clear that it should be a low impedance at all frequencies; i.e. low compared with the diode and its two or more current sources. Ideally the current sink should be resistive in nature and preferably uni-lateral so that the effects of subsequent circuit voltage phenomena cannot react back to the diode. Eminently siutable current sinks may be engineered in several ways such as

i) a shunt fed back common emitter transistor circuit shown in Fig A2.1.1.v .

ii) a common base circuit as shown in Figure A2.1.1.vi

The d.c. conditions may be balanced in various ways. Those shown enable a silicon diode to be biased to the knee by the transistor emitter-base junction potential. If germanium diodes or Schottky barrier silicon diodes are used there is a net forward bias of 0.5 V or o.2 V which will probably be a negligible amount compared with the carrier peak e.m.f. Both sources should have a d.c. path to earth within . If ac coupling is mandatory, it may be possible to arrange some kind of d.c. restoration circuit such as the counterbalance circuit shown in Figure A2.1.1.vii

Charge storage in the PN junction is not excessively troublesome in the Single-ended Mismatched Modulator using sinusoidal carrier waves because the forward current is controlled in magnitude and also because the current is brought smoothly down to zero before reversal occurs. Nevertheless a sluggish diode will spoil the action of the modulator at high frequencies.

A2.1.1.R RESULTS (by photograph)

Photographs page A2.1.1.R.1

The top group are all taken with a carrier frequency of 5 kHz and a signal frequency of 200 Hz synchronised to the carrier so that clear stationary oscillograms are produced. Photographs are referred by number.

16 Signal Wave alone
Carrier and signal waves added

2 Signal wave alone
Carrier and signal waves applied to the Single-ended Mismatched Modulator

3 Similar to photograph 2 but with signal wave shifted to illustrate linearity

1 Signal wave alone.
Wave train of photo 2 or 3 above fed through a band pass filter. It is of interest to note the delay caused by the filter and that the resultant depth of modulation is approximately 91% only (Refer para A2.2 below).

The last pair were chosen to illustrate the apparent linearity of the Single-ended Mismatched Modulator driven by sinusoidal carrier when examined before filtering and the disappointing linearity afterwards.

11 Signal wave alone: 27 Hz triangular wave 8 V pk to pk . Carrier sine 4 V pk plus signal applied to Single-ended Mismatched Modulator having diode of opposite polarity (as between Figures A2.1.1.v and vi for example).

41 Similar to photo 11 but having diode of original polarity. Wave train filtered in band pass filter. Distinct impression of non-linearity at crests and troughs. No low frequency phase error problem.

A2.1.2 Balanced Mismatched Modulator

Many applications of modulators require balance to zero for one or other of the input waves. For instance in the telecommunications multiplex modulating terminals using selective filters to obtain single sideband it is helpful if the carrier is balanced out before application to the filters. In generators of single sideband suppressed carrier by the Phasing Method or the Third (or Weaver's) Method, it is necessary to construct two or four well balanced modulators having a good prospect of remaining balanced in service.

In the light of these well known requirements for good carrier balance, early consideration was given to the development of a Balanced Mismatched Modulator. And the circuit turns out to be very simple, see Figure A2.1.2.i

The source to be balanced out (usually the carrier) must originate from a balanced source. Then cancellation is achieved by connection of the diodes in contrary directions to a common low impedance load. The signal is applied in phase to the diodes so that it may push or pull current at appropriate parts of its cycle, under over-riding control of the larger carrier half cycles. Photograph page A2.1.2.R No 8 shows the phenomenon quite clearly. It is at the extreme condition of signal wave current peak just as large as the carrier current wave peak magnitude. Circuit implementation of the wide band low impedance uni-lateral load can take the form of others found suitable for the Single-ended Mismatched Modulator. There is however some simplification of d.c. biasing considerations due to the natural self balancing form of the counter connected diodes, so a.c. connection to the transistor base is suitable. Figure A2.1.2.ii shows a typical full circuit with typical values. The diagram also shows a technique for fine adjustment of carrier balance, a manoeuvre performed with zero volts at the signal input.

Should strict d.c. coupling be desired, the circuit of Figure A2.1.2.iii may be adopted. Having reliable balance and completely direct coupling the Balanced Mismatched Modulator in this form has the remarkable property of being phase precise right through zero volts input. In other words it may be used as a measurement control in an instrument or in a negative feedback system. The magnitude of the output carrier will follow a gradually descending positive voltage input by a gradually reducing output magnitude until at zero it will continue to follow the input into the negative quadrant by reversal of carrier phase. The Photograph A2.1.2.R No 8 illustrates the effect at sine wave cross-over. Whilst possessing phase precision down to d.c. this modulator still retains high frequency capabilities so that if there should ever be an application for balanced wide band modulation (e.g. balanced television modulation or double sideband television with suppressed carrier*) the circuit shown will perform to design.

Returning to more normal applications, it should be noted that a very interesting behaviour is seen if this modulator is connected to accept balanced audio signals but having larger single phase carrier supplied at the terminal usually allocated to base band. The output wave train at the collector of the amplifier transistor is a remarkably good representation of a modulated carrier wave even though it has not passed through any filter. If the Photograph A2.1.2.R No 6 is examined, the claim can be verified. However it will be noted that some of the carrier wave is missing from areas less than half height. A remedy for such events is explained in chapter A2.2.

* Such as would be necessary to generate phase modulated TV

A2.1.2.R RESULTS

Using an audio tone of 200 Hz synchronised by a carrier of 5 kHz the following waveforms were observed on the Balanced Mismatched Modulator.

Photographs page A2.1.2.R.1

8 Signal wave unbalanced 4 V pk.
Carrier sine wave balanced 2 V pk. Wideband output

9 As photo 8 above but to a slower time base and the output taken through a band pass filter.

By reference to the audio wave at the top it is seen that a double sideband suppressed carrier wave has an outline "beat" appearance containing principally second harmonic (see Part B2.12R page 60).

6 Signal wave balanced, carrier wave balanced sine magnitude 8 V pk. Wideband output.

20 As photo 6 above except that the original signal wave is shown as well. It is brought down close to the modulated train proving that the output is a fairly good reproduction and that there is no delay (as there is no filter). Wideband output.

21 As photo 20 above but reduced carrier wave amplitude to show that amplitude clipping occurs both top and bottom. Wideband output.

17 As photo 8 above with lower trace having the carrier magnitude increased until it squares off, showing the improvement which could come with deliberate square wave carrier input.

Photographs Page A2.1.2.R.2

These show a sequence with the exacting triangular wave modulated upon a 5 kHz sinusoidal carrier wave in which the effect of the magnitude of the carrier wave is gradually reduced to zero.

15 Carrier wave overlarge and being clipped by the modulator to the peak magnitude of the signal

14 Carrier wave only just sufficient

13 Carrier wave distinctly insufficient and cross-over distortion is evident

12 Zero carrier wave ant the signal is cut into two linear parts by the o.4 V forward bias needed by the Schottky diodes.

A2.1.3 Doubly Balanced Mismatched Modulator

Encouraged by the achievement of single balance, thought was given to the establishment of double balance in the mismatched modulator form. This would then be analogous to the RING MODULATOR. Obviously both sources must be balanced to earth, and must feed two normal balanced modulators whose output current pulses are subtracted in a current difference unit of low impedance over a wide band.

A first attempt at such a circuit looked as Figure A2.1.3.i , which when redrawn appears as Figure A2.1.3.ii having a great similarity in appearance to the Ring Modulator see ((31) page 87).

The wide band low impedance difference device may be realised as either

a) the emitter coupled shunt fedback pair shown in Figure A2.1.3.iii

or b) a centre-tapped primary transformer enjoying good balance to carrier frequency band currents, as shown in Figure A2.1.3.iv . Coupling to d.c. will still be achieved since the currents are here carrier waves chopped.

A2.1.3.R RESULTS

The results of this section were measured on a Doubly Balanced Mismatched Modulator comprising the circuit of Figure A2.1.3.ii connected to a circuit of Figure A2.1.3.iii with all resistors designated R chosen to be 33 kΩ .

The effects of the variation of carrier and signal magnitude are shown by the first group of photographs page A2.1.3.R.1 . The second group show in a graphic manner, the excellent low frequency capabilities of this connection of mismatched modulator. The signal and carrier frequencies are brought to almost the same value. Multiple exposure photographs indicate the events at different parts of the cycle.

Photographs page A2.1.3.R.1

All have carrier wave sinusoidal at 5 kHz and the signal frequency is 200 Hz .

43 Signal 11 V pk to pk
Modulation with carrier 30 V pk to pk

44 Signal 3.6 V pk to pk
Signal 14 V pk to pk } Carrier 15 V pk to pk

45 The same as photo 44 above but with expanded time scale

47 As upper trace of 44, after band pass filter.

49 As photo 45 above but after band pass filter.

It should be noted in these photographs how very nearly square wave the carrier component becomes when the signal voltage is small.

Photographs page A2.1.3.R.2

46 Signal 27 Hz unbalanced to earth, applied to one input terminal and earth. Balancing capability of the test circuit is good enough to produce an output apparently of half the magnitude and balanced.

56 Carrier frequency 500 kHz
Signal frequency 20 kHz. Shown at two different time base speeds.

57 As photo 56 with further increased time base speed.

50 Carrier frequency 5 kHz
Carrier frequency 1.5 kHz Signal frequency 200 Hz

51
52 Interaction of 300 Hz and 200 Hz at various times by multiple exposure. "Carrier" 10 times magnitude of "signal".
54

55 Interaction of two sine waves of same frequency 200 Hz. "Carrier" 10 times "signal", slightly slipping phase.

A2.1.4 Conclusions and Comments

At the end of the first experimental section of this thesis comments can already be made concerning the achievements of the simple types of mismatched modulators.

Considering that previously used forms of modulator have been found to suffer from one or more of the following:-

i) uncertainty of precise non-linearity of diode or other active device

ii) difficulties of source addition due to their usually being both unbalanced to earth and single phase

iii) low frequency end of base band being restricted because of the need for transformers to add e.m.f.'s in ii).

iv) uncertainty in regard to balance (and maintenance of balance) both in manufacture and ambience for those required to give carrier cancellation.

.............it may be said that the <u>Mismatched Modulator circuit provides a fair degree of advantage.</u> For instance i) is overcome by the series resistances used to simulate current sources,

ii) is of no difficulty whatsoever as both sources may have one terminal earthed and are actually well isolated from co-influence by virtue of the large resistor in series with each.

iii) base band may extend to d.c. as there are no transformers or capacitors necessary.

iv) balance is retained over considerable ranges of diode characteristic thanks to the overwhelming influence of the high impedance feed and low impedance load.

To sum up, the family of Mismatched Modulators are distinguished in regard to good designability and useful circuit configuration.

A2.2 AM Linearity by Square Wave Carrier

Consideration must now be given to the inherent distortion which arises in all types of Non-linear Modulators including Mismatched modulators but also including the long used Class C modulators.

Referring to the basic waveform diagram in Figure A2.1.1.iv , it will be observed that near the troughs of modulation the remaining fraction of the sinusoid chopped by the non-linearity is not only small in amplitude, it is also small in time-flow. Consequently the energy impulses supplied to the next tuned circuit (the band pass filter) are unnecessarily weak. Thus the troughs sink too fast towards zero.

On the same diagram, it will be seen that near the crests of the modulation, the part sinusoids are both large in amplitude and also large in time-flow. Consequently much of the energy arrives in the wrong phase and actually hinders the fundamental carrier wave see Figure A2.2.i

If a Fourier analysis is made for various sample fractions of chopped sinusoid (Appendix A2.2) the graph shows there will be significant distortion see Figure A2.2.ii . Surprisingly the slope of the centre region is 25% greater than the expected average linear input-output slope. This may cause unexpected undermodulation for if a signal say x volts causes 50% modulation, then a signal of 2x volts will produce only 90% modulation.

Or to express the curvature in a more alarming way, modulation which is almost 100% will contain 20% third harmonic distortion.

These facts explain a) the mysterious undermodulation experienced in most Class C modulators, which has in the past been overcome by modulating a carrier drive stage as well.

b) the necessity for pre-distortion of the video wave in television transmitters.

Having arrived at these values it will be unwise to use sinusoidal carrier waveforms in modulators required to give either deep modulation or very low intermodulation products. Third order components have been long known to be the cause of intermodulation (11).

The obvious way to cure these troubles is to use a carrier waveform which may be correctly amplitude modulated without any accompanying time-flow modulation. Now if square wave carrier is used, there will be (Fig A2.2.iii) a constant time-flow of half cycles and the Fourier component of the chopped waves will be proportional to the amplitude alone. Thus there will be a truly linear modulation characteristic. For these reasons the remainder of the experimental work in this part concerns Mismatched Modulators with square wave carrier input.

Square wave carrier currents have been used before (12). It cannot be claimed that their application is original. However it is possible that the previous practice has been accidental or unintentional. Certainly it came to the author's attention accidentally in 1954 that a much overdriven Class C anode modulated stage gives a highly linear modulation characteristic, and does not suffer from undermodulation. Conceivably such behaviour may be due to the drive waveform becoming squared by a limiting action (for instance excessive grid voltage swallowing up anode current).

Undoubtedly the application of square waves in the White (or was it Blumlein?) oscillator redescribed in (12) by Foss & Sizmur, gives it the ancilliary property of being modulable. Figure A2.2.iv shows an adaption for a student laboratory experiment designed by the author which is so linear that demonstrators have reported that students do not appreciate previous difficulties. A second equipment has been constructed using transistor base characteristic curvature to illustrate the point(1). When the former circuit was used to illustrate a public lecture (19) a considerable impression was made upon a critical audience in regard to its good linearity of modulation.

Furthermore, a successful low power (milliwatt) UHF television modulator has been designed using the concept of a controlled current supplied to an emitter coupled pair of transistors driven by a square wave on one base. The circuit of the complete oscillator-multiplier-modulator chain is shown in Figure A2.2.v . This equipment is in daily use by Aberdeen University Television Service for modulating PAL coded Colour TV signals up to a carrier frequency of 663 MHz. The photographs page A2.2.R shows the quality achieved using the exacting new test card format, and demodulating on a Decca 26" Monitor receiver, with a Grampian TV Test Card as a receiver & photograph comparison check of quality.

A2.2.1 Single-ended Mismatched Modulator with Square Wave Carrier

The circuit components of this are exactly the same as previously described for the sine wave version(A2.1.1).

The description of action is also similar to that described before. However there is now more effect noticable due to capacitive and semiconductor charge storage. The diode must switch very rapidly from forwards to reverse condition of bias, which is the condition most difficult to fulfil. Even so it should be borne in mind that the mismatched modulators of any form have smaller forwards current than many other forms of diode modulator , so charge storage effects may not be too bad. As a precaution the experiments here employed a Schottky diode (a type notable for low values of charge storage).

A2.2.1.R RESULTS

Photographs Page A2.2.1.R.1

4 Carrier wave rectangular 5.6 V pk 5 kHz
Signal sine wave 1.9 V pk 200 Hz

40 Carrier wave square 5 kHz
Signal wave 27 Hz triangular.
Top trace is wideband before filter.
Lower trace is after band pass filter.
It is noticable that there is not so much distortion as there was in the sine wave carrier modulation of this test (see pare A2.1.1.R photo 41)

78 Carrier square wave of 900 kHz
Signal sine of 42 kHz
There is a small reverse current spike near the trough of modulation. The small amount of energy thus indicated represents a very slight distortion.

Photographs Page A2.2.1.R.2

These are Spectrum - Waveform pairs. The top two are spectra with the spectrum analyser horizontal scale set to 4.5 mm per MHz. Thus they show the harmonics of the carrier wave. The first tall reading being the zero frequency mark of the instrument, the others are spaced by intervals of 900 kHz . Signal frequency is 400 Hz .

The two lower spectra are centred at 900 kHz and have a horizontal scaling of 4.5 mm per 50 kHz . Here the signal frequency is 42 kHz (as photo's 78 & 80) and so it is possible to see the carrier and its accompanying sideband set.

The vertical scale of the spectra is logarithmic, being 3.5 mm per 10 dB .

It is possible to see that the band pass filter used in these experiments has not enough bandwidth to pass the 42 kHz sidebands without attenuation.

The waveform oscillograms show on the upper trace the input signal wave, and on the lower trace the waveform at the wideband output terminal of the Single-ended Mismatched Moudulator. They are simultaneous exposures.

Certain dark patches over parts of the spectra are caused by interference between the sweep velocity and the shutter time.

Group of Intermodulation Tests.

Racal Communications Receiver RA 17L used as a refer back voltmeter. Laboratory attenuators used to measure magnitude of the components as number of decibels down.

Composite audio signal comprising added tones of constant amplitude 1 V pkto pk at 4 kHz , with variable amplitude (as marked on table) signal 1.5 kHz. Carrier 900 kHz square wave.

INTERMODULATION

Freq. kHz		900	901.5	902.5	903	904	905.5	906.5	907	908
Input V pk to pk	1.0	-13	0	30	32	0	30	44	44	40
	0.5	-13	6	36	40	0	36	60	56	40
	0.25	-13	12	42	52	0	42	60	60	40

Accompanied by similar set the other side of 900 kHz.

A2.2.2 Balanced Mismatched Modulator with Square Wave Carrier

The Balanced Mismatched Modulator with Square Wave Carrier may have a circuit exactly the same as any one suggested for the sine wave carrier forms. Of course there must be a balanced square wave source of carrier frequency. Such a device is available in the form of an integrated circuit flip-flop driven by a sinusoidal source of double the required frequency. In the experimental work in which this circuit was tested in the laboratory, an oscillator at 1.8 MHz was used to drive a dual D type edge triggered flip-flop type SN 7474 . Output waves were very fast 5 V pk to pk squares , which were a.c. coupled to the high impedance input terminals of the modulator under test.

Since some reverse current spikes were seen in the case of the single-ended mismatched modulator test under square wave conditions , then a capacitive balance arrangement was used . This consisted of the extra capacitor of 4.7 pF from one diode to earth and a variable of 2 - 8 pF from the opposite diode to earth, see Figure A2.2.2.i . Taking fullest precautions against stored charge , Schottky diodes were used. The quoted minority carrier lifetime for the Hewlett-Packard 5082-2800 being typically 100 picosecond. There are more comments concerning diodes and their departure from ideal behaviour in Part B (para B2.1.1).

A2.2.2.R RESULTS

Experiments were performed in two groups. The first group consisted of straightforward waveform and spectrum measurements using photographic records as given for the modulators investigated earlier.

The second group consisted of intermodulation measurements. Here the objective was to test the linearity of modulation in the most perceptive manner. The results are measured by spectrum analysis using either a Hewlett-Packard spectrum analyser type 8558B , or a communications receiver previously carefully checked for gain flatness over the band.

First Group

Photographs page A2.2.2.R.1

Waveform measurements at carrier frequency 5 kHz square.

10 Carrier 5 V pk to pk square balanced.
Signal sine wave 4 V pk 200 Hz . Wideband .

The outline of the wave train follows the shape of the sinusoid even crossing over the zero line with part of the next carrier half square positive and negative.

18 Carrier wave 5 V pk to pk Unbalanced
Signal sine wave 5 V pk to pk Balanced. Wideband.
Signal trace brought down to show time coincidence and fidelity.

The product is a symmetrical normal amplitude modulated wave undelayed because it has passed through no filter. It contains carrier harmonics only. It would be suitable for many applications of amplitude modulated waves.

19 As photo 18 above but with the sine wave peak exceeding the carrier half peak magnitude, to show effect of distortion resulting, and its balance.

Photographs page A2. 2. 2. R. 2

Spectrum measurements at 900 kHz with square wave carrier.

87 Spectrum of sideband components around 900 kHz with signal a sine of 40 kHz. Wideband.
Some indication both on the waveform and the spectrum that the sine wave is too large and is causing third harmonic distortion.

88 As for photo 87 but with reduced signal magnitude. Filtered.

Second Group

INTERMODULATION at 900 kHz

Photographs page A2. 2. 2. R. 3

90 Shows spectrum of sideband components around 900 kHz with carrier modulated by sines of 40 kHz and 65 kHz. Wideband.
Vertical scale 3.5 mm per 10 dB
Carrier balance about -50 dB below Wanted sidebands
Highest intermodulation product about -37 dB

89 As photo 90 above but filtered in band pass filter.

91 As photo 89 but frequencies of modulations 40 kHz & 26 kHz.
Carrier balance seems to have degenerated.
Intermodulation products similar to previous pair.

Composite audio signal comprising added tones of constant amplitude 1 V pk to pk at 4 kHz, with variable amplitude (as marked on table) at 1.5 kHz.

Racal RA 17 L Communications receiver used as refer back voltmeter with laboratory attenuators for magnitudes of dB down.

							Intermodulation		
Freq. kHz	900	901.5	902.5	903	904	905.5	906.5	907	908
Input V pk to pk 1.0	40	0	46	40	0	56	30	30	52
0.5	40	6	52	46	0	56	41	40	52
0.25	40	12	58	54	0	60	46	46	52

With similar set other side of carrier freq.

A2.2.3 Doubly Balanced Mismatched Modulator with Square Wave Carrier

The Doubly Balanced Mismatched Modulator which was used for experiments to verify the expected good linearity and low intermodulation figures was the same circuit as Figure A2.1.3.ii connected to the emitter coupled type of low impedance difference device shown in Figure A2.1.3.iii . The high valued resistors in the Balanced Mismatched Modulator and in the feedback circuits of the transistors were either 33 k ohm or 12 k ohm . There was very little difference in the figures or waveforms either way.

The carrier frequency was 1.6 MHz, and the squarer was an emitter coupled pair switched by a sinusoidal oscillator input. The square wave was 9 V pk to pk

A2.2.3.R RESULTS

Photograph page A2.2.3.R INTERMODULATION TESTS

73 Shows waveform with signals both sinusoidal at 2.2 and 2.9 kHz and equal amplitude.

75 Shows the above two balanced modulations one at a time
Top trace 2.2 kHz
Middle trace 2.9 kHz
Lower trace shows zero audio & is a check on carrier balance adjustment.

76 Shows one of the above single audio balanced, at the time of audio cross over. Expanded time scale.

60 Shows another pair of frequencies of equal amplitude applied to the doubly balanced mismatched modulator
The frequencies are 3 kHz and 2.6 kHz.
Lower trace is a balance check test, showing some error.

77 Shows another pair of sines of equal amplitude.
Frequencies 2.66 & 2.9 kHz.

Composite audio signal comprising added tones of:-

Constant amplitude 2 V pk to pk at 2 kHz.

Variable amplitude as given in table at 3 kHz.

Carrier input square wave at 1.6 MHz

Balance relied on electric shielding and resistor accuracy.

Racal RA 17L Communications receiver used as refer back voltmeter with laboratory attenuators for magnitudes of decibels down.

INTERMODULATION

Freq kHz	1600	001	002	003	004	005	006	007	008	009	010
Input V pk to pk 4	28	28	0	-6	22	36	36	35	23	25	36
2	27	28	0	0	22	45	44	50	55	29	n
1	28	n	0	6	23	56	57	55	56	n	n

Corresponding set the other side of carrier freq.

n indicates no measurement recorded.

Section A3 Developments and Further Work on Mismatched Modulators.

Opportunity was taken during the work to propose a design for the fabrication of a Balanced Mismatched Modulator in the form of an Integrated Circuit.

Having consulted the member of staff who kindly offered to assist in the drawing of masks and diffusion, a design having a quiescent current of 6 mA in the transistor and 12 k ohm resistors for diode feeds and feedback was designed. The circuit is shown in Figure A3.i and a photograph of the resultant slice on page A3.R and below an unmounted slice (failed tests).

The results obtained indicated that the successful samples produced had similar characteristics to those of the discrete component versions more thoroughly investigated. Perhaps a second attempt at the design would avoid some of the mistakes made through no fault of the fabricators, but the diode feed resistors have been packed so close together in the present arrangement that there is an undue amount of unexpected carrier unbalance.

Presumably had there been time and finance available an improved version could have been made.

During the last few months of the project it came to the notice of the author that Gingell (16) had devised a novel new approach to the phasing method of generating SSB in which two similar balanced modulators are of course required. Interestingly enough the inventor himself remarks ((17) page 25) "one of the most important practical problems that has to be overcome is the design of satisfactory modulators".

Out of curiosity a quick design of a polyphase network covering 340Hz to 3.4 kHz was made and tested for the generation of SSB with two of the balanced mismatched modulators of this thesis. There was an immediate indication of the possibilities in the system for a 30 dB wanted to unwanted sideband suppression was measured.

Another problem which is said to be in need of a solution is the cross-modulation of co-sited transmitter signals on to a receiver tuned to a different frequency. With the mismatched modulator as a receiver front end mixer it should be interesting to see if a third input (that is beside local oscillator and r.f. input) can be added in parallel in such a polarity and magnitude to cancel the unwanted cross modulation with some signal picked up on a wideband probe in the transmitter field.

It is confidently anticipated that the mismatched modulators will provide solutions to many circuit situations in which previous modulators have been difficult to trust. For instance it should be possible to include a modulator within a feedback loop. For instance companders could be constructed with characteristics tailored by feedback.

Section A4 Conclusions on Mismatched Modulator Circuits

The mismatched modulator has been thought out from frist principles and fully developed and tested in the course of this Part of the work.

The difficulties and disadvantages of previous forms of modulator circuit have been examined and described. It has been suggested that by comparison the mismatched modulator allows the circuit designer more freedom and at the same time gives him a more reliable design capability.

The non-linearity of the device is controlled and precise because it depends principally on the passive resistors forming the approximated constant current sources.

The mean voltage of the input waves is maintained at the diode bend as there is no charge store in circuit. In fact some balanced versions are so self balancing, that they do not need bias trimming at all.

The signal circuit is directly connected and thus may be used for signals whose potential varies slowly or not all. The modulator may be used as a controlled attenuator.

The sources may be all directly connected to earth. No transformers are necessary and so there need be no problem of phase towards zero frequency.

The sources have minimal mutual effect upon eachother.

This is particularly true in the balanced forms where it is a fact of circuitry that there is always a diode ON and so there is always a low impedance at the centre of the resistor-diode-resistor chain which exists between the si nal source and its carrier source partner. It is for this reason that the outlines of the output wave bears the perfect outline of the constituent signal and carrier waves.

For the above reason a further advantage is gained. The signal and carrier frequencies do not have to be dissimilar. They may even be the same frequency.

The modulation sensitivity of the single-ended square wave driven modulator is precisely calculable:-

$$\text{Depth of Modulation Fraction} \quad m = \frac{i_s}{i_c}$$

Where i_s = peak signal current

i_c = square wave carrier current

Or if approximated current sources have been used:-

$$m = \frac{v_s}{v_c} \cdot \frac{R_c}{R_s}$$

Where v_s = peak signal voltage

v_c = square wave carrier voltage

R_s & R_c are series resistors

Now if, as is usual $R_s = R_c$

Then $m = \dfrac{v_s}{v_c}$

The output voltage is calculated by the following:-

For the shunt fedback amplifier $V_o = R_f \, i_{in}$

But $i_{in} = m.i_c$

$$= \frac{v_s \cancel{V_c}}{\cancel{V_c} R_c}$$

$$= \frac{v_s}{R_c}$$

So $V_o = v_s \frac{R_f}{R_c}$ But often it is arranged that $R_f = R_c$

So $V_o = v_s$

In other words the mismatched modulator is a unity gain modulator (or mixer) provided all the resistors mentioned are the same, and the transistor has high gain at the frequency.

Balanced forms of the mismatched modulator are not dependent upon diode tolerance as such variations are swamped by the current source impedances. So in production and in service adequate balance may be achieved with minimum fuss.

The distortion produced by the residual non-linearity is less than comparable circuits of a less precise nature. Provided square wave carrier drive is employed, it has been indicated that the modulation linearity is as good as can be achieved in an electronic amplifier. The problem of intermodulation will always be present, for the following reason.

If a sinusoidal wave is in process of being linearly modulated in a square wave driven mismatched modulator, and if a third wave (probably sinusoidal) is also added, the third wave will behave as if it had been applied to a sine wave driven modulator (the larger sine acting as "carrier" for the smaller).

From the circuit waveform criterion, a linear modulator subjected to two sines is the same as a piecewise non-linear diode subjected to two sines. Thus it may be assumed that the curve of Figure A2.2.ii obtained from the Fourier analysis, will in these circumstances apply to the square wave driven modulator. So the curvature of the Fourier function will relate the two sinusoids, and the modulation on the larger will be to a small extent transferred to the smaller.

Part B

A M P L I T U D E   D E M O D U L A T I O N

Part B AMPLITUDE DEMODULATION

Section B1 Previous Techniques and Theories

B1.1 Previous Techniques

B1.1.1 Power Conversion Devices

The earliest radio frequency detectors were devices which experienced the power in the radio frequency current and converted the power into heat (to unbalance a bridge for instance) or into light (Hertz & Marconi sought a spark) or into ionic deposition (as in certain pen recorders).

With the arrival of modern semi-conductor techniques such high frequency detection devices have received a new lease of life. They are more sensitive than before , and can be incorporated in negative feedback loops so that their non-linear characteristic behaviour may be straightened out. In the form of the light emitting diode the possibilities are only just being investigated. Perhaps the LED should be categorised in the next section, for whether it is operating by rectified r.f. current or by the conversion of unidirectional pulses of radio frequency power into light, is a debatable question. Unfortunately whatever its novelty and applicability the usefulness of the LED will for ever be restricted by the variability of the sensitivity of the human eye.

Notwithstanding the increased utilisation of power conversion devices for instruments, they are all slow in response. As a result though in principle these detectors would perform the correct function electrically, they are of no use in communications as demodulators

Bl.1.2 Rectification Devices

Communications carrier frequency demodulation is almost exclusively based on the alternating voltage peak rectifier. The carrier frequency voltages bearing the signal modulated upon them as amplitude variations are peak rectified by a diode so as to charge a low pass filter or integrator circuit.

The basic circuit is shown in Figure Bl.1.2.i . For so long its supremacy has remained almost unchallenged. The only variations which have appeared were valve or transistor circuits which imitated the voltage peak rectification idea using the grid or base electrode for the purpose.

It was the tremendous agility of the electron current in changing from forwards conduction to reverse insulation which led

Fleming to suggest the thermionic diode to Marconi for r.f. rectification. In order to produce the greatest possible output voltage in the base band circuit on the right of the diagram, the load resistor is usually chosen to be as great as is found to be useful. In order to by-pass the r.f. current the capacitor is chosen to be as large as possible in view of the requirement that the potential across it must follow the signal changes as rapidly as necessary.

It will be observed that in these circuits the diode is working from a low impedance source into a high impedance load. Presumably the endeavour has always been towards efficiency. Perhaps this is partly because the peak rectifier circuit has been worked-up in parallel with a.c. to d.c. conversion activities in the world of power engineering. Or perhaps there has been the unsaid feeling of fidelity following from efficiency.

Unhappily, under the arrangements developed:-

a) the diode operates near its bend for most of the conduction interval

b) the conduction interval is a very short part of the cycle

c) the diode peak current is large, and so charge storage phenomena become important.

Bl.2 Previous Theories

Most of the published attempts to give a theoretical basis for the understanding of the rectification type devices for the demodulation of amplitude modulated carriers have been either highly descriptive or have been based on the piecewise linear approach to the diode characteristic.

Perhaps the most thorough has been that given pages 427 to 433 of (11). It is noteworthy to find a value calculated for the loading presented to the source by the diode demodulator circuit. Not many texts manage to give their students this important information, since it is needed to calculate the actual Q of the final filter circuit in the r.f. amplifier. It will be shown that it is not necessary to use such a long derivation to deduce a rough loading formula which applies for the generally used situation of load resistance as high as possible. Of course Everitt pre-dates the semi-conductor era, so there is no help concerning reverse current in PN junctions due to minority carrier life time.

Tragically most texts do not warn the designer against using too large a by-pass capacitor. Even a recent text of considerable merit (8) has no mention of the problem of diagonal clipping see Figure Bl.2.i . The investigator who wishes to establish the best possible arrangements for linearity must do his best to avoid the hazard. Perhaps too much reliance has been placed in the assumption that amplitude modulation from broadcast stations will have an average depth of only 30%, so that diagonal clipping will not often be noticed.

Other situations may be considered to be "communications quality" systems, and the signals carried are there said to be so distorted that a little more will not matter! It could now (1974) be argued detection distortion will not be of the form of diagonal clipping since the suppressed carrier systems now built require coherent detectors. The large re-inserted carrier signal at the coherent detector fixes the resultant depth of modulation at a small value, so diagonal clipping is no longer important. However there is another distortion which is to be investigated in paragraph B2.1.2 caused by the single sideband phasor implanting a quadrature component.

Credit must be given for Everitt in the thorough description of the implications of single sideband modulation (page 408 Op cit.). He shows that for distortionless demodulation SSB requires a second order non-linear device whereas normal AM is distorted by second order and therefore requires first order non-linearity, or as will be seen later, odd order linearity.

B1.3 Conclusions on Previous Techniques and Theories.

The principal conclusion which is forced upon the investigator who looks into the literature on Demodulation is one of disappointment. There is so much uncritical usage of the circuit which has always been employed, and so little fresh theoretical investigation. It is hoped to supply some contributions in this work.

Section B2 Linearity of Amplitude Demodulation

B2.1.1 Loading Effect of Peak Rectifying Demodulators

When the peak rectifying demodulator is in use across the final filter of the r.f. amplifier of a radio receiver, it is of interest and importance to know the magnitude of the circuit loading it causes. The analysis in (11) uses a piecewise linear approximation to the diode characteristic and that procedure will be followed here. It will not be necessary to derive a transcendental equation which Everitt evaluates. It can be assumed that in the role of efficient rectifier, the demodulator is passing very short current pulses and that the capacitor is sufficiently large to allow little droop on the output between half cycle charge impulses, see Figure B2.1.1.i .

Assuming t the time during which diode current flows is a very small fraction of a cycle, it may be said that (τ) length of time represented by DP is approximately the cycle time $\tau + \delta t$

But by Similar Triangles $\frac{FO}{OV} = \frac{PD}{DV}$

Or $\frac{\text{Time Constant of Load Resistor-Capacitor}}{\text{Rectified Voltage V}}$

$= \frac{\text{Cycle Time}(\tau + \delta t)}{\text{Ripple Voltage v}}$

Or $v \simeq V \frac{(\tau + \delta t)}{CR}$ A formula of interest to the designer of demodulators for knowing how small may be C .

But occasionally load current I_L is known instead of load resistance R_L

But $R_L = \frac{V}{I_L}$

So the last formula for ripple voltage becomes:-

$$\underline{\underline{v = \frac{I_L}{C}}}$$ which is sometimes of importance.

Further it follows that as the ripple voltage is small,

$X_C \ll R_L$ thus the charging circuit is effectively as shown in Figure B2.1.1.ii

So the charging current i_c is given by

$$i_c = \frac{\hat{v}}{r_d}$$

But from conservation of energy,

Charge input per cycle = Charge drained into load per cycle

$$i_c\, t = I_L(\tau + t)$$

$$\text{So} \quad i_c = I_L\frac{(\tau + t)}{t}$$

$$i_c = I_L\frac{(\tau + t)}{t}$$

But $\hat{v} \simeq V$ So $i_c \simeq \frac{V}{r_d}$

And $I_L = \frac{V}{R_L}$ So the above equation may be written

$$\frac{\cancel{V}}{r_d} = \frac{\cancel{V}(\tau + t)}{R_L\, t}$$

Or $$\frac{R_L}{r_d} = \frac{(\tau + t)}{t}$$

But $$\frac{i_c}{I_L} = \frac{(\tau + t)}{t}$$

So $$\frac{i_c}{I_L} = \frac{R_L}{r_d}$$

Now to work out the effective loading of a demodulator upon its source it is possible to very quickly reach Everitt's value by the following procedure:-

Assume the peak rectifier detector is an efficient device,

Then Power in = Power out

If the effective load the device causes to its source is R_{eff} and if the source voltage is v_{rms}

Then $$\frac{v_{rms}^2}{R_{eff}} = \frac{V^2}{R_L}$$

But V is approximately the peak voltage $\sqrt{2} \times v_{rms}$

So $$\frac{v_{rms}^2}{R_{eff}} = \frac{2v_{rms}^2}{R_L}$$

So $$R_{eff} = \frac{R_L}{2}$$

Now experiments were made on a Q meter to check whether this formula for the loading of a diode detector was born out in practice. It was found that in the case of present day detectors comprising semi-conductor diodes of reasonable or good quality (i. e. fast diodes recommended for h. f. detectors), there was usually too little loading (R_{eff} higher than Everitt theory) and that the loading went up as frequency was increased.

In order to explain these facts, two assumptions are made:-

a) that an adjustment should be made for the diode forward voltage

b) that there will be a frequency dependent loss term due to reverse current caused by charge storage within the junction .

To describe these theories in detail

a) <u>Forwards Voltage Adjustment</u>

Whereas the above theory deduces that the conductance loading caused by the diode is

$$G_{eff} = \frac{2}{R_L} \qquad \text{Siemens}$$

It is proposed to reduce this value by a factor $1/1 + \frac{V_f}{V_{dc}}$

$$\text{So} \quad G_{eff} = \frac{2}{R_L\,(1 + V_f/V_{dc})}$$

Where V_{dc} = DC Voltage across load.

Now when this was done, the agreement between the theory and measurement was quite good at frequencies less than 7 MHz . The measurement procedure and values are given in Appendix B2.1.1.

b) Charge Storage Loss Term

From semi-conductor theory, it is well established that charge stored is proportional to the peak forward current.

Suppose the charge stored is q coulomb

Then from the above argument

$$q \propto i_c$$

But $\quad i_c \propto I_L$

So $\quad q \propto I_L$

Say $\quad q = K\, I_L \qquad$ Where K is a constant (having the dimension TIME)

Now the extra power drawn from the source due to the charge stored may be estimated:-

$$\text{Extra Power} = V_{dc}\, q\, f \qquad \text{Where } f = \text{frequency}$$

$$= V_{dc}\, f\, K\, I_L$$

But $\quad \text{Power} = \dfrac{v_{rms}^2}{R_{eff2}} \qquad$ Where R_{eff2} = new loading term

Or $\quad \text{Power} = v_{rms}^2\, G_{eff2}$

So $\quad V_{dc} f K I_1 = v_{rms}^2\, G_{eff2} \qquad$ But $V_{dc} \doteq \sqrt{2} \times v_{rms}$

Hence $\quad \underline{G_{eff2} = 2fK/R_L}$

The new effective conductance due to charge storage may be added to the effective conductance due to the piecewise or pulse operation of the diode converting alternating current energy into the direct current energy in the load.

Thus Total $G_{eff} = G_{eff1} + G_{eff2}$

$$= \frac{2}{R_L(1 + v_b/V_{dc})} + \frac{2fK}{R_L}$$

$$\underline{\underline{G_{eff} = \frac{2}{R_L}\left(\frac{1}{1 + v_b/V_{dc}} + 2Kf\right)}}$$

Experimental check of this new combined theory was made at frequencies from 29 MHz to less than 8 MHz and the results for several types of diode are plotted in the second part of Appendix B2.1.1

B2.1.2 Single-ended Mismatched Demodulator

The Single-ended Mismatched Modulator circuit can without modification be used as an Odd Order or linear detector. The circuit is shown in Figure B2.1.2.i . The forwards bias from the transistor is used to forward bias the Si diode to the knee and thus linear demodulation right down to the trough of 100% modulated carrier waves is assured. The diode provides a stream of half sine pulses which flow out of the transistor shunt feedback resistor into the transistor load across which the collector voltage changes. The magnitudes of the half sine output voltage changes are exactly the same as the input waves' peak voltages provided R_f is the same as R_S .

When averaged in a low pass filter, the stream of half sines forms aa accurate reproduction of the original modulation. Distortion due to diagonal clipping cannot occur because there is no reaction back to the diode from the averaged voltage across the filter. There cannot be any reaction because the transistor acts as a unilateral device and isolates load and source.

There is no distortion of any kind similar to the chopped sinusoid distortion of the mismatched modulator. All the waves coming from a modulated tuned r.f. amplifier will be of the true half sine duration and they will have a truly constant average value strictly proportional to the peak magnitude of each wave. Thus the average wave existing at the output will be as far as the demodulator is concerned absolutely distortionless at all depths of modulation for all frequencies in which the carrier frequency is sufficiently higher than the modulation frequency to allow reasonable separation of one from the other by filter.

Another notable feature of the Single-ended Mismatched Demodulator is the constant loading of resistance equal to $\frac{1}{2}R_S$ which it throws upon its source. As current flows for the whole of half a cycle, it is the equivalent of $\frac{1}{2}$ the resistance placed across the source for the whole cycle. The peak rectifying demodulator previously used throws upon its source an average resistance loading of $\frac{1}{2}R_L$ (see paragraph B2.1.1) but the loading varies during the cycle of modulation, being more than average during upward slopes and less than average on delines of modulation. Such variations themselves can cause distortion of the incoming wave unless some special precautions are taken.

Turning to the problem of coherent detection, it is found that the Single-ended Mismatched Demodulator is an attractive proposition. The re-insertion oscillator may be switched or off without any disturbance to the actual r.f. circuitry see Figure B2.1.2.iii It may be switched in its power supply only. From the arguments of the last Part of this thesis it may be deduced that for this purpose a square wave carrier will be advantageous from the point of view of lower distortion than sine wave. But the difference will probably be slight when one considers that the distortion that is prevented by the use of square waves is chiefly for waves deeply modulated. The magnitude of the SSB phasor must be considerably smaller than that of the re-inserted oscillator or there will be second harmonic distortion due to the phase modulation implicit in the demodulation of this nature, see below.

To explain the action of SSB phasors as they become demodulated on a single-ended detector of any type, a diagram such as Figure B2.1.2.ii may be used. The SSB phasor and the oscillator add vectorially and the resultant continually shifts in magnitude and phase. The diode peak rectifies the resultant phasor and the low pass filter smoothes the output to an audio component alone.

Designating the SSB phasor as A , and the re-inserted carrier phasor as C , the resultant as R , the analysis proceeds thus:- Audio ρ rad/s Carrier ω rad/s

Resultant Magnitude

$$R = \sqrt{(C + A\cos\theta)^2 + (A\sin\theta)^2} \qquad \text{Where } \theta = \int \rho . dt$$

$$= \left(C^2 + 2AC\cos\theta + A^2\cos^2\theta + A^2\sin^2\theta\right)^{\frac{1}{2}}$$

$$= \left(C^2 + 2AC\cos\theta + A^2\right)^{\frac{1}{2}}$$

$$= \left[C^2 + A^2\right]^{\frac{1}{2}}\left(1 + \left\{\frac{2AC}{C^2 + A^2}\right\}\cos\theta\right)^{\frac{1}{2}}$$

By Binomial

$$= \left[\quad\right]^{\frac{1}{2}}\left(1 + \tfrac{1}{2}\{\;\}\cos\theta + \frac{\frac{1}{2}(-\frac{1}{2})}{2!}\{\;\}^2\cos^2\theta + \cdots\cdots\right)$$

$$= \left[\quad\right]^{\frac{1}{2}}\left(1 + \tfrac{1}{2}\{\}\cos\theta - \tfrac{1}{8}\{\;\}^2\tfrac{1}{2}(1 + \cos 2\theta) + \cdots\cdots\right)$$

Where the wanted audio frequency component is represented by the term in $\cos\theta$ with the unwanted second harmonic being indicated by the $\cos 2\theta$ term.

So Fractional distortion may be calculated:-

$$\frac{\text{2nd. Harmonic Voltage}}{\text{1st. Harmonic Voltage}} = \frac{\frac{1}{8}\frac{1}{2}\{\ \}^2}{\frac{1}{2}\{\ \}}$$

$$= \frac{1}{8}\{\ \} = \frac{1}{8}\frac{2AC}{(C^2+A^2)}$$

Which for less than 1% harmonic distortion for example requires there to be $C \geqslant 25A$, or in other words the peak re-insertion carrier voltage is to be 25 times the normal SSB phasor magnitude!

It is the presence of the implanted phase modulation which is the cause of the distortion term. If the wave had been normal AM it would have consisted of three phasors,

Carrier	$C \cos \omega t$
USB	$\frac{1}{2}A \cos(\omega+p)t$
LSB	$\frac{1}{2}A \cos(\omega-p)t$

and the resultant of such a combination moves up and down the carrier phasor, from the same total amplitude to the same difference amplitude but always on the carrier phase. see Figure B2.1.2.iv .

Here is seen the difference between "BEAT" between two waves and "MODULATION" of a carrier by a low frequency wave. When one wave is much smaller than the other, a beat phenomenon may have certain resemblance to modulation. But when the waves approach one another in magnitude, the outline appearance will be quite different. This can be seen for example on Photo 22 page B2.1.2.R . The top trace shows an 18 kHz wave modulated by 400 Hz to 95% being detected in a single-ended mismatched demodulator. The third trace shows the beat between a wave of 17 kHz of smaller magnitude. The fourth trace shows the beat between equal amplitude 17 kHz and 18 kHz waves. Here the second harmonic is perhaps 50 or 60 percent.

The cure for second harmonic distortion is to use a detector with second order non-linearity as Everitt suggested ((11) page 408). Of course the small signal which has to be applied to find such a portion on any practical diode means that this procedure is not of much use to the equipment designer. In the next paragraph B2.2. there will be given a mismatched demodulator design which will fulfil the need.

B2.1.2.R RESULTS

A test of Single-ended Mismatched Demodulator was performed using the circuit of Figure B2.1.2.v , driven from <u>either</u> a laboratory signal generator

<u>or</u> an IF output from a communications receiver .

Photographs page B2.1.2.R

70 Shows Carrier of 100 kHz modulated to 98% by sinusoidal wave of 1.2 kHz, firstly before detection, then at the demodulator before filtering, then after the filter. It is notable that there is no diagonal clipping. (Marconi 15 kHz to 30 MHz Signal Generator)

65 Waveform at the demodulator before filtering, showing
68 deeply modulated music (piano-accordion) on the IF output at 100 kHz from the communications receiver, tuned to a broadcast station. (Racal RA 17 L).

72 Half sine waves at the demodulator before filtering using much faster time base speed. Top shows unmodulated carrier. Lower shows various occurences during active modulation. Communications receiver again.

22 Shows a group of four traces:-
Top 18 kHz modulated by 400 Hz.
2nd. Unmodulated carrier of 18 kHz.
3rd. Beat between carrier 18 kHz and some 17 kHz.
4th. Beat between carrier 18 kHz and equal 17 kHz.
All waveforms at the demodulator before the filter.

B2.2.1. Balanced Demodulators

As deduced in the last theoretical analysis, there is a requirement during the demodulation of SSB with a re-inserted carrier for there to be a second order non-linearity, in order to cancel second harmonic distortion in the audio output.

If the same nomenclature is used, it may be seen from the diagram Figure B2.2.1.i that a circuit such as Figure B2.2.1.ii (or its equivalent in the mismatched form such as Figure B2.2.1.iii) will fulfil the following analysis.

Diode 1 peak rectifies $R_1 = \sqrt{(C + A\cos\theta)^2 + (A\sin\theta)^2}$

Diode 2 peak rectifies $R_2 = \sqrt{(C - A\cos\theta)^2 + (A\sin\theta)^2}$

$$D\,1 = \left(C^2 + 2AC\cos\theta + A^2\right)^{\frac{1}{2}}$$

$$D\,2 = \left(C^2 - 2AC\cos\theta + A^2\right)^{\frac{1}{2}}$$

$$D\,1 = \left[C^2 + A^2\right]^{\frac{1}{2}} \left(1 + \left\{\frac{2AC}{C^2+A^2}\right\}\cdot\cos\theta\right)^{\frac{1}{2}}$$ by Binomial Expansion

$$D\,2 = \left[C^2 + A^2\right]^{\frac{1}{2}} \left(1 - \left\{\frac{2AC}{C^2+A^2}\right\}\cdot\cos\theta\right)^{\frac{1}{2}}$$

$$D\,1 = \left[\ \right]^{\frac{1}{2}}\left(1 + \tfrac{1}{2}\{\ \}\cos\theta + \frac{\frac{1}{2}(-\frac{1}{2})}{2!}\{\ \}^2\cos^2\theta + \frac{\frac{1}{2}(-\frac{1}{2})(-\frac{3}{2})}{3!}\{\ \}^3\cos^3\theta + \cdots\cdots\right)$$

$$D\,2 = \left[\ \right]^{\frac{1}{2}}\left(1 - \tfrac{1}{2}\{\ \}\cos\theta + \frac{\frac{1}{2}(-\frac{1}{2})}{2!}\{\ \}^2\cos^2\theta - \frac{\frac{1}{2}(-\frac{1}{2})(-\frac{3}{2})}{3!}\{\ \}^3\cos^3\theta + \cdots\cdots\right)$$

The output taken is the difference between these two peak values. It is evident that odd order components add and even order components cancel. Therefore the second harmonic disappears and the major remaining distortion is the third harmonic.

But $\cos^3\theta = \frac{1}{4}(3\cos\theta + \cos 3\theta)$

So there is a small contribution to the fundamental from $\cos^3\theta$

Then $\dfrac{\text{3rd Harmonic Voltage}}{\text{1st Harmonic Voltage}} \simeq \dfrac{\frac{1}{32}\{\ \}^3}{\{\ \} + \frac{3}{32}\{\ \}^3} = \dfrac{\frac{1}{32}\{\ \}^2}{1 + \frac{3}{32}\{\ \}^2}$

Which for less than 1% harmonic distortion for a corresponding example (ref page 58) requires there to be $C \geq 3.16A$, or in other words the peak re-insertion carrier need be only 3.16 times the normal SSB phasor magnitude. Compared with the single-ended devices, this form of detector is obviously the best choice for SSB demodulation.

Such balanced modulators have been used for this purpose for many years. The so called Product Detector being the description given to a commonly used form. A late reference to a square wave carrier type of this form is (13) though the application discussed is frequency changing.

B2.2.1.R RESULTS

An experimental Balanced Mismatched Demodulator has been tested using some simulated SSB signals and the waveforms have been photographed below. In addition a dual purpose demodulator has been made for use with the communications receiver in the laboratory. Having an undetected IF output at 100 kHz a switchable Balanced or Unbalanced Mismatched Demodulator unit plus audio amplifier and speaker unit it has been possible to compare quality of sound for received SSB transmissions. Listening tests have shown that there is a generally agreed improvement, according to a dozen or more persons to whom it has been demonstrated.

Photographs page B2.2.1.R

24 Shows the demodulation of a balanced 18 kHz "SSB phasor" with a 17 kHz "carrier"

31 Top trace shows a situation similar to photo 24 with larger signals.
Lower trace shows the output after one stage of low pass filtering.

30 Shows two overlapping traces. The larger is a simulation of SSB at 40 kHz and the output after filtering.

B2.3 Conclusions on Linearity of Amplitude Demodulation

The subject of the demodulation of amplitude modulated carrier waves has been a fairly stagnant topic for a long time. The peak rectifying detector has reigned unchallenged in the realm of normal AM . The move towards SSB in the radio application has stimulated some activity there, but the interest in linearity has been slight. In the realm of multiplex telephony the need for balance to achieve tolerable quality in demodulation of SSB signals has been well understood and applied. However desire for efficiency has inhibited experimentation with diodes fed from high impedance sources.

It is hoped the Mismatched Modulator technique presented here may be a worthwhile contribution to the subject. Certainly the Mismatched Demodulator is free from the long known "diagonal clipping" distortion of the peak rectifying detector. And certainly for coherent detection it is useful to have the local carrier re-inserted in parallel with the SSB phasor voltages arriving from the modulated source. The energy from the oscillator is very largely isolated from the preceding stage, a fact which in receivers should make the design of an automatic gain control a little more easy. Finally the Mismatched Demodulators can be d.c. coupled so that there is then no difficulty with phase linearity down to zero frequency. So applications to video, facsimile , data, telegraphy, phase modulation and others would be ideal tasks for one of the forms of this versatile device.

It is intended to pursue further investigations towards the improvement of VHF demodulators for TV intermediate frequency amplifiers, or other wideband system There are few requirements for demodulation of frequencies higher than VHF since by application of the superhet principle all higher carrier frequencies may be brought down before demodulation to base band.

P A R T   C

F R E Q U E N C Y   M O D U L A T I O N

Part C FREQUENCY MODULATION LINEARITY

Section C1 Previous Techniques and Theories

C1.1 Previous Techniques

The classical methods of producing Frequency Modulation which are given in the textbooks are:-

i) variable reactance device (valve or transistor)

ii) Armstrong's method (or Phase Modulator)

They will be examined in terms of their ability to produce linear frequency modulation.

i) Variable Reactance Device

The oscillator to be modulated has a resonant circuit part of whose reactive component is formed by the output impedance of the variable reactance under the control of the input signal to be impressed as a frequency change of the carrier. By means of a capacitor-resistor potential divider some of the device output voltage is fedback into its own input terminal at a phase different from the output voltage. Within the device, gain gives to the output a greater or lesser proportion of the wrongly phased energy depending upon the transconductance of the device at that instant set by the instantaneous voltage of the signal. Thus there appears at the output the desired variable reactance for oscillator change.

In terms of linearity, such circuits have three points of criticism.

(a) the transconductance variation of pentode valves was an 'ad hoc' feature of the particular specimen since it depended upon the grid wire spacing progressively in step gradations.

(b) the transconductance of the transistor is different at different temperatures

(c) the variations in reactance are accompanied by simultaneous resistance variations which cause inevitably some amplitude changes as well.

ii) Armstrong's Method

A fraction of the output from a quartz crystal controlled oscillator is modulated in a balanced modulator so that double sideband suppressed carrier results. Addition to the remaining fraction of the original phase shifted 90 gives Phase Modulation. Such phase modulation has a degree of frequency modulation which is linearly proportional to the signal frequency. The desired relationship being independent of signal frequency, there therefore has to be multiplication by an inverse proportionality with respect to signal frequency. Armstrong's achievement was to produce the necessary relationship by interposing a predistortion circuit in the audio input taking the form of a very simple C-R integration circuit.((8) page 65). Betts here adds that for reasonable phase linearity one must not ask for more than about 0.2 rad peak phase deviation of the phase modulator, and this necessitates a X 7500 multiplication for the subsequent stages of the transmitter. In practice X 81 or X 729 is enough to contemplate for otherwise at the higher frequency end of the audio band the input signals will be so small that tremendous amplification will

be obligatory and noise will become troublesome.
In addition multiplication factors of X 7500 may produce phase jitter which also appears as noise to a user.

Present (1974) broadcast transmitters employ a feedback technique not usually described in subject texts, called "Frequency Modulated Quartz" or FMQ. In the FMQ generator there is a reactance controlled oscillator whose output is sampled and demodulated in two parallel frequency discriminators. One of which is narrow band, high Q and stable: being actually a quartz crystal, is used to return the mean frequency to a maintained value.
The other discriminator is wide band and very linear (in the sense of output versus frequency) and is used in an a.c. feedback path to linearise the frequency modulation characteristic of the whole variable reactor - oscillator - discriminator loop.

With careful adjustment and maintained standards it is conceivable that such equipments will produce highly linear FM. Nevertheless it is probably only in the realm of public broadcasting can sufficient investment be provided to accept such complication in construction and such care in maintenance.

Cl. 2 Previous Theories

Nearly all the expounded theoretical presentations have followed the mistaken guidance of the circuit engineers who sought to produce in the variable reactance device a method of producing a linear reactance versus voltage characteristic and then to apply such a component to an oscillator in order to cause a linear frequency versus voltage effect ultimately. Using the principles of differential calculus upon the resonant frequency equation

$$2\pi f_0 = 1/\sqrt{LC}$$

it is shown that for small proportional changes in frequency, the change is almost linearly prortional to reactance, and therby justify the linear variable reactance device provided. The most recent textbook maintaining the same mistaken viewpoint is Betts (8) page 63, though he quotes the modern component namely the varactor diode, remarking that it is a pity it does not have the linear reactance versus voltage characteristic.

Cl. 3 Conclusions on Previous Techniques & Theories

The conclusion of the present position both on technique and theory of FM generation is one of misdirected effort. Only in the case of the Armstrong open loop device and the FMQ closed loop circuit has there been an approach to good linearity. And even these have perceivable limitations.

Section C 2 FM Linearity by Square - Law Reactance and Square Wave Oscillator

C2.1 FM Linearity by Square - Law Reactance

Reconsideration of the resonant frequency equation shows that for an unrestricted change of frequency, the frequency is strictly inversely proportional to the square of reactance.

$$2\pi f_o = 1/\sqrt{LC}$$

In consequence of this simple fact, any reactive component which has a reactance proportional to the square of voltage will produce linear frequency modulation. Thus any oscillator across whose reactive circuit is shunted a square law proportional reactance device will have the desired linear frequency versus voltage characteristic.

Now it is a well known fact of semiconductor electronics that an abrupt junction diode has a capacity proportional to $1/\sqrt{V_R}$ Where V_R = Reverse voltage

i.e. $C_{jn} \propto 1/\sqrt{V_R}$

or $C_{jn}^2 \propto 1/V_R$

Thus a semiconductor junction diode will produce linear frequency versus voltage modulation of an oscillator, without restriction.

C2.1.1 FM Oscillator Circuit

For the sake of completeness an oscillator based on the above principles is shown in Figure C2.1.1

C2.2 FM Linearity by Square Wave Oscillator (Astable)

The Astable Multivibrator may be used as a generator of frequency modulation if slightly modified according to Figure C2.2.i . The signal to be used as frequency control is arranged to add to or subtract from the supply rail potential for the base resistors only. It thereby raises or lowers the aiming voltage (V_{aim}) to which the inside plates of the base capacitors attempt to charge. The charging current they receive is controlled by the base resistors.

Using the designations of Figure C2.2.i and applying these to Figure C2.2.ii it may be shown that the pulse time τ can be calculated from

$$\tau = CR \log_e \frac{V_{aim}}{V_{aim} - V_{rise}}$$

Where V_{aim} is almost exactly $2V_{rise}$

Since the base-collector-capacitor group fall a voltage V_S at each conduction event.

For simplicity in the following analysis

V represents $2V_s$ and also V_{aim} and also $2V_{rise}$

$$\tau \doteqdot CR \log_e \frac{V}{V - V/2}$$

$$\tau \doteqdot CR \log_e 2$$

$$\tau \doteqdot 0.7\ CR$$

But frequency of oscillation f_o is given by

$f_o = 1/2\tau$

So from last page

$f_o \doteqdot 1/\ 1.4CR$

Suppose the signal adds a small increment δv to the aiming potential. Then a new pulse time $\tau + \delta\tau$ arises as follows:-

$$\tau + \delta\tau = CR\ \log_e \frac{V + \delta v}{V + \delta v - V/2}$$

$$= CR\ \log_e \frac{2V + 2\delta v}{V + 2\delta v}$$

$$= CR\ \log_e 2\left(1 + \frac{\delta v}{V}\right)\left(V + \frac{2\delta v}{V}\right)^{-1}$$

which by the Binomial Theorom

$$\tau + \delta\tau \doteqdot CR\ \log_e 2\left(1 + \frac{\delta v}{V}\right)\left(1 - \frac{2\delta v}{V} + \cdots\cdots\right)$$

$$\doteqdot CR\ \log_e 2\left(1 - \frac{2\delta v}{V} + \frac{\delta v}{V}\right)$$

$$\doteqdot CR\ \log_e 2\left(1 - \frac{\delta v}{V}\right)$$

$$\doteqdot CR\ \log_e 2 \quad + \quad CR\ \log_e\left(1 - \frac{\delta v}{V}\right)$$

$$\doteqdot \tau \quad + \quad CR\ \log_e\left(1 - \frac{\delta v}{V}\right)$$

So $\quad \delta\tau \doteqdot CR\ \log_e\left(1 - \frac{\delta v}{V}\right)$ But $\log_e(1-x) = -x - \frac{x^2}{2} - \frac{x^3}{3} - \cdots\cdots$

So $\quad \delta\tau \doteqdot CR\left[-\frac{\delta v}{V} - \cdots\cdots\right]$

Therefore $\quad \dfrac{\delta\tau}{\delta v} \doteqdot -\dfrac{CR}{V}$

Or $\quad \dfrac{\delta\tau}{\delta v} \doteqdot -\dfrac{CR}{2V_s}$

Negative proportionality since τ shorter when V lifted.

Now $f_o = 1/2\tau$

So two deductions follow

i) $\frac{\delta f}{\delta \tau} = -\frac{1}{2\tau^2}$ and ii) $\frac{\delta f}{\delta v} = \frac{\delta f}{\delta \tau} \cdot \frac{\delta \tau}{\delta v}$

$$\doteqdot -\frac{1}{2} \cdot \frac{1}{(CR \log_e 2)^2}$$

Hence

$$\frac{\delta f}{\delta v} \doteqdot -\frac{1}{2} \cdot \frac{1}{(CR)^2 (\log_e 2)^2} \cdot \left(-\frac{CR}{V}\right)$$

$$\doteqdot +\frac{1}{2} \cdot \frac{1}{CR (\log_e 2)^2 V}$$

But $(\log_e 2)^2 = 0.7^2 = 0.5$

So $2(\log_e)^2 = 1$

So $\frac{\delta f}{\delta v} \doteqdot \frac{1}{CRV}$

Or $\underline{\underline{\frac{\delta f}{\delta v} \doteqdot \frac{1}{CR2Vs}}}$ Units are Hertz per volt

Positive proportionality since frequency goes higher when input voltage lifted.

C2.2.1 Experimental Verification

In order to make measurements to check the analysis given above, an astable oscillator for 460 kHz was designed. The circuit was built as the first section of an FM modulation and demodulation experiment of which greater detail is given later in paragraph D2.2.1

Owing to switching time of transistors totalling over 100 ns each and since the normal mid range manual setting of the frequency control would be slightly down from the full supply voltage, the nominal design frequency was raised to 25% above the value desired.

Nominal Design Frequency 460 × 1.25 = 580 kHz

Reasonable value for base frequency 18 kΩ

From $f = 1/\ 1.4CR$

Calculate C value 68 pF

RESULT Mid Manual Control 460 kHz

Modulation Sensitivity

From $\frac{df}{dv} = 1/\ CR\ 2V_S$

Calculate Sensitivity $= \dfrac{1}{68\times10^{-12}\times18\times10^{3}\times2\times9}$ Hz/V

$= \dfrac{10\times10^{6}}{6.8\times1.8\times2\times9}$ Hz/V

$=$ 45.3 kHz/V

But a similar allowance has to be made for switching time etc.

So the expected Modulation Sensitivity will be:-

$= 45.3 \times 1.25$

$=$ 56.6 kHz/V

RESULT Measured Modulation Sensitivity 58 kHz/V

C2.3 Conclusions on Linearity of Frequency Modulation

To summarise the position respecting FM it may be said that with intelligent application of the techniques suggested, very linear modulation is now available in un-complicated circuit arrangements. If there is sufficient component investment much more sophisticated negative feedback systems can be engineered in quite small space. For instance a voltage controlled oscillator of rough linearity may be incorporated into a phase lock loop using a phase sensitive detector and a digital counter. Time does not allow full coverage to be made here.

P A R T D

F R E Q U E N C Y D E M O D U L A T I O N

Part D FREQUENCY DEMODULATION LINEARITY

Section D1 Previous Techniques and Theories.

D1.1 Previous Techniques.

The history of the design of FM Demodulators (or discriminators) has been one of gradual development.

The most naiive circuit design has been the Off-tune Resonant Circuit followed by peak detector (rectifier). Unfortunately as illustrated in Figure D1.1.i there is likely to be considerable distortion.

The form next in sequence takes the curvature of one end and cancels it by use of another off-tune circuit set the other side of the mean frequency. Two peak rectifying detectors are used and their output voltages are subtracted, see Figure D1.1.ii
The arrangement shown whereby energy is fed to the two separate circuits by mutual induction from a single primary is called the Crosby Discriminator. It has an extended straight line region but has approximately the same slope as either single off-tune circuit alone. A great advantage which arises in consequence of the cancellation procedure is the elimination of d.c. output voltage at the mean frequency. Since output d.c. polarity changes at cross-over, the circuit enabled automatic frequency control receivers to be made for the first time.

From the Crosby discriminator there followed the Foster-Seeley Discriminator in which the double secondary evolves into a single tuned circuit with centre tap. Energy is obtained from the primary in two ways simultaneously, see Figure Dl.1.iii . The full voltage of the primary is placed upon the centre tap by the low reactance capacitor (designated v_{in}) but there is also a mutual inductive effect which initiates a circulating current of orthogonal phase (designated i), see Figure Dl.1.iv . Both the primary and secondary inductors are influenced by the mutual inductance but are re-resonated by adjustment elements so that they are each tuned to the centre frequency. For the sake of simplicity in writing the mathematics it is taken that they are both the same value L henry (actually M will increase primary inductance and reduce secondary inductance) and consequently the capacitors of primary and secondary will be also equal, say C farad. In the section to follow a theoretical attack will be made to derive a sensitivity formula and a comment on the likely linearity.

Finally as in so much of modern electronic engineering there are two FM demodulators which depend upon switching techniques and will not be studied extensively in this review.

The Pulse Counter Discriminator is constructed so that an actual pulse count per second is estimated by conversion of each carrier wave into a uni-directional pulse. The resultant train of equal energy pulses is fed via a diode into a leaking integrator which reaches an equilibrium voltage roughly proportional to the number of pulses per second. Whilst the Pulse Counter Discriminator has appealing features, there are so many disadvantages that it will be little used in future. To enumerate:

a) it must be operated at a mean frequency only about $1\frac{1}{2}$ times the peak deviation frequency or the sensitivity is poor.

b) the linearity is an exponential function.

c) due to a) a double superhet receiver is required so that image interference may be avoided.

The Phase Lock Loop may be used as a highly linear FM demodulator. In principle the input frequency modulated carrier is used as one input to a phase sensitive detector comparing the phase error relative to a controlled oscillator responsive to the phase sensitive detector output voltage. Thus during frequency change by the input carrier, there is a change of output voltage due to the detector forcing the oscillator to follow the original. Phase should never slip but will elastically strain back and forth within the plus and minus 90° available. The development of integrated circuits has enabled such complicated circuits to be incorporated in domestic entertainment equipment in recent years (1974).

There will be seen integrated circuits designated "complete FM discriminator" in certain catalogues. These are actually a chain of limiting amplifiers followed by a phase sensitive detector of a switching type. However they require an external tuned circuit to produce a shift of phase linearly proportional to frequency. They are simply integrated circuit Foster-Seeley devices. Their circuit behaviour follows Foster-Seeley theory and restrictions.

Dl.2 Previous Theories.

The standard texts are surprisingly scant in the quantity of useful theory in regard to the explanation of the Fourier spectrum of an FM wave and also in regard to the sensitivity and linearity of practical FM discriminators. It is hoped to fill the gaps in subsequent analysees below.

Dl.3 Conclusions on Previous Techniques and Theories

Undoubtedly there have been a number of very useful FM demodulator circuits devised and implemented with success. To some extent they have been successful because they have been engineered with generous tolerances in regard to bandwidth and cost. The field of FM has been both the sphere of the high-fidelity enthusiast and also of the professional communications engineer with a capital P. Each group has been able to ask for (and usually get) as much bandwidth as he feels he needs. Generally there has been little pressure to design to minimum bandwidth or design to minimum cost. Consequently certain laxity in theory of discriminators has become tolerated.

If a comment is to be made concerning the spectral analysis it can only be said that the theory has been sketched in and very little more. Few texts do more than put forward the Jacobi solution to the puzzling equation containing the strange terms cos(m sin t) and sin(m sin t) which arise quite early in the mathematics of FM. A solution by series which are commonly known is given here, see D2.1 .

Section D2 FM Receiver Bandwidth and Linearity

D2.1 Spectral Components of an FM Wave and their Consequence upon Receiver Bandwidth.

In order to calculate the bandwidth necessary for an FM receiver it has been necessary to analyse the Fourier Spectrum of an FM wave.

Consider a general signal $v = B\cos(pt+\phi)$ which is to be frequency modulated upon a general carrier wave $v = A\cos(\omega_0 t + \theta)$
At all times the instantaneous frequency of the carrier is to be a representation of the instantaneous magnitude of the signal voltage. The amplitude of the carrier wave is to remain unchanged throughout.

Time origins for each wave may be chosen so as to allow phase terms ϕ & θ to be equal zero and dropped out.

So a general FM wave is given by

$$v = A\cos\int(\omega + kB\cos pt)\,dt$$

where k is the modulator sensitivity in rad/s per volt

The mean angular frequency is ω rad/s
The peak angular frequency deviation is kB rad/s which is symbolised ω_d .

Integrating the general wave to obtain instantaneous angle

$$v = A\cos\left(\omega t + \frac{kB}{p}\sin pt\right)$$

which departs plus and minus $\frac{kB}{p}$ radians from the mean angle.

Now the angle $\frac{kB}{p}$ is important in the theory because it is both the peak angle of departure on the phasor diagram, and also because it is the argument of the Bessel functions whose magnitudes determine the component magnitudes.

The term $\frac{kB}{p}$ is named the modulation index and designated m.

Since $\omega_d = kB$ and $m = \frac{kB}{p}$

Then $m = \frac{\omega_d}{p}$

It is convenient to discover that the modulation index has the same numerical value whether the deviation and signal frequencies are expressed in rad/s or in Hertz. The 2π divides out.

So $m = \frac{f_d}{f_s}$ Where f_d is deviation freq. peak in Hz
f_s is signal freq. in Hz

So the general wave can now be written

$$v = A\cos(\omega t + m\sin pt)$$

Or $v = A[\cos\omega t.\cos(m\sin pt) - \sin\omega t.\sin(m\sin pt)]$

Since $\cos(A+B) = \cos A.\cos B - \sin A.\sin B$

Both involve 'cos of sin' or 'sin of sin'

Using Euler's Formula

$$\cos(m\sin pt) + j\sin(m\sin pt) = e^{j(m\sin pt)}$$

But $$e^{jm\sin pt} = e^{\frac{m}{2}2j\sin pt}$$
$$= e^{\frac{m}{2}2j\frac{e^{jpt}-e^{-jpt}}{2j}}$$
$$= e^{\frac{m}{2}e^{jpt}} \times e^{-\frac{m}{2}e^{-jpt}}$$

Which is the product of two series of the type
$$e^{x} = 1 + x + \frac{x^2}{2!} + \frac{x^3}{3!} + \cdots$$

One being $$e^{\frac{m}{2}e^{jpt}} = 1 + \frac{m}{2}e^{jpt} + \left(\frac{m}{2}\right)^2\frac{e^{2jpt}}{2!} + \left(\frac{m}{2}\right)^3\frac{e^{3jpt}}{3!} + \left(\frac{m}{2}\right)^4\frac{e^{4jpt}}{4!} + \cdots\cdots$$

The other $$e^{-\frac{m}{2}e^{-jpt}} = 1 - \frac{m}{2}e^{-jpt} + \left(\frac{m}{2}\right)^2\frac{e^{-2jpt}}{2!} - \left(\frac{m}{2}\right)^3\frac{e^{-3jpt}}{3!} + \left(\frac{m}{2}\right)^4\frac{e^{-4jpt}}{4!} + \cdots\cdots$$

Which are both convergent all m and all e so the product is convergent

Therefore the product is

$$= e^{0}\left[1 - \left(\frac{m}{2}\right)^2 + \frac{\left(\frac{m}{2}\right)^4}{(2!)^2} - \frac{\left(\frac{m}{2}\right)^6}{(3!)^2} + \cdots\right]$$

$$+ e^{jpt}\left[\frac{m}{2} - \frac{\left(\frac{m}{2}\right)^3}{2!} + \frac{\left(\frac{m}{2}\right)^5}{2!3!} - \frac{\left(\frac{m}{2}\right)^7}{3!4!} + \cdots\right] \quad + e^{-jpt}\left[-\left(\frac{m}{2}\right) + \frac{\left(\frac{m}{2}\right)^3}{2!} - \frac{\left(\frac{m}{2}\right)^5}{2!3!} + \cdots\right]$$

$$+ e^{2jpt}\left[\frac{\left(\frac{m}{2}\right)^2}{2!} - \frac{\left(\frac{m}{2}\right)^4}{3!} + \frac{\left(\frac{m}{2}\right)^6}{2!4!} \cdots\cdots\right] \quad + e^{-2jpt}\left[\frac{\left(\frac{m}{2}\right)^2}{2!} - \frac{\left(\frac{m}{2}\right)^4}{3!} + \frac{\left(\frac{m}{2}\right)^6}{2!4!} \cdots\cdots\right]$$

$$+ e^{3jpt}\left[\quad \cdots\cdots \quad\right] \quad + e^{-3jpt}\left[\quad \cdots\cdots \quad\right]$$

$$\vdots \qquad\qquad\qquad\qquad \vdots$$

The expansions in square brackets are the series for the Bessel functions of the first kind $J_n(m)$ see **(23)**.

So the product is $e^{0} J_0(m) + e^{jpt} J_1(m) + e^{2jpt} J_2(m) + e^{3jpt} J_3(m) + \cdots$
$+ e^{-jpt} J_{-1}(m) + e^{-2jpt} J_{-2}(m) + e^{-3jpt} J_{-3}(m) + \cdots$

Or the product is $\sum_{n=-\infty}^{n=\infty} e^{jnpt} J_n(m)$

But $J_{-1}(m) = -J_1(m)$ and $J_{-2}(m) = -J_2(m)$ etc.

And $e^{jpt} = \cos pt + j \sin pt$ and $e^{2jpt} = \cos 2pt + j \sin 2pt$ etc.

But $\cos(-\theta) = \cos(\theta)$ and $\sin(-\theta) = -\sin(\theta)$

So $e^{-jpt} = \cos(-pt) + j \sin(-pt) = \cos pt - j \sin pt$

And $e^{-2jpt} = \cos 2pt - j \sin 2pt$ etc

Therefore the product

$$= J_0(m) + \cos pt . J_1(m) + j \sin pt . J_1(m) + \cos 2pt . J_2(m) + j \sin 2pt\, J_2(m) + \cos 3pt . J_3(m)$$
$$- \cos pt . J_1(m) + j \sin pt . J_1(m) + \cos 2pt . J_2(m) - j \sin 2pt\, J_2(m) - \cos 3pt . J_3(m) .$$

$$= J_0(m) + j 2 J_1(m) \sin pt + 2 J_2(m) \cos 2pt + j 2 J_3(m) \sin 3pt + 2 J_4(m) \cos 4pt + \cdots$$

But the original product came from

$$\cos(m \sin pt) + j \sin(m \sin pt)$$

So equating Re and Im parts

$$\cos(m \sin pt) = J_0(m) + 2\left[J_2(m) \cos 2pt + J_4(m) \cos 4pt + J_6(m) \cos 6pt + \cdots\right]$$

$$\sin(m \sin pt) = +2\left[J_1(m) \sin pt + J_3(m) \sin 3pt + J_5(m) \sin 5pt + \cdots\right]$$

So the prime phasors making up the full resultant are

$$v = A\left[\cos \omega t \left\{J_0(m) + 2 J_2(m) \cos 2pt + 2 J_4(m) \cos 4pt + \cdots\right\}\right.$$
$$\left. - \sin \omega t \left\{2 J_1(m) \sin pt + 2 J_3(m) \sin 3pt + 2 J_5(m) \sin 5pt + \cdots\right\}\right]$$

Which may be sketched at the time when the resultant has maximum phase displacement i.e. when the carrier frequency is equal to the mean frequency. This occurs when $\cos pt$ is zero, or pt is $\pi/2$, $3\pi/2$ etc

At this time $\cos pt = 0$, $\sin pt = 1$

$\cos 2pt = -1$, $\sin 3pt = -1$

$\cos 4pt = +1$, $\sin 5pt = +1$ etc.

see Figure D2.1.i

These prime phasors of alternating length but unchanging angle relative to the carrier phasor, may be transformed into pairs of side frequency phasors of unchanging length but contrary rotational angle relative to the reference carrier phasor by use of the relationship

$$\cos A \cos B = \tfrac{1}{2}\left[\cos(A+B) + \cos(A-B)\right]$$

So
$$v = A\left[J_0(m)\cos\omega t - 2J_1(m)\sin\omega t\sin pt + 2J_2(m)\cos\omega t\cos 2pt - 2J_3(m)\sin\omega t\sin 3pt + 2J_4(m)\cos\omega t\cos 4pt + \cdots\right]$$

Becomes
$$v = A\left[J_0(m)\cos\omega t - J_1(m)\{\cos(\omega-p)t + \cos(\omega+p)t\} + J_2\{\cos(\omega-2p)t + \cos(\omega+2p)t\} - J_3(m)\{\cos(\omega-3p)t + \cos(\omega+3p)t\} + \cdots\cdots\right]$$

Which is a carrier frequency component and a family of pairs of side frequencies. Fortunately in any practical system with ω_d restricted, m is large only when p is small. The values of the Bessel series may be obtained from published tables, and are found to constitute sets of terms which either decay rapidly for small m or remain significant out to the deviation frequency and subsequently decay within another 15% to 20% of that frequency.

For example, in the unlikely event of a loud sound of a 6 kHz tone occuring in a programme a peak deviation of 60 kHz would be allowable on the British standard system for UHF TV Sound.

Then the modulation index $m = \frac{60k}{6k} = 10$

From Tables of Bessel functions it is seen that the sideband terms become less than 3% of the voltage of the original carrier (less than 1% of the power) by the thirteenth order. Consequently 98% of the energy will always be found within twice 13 times the signal frequency band around the mean frequency: i.e. within a band of 156 kHz.

Hence the whole receiving system should be designed to have a pass band of at least 156 kHz. Elementary logic rashly applied would have resulted in a bandwidth of 120 kHz adjudged sufficient.

A working rule which has been distilled from the analyses similar to that given above is to assume that an FM receiver system should be given a bandwidth of 2.8 times the peak deviation frequency, giving here a figure of 168 kHz.

In conclusion Reference (18) also gives a graphical method of evaluating the Bessel functions, but its inclusion here would be an unnecessary diversion from the objective.

D2.2 Foster-Seeley Discriminators

D2.2.1 The Sensitivity of the Foster-Seeley

For the presentation of an approach to a simple formula for the sensitivity of the Foster-Seeley Discriminator, the diagram and designations of Figures D1.1.iii and D1.1.iv wil be used.

It is assumed that both primary and secondary are resonated at the centre frequency including the effect of their mutual inductance. It will further be assumed that each circuit has the same L and C values. It is assumed that the primary has sufficient bandwidth to pass all significant energy. The secondary losses are totalled and shown as two small equal valued series resistors r/2 on opposite halves of the secondary. The diodes peak rectify so all voltages may be considered 'peak' .

Diode 1 Voltage is approximately $v_{in} + \frac{1}{2}v$

Diode 2 voltage is approximately $v_{in} - \frac{1}{2}v$

But $v_{in} = i.(-jX_c)$

And also (Induced voltage due to M) $= i_1.jX_M$

So $i_1.jX_M = i.(jX_L - jX_c + r)$

So $i = \dfrac{i_1 . jX_M}{jX_L - jX_c + r}$

But $i_1 = \dfrac{v_{in}}{jX_L}$ from considerations of primary

So $$i = \frac{v_{in}}{jX_L} \cdot \frac{jX_M}{r + j(X_L - X_c)}$$

But $$M = k\sqrt{L_1 L_2}$$ Where k is coupling factor

$$= k\sqrt{L^2}$$

$$= kL$$

So $$\frac{X_M}{X_L} = \frac{\omega k L}{\omega L} = k$$

But $$v = i(-jX_c)$$

So $$v = v_{in}\frac{\cancel{j}X_M}{\cancel{j}X_L} \cdot \frac{(-jX_c)}{r + j(X_L - X_c)}$$

$$= v_{in}\,k \cdot \frac{-jX_c}{r + j(X_L - X_c)}$$

$$= \frac{v_{in}\,k(-j)}{\omega C\left(r + j(X_L - X_c)\right)}$$

$$= \frac{v_{in}\,k(-j)}{\omega C r\left(1 + j\left(\frac{X_L}{r} - \frac{X_c}{r}\right)\right)}$$ normalising to the mean angular frequency

$$\therefore\quad v = \frac{v_{in}\,k(-j)}{\omega_0 C r \frac{\omega}{\omega_0}\left(1 + j\,\frac{\omega_0 L}{r} \cdot \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \cdot \frac{1}{\omega_0 C r}\right)}$$

But $$\frac{\omega_0 L}{r} = Q = \frac{1}{\omega_0 C r}$$

So $$v = \frac{v_{in}\, k(-j)Q}{\frac{\omega}{\omega_0}\left(1 + jQ\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)\right)}$$

Where $$\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) = X$$

So $$v = -j\,\frac{v_{in}\, kQ}{\frac{\omega}{\omega_0}(1+jQX)}$$

Diode 1 is given $$v_{in}\left(1 + \tfrac{1}{2}(-j)\frac{kQ}{\frac{\omega}{\omega_0}(1+jQX)}\right)$$

Diode 2 is given $$v_{in}\left(1 - \tfrac{1}{2}(-j)\frac{kQ}{\frac{\omega}{\omega_0}(1+jQX)}\right)$$

As shown in Figure D2.2.i

At the mean frequency $\omega = \omega_0$ and so X = zero and each diode has the same resultant voltage applied, as in the Figure D2.2.1.i . If the input frequency changes a small amount $\delta\omega$ from the mean and therefore becomes $\omega_0 + \delta\omega$, the principal effect is the change of sign of X with consequent change of phase of the v term. There is a negligible effect on the magnitude of $\frac{\omega}{\omega_0}$. So the phasor diagram becomes as shown in Figure D2.2.1.i .

Now $$X = \frac{\omega_0 + \delta\omega}{\omega_0} - \frac{\omega_0}{\omega_0 + \delta\omega} \simeq 2\,\frac{\delta\omega}{\omega_0}$$

And $$\theta = \arctan \frac{QX}{1}$$

So in the region of small θ where $\tan\theta \simeq \theta$

Then $$\theta \simeq Q\,\frac{2\delta\omega}{\omega_0} \text{ radian}$$

From Figure D2.2.ii it is clear that the resultant voltages applied to the diodes change by plus and minus $\frac{v_{in} k Q \theta}{2}$. So the effect on the rectified and subsequently differenced output voltage is

$$\delta V_o = \cancel{2} \cdot \frac{v_{in} k Q \theta}{\cancel{2}}$$

$$= v_{in} k Q \theta$$

$$= v_{in} k Q \cdot \frac{Q \, 2\delta\omega}{\omega_o}$$

Therefore $\frac{\delta V_o}{\delta \omega} = v_{in} \frac{2kQ^2}{\omega_o}$ volt/rad/s

Or $\frac{\delta V_o}{\delta f} = v_{in} \frac{2kQ^2}{f_o}$ Volt/Hz

Thus it is seen that the sensitivity of the Foster-Seeley Discriminator is:-

a) proportional to coupling factor

b) inversely proportional to mean frequency

c) proportional to secondary circuit squared.

From practical considerations based on the established normal centre intermediate frequencies of 10 MHz (or 6 MHz) for FM with peak deviation of 75 kHz (or 60 kHz) the secondary values come out to be about 30 .

D2.2.2 The Linearity of the Foster-Seeley

In considering the possible linearity of the Foster-Seeley Discriminator, the most significant approximations are:-

a) the geometry of the phasor diagram Fig D2.2.2.i

b) the equating of $\tan\theta$ and θ.

Firstly :- Geometry

In the approximation, the arc lengths $\frac{V_{in}}{2}kQ\theta$ were taken as correct. From the new Figure it is seen that the diode resultant voltages are precisely

Diode 1 $= \sqrt{V_{in}^2(1-\frac{1}{2}kQ\sin\theta)^2 + V_{in}^2(\frac{1}{2}kQ\cos\theta)^2}$

Diode 2 $= \sqrt{V_{in}^2(1+\frac{1}{2}kQ\sin\theta)^2 + V_{in}^2(\frac{1}{2}kQ\cos\theta)^2}$

which when subtracted are exactly analogous to the Balanced Demodulator (for AM) given in B2.2.1 page 60. Thus even order distortion will cancel (the chief idea of the Foster-Seeley) and thus the third order will be the main distortion left. As in the earlier analysis third order terms are naturally small in triangles with the $\frac{V_{in}}{2}kQ$ phasor less than 1/2 V_{in} and with θ being less than 45° normally.

Secondly:- Tan θ and θ.

The actual series expansion for small angle $\tan\theta$ is $\tan\theta = \theta + \frac{\theta^3}{3} + \frac{2\theta^5}{15} + \cdots$

Now if the approximation of the use of the first term only is employed, the error is 4.6% by 20° and 9.1% by 30°.

If a system of bandwidth 330 kHz at an IF of 10 MHz is assumed to have a suitable Q, it will be 30 . (since $B = f_0/Q$).

For 20° arctan 0.364 comes from $Q\chi$

and $\chi \simeq \frac{2\delta\omega}{\omega_0}$

So $\frac{2\delta\omega}{\omega_0} = \frac{0.364}{30}$

Or $\frac{\delta\omega}{\omega_0} = 0.006$

i.e. $\delta\omega \simeq 0.006\,\omega_0$

Or $\delta f \simeq 0.006\, f_0$

which for 10 MHz gives $\delta f \simeq$ 60 kHz a value less than the peak deviation used for the British VHF FM Service. So the inference from such an estimate must be that for linearity a discriminator should not be expected to produce too much sensitivity.

D2.2.3 Two Variants of the Foster-Seeley

In order to conclude on the Foster-Seeley, it may be of importance to record a useful version which can be called the Modified Foster-Seeley Discriminator disclosed in Ref (10). Now it will be clear that in certain production situations the normal Foster-Seeley circuit is slightly awkward to reproduce because mutual coupling is required between the primary and secondary inductors. Mutual coupling does not supply energy transfer, it is merely to ensure 90° phase relationship between secondary and primary voltages at the mean frequency. Dougharty suggested the phase of the secondary voltage could be set to 90° if some current from the plenitude circulating there is bled away to earth through a high reactance. If the reactance is an inductance it provides a d.c. return path for the diode rectified current and replaces the radio frequency choke of the original Foster-Seeley. The circuit of the Modified Foster-Seeley Discriminator is shown in Figure D2.2.3.i . A typical phase shifting inductor can be 10 times the magnitude of the tuning component. All three coils may be within their own individual screening can, and thus production line variability is diminished.

Very often two other variants of the Foster-Seeley Discriminator have been supplied in domestic FM receivers. One has the circuit form shown in Figure D2.2.3.ii.Here the mutual coupling and phase shift is provided by a few turns of the secondary being tightly coupled to the primary. Perhaps it may be named the Low Inductance Coupled Foster-Seeley Discriminator. It is presumably more reproducible than the elementary form, though there is still a disadvatageous three wire connection between primary and secondary (compared with the last mentioned modification) and the r.f. choke is still needed.

Variant number two is the <u>Ratio Detector</u>, which is a Foster-Seeley having oppositely connected diodes and a large time constant smoothing circuit across one diode-load. The diagram of Figure D2.2.3.iii gives a simplified circuit of the secondary side. The operational behaviour is of the same kind as the other Foster-Seeley discriminators; at mean frequency the diodes peak rectify equal resultant voltages. Equal currents flow in opposite directions towards R_1 and therefore cancel out and flow elsewhere. The resistor R_2 has the same ohmic value but is forced to carry both currents. In consequence it has a d.c. voltage of twice the expected peak value across itself. It is shunted by a large capacitor C_2 which

a) maintains the double d.c. voltage across R_2 unchanging during audio frequency events

b) passes audio currents easily fron diode 2 to appear in R_1

As the carrier swings above or below the mean frequency, the output audio currents flow in opposite directions towards R_1 so there is alternately net current one way and then the other way through it. Thus an audio voltage is available at the terminal shown.

There has been a misconception concerning the ability of this type of discriminator to mitigate amplitude modulated noise . It has often been said that as the voltage across R_2 may vary it will eliminate noise. Unfortunately this is not true because the noise modulation which is a nuisance is audio spectrum noise. The voltage across R_2 is specifically prevented from varying with this kind of rapidity, and therefore the variations due to noise will appear across R_1 just as if they were desirable signals.

Perhaps the misconception has persisted because such discriminators always follow a limiter of the conventional form and it is this limiter which provides the noise reduction attributed to the Ratio Detector. The only advantage of the Ratio Detector (or Ratio Discriminator as it would better have been called) is its ability to perform the output subtraction. In the days of valve receivers this probably meant that a valve could be dispensed with. In the days of semiconductors and I/C electronics such a procedure is unnecessary. One further active component is neither here nor there.

D3 FM Experimental Equipment

Photograph page D3.R.1

In order to make measurements to check the analyses given above, an FM demonstration unit was designed and a pair constructed for student experimentation. A low mean frequency enabled simple oscilloscopes to be employed for observation and simplified the acquisition of filter components. The circuit comprises an astable oscillator with band pass filter which is connectable to a limiter and Modified Foster-Seeley Discriminator. There is also provided a crude spectrum analyser by which the Fourier spectrum may be verified. This has a simple diode mixer circuit and low pass filter. If a signal generator is sewpt slowly through the expected band of a frequency modulated carrier, a low frequency signal will appear at every side frequency. The low pass filter has a cut off

Frequency of 300Hz and so components in the spectra separated by as little as 1 kHz can be observed by use of a sensitive millivoltmeter (a.c.) connected at the output terminal. See Figures D 3.i and D 3.ii

The Astable Frequency Modulated Oscillator design has already been described in paragraph C2.2.1

Modified Foster-Seeley Discriminator

A straightforward Modified Foster-Seeley circuit was fitted following a long tail pair limiter stage. As the phase shifting inductor has a reactance 10 times the secondary tuning inductor , the effective coupling factor k of the design is 0.1 . A reasonable secondary Q for deviation of 16 kHz is calculated from $Q = f_0/f_d$

$$\text{So} \quad Q = 28$$

Load resistors for the diodes chosen as 39 kΩ

From paragraph it is known that the loading conductance of each diode is twice the loading of its own load resistor.

But there are two diodes effectively in parallel

So the total loading conductance is $4 \times 1/39k$

$$= 1024\ \mu S$$

But from $Q = \omega CR$ with $Q = 28$

$$C = 134 + 12$$

$$= 146\ pF$$

$$R = 28 / 2\pi \times 460 \times 10^{3} \times 146 \times 10^{-12}$$

$$= 6.64\ k\Omega$$

So conductance $= 1500\ \mu S$

So extra conductance of $476\ \mu S$ must be provided by a physical resistor of $22\ k\Omega$ wired across the secondary.

Sensitivity of the Discriminator

Measured input voltage v_{in} is 1.66 V pk

So from $$\frac{\delta V_o}{\delta f} = v_{in}\,\frac{2kQ^2}{f_o}$$

Calculate $$\frac{\delta V_o}{\delta f} = \frac{1.66 \times 2 \times 0.1 \times 28^2}{460 \times 10^3}\ \text{V/Hz}$$

$$= \frac{1.66 \times 2 \times 28 \times 28}{460}\ \text{V/Hz}$$

$$\frac{\delta V_o}{\delta f} = 0.056\ \text{V/kHz}$$

The discriminator response is plotted on Figure D3.iii From the graph it is seen that the slope is

4.16 V in 30 kHz i.e. 0.138 V/kHz

Dividing out the measured gain of the amplifier (× 2.55) gives a final value

<u>0.054 V/kHz</u>

which is in good agreement with the design value obtained from the theory presented earlier.

Finally in order to show the linearity of the whole system of modulator-filter-discriminator some active tests are shown in the waveform photographs 81 and 82 page D.3.R.2 .

The photograph No. 81 shows a sinusoidal wave of 4 kHz before and after passage through the equipment. The magnitude of signal level is approximately 4 V pk to pk at the output, with the input wave suitably amplified to give the same sized display magnitude. It may be noted that the output wave is distinguishable only by the 1/6 th wave delay (caused by the filter circuits) and a slight thickening of the trace due to noise.

The photograph No 82 is a 1 kHz triangular wave. The input here being placed above the output trace. Delay is proportionally less. The wave appears to be a faithful reproduction of the input signal.

Part E

FINAL CONCLUSION

Part E Final Conclusion to

Linearity of Some Modulation Circuits

The objective set down at the beginning has been kept in mind at all stages of this work, though a multi-directional approach has been taken. The considerations have included Amplitude and Frequency Modulation and their converse Demodulation processes.

Whilst the direction may have appeared confused, nonetheless it has been held to a pattern. The pattern has constituted the investigation of the present day implementation of each process and the examination thereof to see if there has been achieved the highest possible degree of linear reproduction of the initial signal. In each of the four circuit functions reviewed some points have been found in which present practice could be improved.

In Part A (concerning Amplitude Modulation) the consideration of present methods of modulation (and mixing) revealed that in most cases the design of such circuits has been based largely on empirical methods. Sensitivity has been guessed at. Distortion has been minimised by "recommended procedures" adopted initially as expedients & then handed down without re-examination.

Starting from first principles, a thorough appraisal of the modulation process has been performed and the mismatched modulator is developed and put forward as the best answer to the difficulties of :-

a) modulator gain

b) modulator linearity

c) modulator source interaction

d) modulator l. f. phase response

e) retention of modulator balance

It has been proved that a modulator must be driven with square wave carrier if it is to be linear for deep modulation.

In the realm of amplifier design, negative feedback has usually been relied on to enable non-linearity to be reduced. For communications engineers the use of feedback has been too risky since the uncertainty concerning gain and phase has precluded attempts to enclose amplitude modulators within feedback loops. Perhaps the new circuits suggested will allow designers to contemplate negative feedback in the future.

In Part B (concerning Amplitude Demodulation) it has been explained that the principal distortion of the peak rectifier type of modulator has never been cured. The diagonal clipping has always been hidden by the application of "safety factor" engineering. Since the trouble occurs with deep modulation, for this (and other) reasons deep modulation has always been avoided in broadcasting. Since there is tremendous international demand for allocations of frequency on the AM broadcast radio bands, it has never been possible to transmit frequencies higher than 8 kHz , and therefore the aggravation of the diagonal clipping problem has in practice been prevented.

Analyses are also performed to investigate the linearity possible in the coherent detection of single sideband amplitude modulation with suppressed carrier. Here again commercially there has been little effort directed towards getting good quality, since the microphones and operating circumstances have formed a convenient excuse to accept the present distortion values. It is possible to contend with some justification that operator fatigue could be reduced if "communications quality" were improved.

Convinced that the cure for the problems of <u>modulation</u> would probably be also a cure for the problems of <u>demodulation</u> (for there is much theory to prove that demodulation is a particular kind of modulation), the mismatched modulator has been investigated in the role of demodulator. It has been found to be very useful particularly because the mismatched demodulator eliminates diagonal clipping.

The mismatched modulator and demodulator and variations have been accepted as having some novelty as is shown by the acceptance by the examiners of the Patent Office and the referees of the International Conference "Communications 74" to be held in June 1974 (see Additional Evidence).

In Part C a small amount of new work is given concerning the linearity of elementary frequency modulated oscillators. Of course the professional broadcasting engineering effort has been directed towards the attainment of good linearity in frequency modulation for many years. Thus there are few areas of criticism remaining for research of the type pursued in this project. The items given will be found of interest for small scale systems in which economic considerations demand circuit simplicity together with good performance.

In Part D (concerning Frequency Demodulation) there is a review of the principles of frequency discriminators following an approximately historical sequence, it which it is shown that the Foster-Seeley type is likely to remain the archetype of future open loop forms. It is analysed in detail so that a sensitivity formula is obtained and so that the linearity may be evaluated.

A new approach to the theory of the spectrum of a frequency modulated wave is given. The mathematical derivation can be followed step by step to the full series for the Bessel functions assuming only the series for e^x and Euler's formula. The method has been well received by Honours students over a number of years and was accepted for publication in 1967 (18).

In concluding Part D some figures are given which have been measured during experiments on a demonstration FM Experimental Equipment. The results have given verification to the formulae which have been derived for the sensitivity of both the astable oscillator and of the modified Foster-Seeley discriminator.

Finally it may be said that throughout this work there has been a continuous desire to ensure that the analogue aspect of communications engineering is developed to the full. The draft topic under which the work was originally registered was "The linearity of Modulation Processes". The research work has proceeded closely along these lines and the final title has been made almost the same.

Perhaps it is somewhat ironic that the best analogue modulators discovered have nearly all included square wave phenomena, with their overtones of digital electronic circuitry.

It is hoped that this project will enable communications circuit engineers to have at their command a family of modulator designs for which they may confidently calculate sensitivity, and know that the linearity approaches the best that is possible.

APPENDIX A2.2

Fourier Analysis of Fundamental Component of Chopped Sinusoid by Graphical Integration

Technique

The wave is divided into vertical strips 10° wide and the area standing above the zero line at each strip is multiplied by the sine of the mid angle. Totals of integrals adding energy less totals subtracting energy (due to their occuring when the fundamental is antiphase) up to 9 strips of each kind, are then averaged by division by 9. Thus the equivalent of multiply by 2 and divide by 18 (to take 18 10° strips in half a cycle) is achieved in one step, see attached Figure A2.2.i(graph).

Signal Sine Magnitude

Carrier Sine	.05	.10	.20	.30	.40	.50	INPUT
5°						.007	
15°					.015	.067	
25°				.017	.097	.180	
35°				.097	.218	.330	
45°			.162	.220	.361	.500	
55°		.016	.180	.344	.517	.672	
65°		.096	.281	.463	.644	.823	
75°	.067	.164	.363	.542	.686	.937	
85°	.096	.199	.398	.589	.847	.992	
Total	.163	.475	1.384	2.270	3.385	4.508	
÷9 X10	.181	.528	1.54	2.525	3.76	5.00	
÷10	.018	.053	.154	.253	.376	.500	← PLOT OUTPUT

All adding energy

Contd Appendix A2.2

Signal Sine Magnitude

Carrier Sine	0.60	0.70	0.80	0.90	0.95	1.00
5°	.025	.042	.059	.077	.086	.095
15°	.119	.170	.212	.274	.300	.326
25°	.264	.348	.434	.518	.560	.603
35°	.444	.560	.674	.784	.847	.905
45°	.642	.784	.927	1.067	1.140	1.208
55°	.837	1.000	1.165	1.327	1.410	1.492
65°	1.005	1.185	1.368	1.547	1.640	1.730
75°	1.128	1.320	1.512	1.707	1.805	1.900
85°	1.190	1.395	1.590	1.790	1.888	1.988
Total	5.656	6.804	7.941	9.090	9.676	10.247

INPUT

From these subtract the following

Sine -5°	.060	.026	.048	.060	.071	.078
-15°		.052	.093	.161	.192	.221
-25°			.085	.169	.212	.250
-35°			.017	.126	.242	.299
-45°				.071	.141	.212
-55°					.074	.143
-65°						.082
-75°						.024
-85°						.008
Total	.060	.078	.243	.589	.832	1.317

Giving the following net values

Net	5.596	6.726	7.698	8.504	8.844	8.930
÷9 X10	6.22	7.48	8.55	9.45	9.73	9.94
÷10	.622	.748	.855	.945	.973	.994

← PLOT OUTPUT

Result is plotted Figure A2.2.ii

APPENDIX A2.2.1

NOTE on QUALITY OF OSCILLOGRAMS 78 81 & 82

Oscillograms 78 81 & 82 pages A2.2.1.R & D3.R.2 have been reproduced by Xerography from the negative of the Polaroid instant photographs exposed on the CRO camera.

It has been found that the Xerox machine cannot reproduce oscillograms of the proper positive form. The machine having been designed for reproducing print on white background becomes inhibited by a thin white trace on a black background. However the negative side of the Polaroid film pack which is usually thrown away contains a black trace on white background. Unfortunately the silver halides of the white background quickly darken in the ambient room light. The author preserves these by peeling the negative away in a situation away from strong light and places immediately in a bath of acid hypo fixer. They can then be clipped free of the chemical sachets washed, dried, and stored in a place away from direct light until needed for Xerographic reproduction.

The only remaining snag is that the time base is shown in the reverse direction. As a reminder the author marks his prints with index mark in the top right corner (instead of top left with normal prints).

APPENDIX B2.1.1

Experiments on Loading Effect of Peak Rectifying Demodulators

Taking three typical diodes:-

Germanium AAY 11

Silicon BAX 13 Q Meter Marconi TF 1245

Si Schottky H-P 2800 Lab No. 605/R

Experiments were performed at a series of frequencies in the HF region with two different values of load resistor.

The by-pass capacitor was maintained the same 100 pF, polystyrene.

From measurement of the Q of a tuned circuit before and after attachment of the diode and its load, it was possible to calculate the actual loading caused by the peak rectifying modulator so formed. Inductor $L = 1.0\,\mu H$

Test 1 AAY 11 with 4.7 k ohm load V_f taken as 0.2 V

Freq MHz	X Ω	Qu	Gu μS	ΔC pF	Qd	Gtot μS	Vrf	Gnet μS	Gsub μS	Gf μS	Notes
29	182	320	17	-1.8	13.7	400	.13	383	167	216	
23	145	348	19.8	-1.8	18.8	367	.19	347	207	140	
18	113	325	27	-1.4	24.5	346	.30	319	255	63	
15	94	305	35	-1.6	29	367	.39	332	281	51	
12	75	282	47	-1.6	36	372	.52	324	307	17	
9	56	248	72	-1.4	46	386	.72	313	333	-20	
7.4	47	225	95	-2.4	52	413	.86	308	345	-37	
7.4(½V)	47	232	95	-2.0	59	365	.41	270	286	-16	Check

Symbols:- X = reactance in ohms Qu = Q of unloaded cct

Vrf = RF Voltage Qd = Q when diode on

Gu = Conductance of unloaded cct, Gtot = Conductance loaded cct.

Gnet = Difference between above.

Gsub = Conductance subtracted by forwards voltage adjustment.

Gf = Conductance attributable to Storage Loss.

Appendix B2.1.1 Contd.

Test 2 AAY 11 with 30 k ohm load V_f taken as 0.2 V

Freq	X	Qu	Gu	C	Qd	Gtot	Vrf	Gnet	Gsub	Gf	Notes
29	183	320	17	-2.4	32	172	.57	154	49	106	
23	145	348	20	-0.5	48	142	.95	122	55	67	
18	113	325	27	-0.6	68	129	1.36	102	58	44	
15	94	303	35	-0.8	82	129	1.75	95	60	34	
12	75	280	47	-0.6	135	135	2.16	87	61	26	
9	56	245	72	-0.8	115	154	2.61	82	62	20	
7.4	46	225	95	-1.0	123	175	2.82	80	67	14	
7.4($\frac{1}{2}$V)	46	225	95	-1.0	124	175	1.33	80	58	22	Check

Test 3 BAX 13 with 4.7 k ohm load V_f taken as 0.7 V

Freq	X	Qu	Gu	C	Qd	Gtot	Vrf	Gnet	Gsub	Gf	Notes
29	183	324	17	-1.8	27	202	.18	185	87	97	
23	145	348	20	-1.8	32	215	.29	194	125	70	
18	113	330	27	-1.8	38	235	.41	207	157	50	
15	94	306	33	-2.2	43	249	.52	216	184	32	
12	75	284	47	-3.0	50	266	.67	220	208	12	
9	57	248	71	-3.4	58	305	.88	236	237	2	
7.4	47	225	96	-3.6	65	321	1.04	225	254	-25	
7.4($\frac{1}{2}$V)	47	232	96	-3.6	80	267	.46	172	169	-3	Check

Test 4 BAX 13 with 30 k ohm load V_f taken as 0.5 V

Freq	X	Qu	Gu	C	Qd	Gtot	Vrf	Gnet	Gsub	Gf	Notes
29	183	322	17	-1.8	54	101	.95	85	44	41	
23	145	338	20	-1.8	72	96	1.39	76	49	27	
18	113	324	27	-1.4	92	97	1.92	70	53	17	
15	94	312	33	-1.2	106	100	2.31	67	55	12	
12	75	282	47	-1.2	121	110	2.70	63	56	7	
9	57	250	71	-1.4	135	131	3.08	60	57	3	
7.4	47	225	96	-1.5	139	155	3.20	60	58	2	
7.4($\frac{1}{2}$V)	47	230	94	-1.5	150	144	1.47	50	50	0	Check

Test 5 H-P 2800 with 4.7 k ohm load V_f taken as 0.7 V

Freq	X	Qu	Gu	C	Qd	Gtot	Vrf	Gnet	Gsub	Gf	Notes
29	183	324	17	-1.8	27	202	.18	185	87	98	
23	145	348	20	-1.8	32	215	.29	194	125	69	
18	113	330	27	-1.8	38	235	.41	207	157	50	
15	94	306	33	-2.2	42.6	249	.52	216	184	32	
12	75	284	47	-3.0	50	268	.67	220	208	12	
9	57	248	71	-3.4	58	305	.88	236	237	0	
7.4	47	225	96	-3.6	65	321	1.04	226	244	-22	
7.4($\frac{1}{2}$V)	47	232	96	-3.6	81	268	.46	172	169	3	Check

Appendix B2.1.1 Contd.

Test 6 H-P 2800 with 30 k ohm load V_f taken as 0.6 V

Freq	X	Qu	Gu	C	Qd	Gtot	Vrf	Gnet	Gsub	Gf	Notes
29	183	325	17	-1.5	23	240	?	223	?	?	
23	145	350	20	-1.4	28	246	0.4	225	170	55	
18	113	332	27	-1.4	33	270	0.54	242	201	41	
15	94	305	33	-1.7	37	286	0.64	253	219	34	
12	75	282	47	-1.8	43	310	0.78	263	240	22	
9	57	245	71	-1.8	52	340	0.99	270	264	6	
7.4	47	228	96	-2.0	58	371	1.14	275	278	-3	
7.4($\frac{1}{2}$V)	47	232	96	-2.0	67	321	0.56	225	224	1	Check

APPENDIX D2. 2. 1.

Table D2. 2. 1. T

Foster-Seeley Discriminator

Output voltage after d. c. amplifier with Gain X 2. 55

Measurements made at d. c. output terminal on AVO 8,

Frequency on Dawe digital frequency counter.

Volts		Frequency kHz
5. 00		466. 4
5. 50	UP	470. 0
6. 00		473. 7
6. 50		477. 4
7. 00		481. 4
7. 50		485. 8
8. 00		490. 7
8. 50	TURN	497. 7
8. 25	DOWN	493. 1
7. 75	AT INTERMEDIATE	487. 5
7. 25	VALUES	483. 1
6. 75		479. 0
6. 25		475. 2
5. 75		471. 6
5. 25		468. 1
4. 75		464. 5
4. 25		460. 9
3. 75		457. 5
3. 25		453. 9
2. 75		450. 4

DOWN CONTD

Routine established for check of both randomness in reading and of drift with time. The start is made in mid scale, moving to top return via intermediate values to bottom turn round & up to centre finishing on the starting value. See graph Figure D2. 2. 1. iii

Contd:-

Volts		Frequency kHz	
2. 25	DOWN	446. 4	
1. 75	TURN	441. 86	
1. 50	UP	439. 0	
2. 00		444. 1	
2. 50		448. 3	
3. 00		452. 1	
3. 50		455. 9	
4. 00		459. 5	
4. 50		463. 0	
5. 00	FINISH AT STARTING VALUE	466. 7	DRIFT ERROR ONLY 0.3 kHz

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Communications Lab., R.G.I.T., 1972

(2) V. Belevich — Corres.
Wireless Engineer 1949 Vol 26 p 177

(3) " — Corres.
ibid 1950 Vol 27 p 130

(4) " — Corres
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(17) " — Single Sideband Modulation using Sequence Asymmetric Polyphase Networks
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(18) M.C.Hately Frequency Modulation & Bessel Functions
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(20) " Mismatched Modems
British Patent No 1,339,608 Dec 1973
Application Date 23 April 1971
Royal Seal Dated 3 April 1974

(21) E.Hood Video Electronics for a Weather Satellite
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(22) D.P.Howson & D.G.Tucker Rectifier Modulators with
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(23) ITT Corpn. "Radio Engineers Reference Book"
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ADDITIONAL EVIDENCE

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" Part II May 1962

" Part III June 1962

Journal of Radio Soc. of Gt Britain.
(Awarded Bevan Swift Memorial Premium for these papers)

2) Frequency Modulation and Bessel Functions

Internat.Journal of Elec.Eng.Education 1967 Vol 5 p 459-470

3) A Converter-Reverter Chart for Impedance Calculations

ibid 1971 Vol 9 p 368-369

5) Mismatched Modems

British Patent No. 1,339,608 Dec 1973

6) Paper Accepted for Presentation at

International Conference " Communications 74"

at Brighton June 1974

Title "Mismatched Diode Modulator-Demodulator"

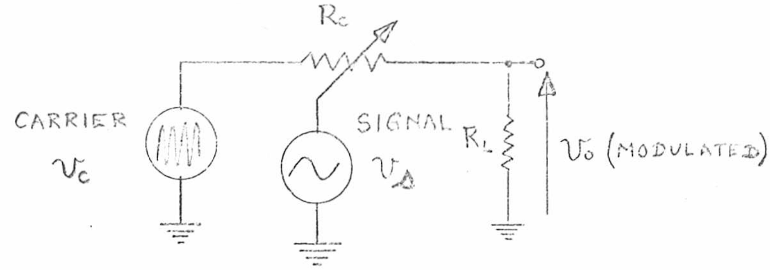


FIGURE A1.1.1

RESISTIVE MODULATION

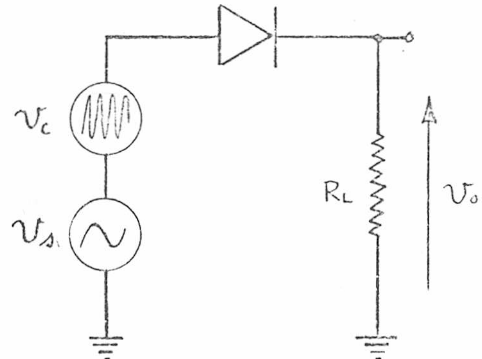


FIGURE A1.1.2.i

DIODE MODULATOR

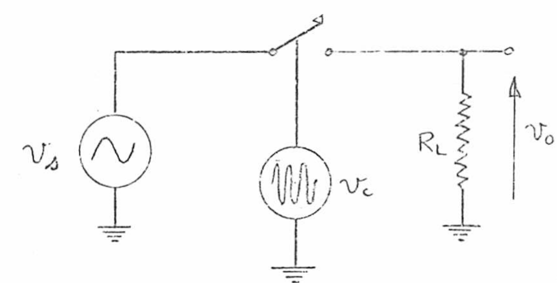


FIGURE A1.1.2.ii

SWITCH MODULATOR

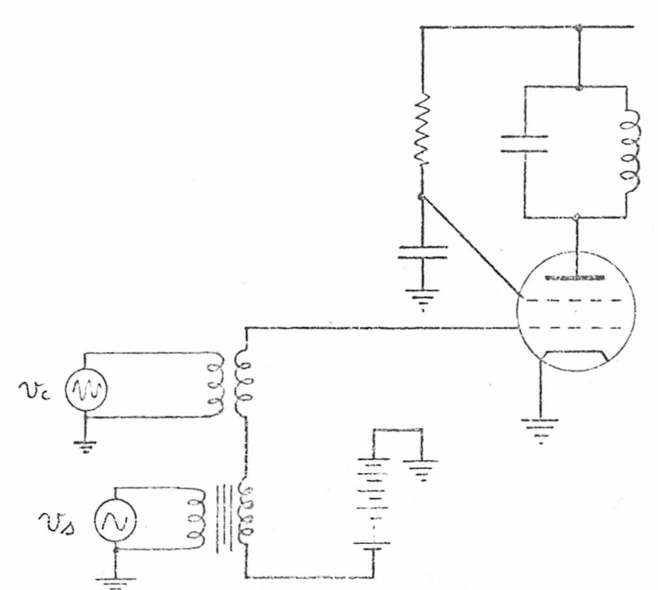


FIGURE A1.1.2.iii

GRID MODULATED VALVE

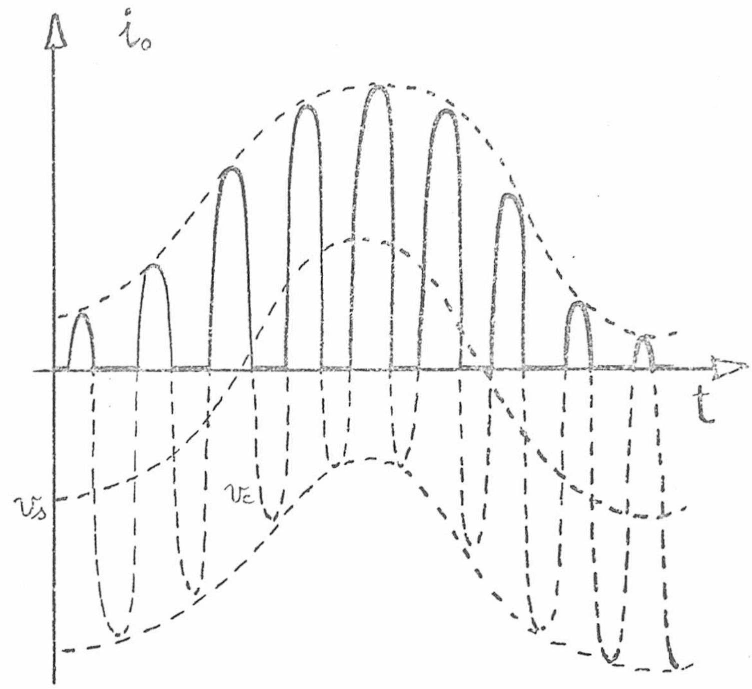


FIGURE A1.1.2.iv

NON-LINEARITY MODULATION

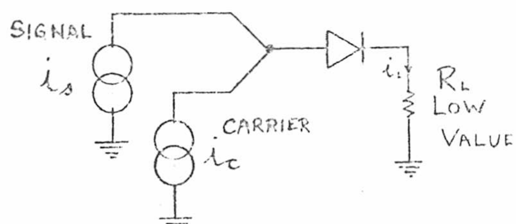


FIGURE A2.1.1.i

MISMATCHED MODULATOR

(BASIC)

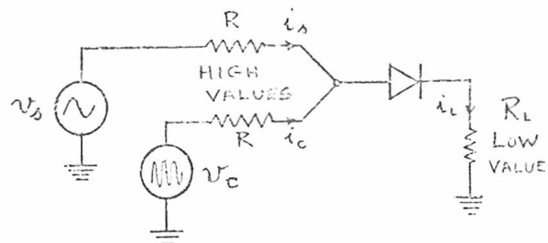


FIGURE A2.1.1.ii

MISMATCHED MODULATOR

(VOLTAGE DRIVE)

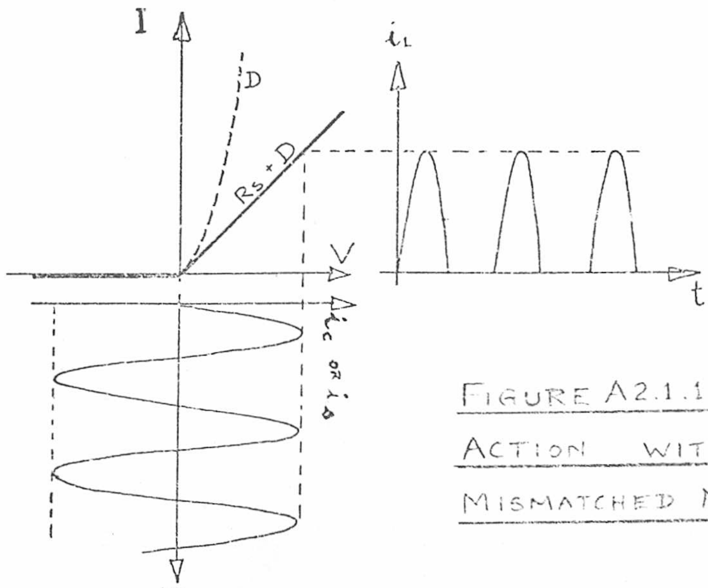


FIGURE A2.1.1.iii

ACTION WITHIN

MISMATCHED MODULATOR

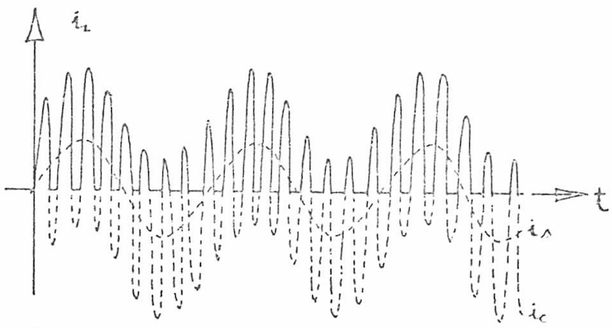


FIGURE A2.1.1.iv

CURRENT SUM IN LOAD R_L

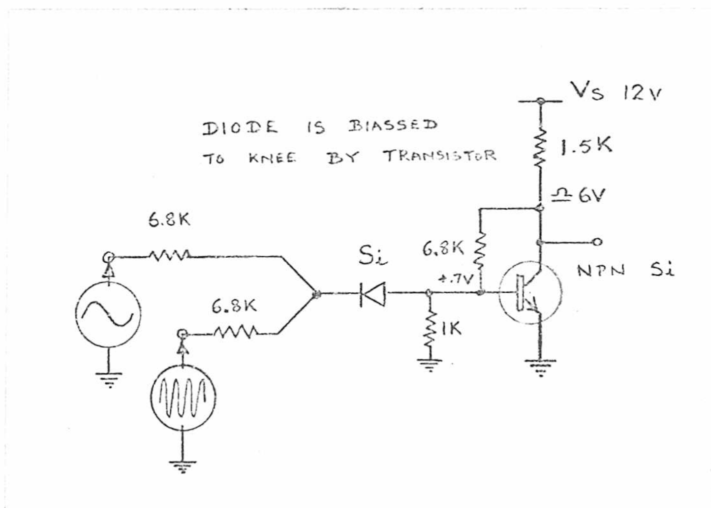


FIGURE A2.1.1.v

SHUNT FED-BACK TRANSISTOR AS R_L

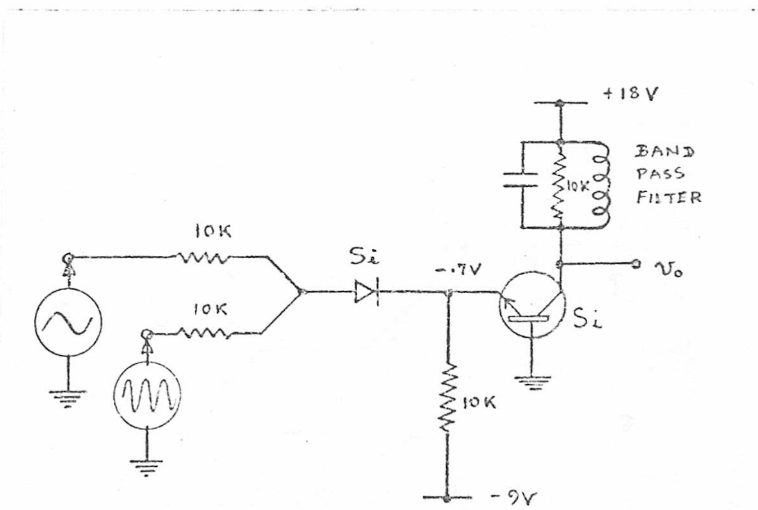


FIGURE A2.1.1.vi

COMMON BASE TRANSISTOR AS R_L

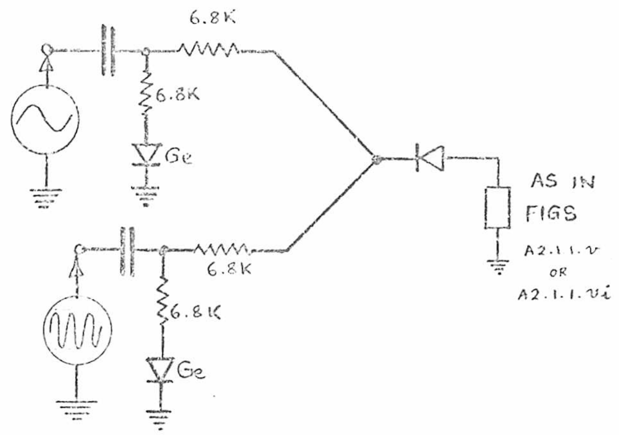


FIGURE A2.1.1.vii

D.C. COUNTERBALANCED SOURCES

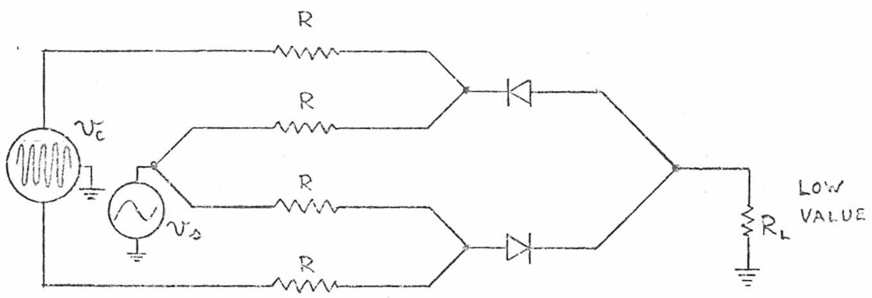


FIGURE A2.1.2.i

BALANCED MISMATCHED MODULATOR

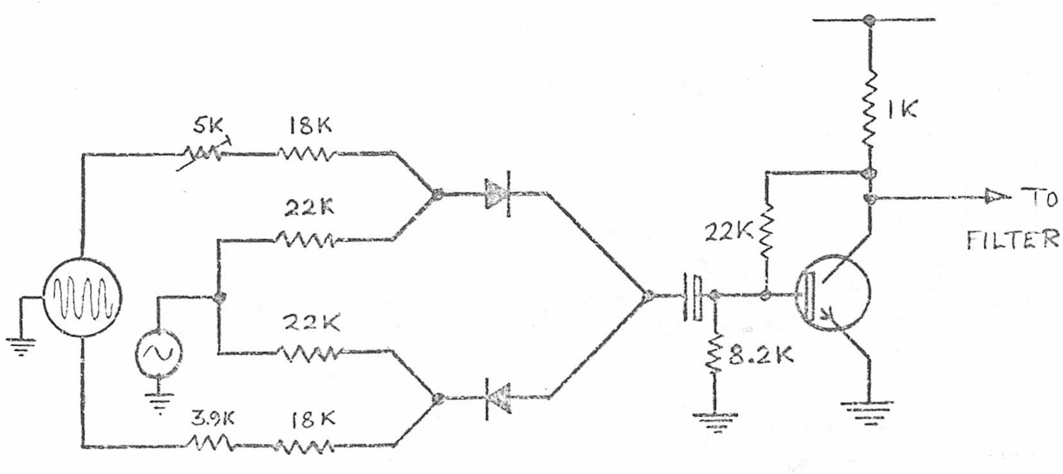


FIGURE A2.1.2.ii

BALANCED MISMATCHED MODULATOR

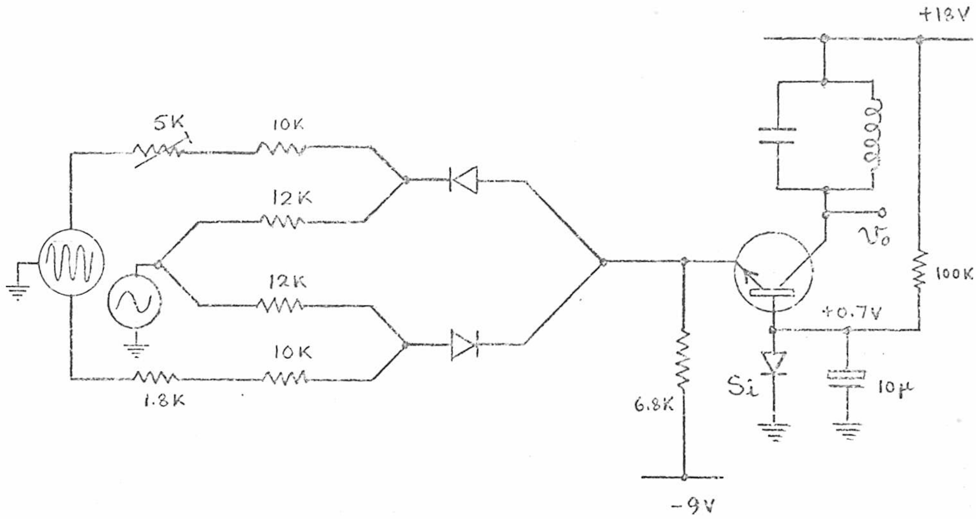


FIGURE A2.1.2.iii

D.C. COUPLED BALANCED MISMATCHED MODULATOR

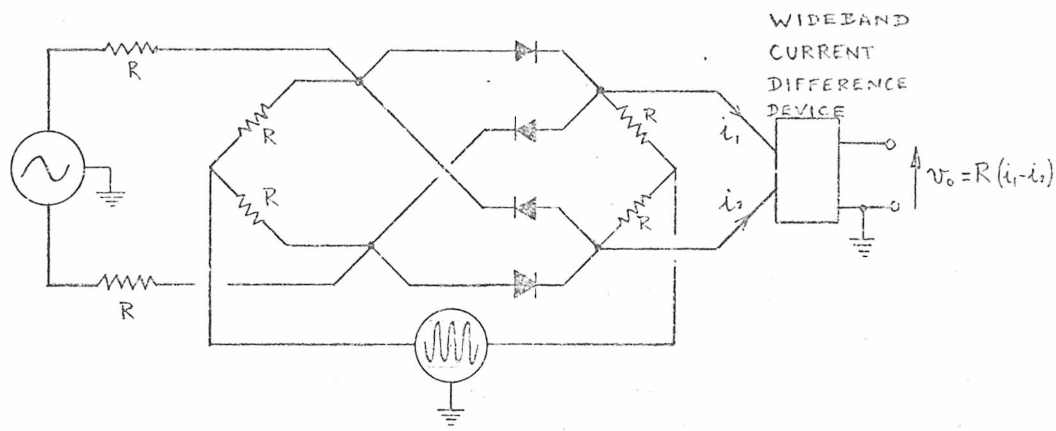


FIGURE A2.1.3 i

DOUBLY BALANCED MISMATCHED MODULATOR

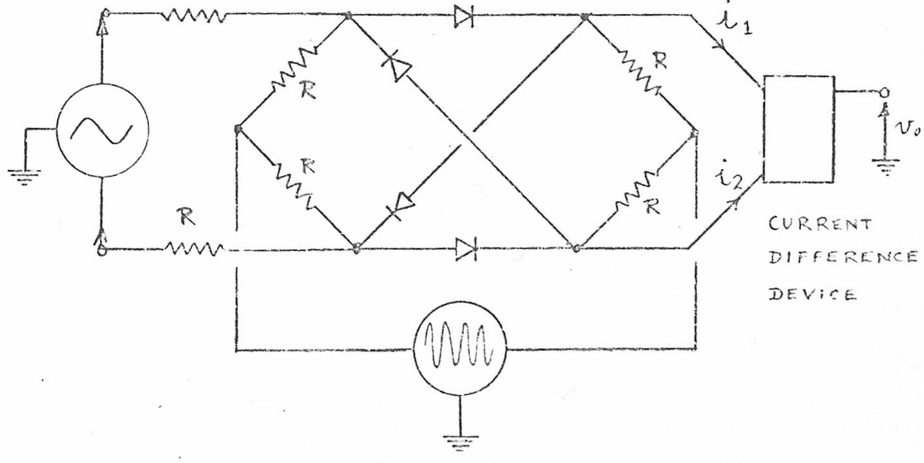
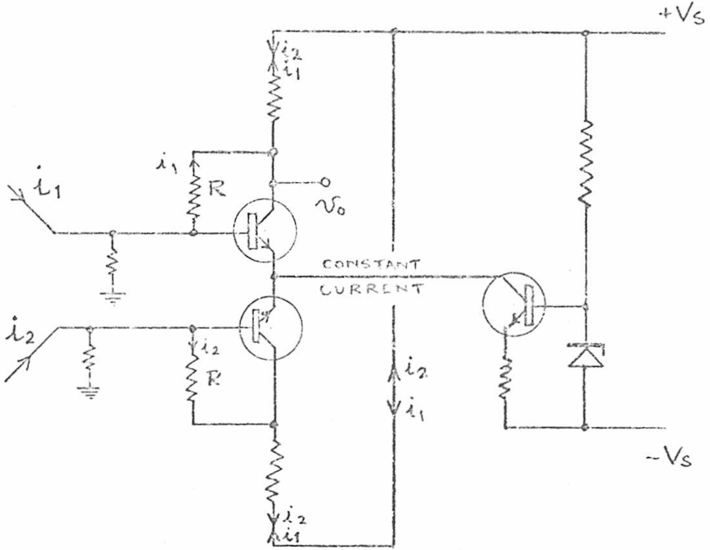


FIGURE A2.1.3.ii

DOUBLY BALANCED MISMATCHED MODULATOR



+Vs
-Vs
R
R
CONSTANT
CURRENT

FIGURE A2.1.3.iii

EMITTER COUPLED SHUNT FEDBACK PAIR

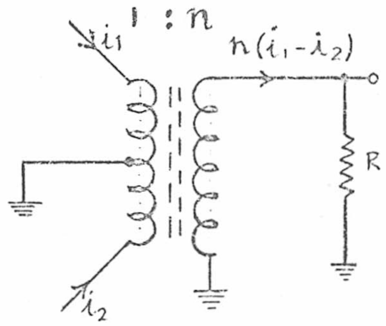


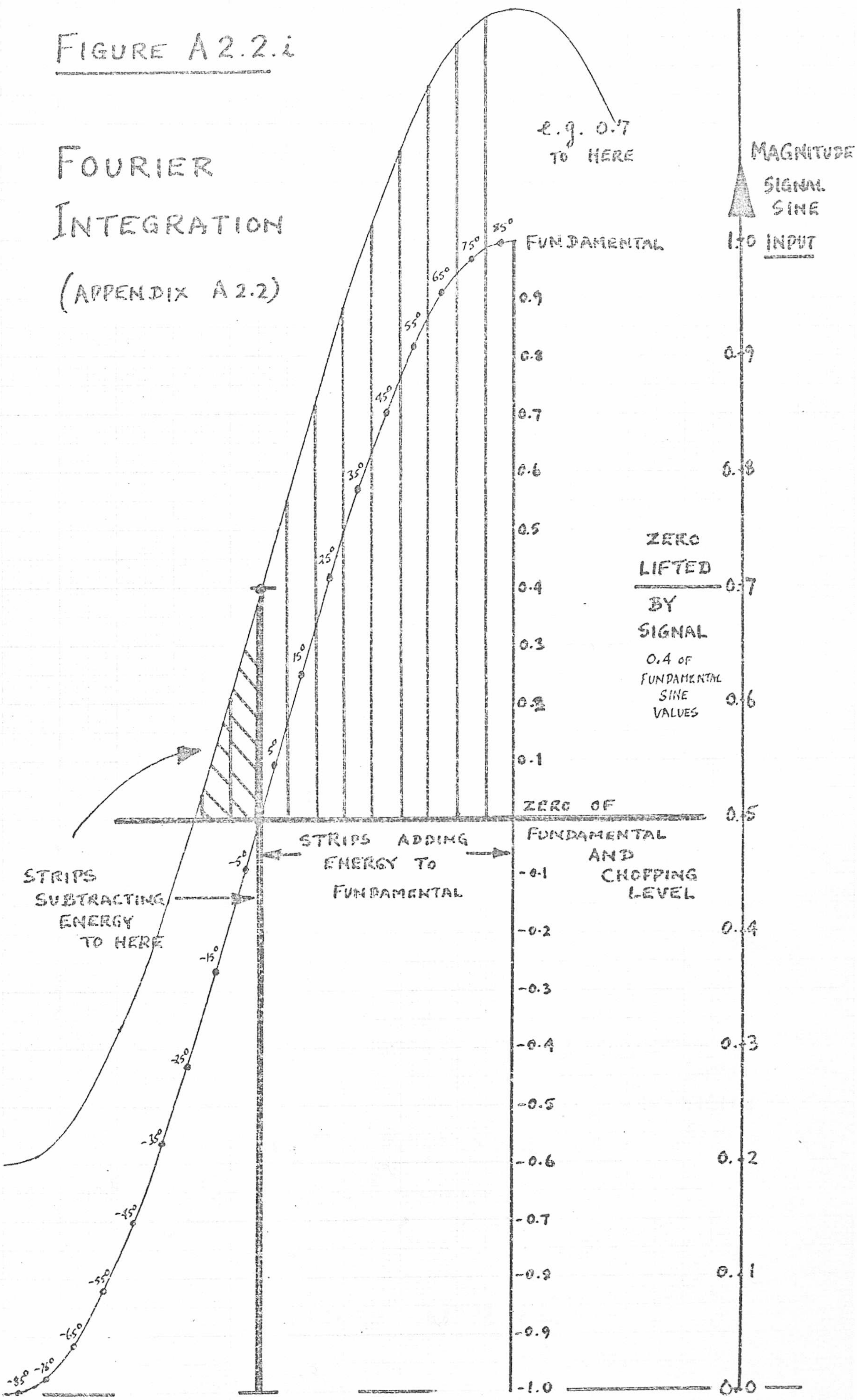
FIGURE A2.1.3.iv

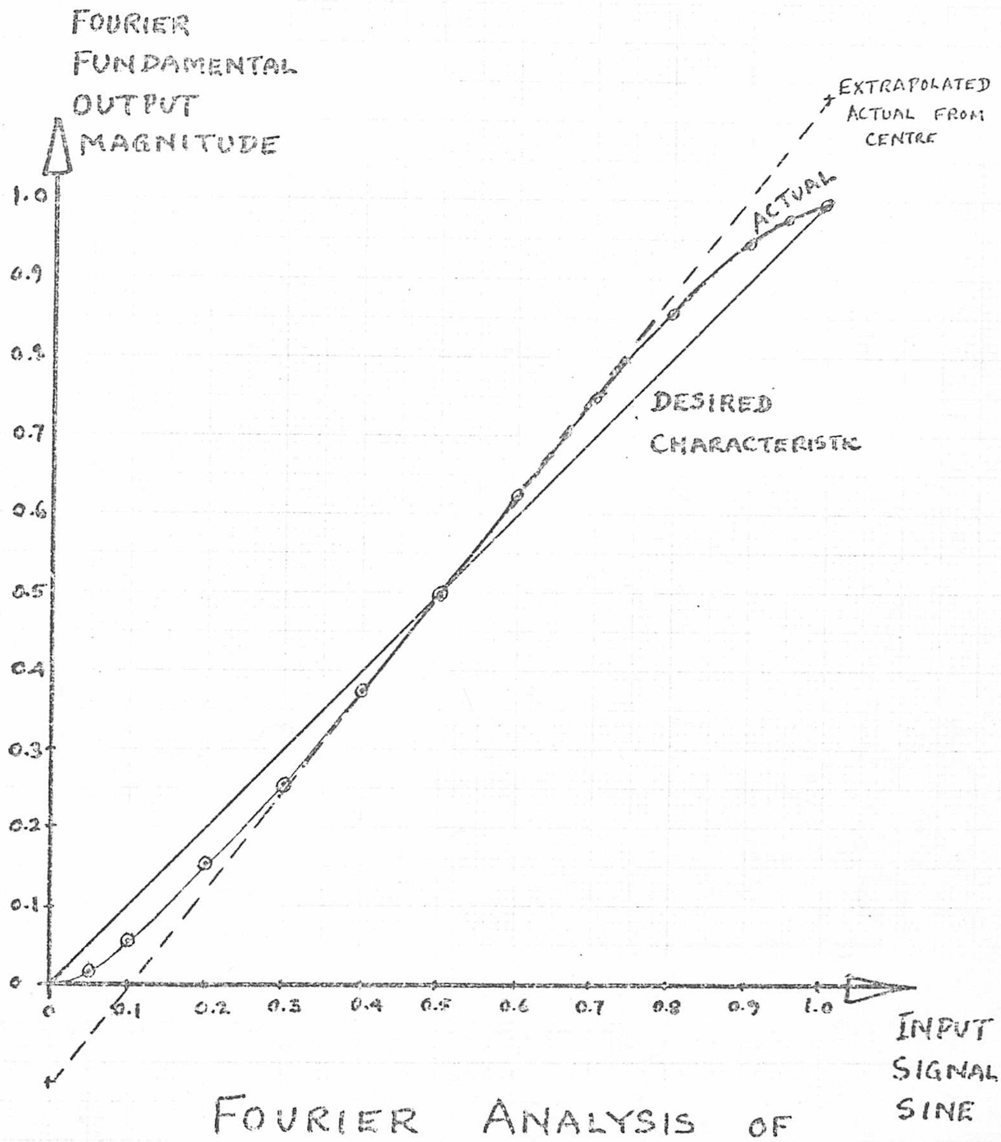
CURRENT DIFFERENCE TRANSFORMER

FIGURE A2.2.i

FOURIER INTEGRATION

(APPENDIX A2.2)





FOURIER ANALYSIS OF CHOPPED SINUSOID

FIGURE A2.2.ii.

Third Harmonic Estimation:-

Sine wave centred at 0.5, of peak amplitude 0.625 output will be reduced to 0.50 output. So $\frac{\text{Third}}{\text{Fundamental}} = \frac{.125}{.625} = 0.2$

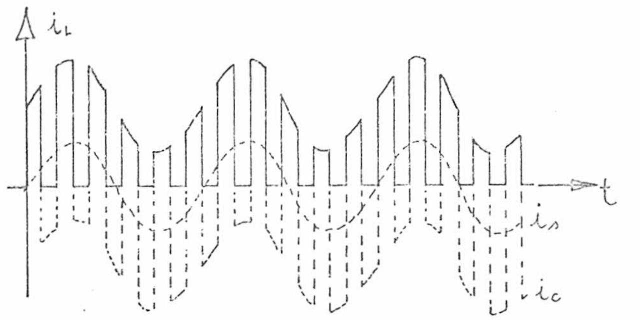
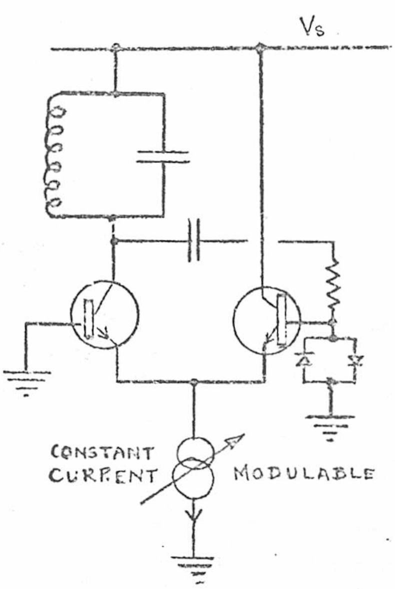


FIGURE A2.2.iii
SQUARE WAVE CARRIER MODULATION

EXPT. AMPLITUDE MODULATION

(i) Diagram of Basic Circuit



(ii) Detailed Circuit

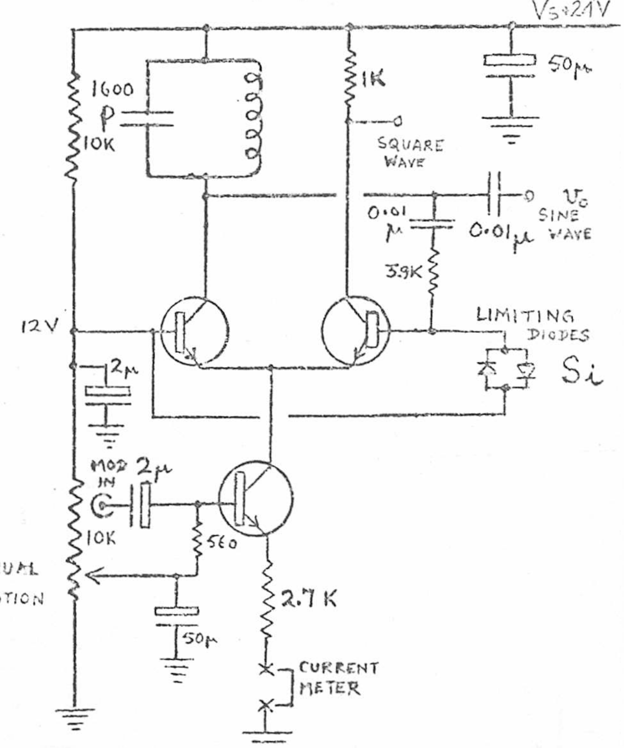


FIGURE A2.2.iv
ADAPTED FOSS & SIZMUR MODULATED OSCILLATOR

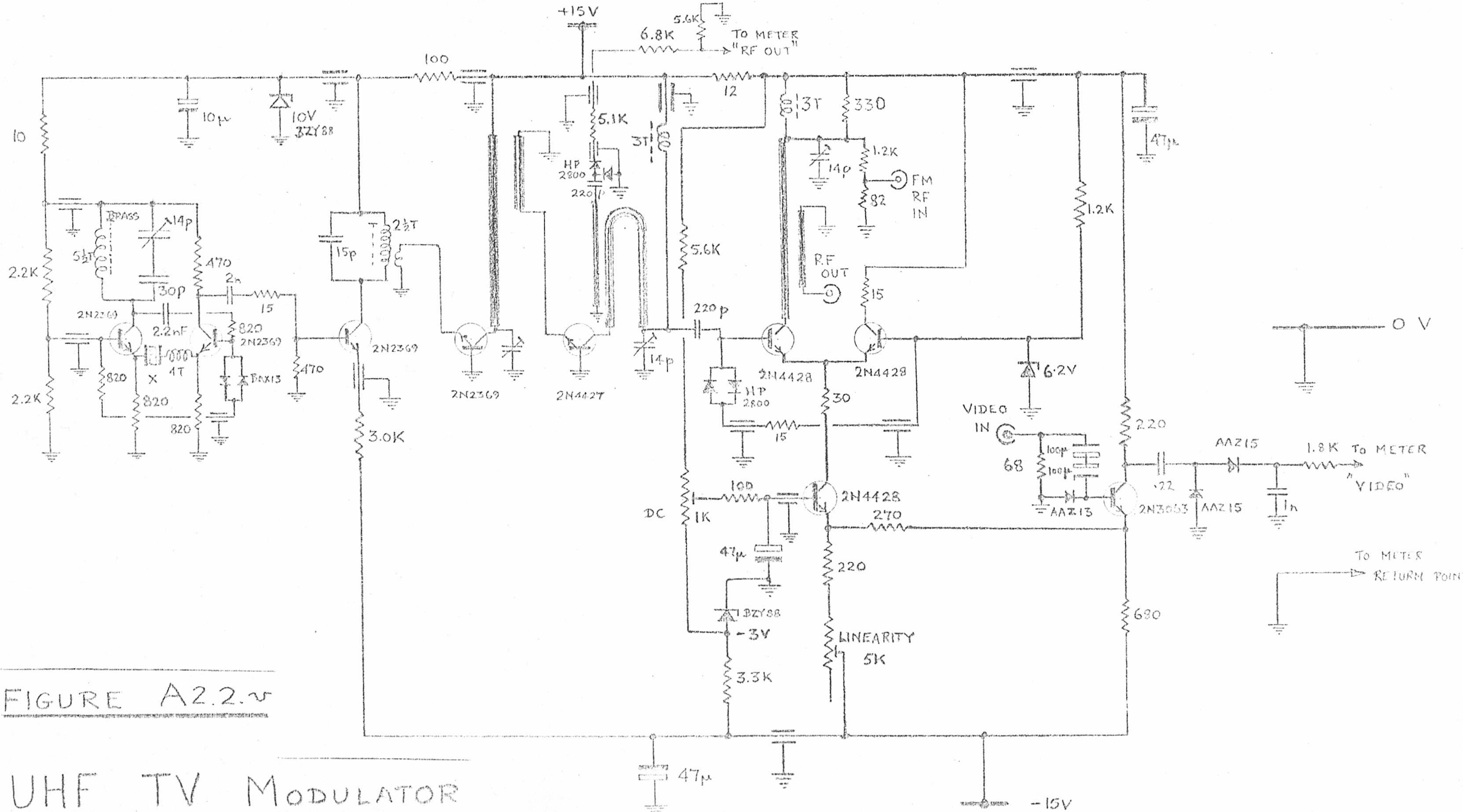


FIGURE A2.2.v

UHF TV MODULATOR

ALL FEED THROUGHS ARE 1nF

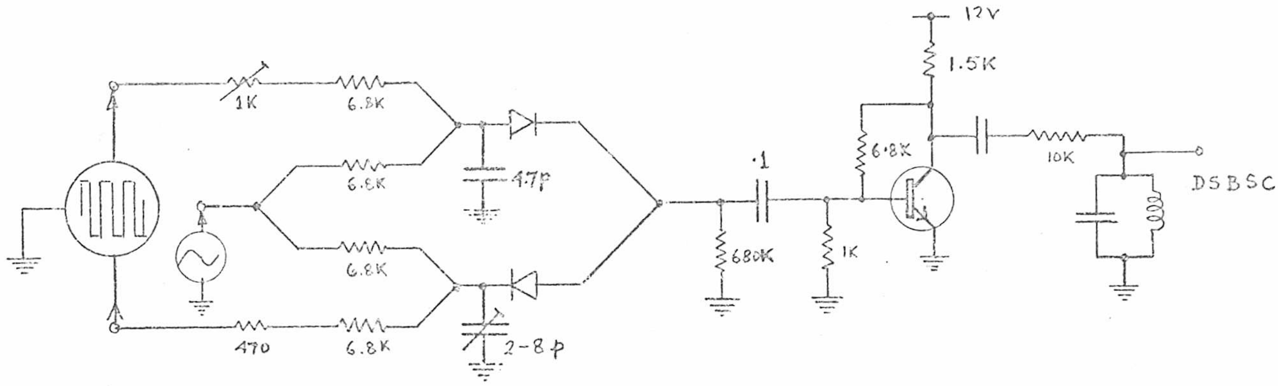


FIGURE A2.2.2.i

BALANCED MISMATCHED MODULATOR WITH CAPACITIVE BALANCE

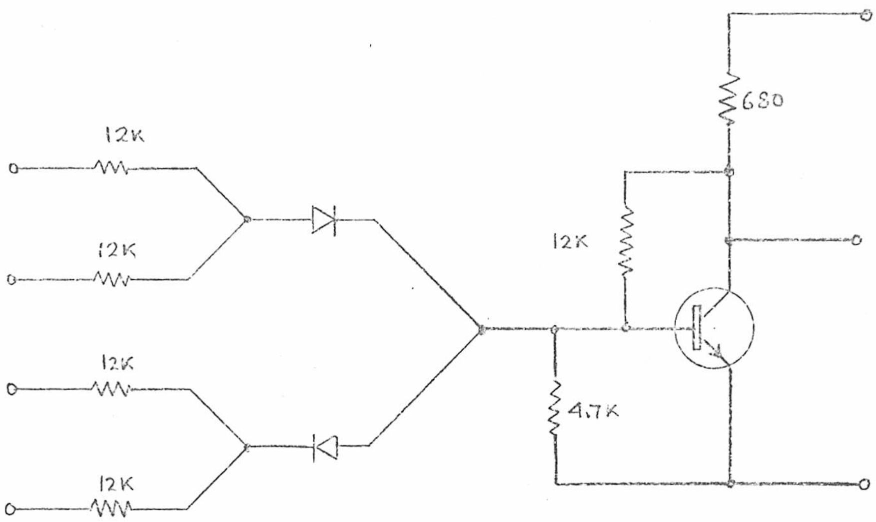
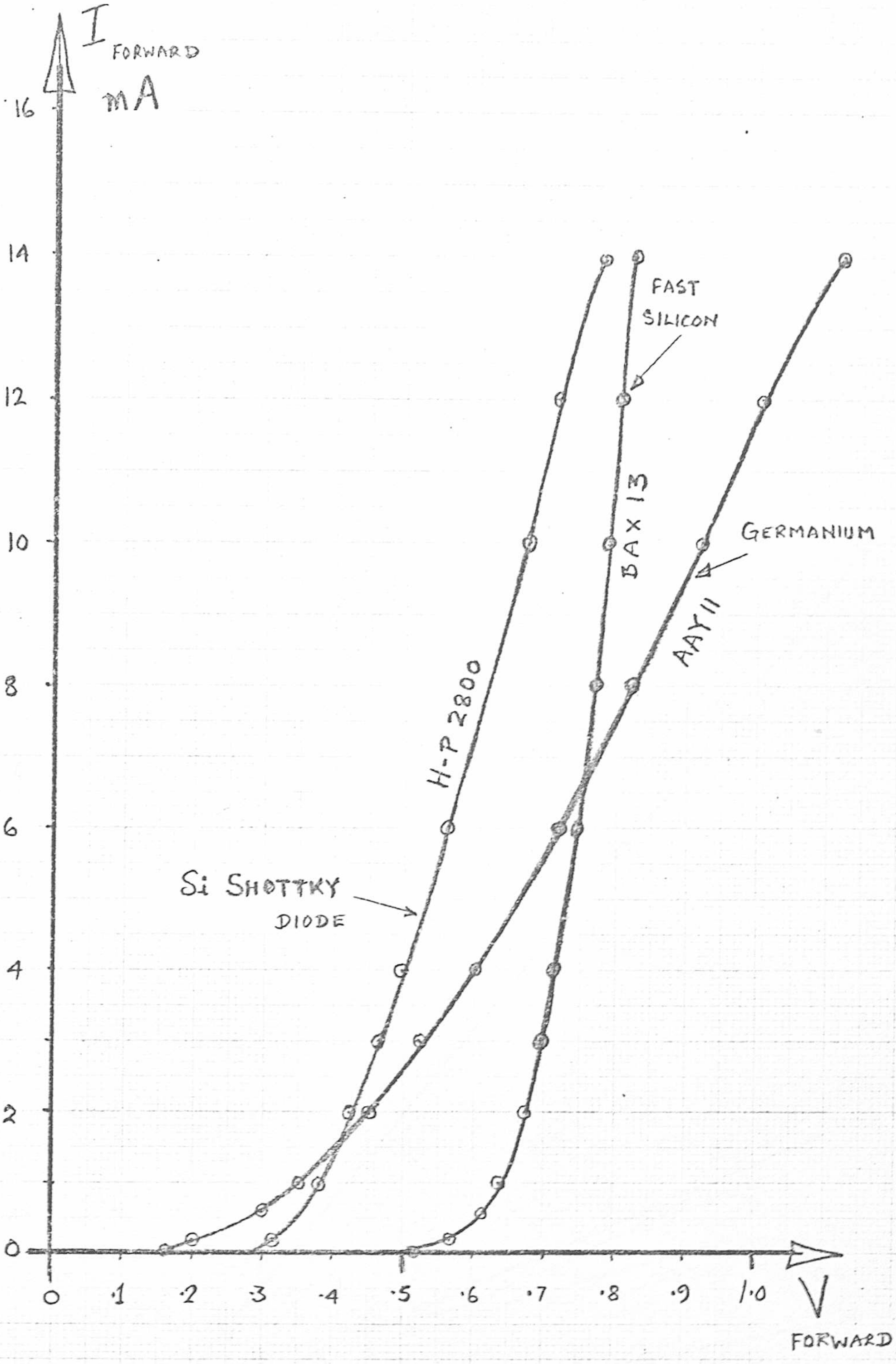


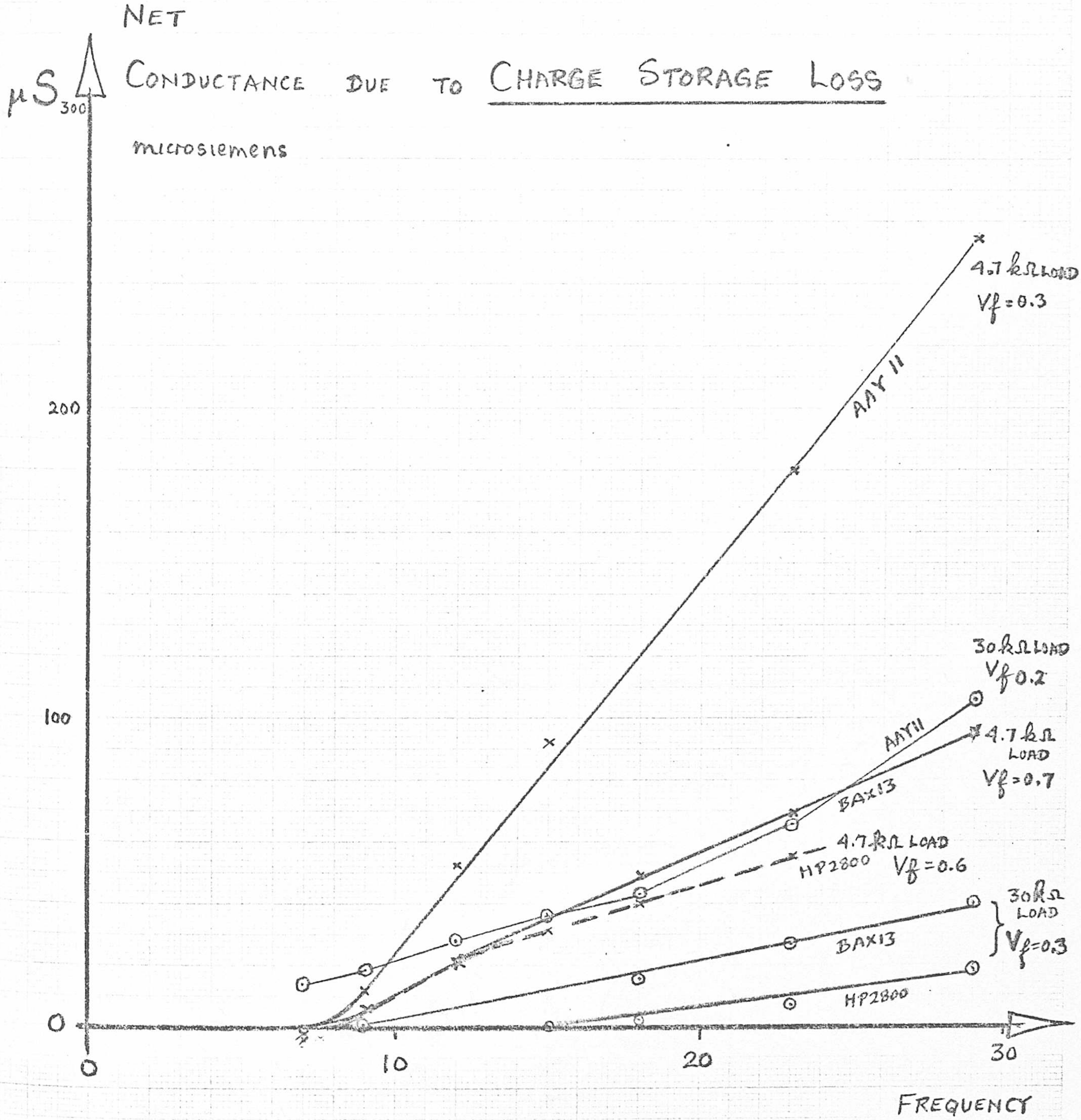
Figure A3.i

I/C Form of Balanced Mismatched Modulator or Demodulator



D·C CHARACTERISTICS OF DIODES

APPENDIX B2.1.1 GRAPH 1



APPENDIX B2.1.1. GRAPH 2

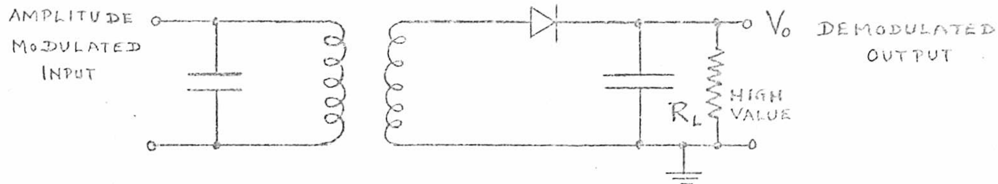


FIGURE B 1.1.2.i

PEAK RECTIFICATION DEMODULATOR

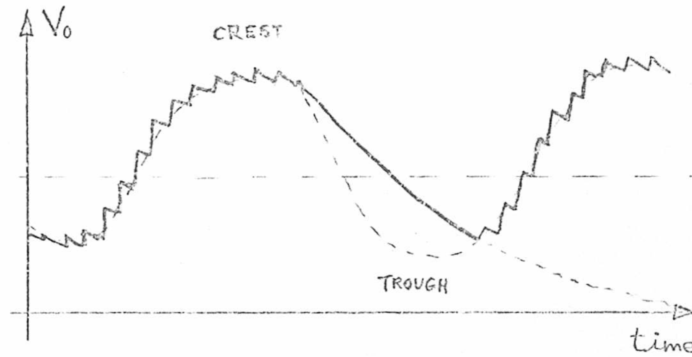


FIGURE B 1.2.i

DIAGONAL CLIPPING

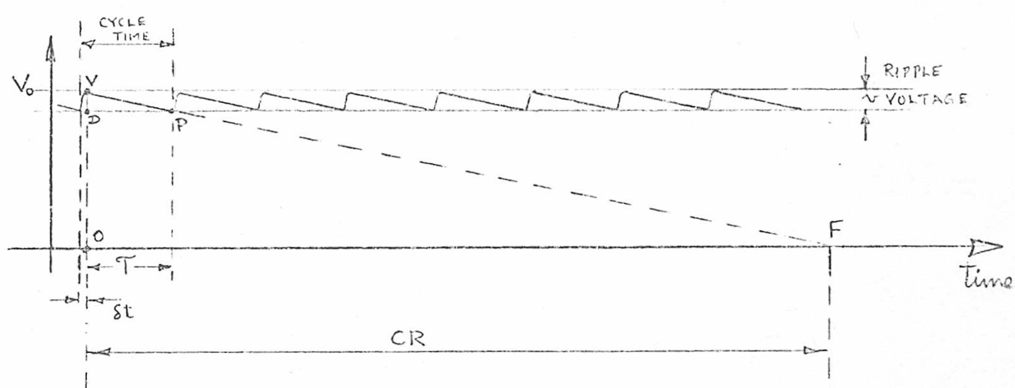


FIGURE B2.1.1.i

SMOOTHING BY CAPACITOR

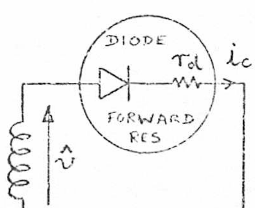


FIGURE B2.1.1.ii

CHARGING CIRCUIT

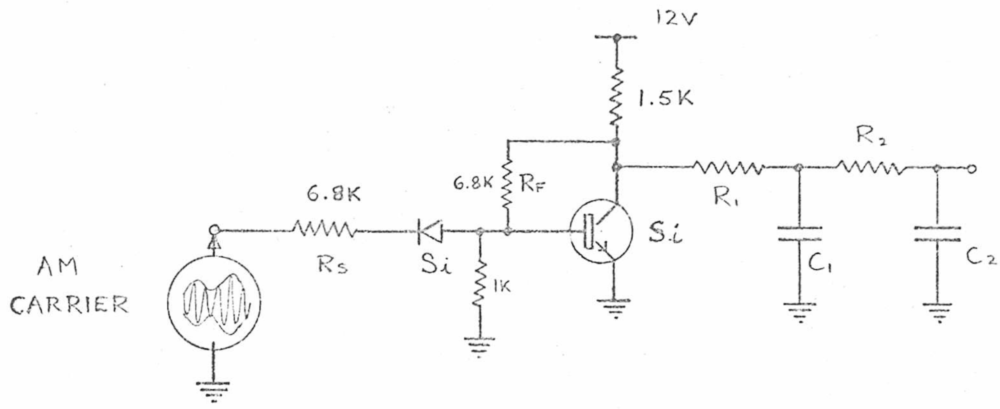


FIGURE B2.1.2.i

SINGLE-ENDED MISMATCHED DEMODULATOR

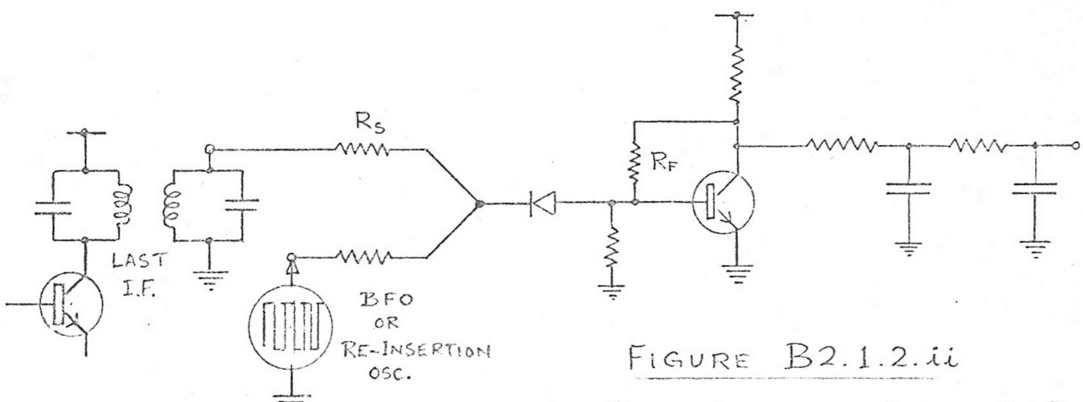
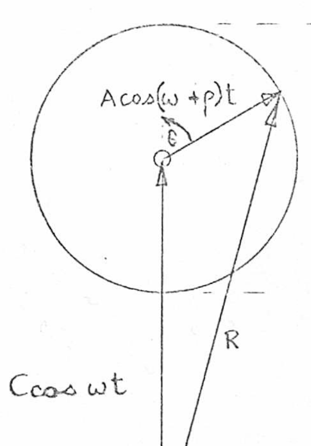


FIGURE B2.1.2.ii

COHERENT DETECTION FOR CW OR SSB



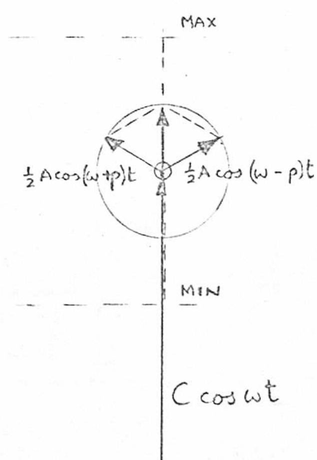


FIGURE B2.1.2.iii

SSB DEMODULATION

FIGURE B2.1.2.iv

NORMAL AM

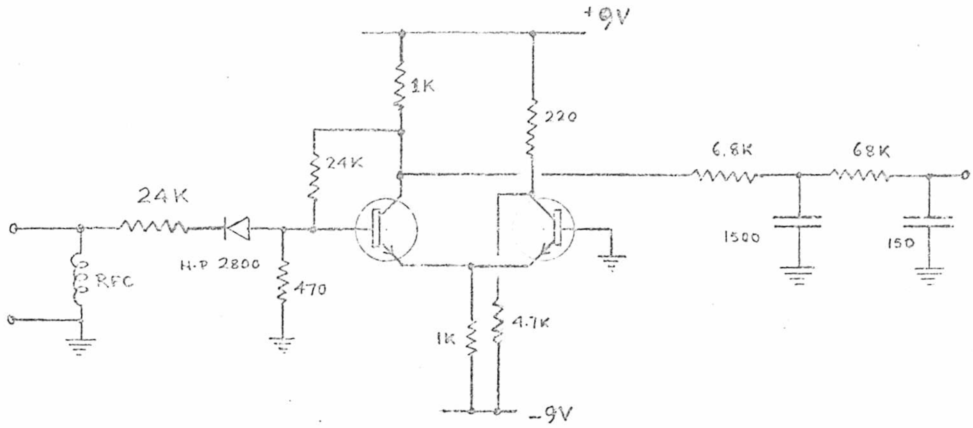


FIGURE B2.1.2.v

SINGLE-ENDED MISMATCHED DEMODULATION TEST CIRCUIT

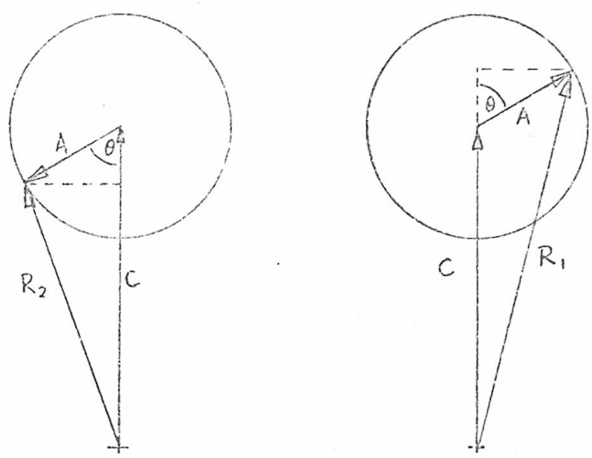


FIGURE B2.2.1.i

BALANCED SSB DEMODULATION

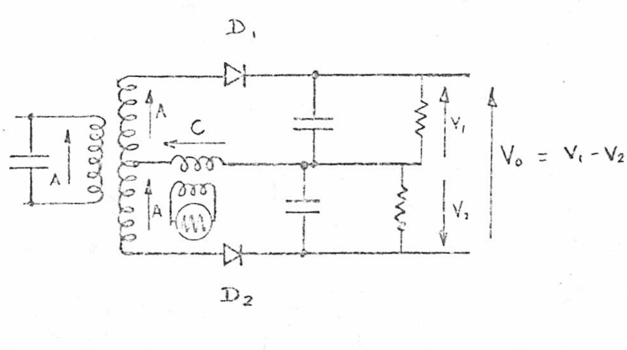


FIGURE B2.2.1.ii

OLD STYLE BALANCED DEMODULATOR

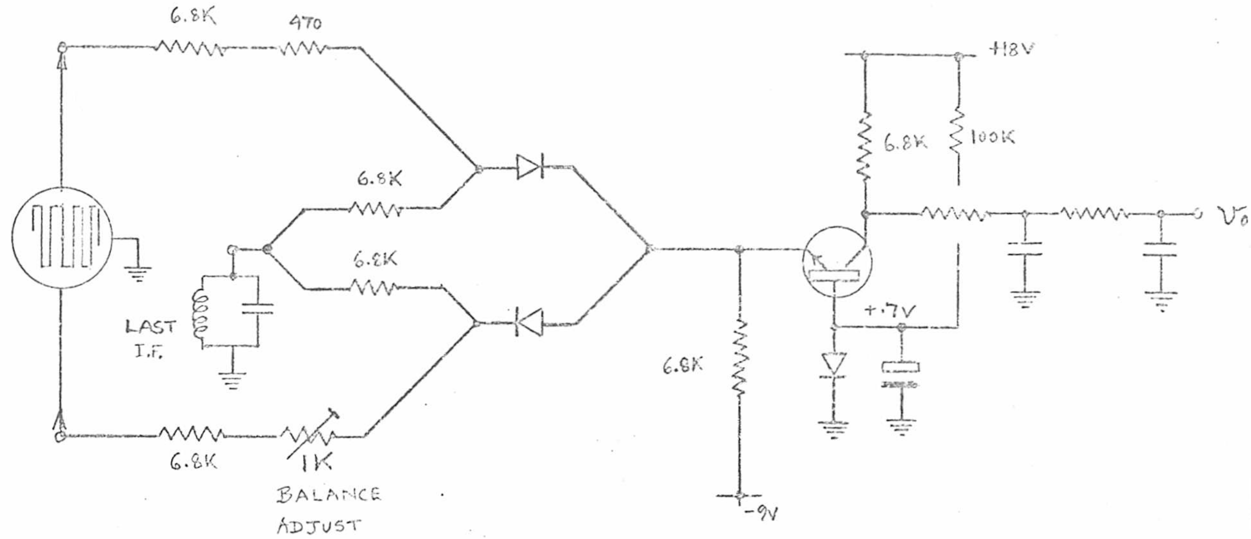


FIGURE B2.2.1.iii

BALANCED MISMATCHED DEMODULATOR

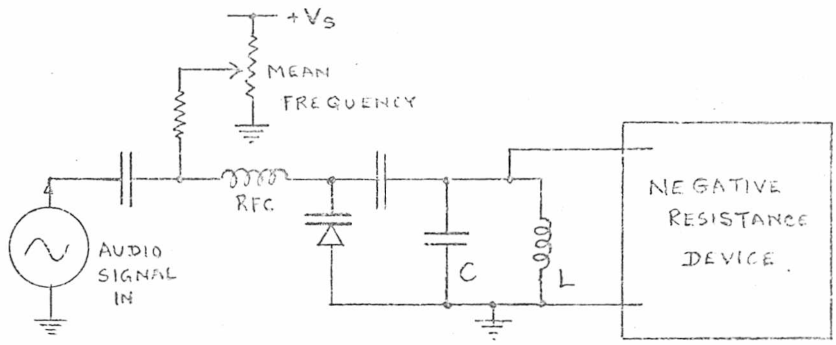


FIGURE C2.1.1
FM OSCILLATOR CIRCUIT

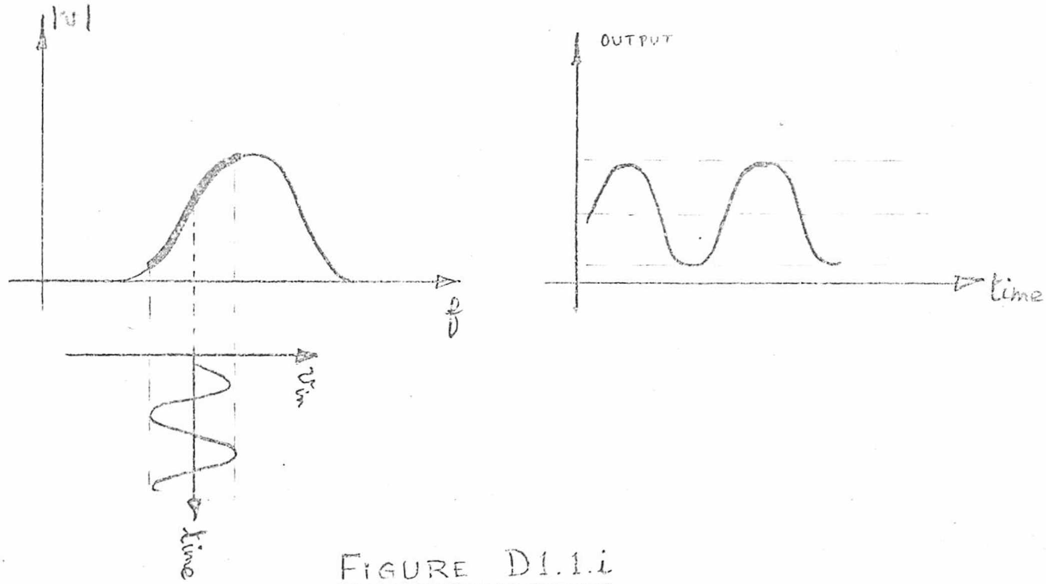


FIGURE D1.1.i
OFF-TUNE RESONANT CIRCUIT

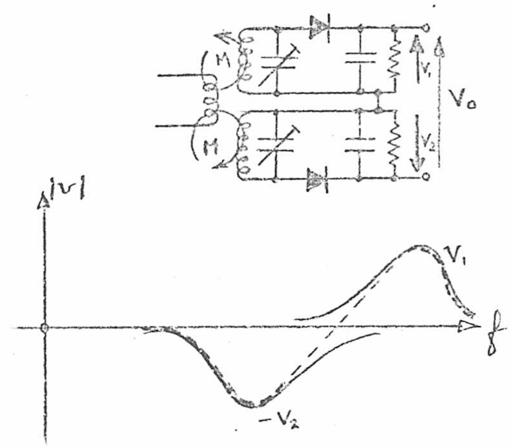


FIGURE D1.1.ii
CROSBY DISCRIMINATOR

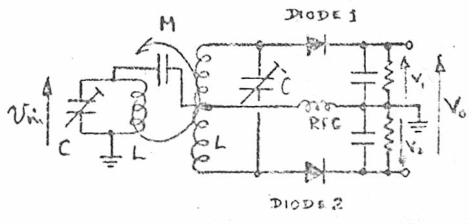


FIGURE D1.1.iii
ENERGY FEED OF
FOSTER-SEELEY

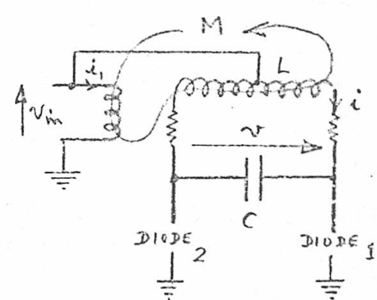


FIGURE D1.1.iv
CURRENT FLOW OF
FOSTER-SEELEY

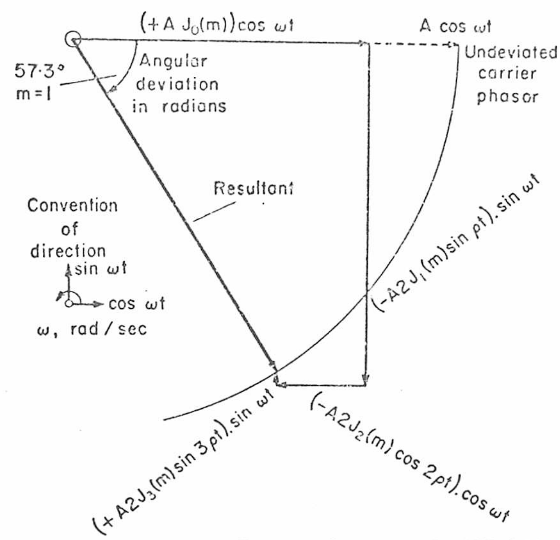


FIGURE D2.1.i
FM PHASOR AT PEAK PHASE ANGLE

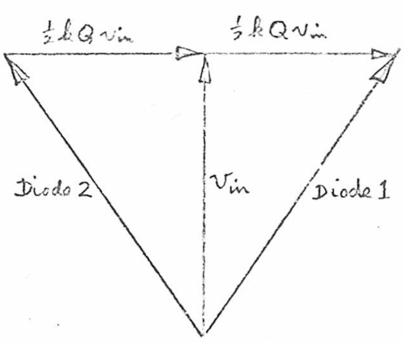


FIGURE D2.2.1.i
RESULTANTS AT MEAN f

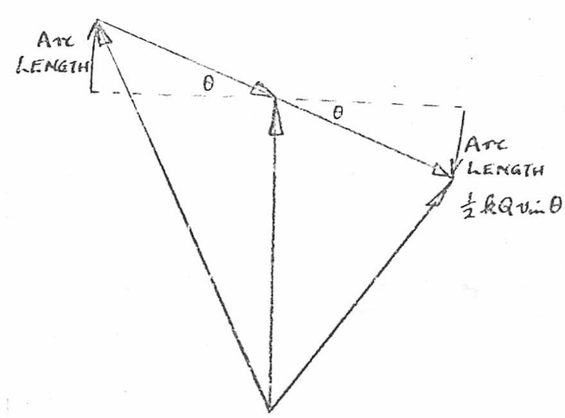


FIGURE D2.2.1ii
RESULTANTS OFF TUNE

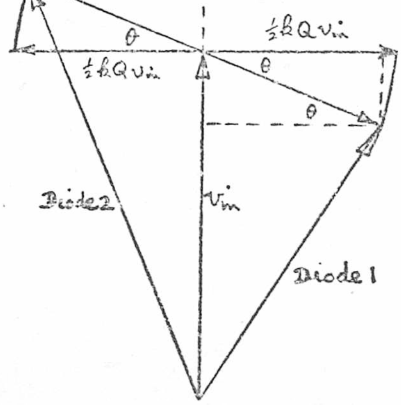


FIGURE D2.2.2.i
LINEARITY GEOMETRY

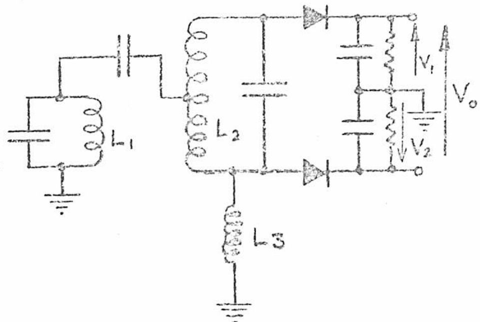


FIGURE D2.2.3.i

MODIFIED FOSTER-SEELEY

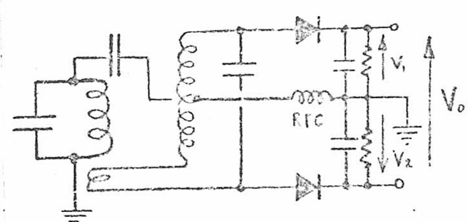


FIGURE D2.2.3.ii

LOW INDUCTANCE COUPLED F-S

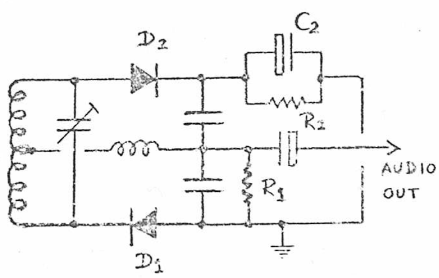


FIGURE D2.2.3.iii

RATIO DETECTOR

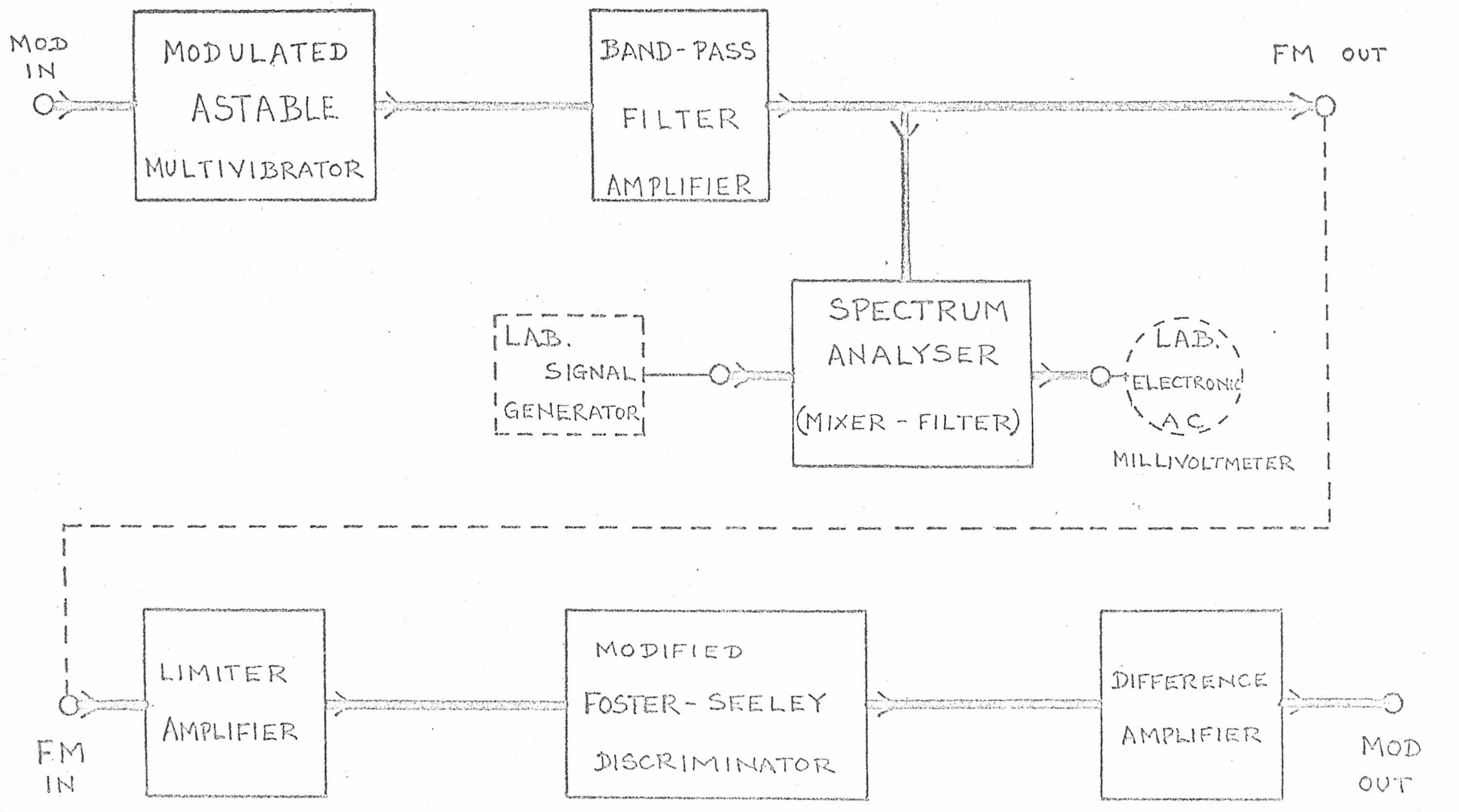


FIGURE D3.i BLOCK SCHEMATIC OF FM DEMONSTRATION

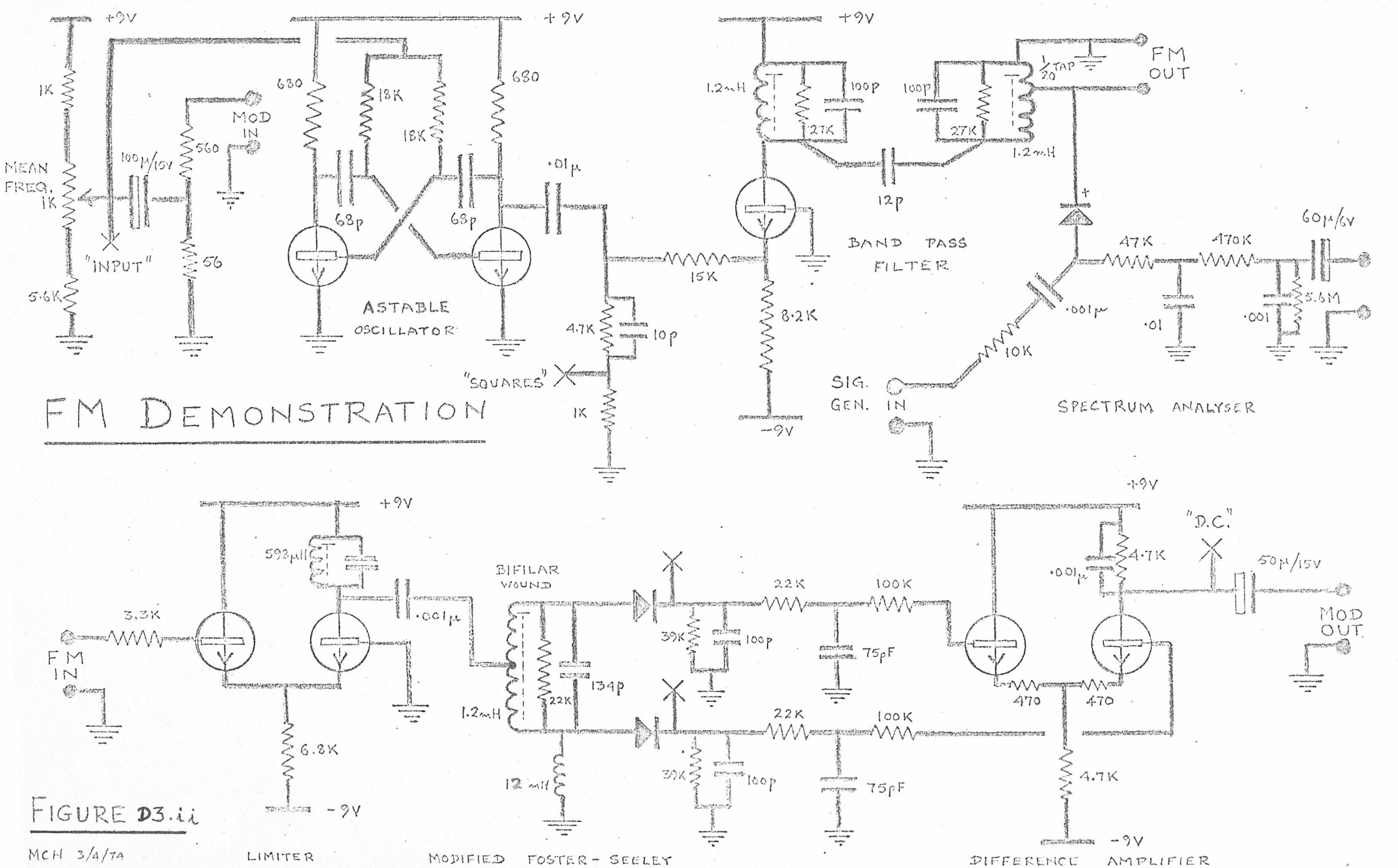

FM DEMONSTRATION
+9V
+9V
+9V
1K
MEAN FREQ. 1K
5.6K
"INPUT"
100μ/15V
560
56
MOD IN
680
680
18K
18K
68p
68p
ASTABLE OSCILLATOR
·01μ
15K
4.7K
10p
"SQUARES"
1K
1.2mH
27K
100p
8.2K
-9V
12p
BAND PASS FILTER
100p
27K
1.2mH
1/20 TAP
FM OUT
47K
470K
60μ/6V
·001μ
10K
·01
·001
5.6M
SIG. GEN. IN
SPECTRUM ANALYSER
+9V
593μH
3.3K
FM IN
6.8K
-9V
·001μ
BIFILAR WOUND
1.2mH
22K
134p
12 mH
39K
100p
22K
100K
75pF
39K
100p
22K
100K
75pF
+9V
"D.C."
4.7K
·001μ
50μ/15V
MOD OUT
470
470
4.7K
-9V
MCH 3/4/74
LIMITER
MODIFIED FOSTER - SEELEY
DIFFERENCE AMPLIFIER

FIGURE D3.ii

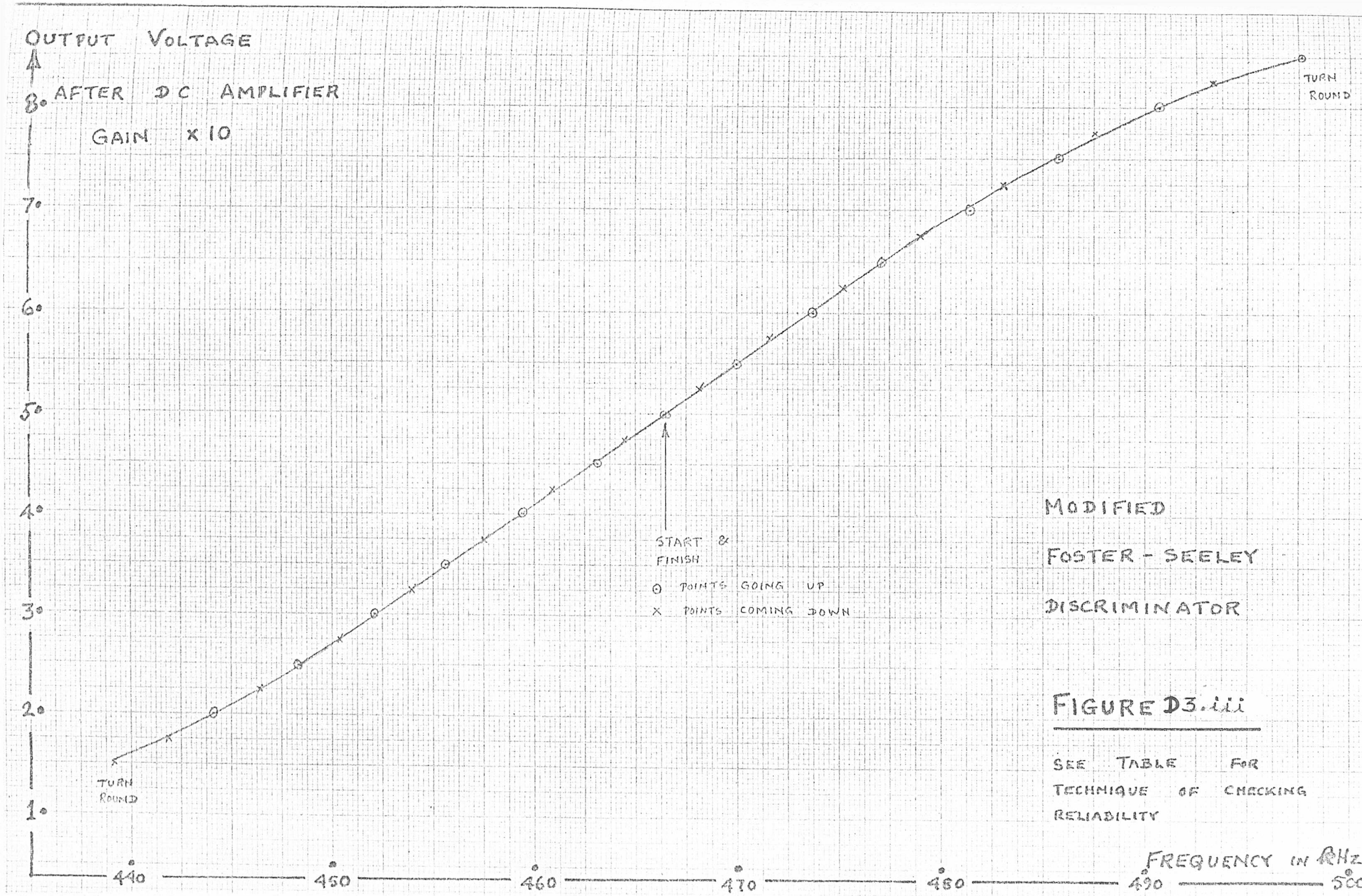
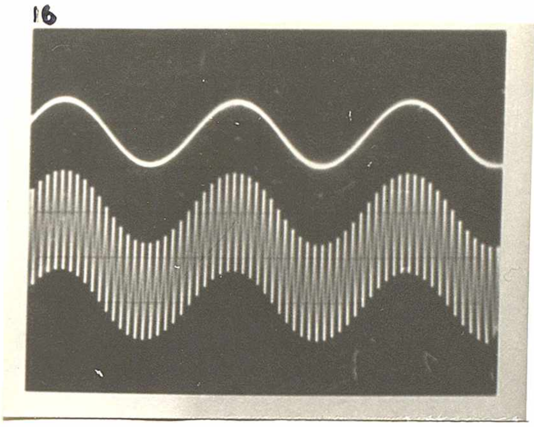
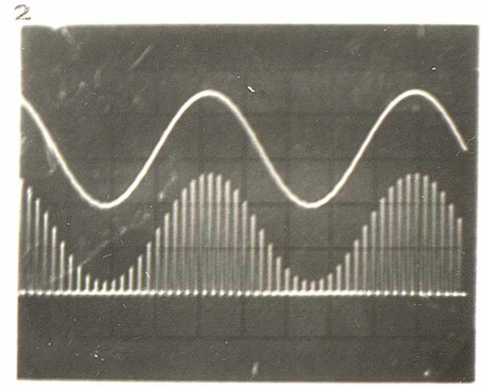
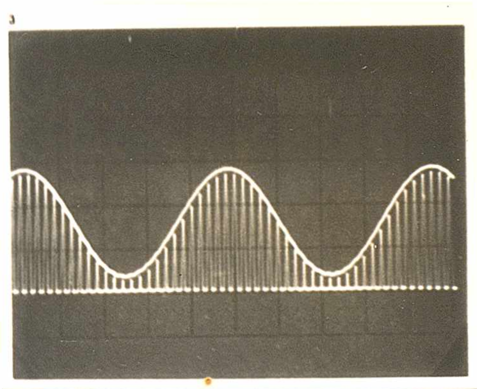


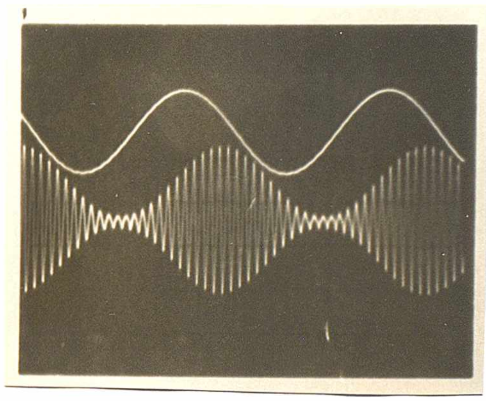
FIGURE D3.iii

SEE TABLE FOR TECHNIQUE OF CHECKING RELIABILITY



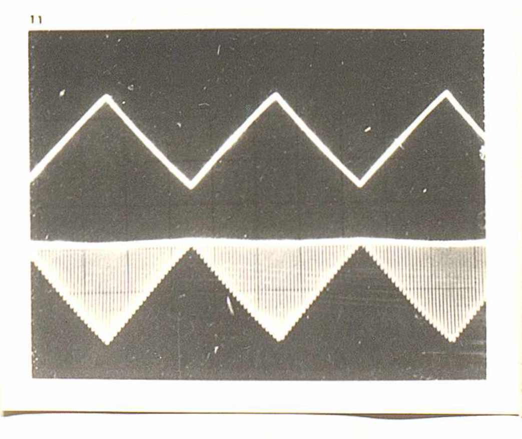


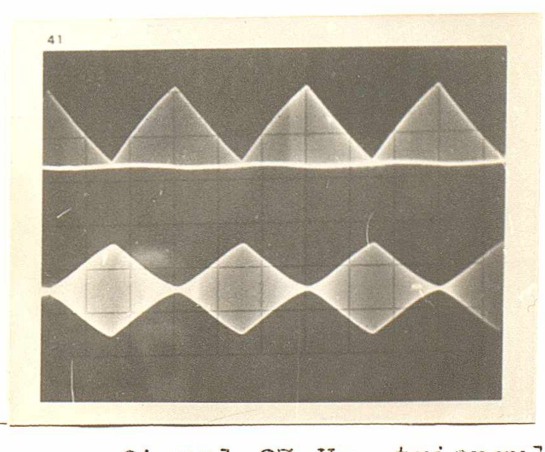




Filtered

Carrier 5 kHz sine 4 V pk Signal 200 Hz 3.6 V pk



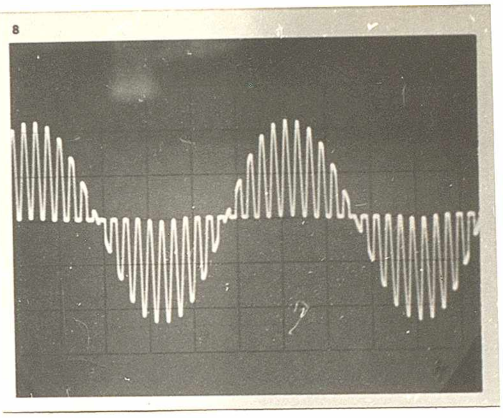


Filtered

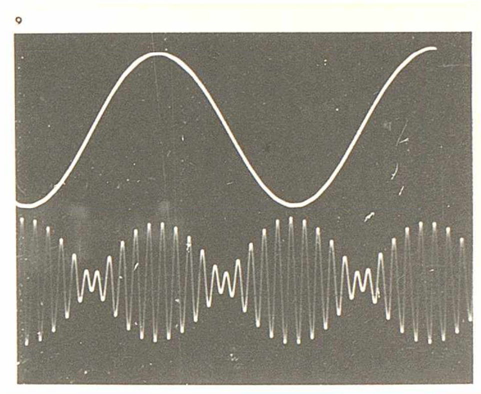
Carrier 5 kHz sine Signal 27 Hz triangular wave

BALANCED MISMATCHED MODULATOR

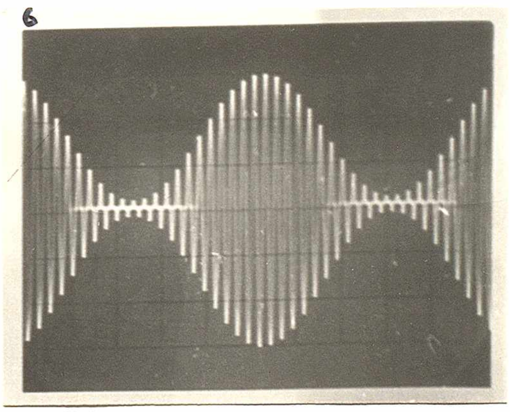
Carrier 5 kHz sine



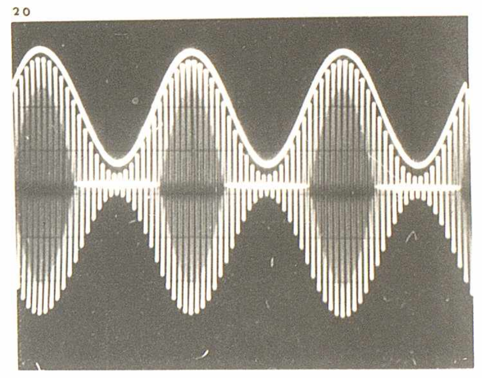
Signal 200 Hz

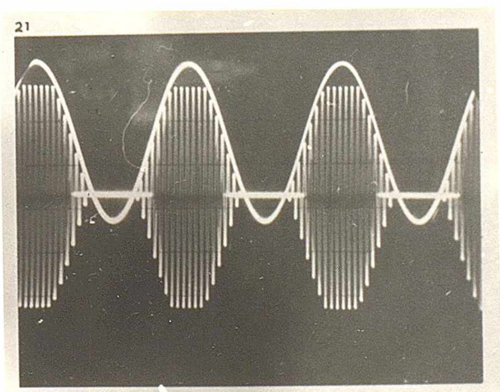


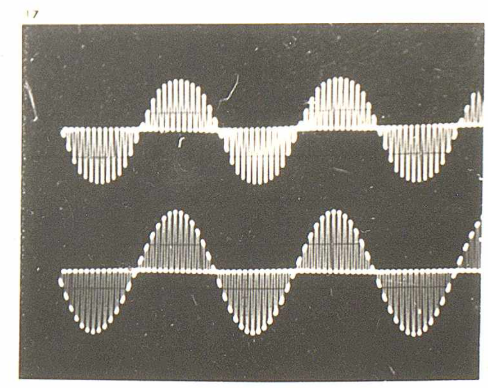
Carrier Balanced
Wideband



Carrier Balanced
Filtered





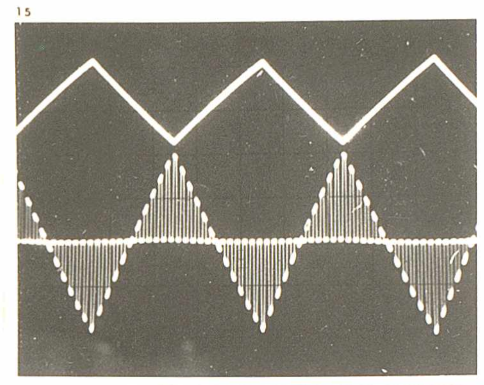


All three Signal Balanced
Wideband

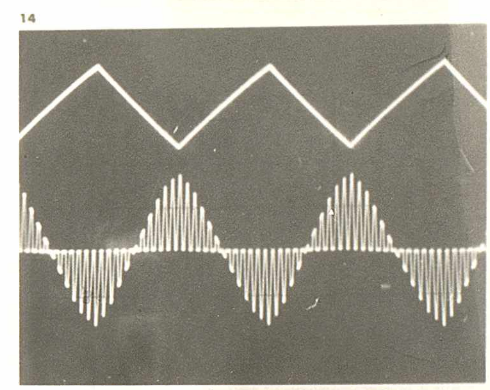
Page A2.1.2.R.2 BALANCED MISMATCHED MODULATOR

Sequence of gradually reducing carrier magnitude.

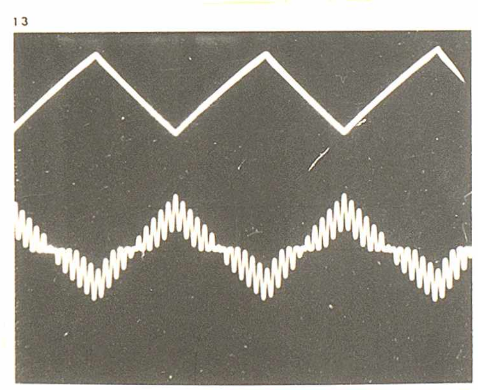
Carrier 5 kHz sine wave Signal 27 Hz triangular wave



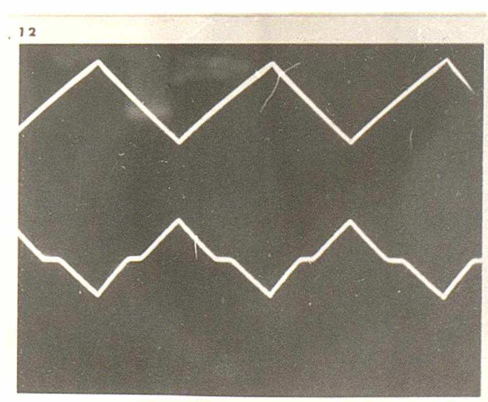
Carrier over large



Carrier just sufficient



Carrier insufficient

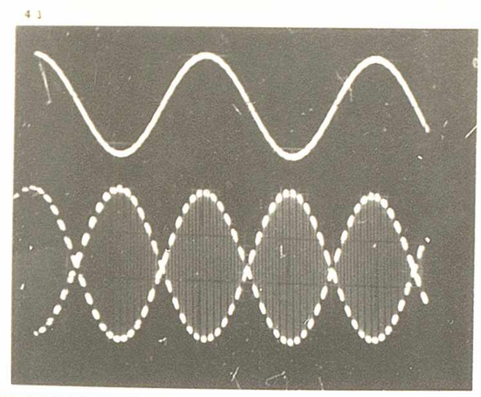


Carrier Zero

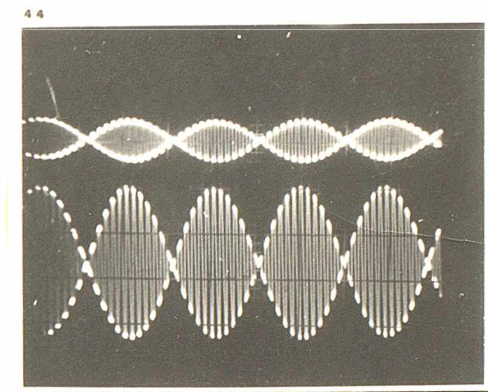
Page A2. 1. 3. R. 1 DOUBLY BALANCED MISMATCHED MODULATOR

All photographs are with Carrier sinusoidal wave 5 kHZ

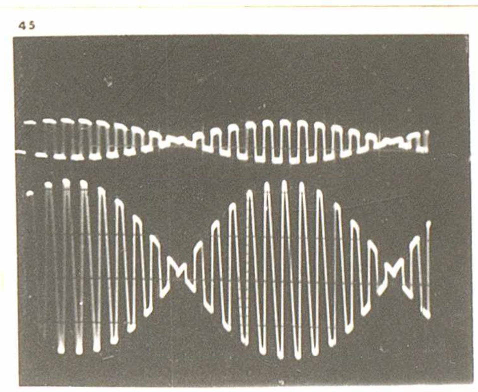
Signal 200 Hz



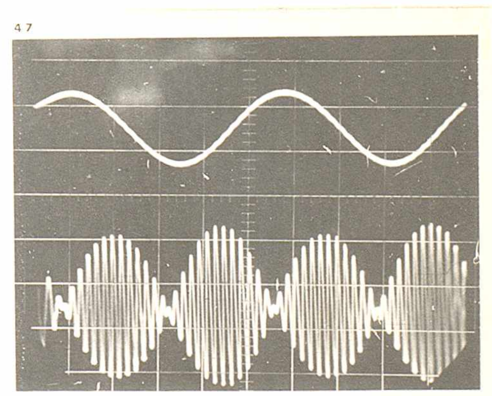
Signal 11 V pk to pk
Carrier 30 V pk to pk



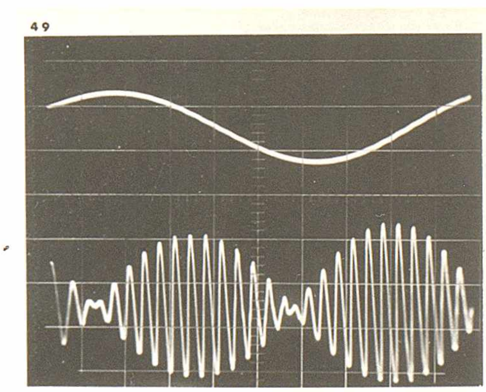
Sig. 3.6 V ptp
Sig 14 V ptp
Carrier 15 V ptp



As for 44 but
faster time base



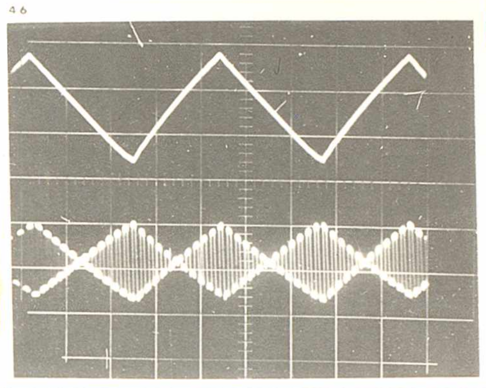
As upper tr of 44
but filtered



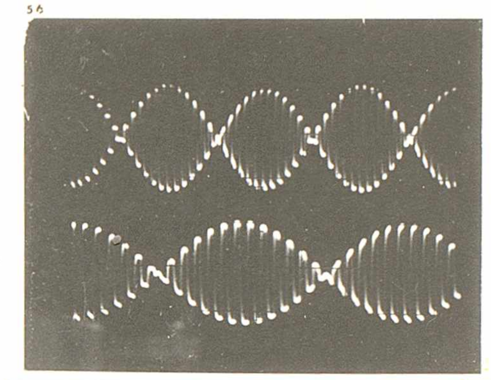
As lower tr of 45
but filtered

 DOUBLY BALANCED MISMATCHED MODULATOR

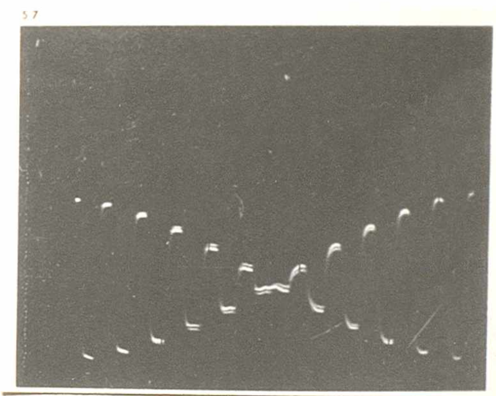
All photographs of wideband waveforms



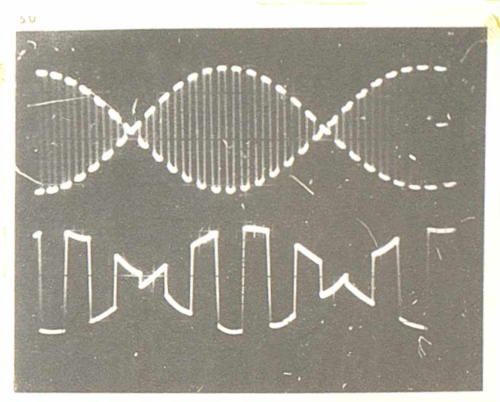
Carrier 5 kHz; Signal 27 Hz
Unbalanced



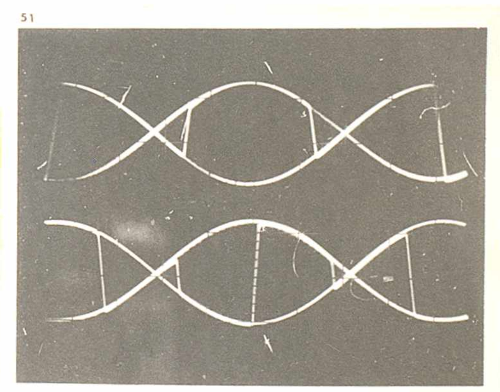
Carrier 500 kHz Signal 20 kHz

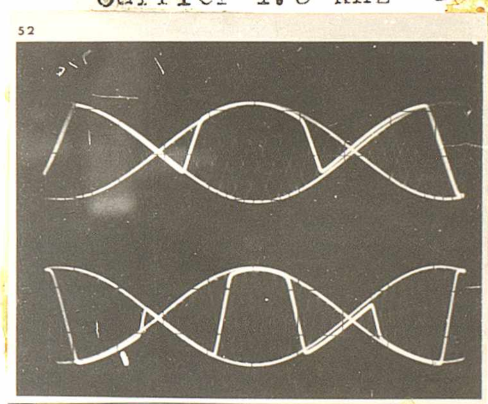


As No. 56 Time base faster

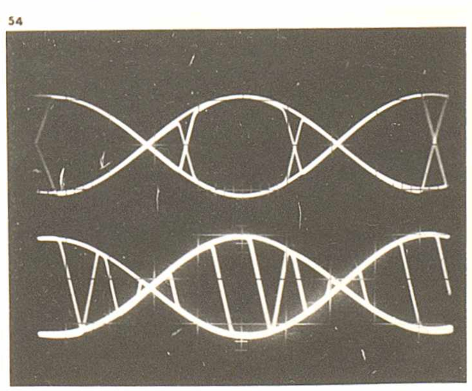


Carrier 5 kHz
Carrier 1.5 kHz Signal 200 Hz

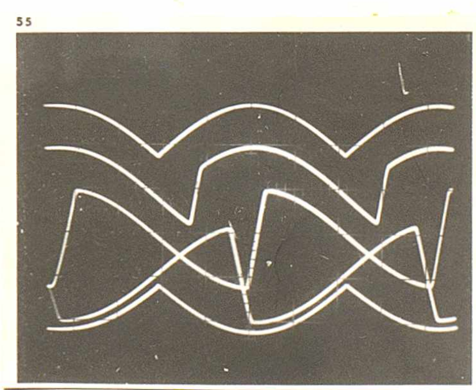




"Carrier" 300 Hz 10 times "Signal" 200 Hz



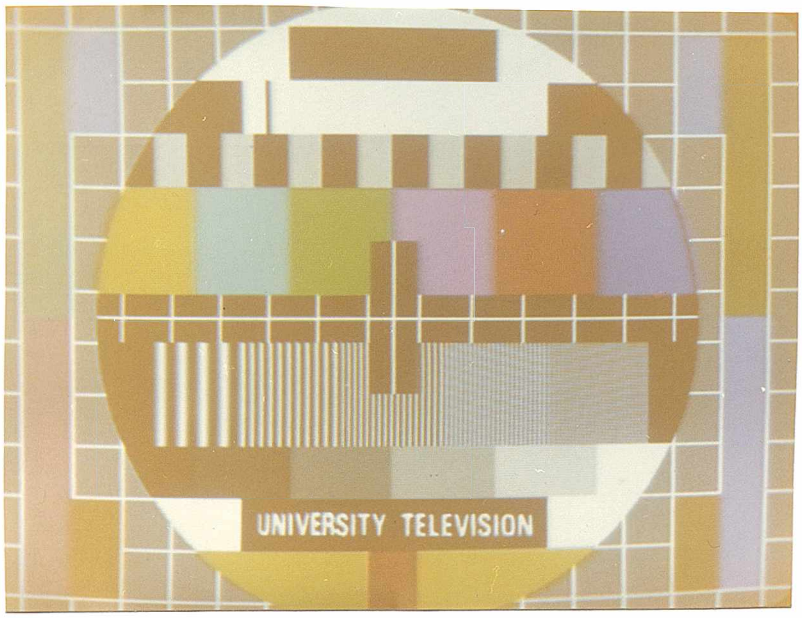
As 51 and 52



Both same frequency

Photograph of Philips Electronic Test Card

Via UHF TV Modulator

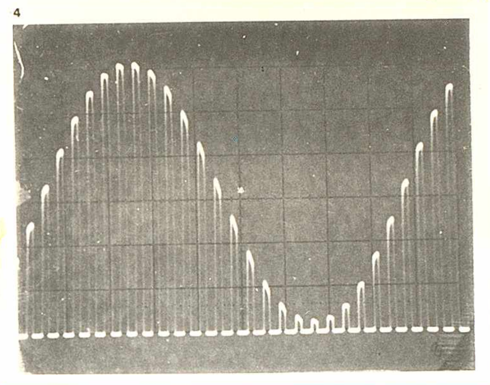


Comparison Photograph of Grampian TV

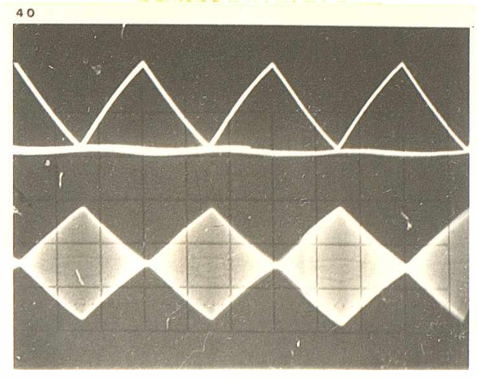
Test Card Received

Page A2.2.1.R.1 SINGLE-ENDED MISMATCHED MODULATOR

with SQUARE WAVE CARRIER

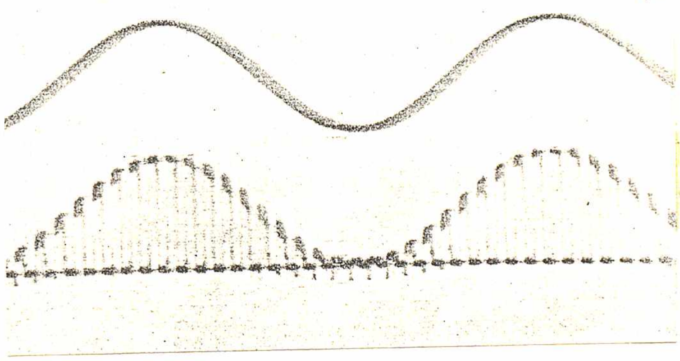


Carrier rectangular
5 kHz
Signal 200 Hz



Carrier square wave
5 kHz
Signal 27 Hz triangular

78



Carrier 900 kHz square wave 5 V pk to pk
Signal 42 kHz sine 2.4 V pk

(some synch jitter)

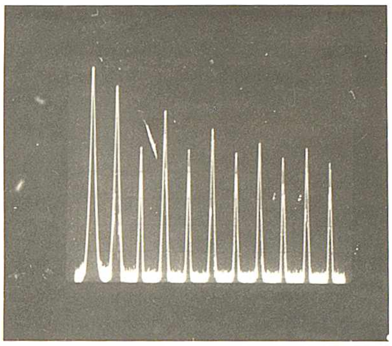
For note on Quality of this oscillogram

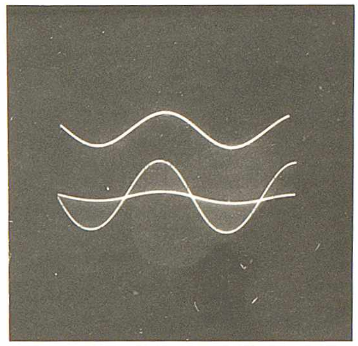
see Appendix A2.2.1

SINGLE-ENDED MISMATCHED MODULATOR with SQUARE WAVE CARRIER

Wideband

83



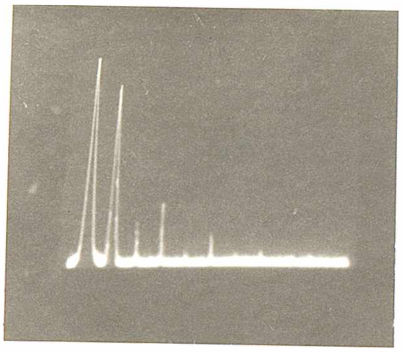


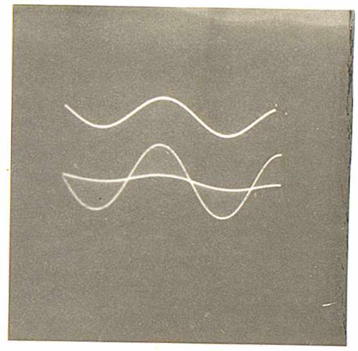
DC to 10 MHz

Signal 400 Hz

Band Pass Filtered

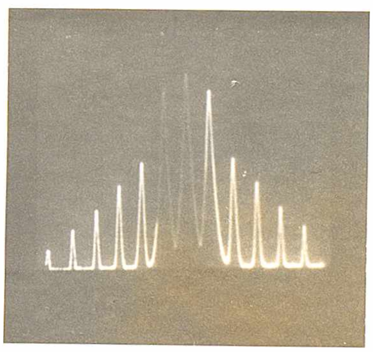
84

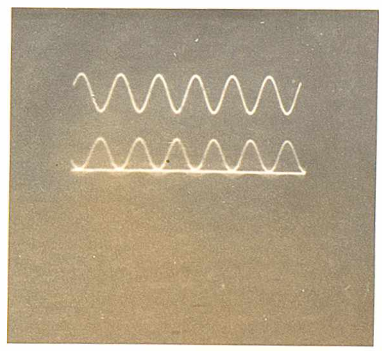




Wideband

86



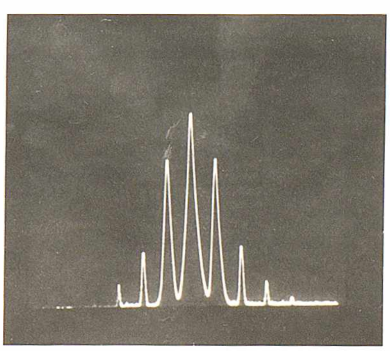


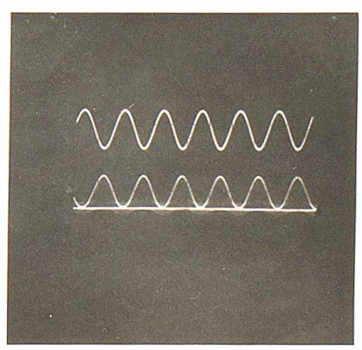
900 kHz ± 250 kHz

Signal 40 kHz

Band Pass Filtered

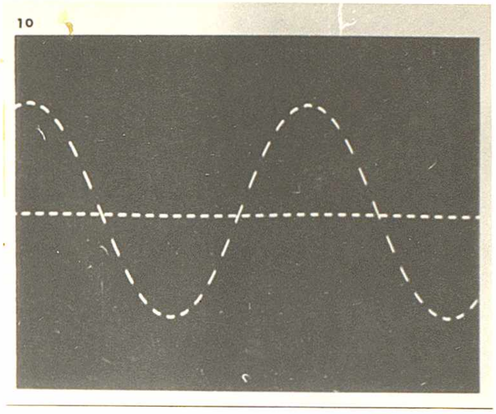
85



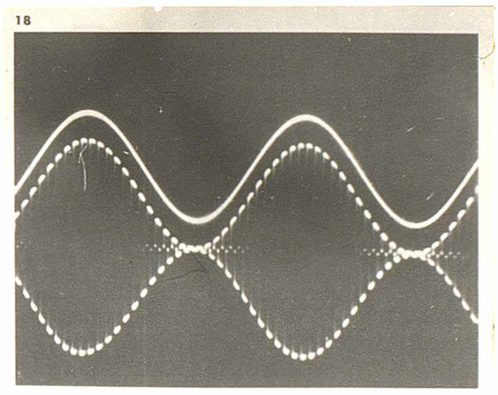


Page A2.2.2.R.1 BALANCED MISMATCHED MODULATOR

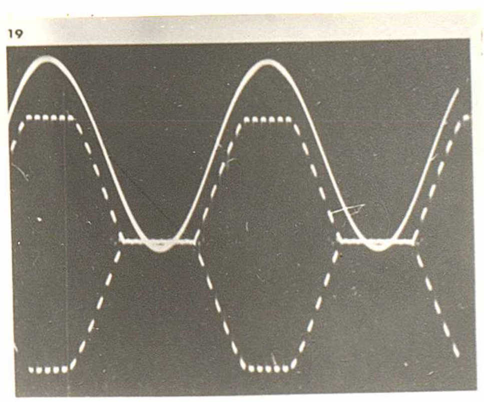
with SQUARE WAVE CARRIER



Carrier 5 kHz Square Balanced 5 V pk to pk
Signal 200 Hz sine 4 V pk



Carrier 5 kHz Square Unbalanced 5 V pk to pk
Signal 200 Hz sine 2.5 V pk BALANCED



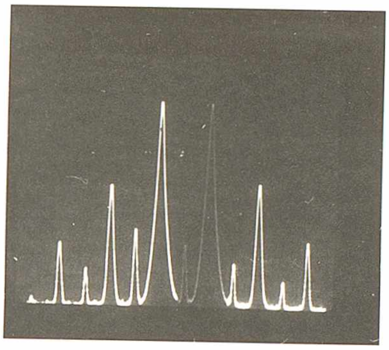
As 18 above but excessive Sinusoid

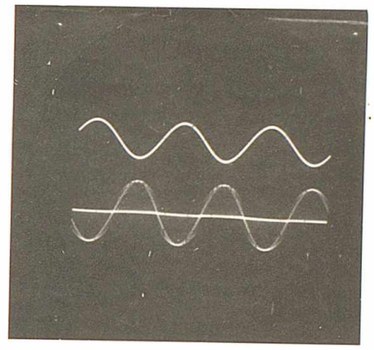
Page A2.2.2.R.2 BALANCED MISMATCHED MODULATOR

with SQUARE WAVE CARRIER

87

Wideband

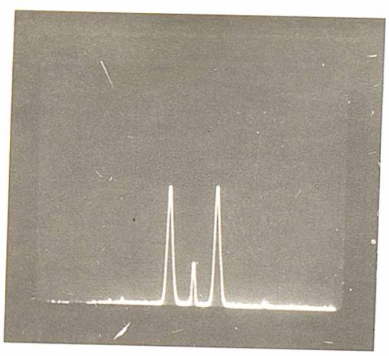


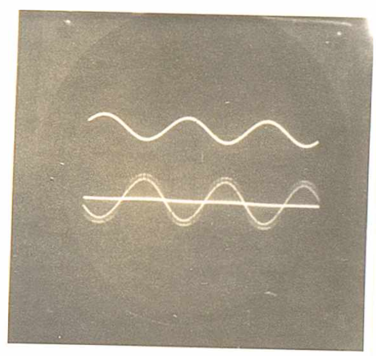


900 kHz ± 250 kHz Signal 40 kHz

88

Filtered and smaller signal input

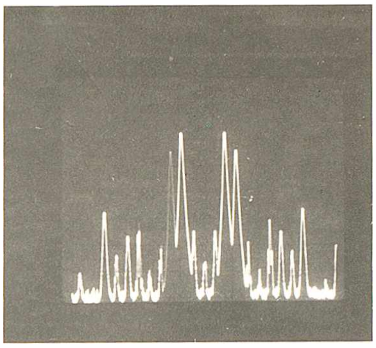




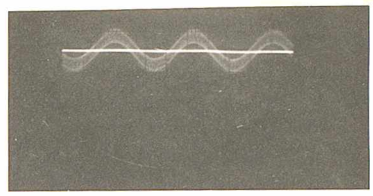
Spectrum Analyser Vertical Scale 3.5 mm per 10 dB

Page A2.2.2.R.3 BALANCED MISMATCHED MODULATOR

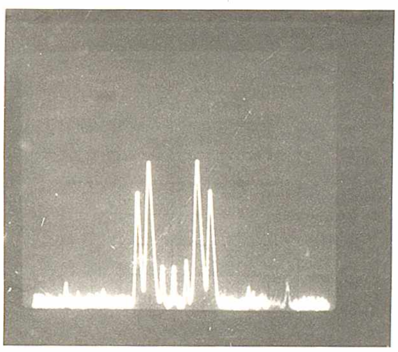
with SQUARE WAVE CARRIER



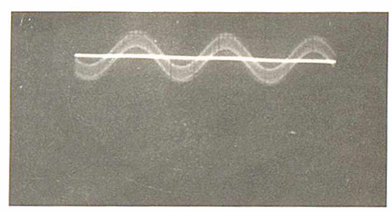
90



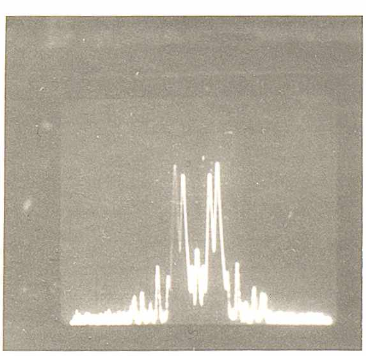
900 kHz ± 250 kHz Sines 40 & 65 kHz
Wideband



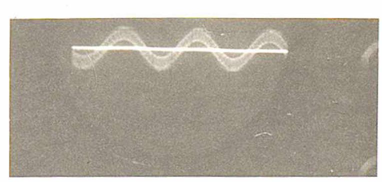
89



As photo 90 above but band pass filtered

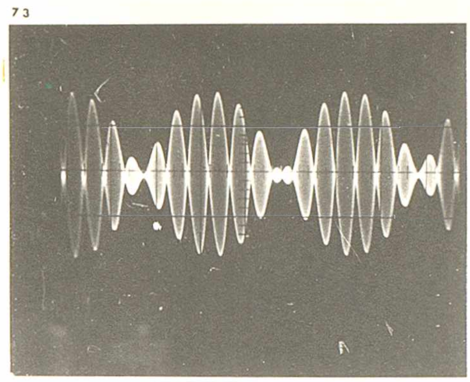


91



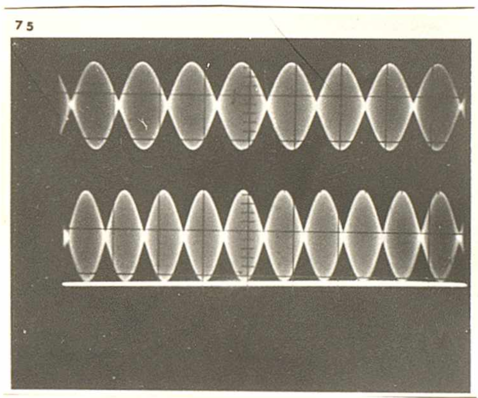
As others above but Sines 40 & 26 kHz

DOUBLY BALANCED MISMATCHED MODULATOR with SQUARE WAVE CARRIER



Two equal sines as signal

2.2 & 2.9 kH

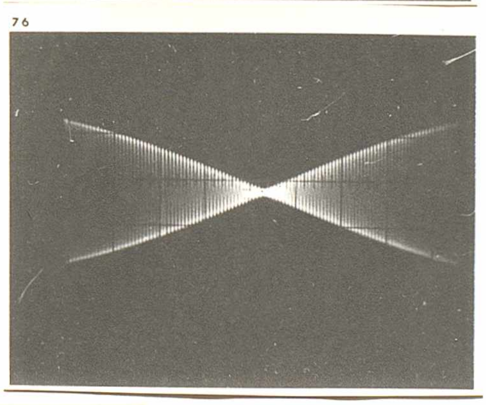


Above signals one at a time

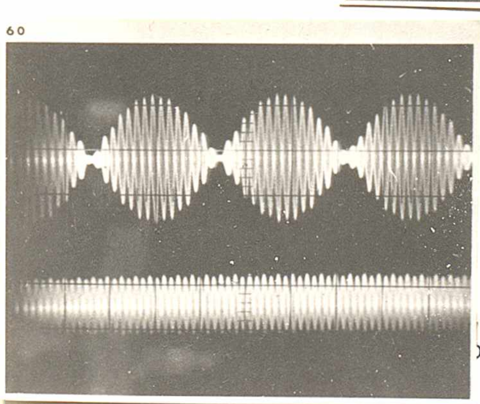
Top 2.2 kHz

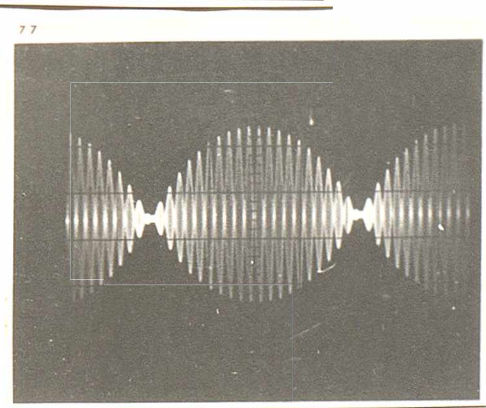
Middle 2.9 kHz

Lower carrier, no signal.



Expanded time scale of zero crossing time of one of above

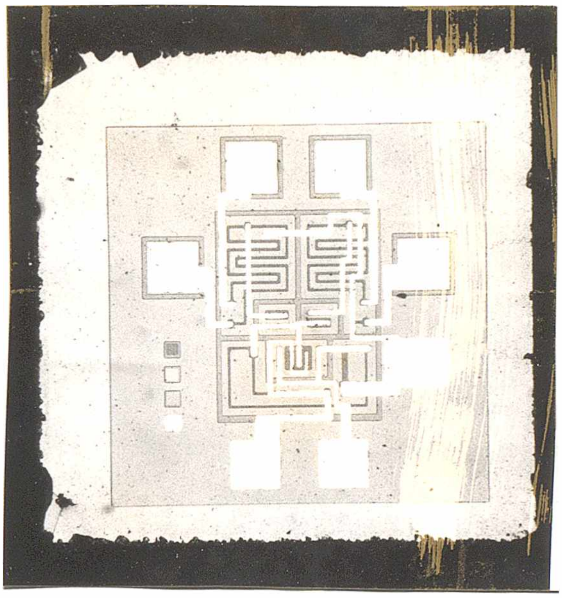




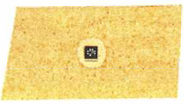
Two equal sines as signal.

2.6 & 3.0 kHz

2.6 & 3.0 kHz   Equal amplitude sines   2.66 & 2.9 kHz



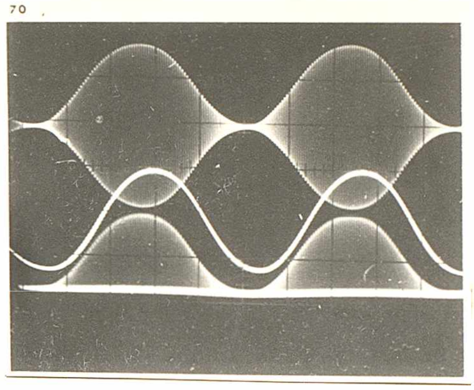
Microscope Photograph of Slice
of Balanced Mismatched
Modulator/Demodulator



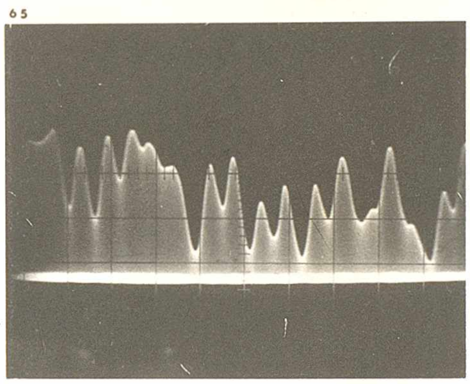
Actual Specimen

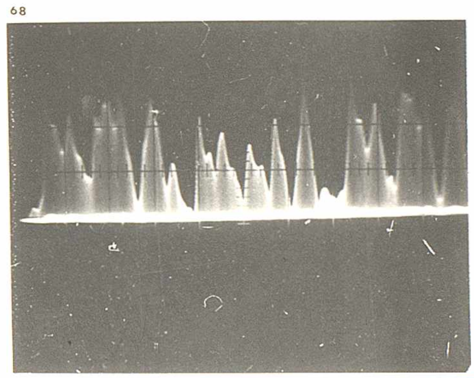
(Failed Tests)

SINGLE-ENDED MISMATCHED DEMODULATOR with NORMAL AM APPLIED

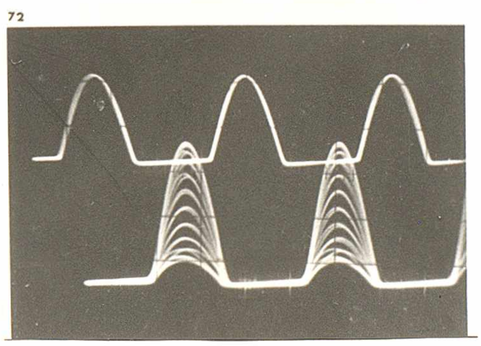


100 kHz AM by 1.2 kHz

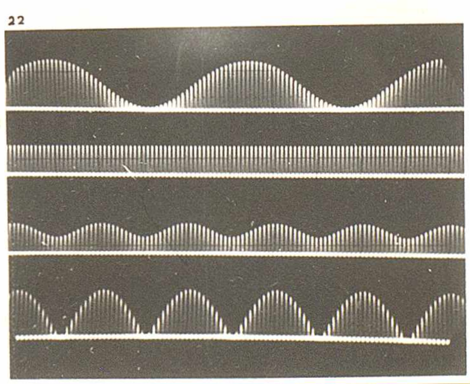




100 kHz modulated by music(from broadcast)



100 kHz at demodulator

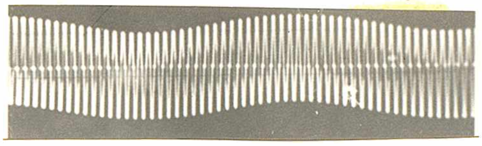


18 kHz AM by 400 Hz
and beats with 17 kHz

BALANCED MISMATCHED DEMODULATOR

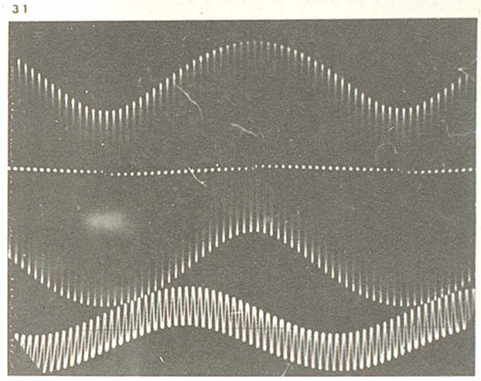
with SIMULATED SSB APPLIED

24



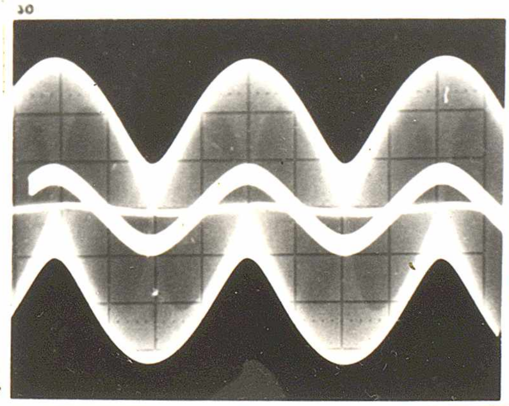
17 kHz Carrier, 18 kHz "SSB

31

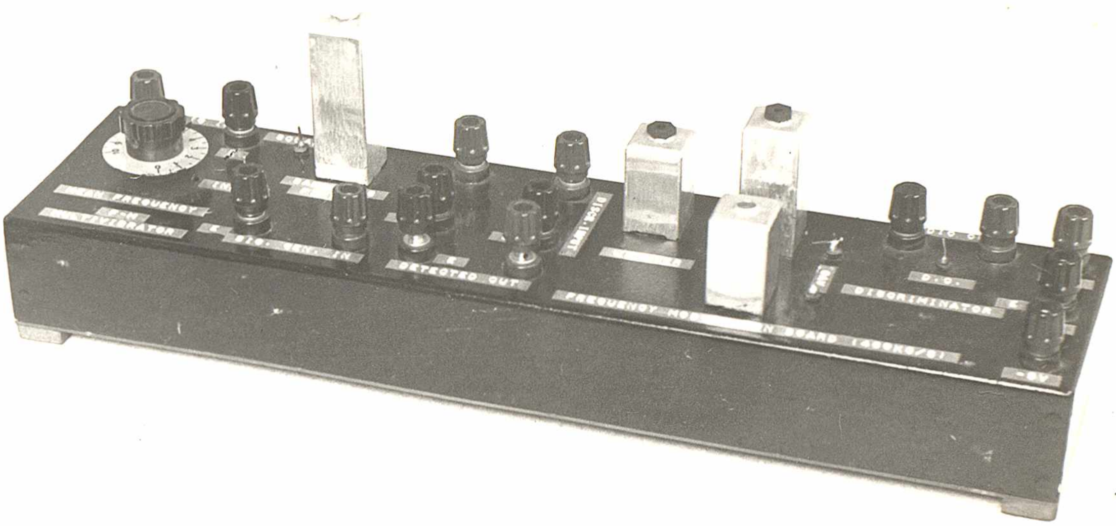


Similar to 24 above
Larger Signals
Lower trace filtered

30



40 kHz Carrier, 42 kHz "SSB"
Output filtered also

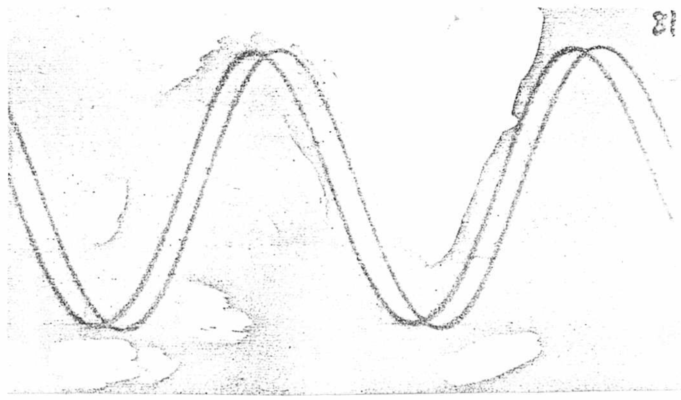


Knob on left is manual control of mean frequency.

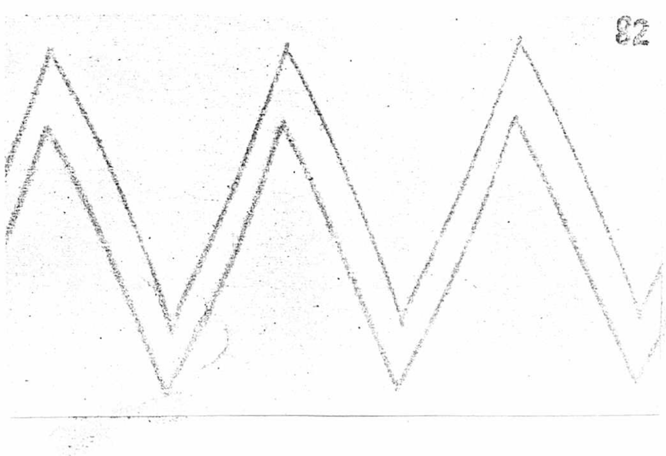
Sceening can on left is band pass filter.

Three screening cans on right are the three independent coils comprising the Modified Foster-Seeley discriminator.

Length is about 38 cm .



4 V pk to pk Sine of 4 kHz



4 V pk to pk Triangular Wave 1 kHz