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"To promote the advancement of radio, electronics and kindred subjects by the exchange of information in these branches of engineering."

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LOOKING AHEAD

THE first month of the year is traditionally a time of renewed hope and resolution. For most of us it is also the time to review the past year and to make new resolutions.

Members may look back with some satisfaction on the Institution's activities during 1957; outstanding of course, was the Convention, at which we enjoyed an international attendance of many radio and electronics engineers. The Convention was one of the most successful the Institution has ever held and did much to emphasize future possibilities in the application of electronics to automation.

We may also look back with *reasonable* satisfaction on the increased membership and status of the Institution. I emphasize *reasonable* because I do not think that we should be wholly satisfied with the rate of increase in membership. I do realize that any disappointment we may have in this connection is shared by several professional bodies in other branches of engineering. In the technical press there has been much comment about the decrease in studentship registration of most engineering institutions. In our own Institution we have this particular problem to tackle, for unless we secure the early interest of students taking their qualifying courses, we have less opportunity for forward planning of Institution activities.

Apart from the principal object of disseminating information—which is done by meetings and by such publications as the *Journal*—all professional bodies have a particular responsibility in encouraging the extension of facilities for technical education.

This is probably one of the most important contributions to the well-being of industry, for without properly trained engineers, industry must inevitably fall behind in the supply of new ideas and the application of scientific research.

It is important, therefore, that the future membership of professional institutions should include men who are prepared to serve their profession and undertake the responsibilities of leadership. We must also not overlook the fact that the future leaders in our profession must come from the present younger members.

Certainly the future for the young member is very promising. Information gained during the International Geophysical Year affords countless examples of the work still to be fulfilled by the radio and electronics engineer in the fields of communication and instrumentation.

Components used in the radio and electronics industry now have application far beyond the original communications field. There is still much to do in design work, as well as in the technique of producing consistently reliable equipment. I hardly need mention that development in the field of solid state physics is at present only in its infancy and thus provides further opportunity for engineers.

There also remains a world-wide demand for the expansion of all forms of radio and television services. The production, installation, operation and maintenance of such equipment calls for an increasing number of engineers qualified in radio and electronic engineering.

I believe that the future of our profession and industry is bright, and that there is every reason why my New Year wishes to every member—that he may achieve his ambition and have satisfaction in his work in 1958—will be fulfilled.

G. A. Marriott.

INSTITUTION NOTICES

Members in the New Year Honours List

The Council of the Institution congratulates the following members:

Commander Richard Lewis, R.N., (Associate Member), on his appointment as an Officer of the Military Division of the Most Excellent Order of the British Empire. Commander Lewis is Training Commandant of the Royal Naval Electrical Training Establishment, H.M.S. *Collingwood.*

Squadron Leader William Lawrence Price, R.A.F., (Associate Member), on his appointment as an Officer of the Military Division of the Most Excellent Order of the British Empire. Squadron Leader Price is Officer Commanding the Electronics Division, R.A.F. Technical College, Henlow. He is a member of the Institution's Education and Examinations Committee.

Henry William Akester (Associate Member), on his appointment as a Member of the Civil Division of the Most Excellent Order of the British Empire. Before his retirement in June 1957, Mr. Akester was Chief Technical Superintendent of the Gold Coast Government Broadcasting Department, and he is at present with the South African Broadcasting Corporation engaged on reception and propagation research.

Institution Meeting in Cambridge

Institution activities in Cambridge have hitherto been confined to holding Conventions in the Cavendish Laboratory during the University vacation, since it has been represented to the Institution that regular meetings of a local section might add unduly to extra-mural activities for undergraduates. Recently, however, the Institution was approached by the Cambridge Section of another Institution with a view to holding a joint meeting to show the Technical Films which were screened at a Brit.I.R.E. meeting in London during December. The arrangements for the joint meeting have subsequently been cancelled, but in view of the considerable interest shown in this programme, it has been decided to go ahead with the meeting for the benefit of Brit.I.R.E. members in the Cambridge area.

Six films will therefore be shown, on Radio Propagation, Computers, Guided Weapons and Nuclear Power, and the meeting will be held in the Cavendish Laboratory on Tuesday, February 11th, at 7.30 p.m. Details are being sent to all members in the Cambridge area.

Frequency Allocation Committee

The Postmaster-General has announced that Sir Lawrence Bragg, O.B.E., M.C., F.R.S., will be chairman of a committee which will advise on broad aspects of radio frequency allocation. The P.M.G. is also appointing to the committee members representative of the radio industry, users' organizations, and Government departments. The committee will enable both users and industry to be associated more closely with frequency planning and will be of assistance to the radio industry in its development work. It is expected that the first meeting of the committee will be held shortly.

Sir Lawrence Bragg is Director of the Royal Institution and it will be recalled that he gave the Third Clerk Maxwell Memorial Lecture at the 1957 Convention in Cambridge.

Retirement of R.I.C. Director

On doctor's advice, but not with effect until October 31st, 1958, Vice-Admiral J. W. S. Dorling, C.B., (Member), is to retire from his appointment as Director of the Radio Industry Council. He was the Council's first Director, being appointed in 1946, immediately after the formation of the Council as the co-ordinating body of four associations in the industry.

Congress on U.H.F. Circuit and Antennas

The papers read at the International Congress on Ultra High Frequency Circuits and Antennas held in Paris under the auspices of the Société des Radioélectriciens in October 1957 are being published as a single volume in July. The subscription price will be 5,000 francs (non-subscribers 7,500 F.) per copy, and further information may be obtained from Société des Radioélectriciens, 10 Avenue Pierre-Larousse, Malakoff (Seine), France.

Back Copies of the *Journal*

The Publications Department of the Institution wishes to acquire copies of the *Journal* dated October and December 1956. Members willing to dispose of their copies are invited to return them to the Institution. Please note that these are the *only* issues which are at present required.

RECOMMENDED METHOD OF EXPRESSING ELECTRONIC MEASURING INSTRUMENT CHARACTERISTICS

1. Amplitude Modulated or Frequency Modulated Signal Generators*

Prepared by the Technical Committee of the Institution

PREFACE

The users of radio and electronic measuring instruments often find it difficult to assess the suitability of any particular instrument for a specific purpose from the information which is given. There is no convention governing the presentation of information that is required, and the different methods make comparison of instruments difficult.

The Technical Committee of the Institution has compiled the first of a series of recommendations as the basis for a common standard. This deals with Amplitude or Frequency Modulated Signal Generators. In it are listed and, where necessary, defined, all the principal characteristics of which the users of this class of instrument might be likely to require details; it is not implied that all these characteristics need be specified for all grades of instrument.

The form in which the characteristic and, where applicable, its standard of accuracy, should be expressed, is given. Performance limits are not specified because the establishment of actual performance standards is not the objective of these recommendations.

During the period in which this subject has been receiving the Technical Committee's attention, the work of the Radio and Electronic Measurements Committee of the Ministry of Supply came to its notice. The recommendations below are generally parallel to the findings, published as R.E.M.C./24/FR (Issue 1, May 1956), and use the special definitions of that Committee. It should, however, be noted that the approach of the Ministry of Supply document has the designer in mind, while the present recommendations are drawn up primarily from the point of view of the user.

The report is divided into four parts: an introductory section giving general data, and three main sections, dealing with frequency, radio-frequency output, and modulation characteristics respectively. An appendix of definitions is given for those terms which have a particular meaning when applied to signal generators.

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* Approved by the Council for publication on 16th July 1957. (Report No. 13.)
Based on a report compiled by Commander K. W. Pilgrim, R.N. (Member).
U.D.C. No. 621.373.42:621.376.22:621.376.32

Part 1—GENERAL DATA

	Feature	Method of expression	Remarks
1.1.	Power Supply Requirements (Voltage Change ± ... volts)	... volts d.c./a.c. ... c/s, ... watts	Maximum mains voltage variation for which the stated accuracies hold good must be given.
1.2.	Temperature Range	... °C to ... °C	Maximum ambient temperature range for which the stated accuracies hold good.
1.3.	Dimensions (over projections)	Height ... in. (... cm) Width ... in. (... cm) Depth ... in. (... cm)	
1.4.	Weight	... lb. (... kg)	
1.5.	Construction and Finish		Where the construction has been to a defined specification, it should be quoted.
1.6.	Accessories		Details of any connectors and adaptors included with the instrument should be stated.
1.7.	Valve Complement		

Part 2—FREQUENCY CHARACTERISTICS

Note: Definitions of expressions printed in italic type are given in the Appendix.

	Characteristic	Method of expressing the characteristic	Remarks
2.1.	**Range**	From ... c/s to ... c/s in ... bands	The frequency range of each band should be stated. Where auxiliary coils are provided the frequency range of each coil should be given. Where use is made of the harmonic content of the oscillator to extend the frequency range this should be stated.
2.2.	**Calibration Accuracy**	... %	This is the maximum error at any value of the output frequency in relation to the calibration of the main frequency control. The type and effective scale length of the main frequency control should be given. Where a crystal oscillator is fitted to check the calibration of the main frequency control this should be stated together with the fundamental frequency of the crystal, the effective range of harmonics and the calibration accuracy.
2.3.	**Re-setting Accuracy**	... %	This is the re-setting accuracy of the main frequency control.

	Characteristic	Method of expressing the characteristic	Remarks
2.4.	**Incremental Frequency** (where incremental tuning is provided)		
2.4.1.	Range	± ... % at ...c/s	This is the frequency range of the incremental control expressed as a percentage of the main dial calibration. The method used to provide incremental tuning should be stated.
2.4.2.	Setting Accuracy	... % at ... c/s	This is the maximum error in any value of Δf expressed as a percentage of the output frequency.
2.5.	**Drift**		
2.5.1.	Short Term	... % or ... c/s	This is the maximum change in frequency over any period of 10 minutes within a 7-hour period commencing 60 minutes after switching on. During the 10-minute period the mains supply voltage and the temperature are assumed to be constant.
2.5.2.	Long Term	... % or ... c/s	This is the maximum change in frequency over a period of 7 hours commencing 60 minutes after switching on. During this 7-hour period the mains supply voltage and the temperature are assumed to be sensibly constant. (See 1.1 and 1.2.)
2.6.	**Pulling**		
2.6.1.	Load Reaction	... % at ... c/s	This is the maximum change in frequency between open-circuit condition of the outlet and when loaded with a resistance equal to the *nominal source impedance*, with the attenuator at the *reference level* setting. The carrier frequency at which the maximum change occurs should be stated.
2.6.2.	Attenuator Reaction	... % at ... c/s	This is the maximum change in frequency over the range of the attenuator up to the *reference level setting*, when the outlet is loaded with the specified load impedance. The carrier frequency at which the maximum change occurs should be stated.
2.7.	**Modulation Reaction**		
2.7.1.	A.M. Reaction	... % at ...% depth of modulation at ... c/s	This is the maximum change in the mean carrier frequency due to the application of a.m. up to the maximum depth of modulation for which the instrument is designed. The carrier frequency at which the maximum change occurs should be stated.

	Characteristic	Method of expressing the characteristic	Remarks
2.7.2.	F.M. Reaction	... % at ... c/s deviation at ... c/s	This is the maximum change in the mean carrier frequency due to the application of f.m. up to the maximum frequency deviation for which the instrument is designed. The carrier frequency at which the maximum change occurs should be stated.
2.8.	**Unwanted Modulation** (under c.w. conditions)		
2.8.1.	Amplitude Modulation	... % at ... c/s	This is the maximum value of the *r.m.s. modulation factor (a.m.).* The carrier frequency at which it occurs should be stated.
2.8.2.	Frequency Modulation	... $\times 10^{-x}$ at ... c/s	This is the maximum value of the *r.m.s. frequency swing (f.m.)* expressed as a fraction of the carrier frequency. The carrier frequency at which it occurs should be stated.
2.9.	**R.F. Distortion and Unwanted Outputs** (under c.w. conditions)		
2.9.1.	Minimum ratio of the carrier amplitude to the amplitude of any unwanted frequency within $\pm 20\%$ of the carrier frequency.		... db
2.9.2.	Minimum ratio of the carrier amplitude of any other unwanted frequency between one-third and three times carrier frequency.		... db
2.9.3.	Minimum ratio of the r.m.s. value of the carrier at the *reference level* setting to the value of any d.c. component present in the output signal.		... db

Part 3—R.F. OUTPUT CHARACTERISTICS

Note : Definitions of expressions printed in italic type are given in the Appendix.

	Characteristic	Method of expressing the characteristic	Remarks
3.1.	**Range**	The generator is calibrated for use under open-circuit conditions/terminated conditions. The r.f. output level is continuously variable/variable in ... db steps from ... db to ... db relative to ... volts or ... watts at a *source impedance* or *nominal source impedance* of ... ohms.	0 db is assumed to be the *reference level* setting. The means used to produce attenuation, e.g. slide wire, resistive attenuator or piston attenuator should be stated. Where additional outputs are provided, e.g. high level output, details should be stated.

Characteristic		Method of expressing the characteristic	Remarks
3.2.	**Calibration Accuracy**		
3.2.1.	Reference Level	... db	In the case of generators calibrated in terms of voltage (r.m.s.) and for use under open-circuit conditions, this is the maximum error in the value of the *source e.m.f.* at the *reference level* setting. In the case of generators calibrated in terms of voltage (r.m.s.) and for use under terminated conditions, this is the maximum error in the value of the *equivalent source e.m.f.* at the *reference level* setting. In the case of generators calibrated in terms of power, this is the maximum error in the value of the *equivalent available power* at the *reference level setting.*
3.2.2.	Attenuation	... db	This is the maximum inaccuracy of the attenuator system over any part of its range.
3.2.3.	Source Impedance:		
3.2.3.1.	Resistance and Phase Angle	Departure from nominal not greater than ... ohms, not greater than ... °	Applicable to generators for use under open-circuit conditions.
3.2.3.2.	V.S.W.R.	Not less than ... v.s.w.r.	Applicable to generators for use under terminated conditions. The value of the *source impedance* is expressed as a v.s.w.r. in relation to the *nominal source impedance*.
3.3.	**Drift**		
3.3.1.	Short Term	... db	This is the maximum change in the *reference level* over any period of 10 minutes in 7 hours commencing 60 minutes after switching on, without re-setting the level monitor. A constant mains supply voltage is assumed.
3.3.2.	Long Term	... db	This is the maximum change in the *reference level* over a period of 7 hours commencing 60 minutes after switching on, without re-setting of the level monitor. A sensibly constant mains supply voltage is assumed. (See 1.1 and 1.2.)
3.4.	**Modulation Reaction**		
3.4.1.	A.M. Reaction	... db at ... % depth of modulation	This is the maximum change in the *mean carrier voltage level* between the unmodulated condition up to the maximum depth of modulation for which the instrument is designed. The depth of modulation at which the maximum change occurs should be stated.

	Characteristic	Method of expressing the characteristic	Remarks
3.4.2.	F.M. Reaction	... db at ... c/s deviation	This is the maximum change in the output power level due to the application of f.m. up to the maximum frequency deviation for which the instrument is designed. The frequency deviation at which the maximum change occurs should be stated.
3.5.	**Carrier Frequency Reaction**	... db at ... c/s	This is the maximum change in the r.m.s. voltage output level for any variation of 5% in carrier frequency without re-adjusting the carrier level. The carrier frequency at which the maximum change occurs should be stated.
3.6.	**R.F. Leakage**	... V/m	This is the maximum value of the stray field produced by the generator under any condition of normal usage, at a distance of one metre with the outlet screened and terminated by the specified load impedance.

Part 4—MODULATION CHARACTERISTICS

Note: Definitions of expressions printed in italic type are given in the Appendix.

	Characteristic	Method of expressing the characteristic	Remarks
4.1.	**Internal Amplitude Modulation** (Sinusoidal)		
4.1.1.	Modulation Frequencies	... c/s, ... c/s, etc.	If monitoring facilities are provided, these should be stated.
4.1.1.1.	Modulation Frequency Calibration Accuracy	... %	This is the maximum error in the value of the modulation frequency in relation to the calibration of the modulation frequency control.
4.1.2.	Modulation Depth Range	Fixed at ... %/Variable up to ... %	Where there is any limitation due to the value of the carrier frequency, details should be stated.
4.1.2.1.	Modulation Depth Calibration Accuracy	... % at ... % depth of modulation	This is the maximum error in the value of the modulation depth in relation to the calibration of the modulation depth control. The modulation depth at which it occurs should be stated.

	Characteristic	Method of expressing the characteristic	Remarks
4.1.2.2.	Amplitude/ Frequency Characteristic (Where more than one modulating frequency is provided.)	... %	This is the maximum change in the depth of modulation over the specified modulating frequency range.
4.1.3.	Unwanted F.M.	... $\times 10^{-x}$ at ... % depth of amplitude modulation	This is the maximum value of the *r.m.s. frequency swing (f.m.)* under any condition of wanted a.m. up to the maximum depth of modulation for which the instrument is designed, expressed as a fraction of the carrier frequency. The modulation depth at which this occurs should be stated.
4.1.4.	Harmonic Distortion	... % at ... % depth of modulation	This is the maximum value of the harmonic distortion after linear detection of the modulated wave up to the maximum depth of modulation for which the instrument is designed. The depth of modulation at which this occurs should be stated.
4.2.	**External Amplitude Modulation** (Sinusoidal)		
4.2.1.	Modulation Frequency Range	From ... c/s to ... c/s	Where there is any limitation due to the value of the carrier frequency, details should be stated.
4.2.2.	Modulation Depth Range	From ... % to ... %	Where there is any limitation of the value of the carrier frequency, details should be stated.
4.2.2.1.	Amplitude/ Frequency Characteristic	... %	This is the maximum change in the depth of modulation over the specified modulating frequency range.
4.2.3.	Input Voltage (r.m.s.)/Depth of Modulation	... volts (r.m.s.) into ohms are required to produce ... % depth of modulation	
4.2.4.	Harmonic Distortion	... % at ...% depth of modulation	This is the maximum value of the harmonic distortion after linear detection of the modulated wave up to the maximum depth of modulation for which the instrument is designed. The depth of modulation at which this occurs should be stated.
4.2.5.	Non-linearity	Not greater than ...	This is the value of the *non-linearity* at any modulating frequency within the specified range.

	Characteristic	Method of expressing the characteristic	Remarks
4.3.	**Internal Frequency Modulation** (Sinusoidal)		
4.3.1.	Modulation Frequencies	... c/s, ... c/s, etc.	If monitoring facilities are provided, these should be stated.
4.3.1.1.	Modulation Frequency Calibration Accuracy	... %	This is the maximum error in the value of the modulation frequency in relation to the calibration of the modulation frequency control.
4.3.2.	Deviation Range	Fixed at ... c/s/ Variable up to ...c/s	Where there is any limitation due to the value of the carrier frequency, details should be stated.
4.3.2.1.	Deviation Accuracy	... % of the frequency deviation	This is the maximum error in the value of the deviation in relation to the calibration of the deviation control.
4.3.2.2.	Amplitude/ Frequency Characteristic	... %	This is the maximum change in the frequency deviation over the specified modulating frequency range.
	(Where more than one modulating frequency is provided.)		
4.3.3.	Unwanted A.M.	... at ... c/s deviation	This is the maximum value of the *r.m.s. modulation factor (a.m.)* under any condition of wanted f.m. up to the frequency deviation. The deviation at which this occurs should be stated.
4.3.4.	Harmonic Distortion	... % at ... c/s deviation	This is the maximum value of the harmonic distortion after linear detection of the modulated wave up to the maximum frequency deviation for which the instrument is designed. The deviation at which this occurs should be stated.
4.4.	**External Frequency Modulation** (Sinusoidal)		
4.4.1.	Modulation Frequency Range	From ... c/s to ... c/s	Where there is any limitation due to the carrier frequency, details should be stated.
4.4.2.	Deviation Frequency Range	From ... c/s to ... c/s	Where there is any limitation due to the carrier frequency, details should be stated.
4.4.2.1.	Amplitude/ Frequency Characteristic	... %	This is the maximum change in the frequency deviation over the specified modulating frequency range.
	(Where more than one modulating frequency is provided.)		
4.4.3.	Input Voltage (r.m.s.)./ Frequency Deviation	... volts (r.m.s.) into ... ohms are required to produce ... c/s deviation	

	Characteristic	Method of expressing the characteristic	Remarks
4.4.4.	Harmonic Distortion	... % at ... c/s deviation	This is the maximum value of the harmonic distortion after linear detection of the modulated wave up to the maximum frequency deviation for which the instrument is designed. The deviation at which this occurs should be stated.
4.4.5.	Non-linearity	Not greater than ...	This is the value of *non-linearity* at any modulating frequency within one specified range.

APPENDIX OF DEFINITIONS

These new definitions have been taken from R.E.M.C./24/FR, Issue 1, May 1956

1. *Source Impedance*

(*a*) For signal generators having coaxial or two-terminal outlets the source impedance is the impedance presented at a specified plane in the outlet.

(*b*) For signal generators having waveguide outlets the source impedance is the impedance presented at the outlet in a specified plane for the mode for which the signal generator is designed.

2. *Nominal Source Impedance*

(*a*) For signal generators having coaxial or two terminal outlets the nominal source impedance is the specified source impedance for which the signal generator is designed.

(*b*) For signal generators having waveguide outlets the nominal source impedance is the wave impedance of the guide for the mode for which the signal generator is designed.

Note.—In accordance with current practice it is assumed that the nominal source impedance is non-reactive.

3. *Source E.M.F.*

The source e.m.f. of a signal generator is the r.m.s. value of the open-circuit voltage at the outlet in a specified plane.

Note.—This definition is applicable to signal generators of low *nominal source impedance* for use under substantially open-circuit conditions.

4. *Equivalent Source E.M.F.*

The equivalent source e.m.f. of a signal generator is twice the r.m.s. value of the voltage produced across a non-reactive load equal to the *nominal source impedance* at the fundamental frequency when measured at a specified plane in the outlet.

Note.—(i) This definition is applicable to signal generators for use with an external load substantially equal to the *nominal source impedance* at the fundamental frequency.

(ii) This definition is an alternative to *equivalent available power.*

5. *Equivalent Available Power*

The equivalent available power of a signal generator is the power dissipated in a non-reactive load equal to the *nominal source impedance* at the fundamental frequency when measured at a specified plane in the outlet.

Note.—(i) This definition is applicable to signal generators for use with an external load substantially equal to the *nominal source impedance* at the fundamental frequency.

(ii) This definition is an alternative to *equivalent source e.m.f.*

Comment: For signal generators having a waveguide outlet the conditions for the measurement of *equivalent available power* will obtain when the resistive load has a waveguide inlet of the same internal dimensions as the outlet of the signal generator and a standing wave ratio of unity at the fundamental frequency for the mode for which the signal generator is designed.

6. *Reference Level*

The reference level is the output level at which the *source e.m.f.*, the *equivalent source e.m.f.* or the *equivalent available power* of a signal generator is measured and to which measurements on the attenuator system are referred.

7. *Mean Carrier Voltage Level*

The mean carrier voltage level of an amplitude-modulated oscillation is the mean amplitude of the wave over one period of the modulation.

8. *Effective Modulation Factor*

(*a*) The *effective modulation factor* of an *f.m.* oscillation or *p.m.* oscillation is the effective frequency swing divided by the frequency deviation.

Note.—This definition involves the use of an arbitrary value representing deviation for 1·0 modulation factor.

(*b*) The *effective modulation factor* of an *amplitude modulated* oscillation for a sinusoidal modulating oscillation is the amplitude of the component of the rectified modulated oscillation, having the frequency of the modulating oscillation divided by the value of the d.c. component of the rectified modulated oscillation.

Note.—*Effective modulation factor* may be expressed as a percentage.

Comment: The definition of *effective modulation factor* is useful in that it provides a means for uniform assessment of modulation capability in cases where the important criterion is the power in the wanted sideband.

9. *R.M.S. Modulation Factor (A.M.)*

The r.m.s. modulation factor of an amplitude-modulated oscillation is $\sqrt{2}$ times the r.m.s. value of the rectified modulated oscillation divided by the value of the d.c. component of the rectified modulated oscillation.

Comment: This definition is needed in order to assess the magnitude of unwanted amplitude modulation of an otherwise unmodulated carrier or in the presence of wanted frequency or phase modulation.

10. *R.M.S. Frequency Swing (F.M.)*

The r.m.s. frequency swing of a frequency-modulated oscillation or a phase-modulated oscillation is the frequency swing of a modulated oscillation having a sinusoidal modulated response, and of the same r.m.s. value after linear demodulation as the modulated oscillation.

Comment: This definition is needed in order to assess the magnitude of unwanted frequency or phase modulation of an otherwise unmodulated carrier or in the presence of wanted amplitude modulation.

11. *Non-linearity*

The non-linearity in the modulation characteristic of a system employing amplitude modulation, frequency modulation or phase modulation for a sinusoidal modulating oscillation is the maximum departure of the *effective modulation factor* (*a.m.*, *f.m.*) from the value corresponding to a linear characteristic between zero and the reference effective modulation factor; the reference effective modulation factor being 0·9 for amplitude modulation and 1·0 for frequency modulation or phase modulation.

Note. — Expressed mathematically non-linearity is the maximum value of L given by

$$L = \pm\left(M_a - \frac{M_r}{V_r} \cdot V_a\right)$$

where M is the effective modulation factor

V is the amplitude of the external sinusoidal modulating oscillation

V_r is that value of V required to produce the reference effective modulation factor of $M_r = 0{\cdot}9$ for a.m. or $M_r = 1{\cdot}0$ for f.m. or p.m.

V_a is any other value of V less than V_r which produces an effective modulation factor of M_a.

Comment: Non-linearity in the modulation characteristic of a signal generator is an unwanted effect under the condition of an externally applied modulating oscillation. For an internally applied modulating oscillation the characteristic of the modulating system is taken into account by the calibration of the scale indicating effective modulation factor or effective frequency swing.

RADAR SIMULATORS*

by

L. J. Kennard† and C. H. Nicholson (Graduate)†

A paper presented at the Convention on "Electronics in Automation" in Cambridge on 29th June 1957. In the Chair: Professor D. G. Tucker (Member)

SUMMARY

The need for radar simulators for personnel training and evaluation purposes is discussed. The equipment must faithfully reproduce the characteristics of the target as well as of the radar, and the performance specification on the synthetic targets is laid down. Radar pulse widths and repetition rates present no problem. The necessity for a true antenna pattern function generator is considered in some detail. Various methods of analogue computation to determine the position of the targets are discussed. Two computing systems operating from 400 c/s supplies are described.

1. Introduction

This paper deals chiefly with ground- or ship-to-air radars of the early warning and surveillance type but mention is made of fire control and air-to-air installations.

All simulators aim at reproducing, under controllable conditions, situations where human decisions are of great value. The greatest effort has gone into simulating those situations which are fraught with danger, loaded with consumable cost and which cannot be stopped, reversed and repeated.

Air traffic, air warfare, surface vessel movement and naval warfare are all covered by the above definition, and since they are all tracked, and in some cases controlled, by radar, a simulator which presents radar information for the tactical marshalling of aircraft or ships is a very useful tool.

As in all simulators, the radar simulator can be used for training and, if reasonably realistic, for evaluation also.

Typical sub-divisions of these uses are:—

Training in radar observation and interpretation of data;
in ground to flight control;
in tactical handling of aircraft (when used in conjunction with a flight simulator).

Evaluation of personnel;
of tactical methods;
of radar equipment.

2. General Considerations

In many of the applications mentioned above where the emphasis of the complete exercise is on tactics, there is a voice link which may be from one to many, as, for example, from one ground controller to many aircraft. This may imply using unskilled personnel as target pilots and a high degree of realism in the target controls then becomes a nuisance—simplicity is the keynote.

On the other hand, where the simulator is used to evaluate equipment, human control is probably better dispensed with and a repeatable programming device used. The value of the simulator here is that the parameters of an aircraft, ship or radar installation which only exist on paper may be quickly and cheaply set up and tried under near operational conditions.

In both applications, the resulting output appearing on the presentation display, be it p.p.i., range and height indicator (r.h.i.), A-scope, etc., must approach realism if the device is to be truly useful. All simulators however, stop somewhere short of realism, although flight simulators go far, reproducing air buffeting, vibration, noise, and control feeling.

Some of the earlier radar simulators lacked realism where realism was tactically needed.

* Manuscript received 31st May 1957. (Paper No. 433.)
† Solartron Electronic Group Ltd., Dorking, Surrey.
U.D.C. No. 621.396.96:681.142

The reduction of flight manoeuvres to the polar co-ordinates of a radar p.p.i. screen is, in its simplest form, an easy challenge to the computer engineer. However, a simulator producing a collection of bright dots which trail about the screen, more uniform in magnitude than the stars and planets, regardless of coverage zones, antenna patterns and target manoeuvre parameters is tactically unrealistic, and when electronic countermeasures (ECM) are brought in, its value becomes virtually nil.

To sum up the general requirements therefore:

(*a*) The target controls do not need to be very realistic since this is not a flight simulator.

(*b*) The target manoeuvres must be realistic and at the same time the synthetic ship or aircraft should be incapable of exceeding the parameters of manoeuvrability of their real counterparts.

(*c*) The equipment should reproduce faithfully the characteristics of a specific radar installation with preferably sufficient flexibility to allow different radar types to be simulated.

These points will now be considered in greater detail.

3. Target Control Requirements

To be able to simulate modern fighter aircraft, a maximum speed of 1000 knots with rates of turn, port or starboard, up to 15°/sec (alternatively 7·5 *g*) is required; coupled with these a rate of dive or climb of 15000 ft/sec is needed. Wind speeds up to 100 knots in any direction is a basic requirement, but this may well be elaborated into a number of wind layers of various heights and "widths" with jet stream speeds up to 600 knots.

The ceiling height of the aircraft should be at least 50,000 feet and the range capabilities of the target are, in the case of a single radar installation, basically dependent on the range of the radar—200 nautical miles is usually adequate; this obviously does not apply where a large exercise area with several widely spaced radar sites are involved, and in any case different computation techniques are used.

Missile targets come into a category of their own in which the velocity, rates of turn, dive and climb are doubled or even trebled and the ceiling height would be 100,000 feet or more.

Surface vessel targets should be capable of a maximum speed of 60 knots with a rate of turn of up to 6°/sec.

The controls for velocity, rate of turn and rate of dive/climb can be straightforward calibrated knob type dials of 3 in. diameter, and each should carry some mechanical device for limiting the actual movement of the dials so that the performance parameters of low speed targets cannot be exceeded.

The outputs from these controls for a training device do not need to be particularly accurate—2 to 3 per cent. is adequate, since no actual pilot can hold his rate of turn or rate of dive/climb to better than a few per cent. Airspeed, too, is subject to wind buffeting and errors in setting, so that output accuracies better than 1 per cent. are really a waste of high quality components. Similar limits are applicable to missile and surface vessel control units.

The layout of a suitable control panel for a two-dimensional controlled aircraft (i.e. no height), is shown in Fig. 1. It will be seen that there are other switch controls on this panel for the target to hover or fly, auto pilot, "window" release, parachuted noise jammer release, and noise jammer aircraft; in addition there are three indicator lamps showing target visible, off scale and boundary, and two drum dials giving the cartesian co-ordinates of position. These switches and indicators carry out the following functions:

(*a*) *Hover/fly switch*—although this enables the operator to simulate a helicopter, its prime use is for "parking" targets beyond the range of the display so that the controller can call upon an invisible attacking force as and when required—the indicator lamp "off scale" is then illuminated automatically.

(*b*) *Auto pilot*—this switch removes the rate of turn voltage from the heading angle integrator, thus allowing the target to fly on a straight course at some fixed heading angle.

(*c*) *"Window" release*—releases a blob of "window" from the parent aircraft (or shell fired "window" from a surface vessel) which then drifts with the wind, expands in three dimensions and sinks slowly to the ground. ("Window" is the code name for a radar countermeasure which consists of lengths of metal foil, resonant at or about the frequency of the radar.)

(*d*) *Parachuted noise jammer release*—switches on a wide-band noise jammer which also drifts with the wind and loses height at some predetermined rate.

(*e*) *Noise jammer aircraft*—switches on a wide-band noise jammer which is carried along with the aircraft; this effectively obliterates its position on the radar screen by saturating the receiver.

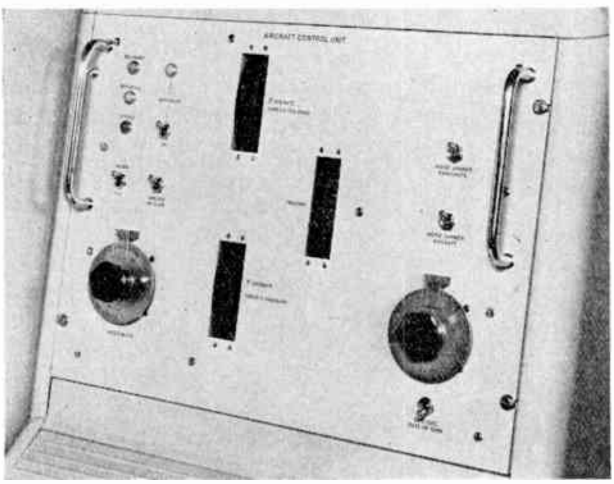

Fig. 1. Control panel for a two-dimensionally controlled aircraft.

(*f*) *Indicator lamps*—"visible" and "off-scale" are self-explanatory; the boundary lamp indicates that the maximum range of the computer has been reached. It hunts in this position and flashes the appropriate lamp to attract the attention of the operator.

(*g*) *Cartesian co-ordinate dials*—these are not normally used during an exercise, their prime purpose being for setting the initial position of the target.

A three-dimensional aircraft control panel, which is a standard item, would also carry a rate of dive/climb control and a height indicator.

The above controls are considered to be the minimum required for a tactical training device but there are many possible additional refinements, including the following:

(*a*) Fuel computation indicators.

(*b*) Desired heading and height controls whereby, once set, the target will climb to the required height at the correct rate and stay there; similarly it will turn to the desired heading angle at the correct rate of turn and stay on that course.

(*c*) Speed control calibrated in Mach number so that speed varies with height up to approximately 36,000 feet.

(*d*) Aircraft to go gradually into a climb or dive with limited acceleration ("*g*").

(*e*) Loss of speed during turn.

(*f*) Automatic speed, turn and climb limitations with height according to characteristics of aircraft.

(*g*) Computed ground speed dependent on climb or dive angle.

These additional refinements tend to be rather expensive and although desirable from a pilot's point of view, they add only slightly to the overall realism of the equipment.

The position of the control panels in the main equipment is dependent to a large extent on the requirements of the customer. One school prefers to keep all the control panels for the complete installation together with the computing equipment as a separate entity, whilst others prefer to combine two control panels and computers together to make a console arrangement which is repeated several times. An example of this latter arrangement, which is preferred by the authors, is shown in Fig. 2, and the layout of an operations room using the individual console arrangement, together with a flight simulator section is shown in Fig. 3.

4. Radar Characteristics

Most radar simulators operate at video frequency and the equipment with which the authors have been concerned is no exception. Most radar anti-clutter devices such as "moving-target indication," "fast time-constant," "instantaneous a.g.c.," "detector balance bias," etc., can be simulated at video frequency and it is only when very special and difficult circuits are used for some specific purpose that it is necessary to utilize the i.f. strip of the radar receiver to ensure realistic presentation.

The basic radar p.r.f. is easy enough to produce but the simulated pulse which feeds the display is not simply the transmitted square pulse but one that is shaped by the bandwidth (often restricted) of the receiver i.f. strip—this is normally reproduced by limiting the bandwidth of the video amplifiers.

Having obtained a video p.r.f. of the correct width and rise time, it is necessary to gate these pulses through an antenna function generator.

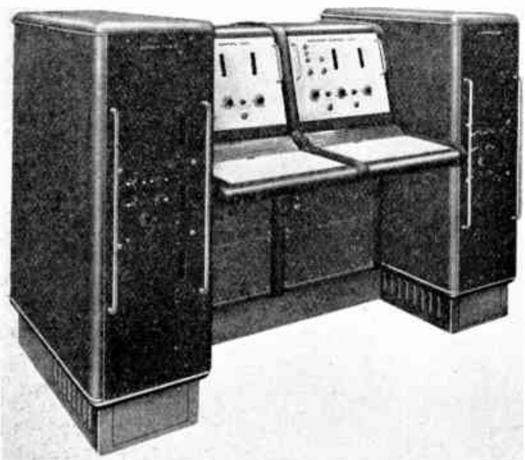

Fig. 2. Console type of control desk including computers for two aircraft.

A straightforward electronic gate is not good enough especially where active jamming sources are likely to be used. Every antenna system has minor lobes and it is a basic fundamental of antenna design that the finer the main beam the greater the number of minor lobes, albeit they are very much reduced in level. However, these minor lobes, especially those adjacent to the main beam can be "lit up" on a strong passive signal return, and when an active jammer is in operation these minor lobes, as well as the main beam, can be saturated. This then produces on the p.p.i or r.h.i. a very much wider arc of noise than would be expected from the main beam alone.

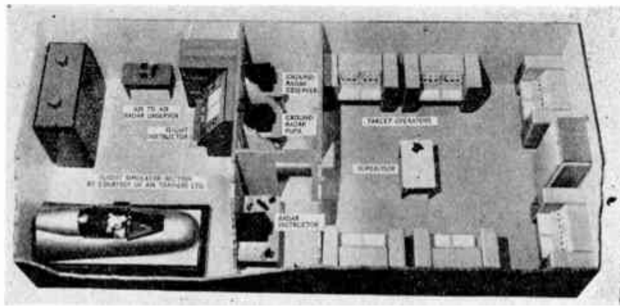

Fig. 3. Model operations room for radar simulator and flight simulator.

In addition, to be truly realistic the antenna pattern produced for passive echoes (i.e. echoes from non-transmitting targets) should be different to that produced for active signals since the gain of the antenna system for passive signals is the square of that for active ones; the minor lobes therefore are more predominant on one way (active) signals as opposed to two way (passive) signals. To achieve realistic antenna simulation using purely electronic methods is not impossible but the equipment would be very complex, very expensive, and subject to much adjustment. Simple electronic gates, being switching devices, are not good enough since they do not reproduce the build-up of the antenna pattern and so cannot be considered.

A suitable electro-mechanical antenna function generator which has proved to be very successful consists of a metal disc rotating between two capacitive electrodes. Holes and slots of suitable shape and size to represent the antenna azimuth main and important minor lobes are cut in the disc; transfer of energy occurs between the electrodes when a slot is opposite them and complete attenuation is achieved when the blank disc is present. To obtain a reasonable transfer between the probes it is necessary to rotate the disc at some multiple frequency of its nominal since the slot for a 1 deg. beam is only 0·035 in. wide on the periphery of a 4-in. disc. By rotating the disc at 30 times its nominal rate it is possible to increase the size of the holes by 30 times but it is then necessary to gate 29 revolutions out of every 30 to obtain the correct one.

The actual video p.r.f. is fed into these probes and triggered at the correct instant in time by the ranging circuit. The disc is driven through a mechanical differential gear at a speed which is the difference between the antenna rotation rate and the rate of change of bearing, so that it is only when the bearing of the target and the bearing of the antenna coincide that any transfer of energy occurs between the probes. The signals from this azimuth disc are amplified to maintain the signal level and are passed on to a similar disc which has slots in it to represent the elevation pattern of the antenna system. This disc is mounted on a shaft which turns through the elevation angle of the target, so that if the target is flying above or below the vertical coverage of the antenna, no signals are transferred through this disc and hence nothing appears on the display. Unlike the azimuth disc, the elevation disc moves at single speed as the antenna pattern is very much broader in the vertical plane.

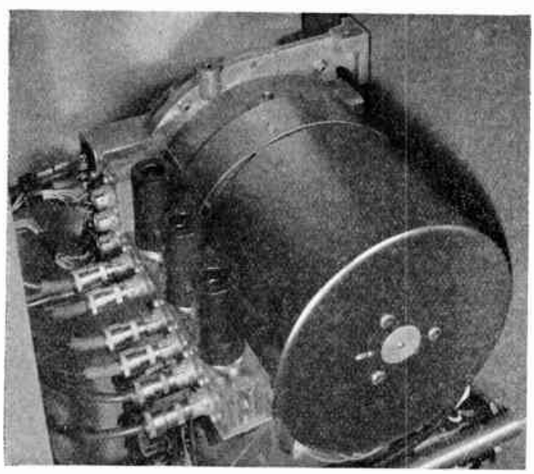

Fig. 4. Azimuth antenna pattern function generator, showing slots for major and minor lobes.

Several discs are mounted on both the azimuth bearing shaft and elevation angle shaft to accommodate back-to-back antenna systems, high, medium and low vertical beams, active signal antenna characteristics, i.f.f., noise, etc., and it will be appreciated that these discs can easily be changed to obtain different antenna parameters, thus making a very flexible system. The discs being used at present are in printed circuit form so that identical antenna characteristics are assured for a large number of targets. Fig. 4 shows an earlier model of an antenna function generator in which a drum was used—the slots for the main and minor lobes being clearly seen.

The output from the elevation discs are then fed into a mixing stage where all the other synthetic target signals are mixed together. They are then modulated with noise and also with a random low frequency to simulate fading and echo "glint."

The noise jammer channels are mixed in a similar manner and i.f.f. signals likewise, and these three sets of signals are then combined in a composite mixer which feeds the requisite number of cathode follower output stages. General receiver noise is also fed in at the final mixing stage and if a swept-gain receiver is being simulated the noise output is also swept to give increase in noise with range.

This method of mixing allows the jamming channels to be separated from the signal channels so that the instructor's display can be fed through a separate composite mixer. The jamming signals can then be cut out by the instructor on his display alone.

Ground clutter and video mapping can be fed into the final mixing stage from a film recorder thus giving an overall picture which approaches the real one very closely. In addition, radar signals from a live radar can also be fed in at the final mixing stage, thus allowing synthetic targets to chase real aircraft! This latter arrangement requires synchronization of the simulator parameters with the antenna rotation rate, p.r.f., and pulse width of the live radar, but this, in general presents little difficulty.

5. Target Position Computing

There are numerous methods of changing the polar co-ordinate velocities of the aircraft to the polar co-ordinates of position relative to the radar. D.c. electronic computer techniqes involve tiresome capacitor problems due to the long time-constants involved, and since shaft rotations are needed anyway for the antenna function generators used, electro-mechanical computing techniques seemed the obvious choice. Flight simulator computers have gradually changed over to a 400 c/s system

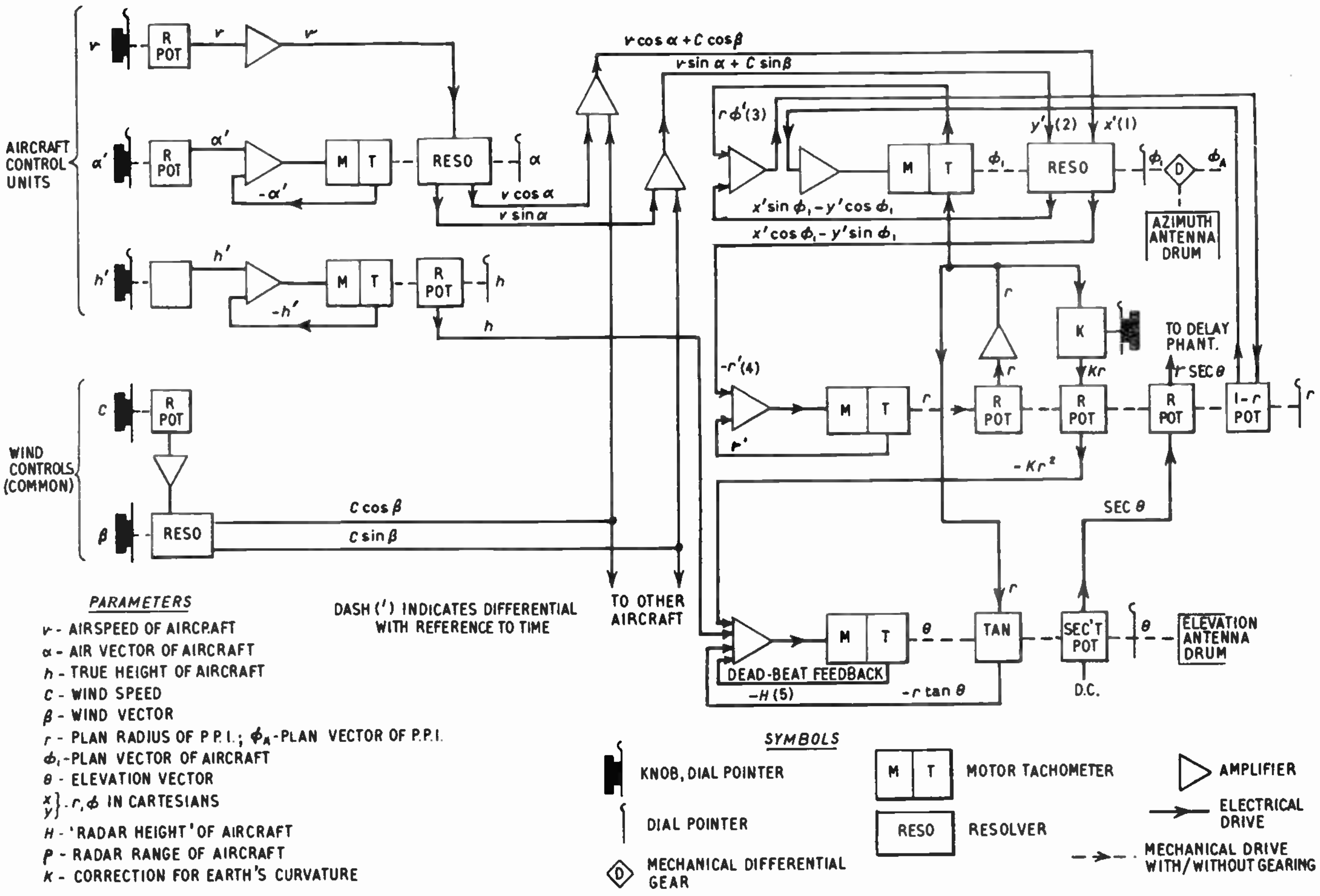

Fig. 5. Block diagram of computer of the first system described in the paper.

and since radar simulators may well be used with flight simulators, a 400 c/s system for the computation chain gives immediate co-ordination of supplies and positional voltages; in addition 400 c/s resolvers and motor tachometers are conveniently small in size compared with 50 c/s and d.c. machinery, and the accuracy of modern 400 c/s resolvers, synchros, etc., is quite acceptable for the specification of a radar simulator system.

A system which avoids the use of precision sine and cosine potentiometers is preferable, due to their large size and high cost. The two systems to be described meet this requirement and each has its own merits.

The first system describes a method of computing direct from the target velocity, rate of turn, and climb to the polar co-ordinates of position, whilst the second system converts the target rates into cartesian co-ordinates of position and then into polar co-ordinates of position.

The first system is suitable for a single radar base simulator where there is no likelihood of a second distant radar base having to "see" the same targets, whereas the second system can be married into as many radar bases as one cares to scatter over an arbitrary exercise area.

5.1. *First System of Computing*

Figure 5 shows the computer block diagram of the first system.

A stable source of voltage at 400 c/s feeds the velocity, rate-of-turn and rate-of-dive potentiometers. The output from the velocity potentiometer feeds the input winding of an electrical resolver. The rate-of-turn voltage feeds an amplifier and motor-tachometer rate servo loop, in which the tachometer output is fed back in phase opposition to the input voltage of the servo-amplifier. This functions as an integrating servo loop in which the output shaft of the motor-tacho turns through an angle proportional to the integral of the input rate, namely, the heading angle. By using a suitable ratio of step down gearing, the output shaft of the gearbox turns through the actual heading angle: this shaft drives the input shaft of the electrical resolver, already mentioned, and the output windings of the resolver produce electrical voltages analogous to the cartesian co-ordinates of velocity of the target. These voltages are mixed in two adding stages with the velocity cartesian co-ordinates of the wind, which are obtained from another electrical resolver arrangement similar to that already described. These outputs then, are the target cartesian velocities wind corrected.

The rate of dive or climb control potentiometer feeds a similar integrating servo to the rate-of-turn, and the output shaft of the motor-tacho drives via a suitable gearbox, a precision potentiometer to produce a voltage analogous to height.

The three voltages so far produced are fed into three cross-coupled servo loops which compute the bearing angle, the slant range and the elevation angle of the target from the radar.

The cartesian velocities feed the input windings of another electrical resolver, one output of which feeds a servo amplifier driving a motor-tachometer and gearbox; this gearbox drives the shaft of the electrical resolver and the loop stabilizes when the resolver shaft turns to the bearing angle of the target. The shaft also drives an input shaft of a mechanical differential gear, the second input being the antenna rotation rate; the output shaft, which is a rotation difference between target bearing rate and antenna rate, drives the antenna azimuth pattern discs already described.

The second electrical output of this resolver is a voltage proportional to the rate of change of range; this is fed into an integrating servo to produce a shaft rotation proportional to range. The shaft carries four potentiometers, the functions of which are more easily understood by reference to the Appendix—one produces a d.c. voltage proportional to the slant range required for the ranging circuits, a second is part of the bearing angle integrating servo loop and is inserted to maintain the loop gain at small ranges, a third feeds the range voltage into the bearing angle tachometer and also feeds the fourth potentiometer to produce a correction factor for the curvature of the earth.

The aircraft's height above ground as an analogue voltage is fed into the servo amplifier of a positional servo loop together with the earth's curvature correction voltage, the radar height voltage and some dead-beat feedback (to prevent overshooting). This servo loop produces a shaft rotation proportional to the elevation angle of the target, which drives the antenna vertical coverage pattern discs and two potentiometers. One of these potentiometers feeds a similar component on the range shaft, to produce a voltage proportional to the slant range, the other is part of the elevation angle servo loop.

It will be noted from this diagram that the correction factor K for the curvature of the earth has a preset control; this enables the effective curvature of the earth to be varied so that sub-normal, normal and superrefractive propagation conditions can be simulated.

One of the problems associated with this type of computer in which cartesian velocities are integrated to give polar positions, is the target which insists in flying dead overhead. This puts a grave demand on the bearing angle servo-amplifiers because they are given a sudden reversal of information and become grossly overloaded. In a positional servo loop this may not matter because the error voltage will eventually fall to zero provided the input information is not fed in at a greater rate than the response time of the servo. The servos in question however, are integrators and do not have a chance of "catching up," so that an integration error is bound to be produced. This is very serious in the bearing angle computation because the target is required to fly into the centre at some known heading and come out the other side at an angle 180 deg. opposed; in actual fact it would come out at the angle at which the resolver shaft happens to be when the overload voltage in the amplifier disappears.

This can be calculated but it does not help solve the problem since it is not practical to have an amplifier capable of handling infinitely large signals: in any case the response time of the overall electro-mechanical servo loop is bound to be too long even if the amplifier is incapable of overload. There are several ways of overcoming this difficulty, two of which have been tried and proved successful. The first is to prevent the overload signal from occurring by deviating the target from its course through the centre, the second is an integrator with a memory which allows the integrand to catch up.

5.1.1. "Dodging the pole"

In the first arrangement a system has been developed known as "dodging the pole"—this is a mathematical exercise in polar co-ordinates. When the target reaches a point on a circle of 2 miles radius from the centre, its cartesian velocity input information from the control panel is removed and substituted by information which will make it fly a semi-circle. At the same time its velocity is increased by $\pi/2$ so that it reaches its diametrically opposed entry point at the same time as its direct course through the centre would have taken; the target is then reconnected to its control information. This system has one minor disadvantage in that an aircraft hitting the 2 mile radius "guard ring" at a tangent is also pushed round a semi-circle so that it comes out the other side 4 miles offset but parallel to its original course.

This is considered to be a better fault (if one can grade faults as better or worse) than a system which produces an erroneous bearing angle. However, the centre of a p.p.i. display is a blank zone due to ground clutter and lack of vertical coverage so that it does not really matter what happens to the target when it is in this zone provided it is in the correct position when it re-appears—a 4 mile change of position on a 200 mile p.p.i. is only just discernible.

5.1.2. Integrator with memory

The second device, an integrator with a memory, is an electro-mechanical arrangement using an M motor. The M motor windings are fed with driving potential via high speed relays which in turn are transistor energized. By a suitable choice of relay energization, it is possible to drive the M motor forwards or backwards and the shaft rotation becomes the required integrand.

The voltage to be integrated is fed to a Miller integrator normally working at the middle of its range, with a senser to observe the output rising or falling. This senser drives the input of the motor relays. The motor as it rotates charges a capacitor C to a datum voltage at each step, and discharges it on to the Miller grid. The Miller anode comes down by a fraction fixed by the ratio of two precision capacitors (C) and no output will be supplied to the motor input circuit until it rises again. When an overload occurs the Miller will accumulate back pulses up to the limit of its sweep and the motor will just clock these off as fast as it can and thus retain the integrand. The system as described is capable of handling overload of ten times and by introducing "snubber" circuits which increase the Miller time-constant in the top and bottom regions of its sweep, the overload figure can be considerably increased.

As already mentioned, this computation system is only suitable for single platform radars since the cartesian co-ordinates of position are not available in the chain for exchange of positional information with another platform. It is, however, economical in equipment because this stage of computing machinery is eliminated.

The computer cabinet housing the electronic units, gearboxes etc., for such a system is shown in Fig. 6.

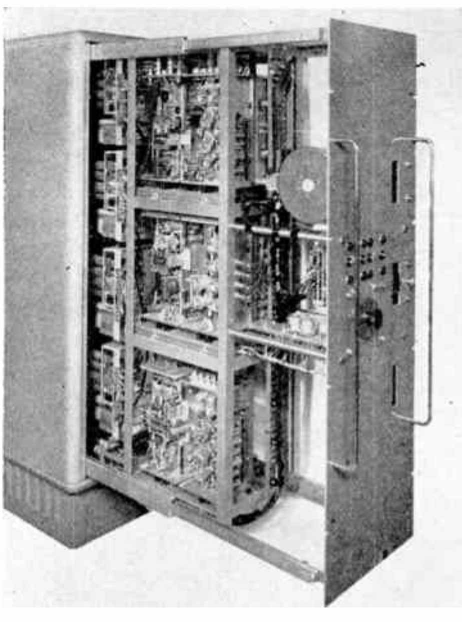

Fig. 6. Polar computer and antenna function generator unit showing the main electronic circuits.

5.2. *Second System of Computing*

Figure 7 shows the computer block diagram of a second and more flexible system.

Like the first system a stable source of voltage at 400 c/s frequency feeds the target velocity, rate of turn, rate of dive/climb and wind velocity control potentiometers.

The target velocity voltage feeds an electro-mechanical integrator consisting of servo-amplifier, motor tachometer and gearbox. The target velocity as a voltage is therefore integrated to a distance given as a shaft rotation. This then drives the input shaft of a mechanical ball resolver.

The angular input ring of the ball resolver is turned by the heading angle indicator (compass) which is derived from the integration of the rate of turn. The outputs from the ball resolver are two shafts which give the sine and cosine functions of the heading angle with respect to the input shaft—these are the cartesian co-ordinates of position as shaft rotations.

In a similar manner the wind velocity and direction is fed to another ball resolver to produce the wind cartesians of position, again, as two shaft rotations. These shaft positions are transmitted by M transmitters and receivers to all the control panels and in each one they are mixed in a pair of differential gears with the target cartesian shafts to give the wind-corrected co-ordinates of position. These wind-corrected shaft rotations are then stepped down through gears of suitable ratio to drive through clutches a pair of precision potentiometers; the potentiometers give a voltage analogous to the cartesian co-ordinates of position, and the clutches allow the potentiometers, and thus the position of the target, to be set to any position prior to an exercise.

The rate of dive/climb potentiometer also feeds an electro-mechanical integrating servo which produces a shaft rotation proportional to the rate integral, namely height; the shaft then drives a precision potentiometer to produce a voltage analogous to height.

Having produced the cartesian co-ordinates of position and height as voltages, they are then fed to a triangle solving computer which converts the cartesians to polar co-ordinates, range and bearing, and the height to the elevation angle of the target.

The x and y voltages feed, via a pair of unity gain drivers, an electrical resolver which is in a positional servo loop. One output of the resolver feeds a servo-amplifier, motor tacho-meter and gearbox which in turn is coupled to the shaft of the resolver; the loop stabilizes

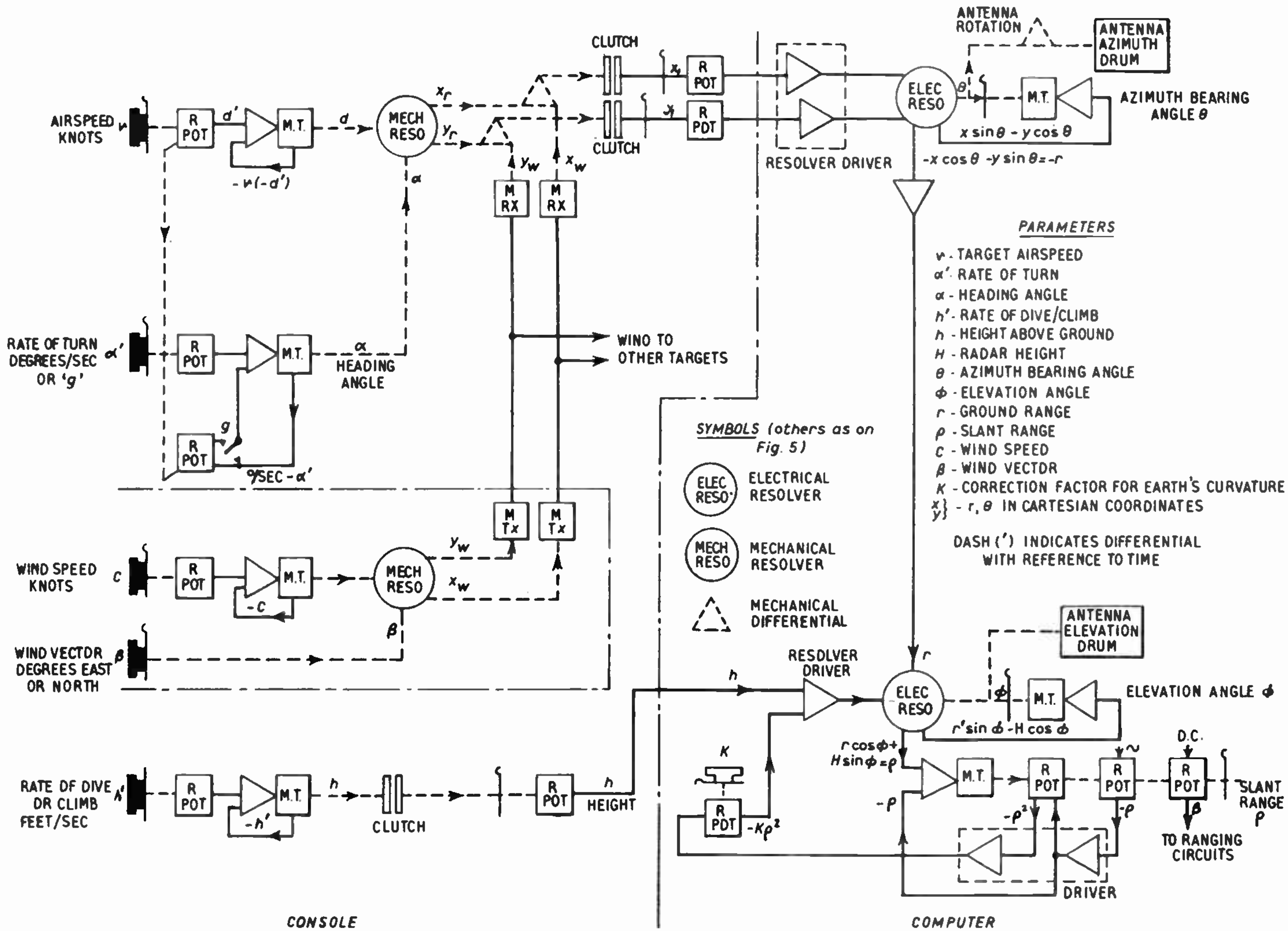

Fig. 7. Block diagram of computer of the second system described.

when the resolver shaft turns to the bearing angle of the target. The bearing angle shaft is also coupled to one input shaft of a mechanical differential gear, the other input being the antenna rotation rate; the output of the differential then drives the antenna azimuth discs.

The other output from the bearing angle electrical resolver is a voltage proportional to ground range and this is fed to a second electrical resolver via a driver stage. This second resolver also receives an input proportional to the radar height of the target and two output voltages are given. One feeds the amplifier of a positional servo loop in which the resolver shaft is the final output. The servo stabilizes when the resolver shaft turns to the elevation angle of the target and this in turn drives the antenna elevation coverage discs and an indicator dial. The second output from the resolver is an a.c. voltage proportional to slant range but, as a d.c. voltage is required for the ranging circuits, a further positional servo is needed to produce a shaft rotation to drive a precision potentiometer.

This slant-range servo shaft carries two other potentiometers besides the one used for triggering the ranging circuits; one is used to produce the opposing voltage to stabilize the servo whilst the second potentiometer receives the input of the first to produce a voltage proportional to the slant range squared. This is applied to a potentiometer with a preset control for curvature of earth correction. The output voltage from this potentiometer is then applied to a subtracting stage where the height voltage is also injected; the output from this stage, the difference of the two inputs, is the true radar height of the target and this feeds one of the input windings of the elevation angle resolver.

This arrangement will produce accuracies better than 0·5 deg. on bearing and elevation angles and ranging d.c. voltage is within 0·1 per cent. The accuracy of the position of the "blip" on the c.r.t. depends on the ranging circuitry and 1 per cent. is adequate for normal purposes. Evaluation equipment will need 0·1 to 0·2 per cent. ranging accuracy and this can be achieved with more complex circuitry.

This second system of computation allows positional information to be exchanged between various radar platforms and the system also allows moving platforms to be incorporated for ship or airborne search radars.

6. Other Types of Radar

The basic problem of computing the polar co-ordinates of position of a moving target to a fixed or moving radar platform have been elaborated and a realistic system of producing on a p.p.i. a true time series echo giving genuine antenna characteristics has been explained. These same principles can be used for nodding and vee beam height-finders and, with complication, conical scan fire control and guided missile beam-riding radars can be simulated. Air-to-air radar simulation presents problems especially if the scanner is not stabilized; the yaw, pitch and roll of an aircraft bring in three more variables but these in themselves are not formidable obstacles. Some of the more complicated spiral scanning methods, however, can be tiresome if a fully realistic presentation is required.

7. Conclusions

The authors of this paper have endeavoured to give a general description of some radar simulation equipment with which they have been concerned and at the same time show some of the difficulties in presenting a truly realistic picture without undue complication of equipment. The design problems of individual units have purposely been omitted in order to keep the paper down to a reasonable length. The mathematics of the two computer systems described are given as an Appendix.

8. Acknowledgments

The authors wish to thank the Directors of the Solartron Electronic Group Limited, for permission to publish this paper.

9. Appendix: Computer Mathematics of the two Systems Described in the Paper.

V = target airspeed
α = heading angle
α' = rate of turn $(= \mathrm{d}\alpha/\mathrm{d}t)$
h = true height of target
h' = rate of dive/climb $(= \mathrm{d}h/\mathrm{d}t)$
C = wind speed
β = wind vector
r = ground range of target
ρ = radar (slant) range of target

θ = target azimuth bearing angle
φ = target elevation angle
d = distance travelled by target
Δ = wind "distance"
x, y = cartesian co-ordinates of position
H = radar height of target
K = correction factor for earth's curvature
$'$ denotes a differential with respect to time, e.g. $\alpha' = d\alpha/dt$.

9.1. *System One* (Fig. 5)
Equations

(1) $x' = V\cos\alpha + C\cos\beta$
(2) $y' = V\sin\alpha + C\sin\beta$
(3) $r\theta' + x'\sin\theta - y'\cos\theta = 0$
(4) $r' - x'\cos\theta - y'\sin\theta = 0$
(5) $h - Kr^2 - H = 0$
(6) $\rho = r\sec\theta$
(7) $H = r\tan\theta$

9.1.1. Target controls

The target velocity V, is fed into a resolver whose shaft follows the air vector of the target, α. α is obtained by integrating the rate of turn α' in a servo amplifier, motor-tachometer arrangement.

The output from the resolver is two electrical voltages

$$V\cos\alpha \text{ and } V\sin\alpha.$$

These voltages are mixed with similar ones derived from the wind velocity and direction

$$C\cos\beta \text{ and } C\sin\beta$$

to give $V\cos\alpha + C\cos\beta$ and $V\sin\alpha + C\sin\beta$

Let $V\cos\alpha + C\cos\beta = dx/dt = x'$
and $V\sin\alpha + C\sin\beta = dy/dt = y'$

9.1.2. Bearing angle, θ

x' and y' are fed into the θ resolver, the shaft of which resolves the bearing angle, θ. The resolver output is $x'\sin\theta - y'\cos\theta$ and $-x'\cos\theta - y'\sin\theta$.

$x'\sin\theta - y'\cos\theta$ feeds the input to the θ servo-amplifier where it is phase opposed by a voltage $r\theta'$ from the θ tachometer. These two components satisfy equation (3) when the resolver shaft turns to θ, the bearing angle.

9.1.3. Range, r

The second output from the θ resolver, $-x'\cos\theta - y'\sin\theta = -r'$ feeds the range servo amplifier together with a voltage r' from the tachometer in opposition to it (eq. (4)). This integrator produces a shaft rotation proportional to the ground range, r.

The r shaft carries four potentiometers which serve the following functions:

(1) Produces a voltage r to feed to (*a*) the tachometer on the θ shaft to produce $r\theta'$, (*b*) the K potentiometer to produce Kr for earth's curvature correction.
(2) Receives the input Kr and gives Kr^2 which is the final value required for curvature correction (eq. (5)).
(3) Receives a d.c. voltage $\sec\varphi$ from the elevation angle shaft and produces $r\sec\varphi = \rho$, the slant range, which triggers the ranging circuit (eq. 6).
(4) This potentiometer is part of the rate servo loop on the θ shaft. It will be seen that one of the input voltages to the θ servo is $r\theta'$, so that as r decreases the loop gain tends towards zero. To overcome this the amplifier is split and a gain control giving $1 - r$ inserted. This needs to be $1/r$ to give constant gain with range; a $1/r$ potentiometer however is a practical impossibility and $1 - r$ is substituted.

9.1.4. Elevation angle, φ

The height, h, which is obtained by integrating the rate of dive/climb, h', feeds the φ servo-amplifier, together with $-Kr^2$ from the r shaft, and positional feedback $-r\tan\varphi = -H$ from a tangent potentiometer on the elevation shaft.

This servo loop stabilizes when these three inputs equate to zero (eq. (5)) and in this condition the mechanical output shaft turns to φ, the elevation angle.

This computation chain therefore produces θ, bearing angle; φ, elevation angle; ρ, slant range from V, target velocity; α', rate of turn; and h', rate of change of height, by solving two simultaneous differential equations and one triangulation equation.

9.2. *System Two* (Fig. 7)

(1) The air velocity V (or d') is derived from a potentiometer and fed into a motor-tachometer integrator servo loop. The electrical output from the tachometer, $-V$, is fed back in phase opposition to the input of the servo. The velocity, V, (or rate of change of distance d') is therefore integrated to give a distance, d, as a shaft rotation, which drives the input shaft of a mechanical ball resolver.

(2) The rate of turn, α', feeds a similar integrating arrangement, and the mechanical output as a shaft rotation is the heading angle, α. This shaft position drives the angular input ring of the ball resolver.

(3) The outputs from the ball resolver are two shaft rotations giving sine and cosine functions of heading angle with respect to d, namely,

$$d \sin \alpha \text{ and } d \cos \alpha$$

These are the cartesian co-ordinates of change of position.

(4) The wind velocity C is integrated to give a distance Δ as a shaft rotation and this drives the input shaft of another ball resolver. The angular input shaft of the ball resolver is turned by the wind direction control β to the wind vector.

The outputs from the ball resolver are two shaft rotations giving sine and cosine functions of wind vector with respect to C, namely,

$$C \sin \beta \text{ and } C \cos \beta$$

These are the cartesian co-ordinates of change of wind "position" and are transmitted to all targets by an "M" motor system.

(5) The outputs from the target ball resolver are mixed with the outputs from the wind ball resolver in mechanical differential gears and these then drive precision potentiometers which give the targets cartesian co-ordinates of position (wind corrected), as a.c. voltages, x and y.

(6) The rate of dive or climb control feeds a voltage h', proportional to the rate of change of height, into a motor-tachometer integrator servo loop, and the output shaft drives, through a suitable step-down gearbox assembly, a precision potentiometer which gives an a.c. voltage analogous to the height, h. This voltage, h, together with the cartesian positional voltages, x and y, are then fed to the main computer where the following equations are solved.

(7) (A) $x \sin \theta - y \cos \theta = 0$

(B) $r \sin \varphi - H \cos \varphi = 0$

(C) $r \cos \varphi + H \sin \varphi = \rho$

where $x = r \cos \theta$

$y = r \sin \theta$

$H = h - Kr^2 \cong h - K\rho^2$

(8) The cartesian co-ordinates of position, x and y, are fed into the windings of an electrical resolver and outputs of $x \sin \theta - y \cos \theta$ and $x \cos \theta + y \sin \theta$ are obtained. $x \sin \theta - y \cos \theta$ is fed back to the input of a servo amplifier and motor-tachometer which drives the resolver shaft.

This constitutes a positional servo and equation (A) $x \sin \theta - y \cos \theta = 0$ is solved when the resolver shaft turns through θ. Hence the bearing angle, θ, is produced from x and y.

(9) The other resolver output, $x \cos \theta + y \sin \theta$ feeds one input of second electrical resolver, and is equal to the ground range, r, i.e.

$$r = x \cos \theta + y \sin \theta$$

This resolver also receives a second input, H, the radar height, and outputs of

$$r \cos \varphi + H \sin \varphi$$

and $r \sin \varphi - H \cos \varphi$ are obtained. $r \sin \varphi - H \cos \varphi$ is fed into a positional servo loop amplifier, and the output shaft of the motor-tachometer drives the resolver shaft.

The servo loop reaches equilibrium when equation (B) is solved, namely

$$r \sin \varphi - H \cos \varphi = 0$$

and this occurs when the resolver shaft turns through the elevation angle, φ.

(10) The second output of the elevation angle resolver is $r \cos \varphi + H \sin \varphi$ and this solves equation (C) as it stands, i.e.

$$r \cos \varphi + H \sin \varphi = \rho$$

Unfortunately, this voltage cannot be used for driving the ranging unit, a d.c. voltage being required, so that another positional servo is called for which carries three precision potentiometers on its output shaft. One of these potentiometers feeds $-\rho$ in phase opposition to the input voltage ρ, and the servo stabilizes when the output shaft turns through an angle proportional to ρ.

(11) The second potentiometer is fed with d.c. and its output triggers the ranging unit, whilst the third potentiometer takes the output, ρ, from the first potentiometer and hence gives an output of ρ^2.

(12) This is then fed through a further independent potentiometer, K, which gives a variable setting for the correction factor for the curvature of the earth, the output being $K\rho^2$. It will have been noticed that the

second input to the φ resolver was H, the radar height. Now $H=h-Kr^2 \cong h-K\rho^2$. $K\rho^2$, therefore, is subtracted from the height, h, to give H, the radar height.

The azimuth bearing angle, θ, the elevation angle, φ, and the slant range, ρ, of the target are thus produced from the target velocity, V, its rate of turn, α', and its rate of dive or climb, h'.

9.3. *Correction for the curvature of the earth, Kr^2*

Consider Fig. 8 (which is out of proportion due to the small radius shown), where

H = radar height
ρ = slant range
r = ground range
h = true height
$= a+H$
R = radius of the earth.

The control unit gives the true height, $h=H+a$, so to get the true elevation angle φ, we need to know the variable, a.

$$\cos\Phi = \frac{R-a}{R}$$

Hence $R\cos\Phi = R-a$

$$\sin\Phi = \frac{r}{R}$$

Hence $\sin^2\Phi = \frac{r^2}{R^2}$

Now $\cos^2\Phi = 1-\sin^2\Phi$

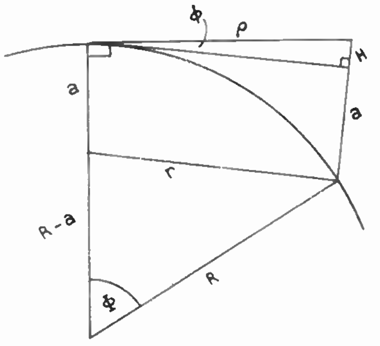

Fig. 8. Correction for curvature of the earth.

Therefore $\cos\Phi = \left[1-\frac{r^2}{R^2}\right]^{\frac{1}{2}} = \frac{R-a}{R}$

$$R-a = R\left(1-\tfrac{1}{2}\frac{r^2}{R^2}+\ldots\right)$$

which, by the Binomial Theorem

$$= R-\tfrac{1}{2}\frac{r^2}{R}$$

Thus $a=\frac{1}{2}\frac{r^2}{R}=Kr^2$ where $K=\frac{1}{2R}$

R varies between infinity and $\frac{1}{2}$ the actual radius of the earth to simulate different propagation conditions which bend the radar beams and effectively change φ.

From this we get the equation,

$$h-Kr^2-H=0$$

DISCUSSION

M. E. H. Downer: Some brief remarks on the Mullard Radar Trainer, now in service with the R.A.F. may be of interest. In this trainer several types of radar are simulated, necessitating a simple, symmetrical change of beamwidth. This is achieved by the use of a magslip on the aircraft bearing shaft fed from a transmitter magslip on the aerial shaft. A cross-over waveform is produced at coincidence which, when rectified, provides a suitable gate pulse, a frequency of 400 c/s providing sufficient cycles during the gate for most purposes. A variable bias to the valve amplifying the gate pulse provides the beam-with control. The ambiguous gate is eliminated by means of a cross winding.

My second point arises from the paper by Messrs. Kennard and Nicholson and concerns correction for the curvature of the earth. The usual expression for correction is the square of the range divided by twice the radius of the earth, R_e, the value of range being derived from the slant or plan range shaft, whichever is present, as shown in the paper.

Referring to Fig. A, the true expression for height is BE.

Taking slant range as R_s and elevation as ε, an expression may be derived for h which is innocent of approximation.

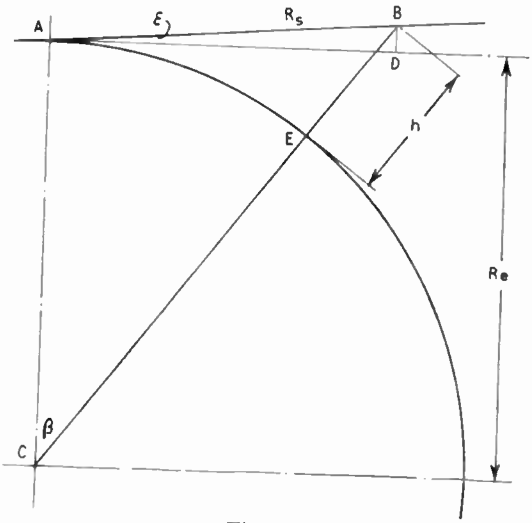

Fig. A.

Thus in the triangle ABC—

$$(R_e + h)^2 = R_s^2 + R_e^2 - 2R_sR_e \cos(90 + \varepsilon)$$

$$R_e^2 + 2R_eh + h^2 = R_s^2 + R_e^2 + 2R_sR_e \sin \varepsilon$$

Therefore, $$h = R_s \sin \varepsilon + \frac{R_s^2}{2R_e} - \frac{h^2}{2R_e} \quad \text{......(i)}$$

$R \sin \varepsilon$ is BD, the radar height.

$h^2/2R_e$ is normally very small, but serves to give the error in the close approximation for true height

$$h_1 = R_s \sin \varepsilon + \frac{R_s^2}{2R_e} \quad \text{.........(ii)}$$

which may easily be computed from the radar information. This error may become important when an accurate radar theodolite is used on high flying objects at long range.

Dr. J. H. Westcott: In attempting to allow for a curved earth it is indeed customary to use the relation of eqn. (ii) derived by Mr. Downer, where the first term is the radar height and the second term is a correction term. The relationship is an approximate one and, as Mr. Downer has pointed out, the true value of h involves a further term, $h^2/2R_e$.

To include this term in the calculation of h would involve making successive approximations to the value of h. This of course is an unpleasant business, but not strictly necessary. Equation (i) is a quadratic in h which is easily solved for its two values of h, one of which is clearly inapplicable. This gives:

$$h = R_e \left[\sqrt{\left(1 + \frac{2R_s \sin \varepsilon}{R_e} + \frac{R_s^2}{R_e^2}\right)} - 1\right]$$

If the term under the root is expanded by the binomial expansion, and terms of order $>R_s^2$ are neglected, there results the following relation:

$$h \cong R_s \sin \varepsilon + \frac{R_s^2}{2R_e} \cos^2 \varepsilon \quad \text{.........(iii)}$$

which is very little more complicated than the usual relation used, and is much more accurate.

L. J. Kennard and C. H. Nicholson *(in reply)*: In reply to Mr. Downer's remarks concerning the Mullard Radar Trainer, we would suggest that the method of varying the bias on the valve amplifying the gate pulse is a satisfactory method of beamwidth variation for a single target, but is unsatisfactory where a number of targets are used, because matched valves would be required to ensure a common beamwidth; in addition, the calibration of the beamwidth control is liable to vary with valve ageing.

Concerning his remarks on correction for the curvature of the earth it is true that eqn. (i) is the correct one to use, but the last term can be neglected, and this leaves an expression which is the same as we have used (Sect. 9.3), i.e.,

$$h = a + H \text{ where } a = \frac{r^2}{2R} \text{ and } H = \rho \sin \theta$$

Neglecting the term $h^2/2R_e$ introduces an error of only 0·25 per cent. for a target flying at 50,000 feet at 200 miles range.

With reference to Dr. Westcott's reasoning, it is again true that his final expression (eqn. (iii)) is more accurate than Mr. Downer's final result (eqn. (ii)), and it is relatively easy to obtain the $\cos^2 \varepsilon$ function. However, since the omission of the last term from the exact relation introduces such a small error, the approximate expression used has been found to be satisfactory for all practical purposes.

THE RADIO TRADES EXAMINATION BOARD

Thirteenth Annual Report—1956-7

This Report is published for the interest of members in view of the active work of the Institution in the formation and operation of the Board.

THE BOARD has expressed deep regret on the loss it has sustained by the death in May 1957 of its second Chairman, Mr. E. J. Emery. As a representative of the Radio Industry Council, Mr. Emery had served on the Board since its formation in 1942 and his keen interest in the work of the Board had a great influence upon its development.

This report covers the year ended 31st August, 1957, and the analysis of examination results shows the continuing attraction of the examination to candidates wishing to obtain recognition of their ability as mechanics and technicians. There are now in Great Britain 65 colleges providing a definite course of instruction specifically for the Board's Radio Servicing Certificate Examination, and thirty of these schools also run courses for the Television Servicing Certificate Examination. All these courses have been approved jointly by the Board and the City and Guilds of London Institute through Ministry of Education Inspectors.

The standard set by the Board is high, and any criticism received has been from technical colleges which are inclined to the view that especially in the practical test the Board's requirements are, if anything, higher than that required in other trade examinations. Nevertheless, the number of candidates submitting themselves for examination has shown a steady increase.

Comparison of 1956 and 1957 examination results.—The table shows that a considerable percentage of candidates are referred in the Practical Test. An analysis of the entries indicates that some 50 per cent. of the candidates are under 19 years of age and this would appear to confirm the lack of practical experience in many candidates.

Appointments Register.—During the year the Board acquired a licence from the London County Council to operate an appointments service. No charge is being made either to employers or to candidates, but this service will, of course, involve the Board in some extra expense.

Finance.—Obviously it is not possible to conduct the work of the Board entirely on revenue received from examination fees, which the Board has always wished to keep within the financial ability of prospective candidates. This has been made possible by the subsidy provided over the last 15 years by the Brit.I.R.E., the Radio Industry Council, the Radio and Television Retailers' Association, and the Scottish Radio Retailers' Association.

Acknowledgments.—This is the only opportunity the Board has of recording appreciation for the services of the examiners, the co-operation afforded by the City and Guilds of London Institute, and all who have assisted in bringing to the Board such satisfactory results for its work.

The Board is also appreciative of the continued assistance rendered by the British Institution of Radio Engineers in providing the increasing secretarial facilities so necessary for the execution of the Board's expanding work.

Analysis of Examination Results

	Year	Entered	Sat	Passed	In one sitting	Having been Referred	Referred	Failed	% Pass
Radio	1956	822	802	322	261	61	185	295	40
	1957	1140	1117	466	384	82	336	315	42
Television	1956	138	134	60	54	6	51	23	51
	1957	247	237	106	83	23	48	83	45

THE USE OF RADAR SIMULATORS IN THE ROYAL NAVY *

by

P. Tenger, B.Sc., B.Sc.(Eng.)†

A paper presented at the Convention on "Electronics in Automation" in Cambridge on 29th June 1957. In the Chair: Professor D. G. Tucker (Member)

SUMMARY

The paper describes a synthetic training system developed for the Royal Navy to provide a means of semi-realistic study of tactical problems involving ships and aircraft. The requirements of realism and accuracy necessary in such a trainer are considered and the paper goes on to describe the radar simulator on which the system is based, together with other circuits devised to satisfy these requirements. The particular system described has been developed to meet somewhat specialized requirements for the Royal Navy and details of the number of targets used, speed, height and radar range have been omitted for security reasons.

LIST OF SYMBOLS

The "mile" used in this paper is the tactical mile of 2,000 yards.

v = velocity of target (knots)
V_o = output voltage
V_i = input voltage
f = frequency of operation (c/s)
S_p = scale of position (volts/yard)
S_v = scale of velocity (volts/yard/sec)
R = rate of fuel consumption (lb. or gallons/min)
h = height in feet.

1. Introduction

During the past decade the cost and complexity of modern aircraft and their operation has increased so much that it is now no longer an economical proposition for large scale exercises to be mounted as often as desirable. Hence the study of tactical problems posed by the use of such weapons tends to be carried out only during comparatively infrequent exercises or on paper.

The trainer to be described in this paper has been designed to enable the semi-realistic study of such problems involving ships and aircraft disposed over a large area.

It is a relatively straightforward matter to design on paper a synthetic radar trainer to meet any desired requirements of complete realism using well-known electro-mechanical means‡ but in general the cost and complexity of such a system is prohibitive. The trainer described in this paper is a practicable proposition because it uses relatively inexpensive electronic circuits, and at the same time the opportunity has been taken to cut down the degree of realism where it is not considered vital.

1.1. *Requirements of the Trainer*

The trainer to be described has been designed to provide the following facilities:—

(i) An exercise area 500 miles × 500 miles.
(ii) Aircraft simulators with course, speed and height controls.
(iii) Ship simulators with course and speed controls.
(iv) Both (ii) and (iii) to be capable of starting from any position in the exercise area.
(v) Provision of three-dimensional radar information for each ship, complete with noise background and echoes which

* Manuscript received 2nd May 1957. (Paper No. 433.)

† Admiralty Signal and Radar Establishment, Portsmouth.

U.D.C. No. 621.396.96:681.142

‡ F. W. Cook, "Synthetic radar trainer," *A.T.E. Journal*, **12**, No. 2, pp. 89-100, April 1956.

appear once per aerial revolution.

(vi) An overall picture of the whole exercise area completely free from radar background noise and interference showing the position of each target in the area, together with a group of numerals to represent information known about it. This picture is generated electronically and brought up to date at 2 sec. intervals.

(vii) Fuel consumption meters for fighter type simulators.

(viii) Circuits to provide realistic handling times for ship-borne aircraft.

(ix) Provision of typical control room equipment for each of the ships.

(x) Provision of synthetic ship-air, ship-ship and ship-shore communication.

The last two items will not be considered in this paper, since they do not form part of the main electronic equipment.

2. Realism and Accuracy

2.1. *Realism*

It has been stated earlier that the design of a synthetic trainer to provide complete realism presents no insuperable problems given unlimited space and money. It is a feature of the trainer described that it achieves economy where possible by departing from complete realism when this can be done without significantly reducing its value.

There are three forms of realism involved—

(i) The appearance of the radar display, i.e. degree of correct simulation of beam width, pulse length, noise, etc

(ii) The motion of the simulators, i.e. velocities and accelerations.

(iii) Control of the simulators.

The degree of realism required for the operator of the display depends on their needs.

For this trainer there are three classes of operator:

(i) Those concerned with reporting the presence and position etc. of an echo on the radar display. They are vitally interested in the realistic display of information—e.g. presence of noise, size of echo, fading etc., anything in fact which makes the extraction of accurate position and height data from the displayed information more difficult.

(ii) Those concerned with directing intercepting fighters. This class is interested in relative positions of fighter and target and complete realism of the display is not of great importance, particularly with modern high definition radars. They are, however, vitally interested in the movement of the echoes, which should include acceleration.

(iii) The exercise controllers who use the facility of 1.1. (vi) and are only concerned with position, course and speed.

The degree of realism necessary thus depends on the use to which the completed trainer will be put. The prime concern of this trainer is for the study of tactical problems and hence the high degree of realism required for class (i) of the user is not necessary.

The question of realism of control is discussed in Sect. 2.3.

2.2. *Accuracy*

One of the requirements (1.1. (v)) implies as many separate radar pictures as there are ships. In addition there is the requirement for an overall picture (1.1. (vi)). A fundamental necessity is that all these pictures should match up. However, the accuracy required is rather a "relative" accuracy than an absolute accuracy, i.e. the relative positions of echoes on any one of the pictures produced should correspond with those on any other. It does not matter if the actual position with respect to a fixed reference grid is slightly different from radar to radar or radar to overall picture. This difference corresponds in practice to the uncertainty of own ship's position and the difference between slant range and ground range.

2.3. *Control of the Simulators*

This again can be considered in the light of "realism" and "accuracy" requirements.

2.3.1. Realism of control

The basic requirement has been shown to be one of reproducing realistic movement on the various pictures generated, particularly in the case of simulators representing fighter aircraft where accelerations are required. The trainer is not concerned with training pilots and hence the physical controls of the simulator need not bear any resemblance to those of an actual aircraft. They should be as simple as possible consistent with providing the necessary movements on the display.

2.3.2. Accuracy of control

A high degree of accuracy of control, i.e. setting course, speed, etc., is in general unrealistic in that it eliminates the causes of error met with in practice, e.g. if the course flown with respect to a fixed grid is slightly different to that indicated on the simulator "compass" the error is one often met with in practice when navigating by dead reckoning and could for example be due to compass error, wind, etc. As long as the errors introduced into the system are comparable with those met with in practice they can be accepted.

The foregoing paragraphs show that for a trainer aimed at tactical studies a high degree of realism is not required. Electronic circuits to meet the main requirements of the trainer have been designed and are described in the subsequent sections.

3. The Radar Simulator

The radar picture generated is that of an idealized narrow beam radar providing gapless cover to any desired range.

The problem of providing three-dimensional radar information can be conveniently divided into two sections:—

(*a*) The plan picture,

(*b*) The range-height picture,

both of which are again conveniently divided into two sections:—

(*c*) Co-ordinate generation.

(*d*) Echo generation.

3.1. *The Plan Picture*

3.1.1. Aircraft co-ordinate generation

A block diagram of the co-ordinate generator is shown in Fig. 1 and consists of a velocity control, a means of resolving this velocity into components in the North-South and East-West directions (referred to in future as y and x axes respectively), an integrator to provide a voltage output representing instantaneous position and a means of setting the integrator to any desired starting position.

The exercise area is 500 miles × 500 miles giving a range swing of ±250 miles. The scale

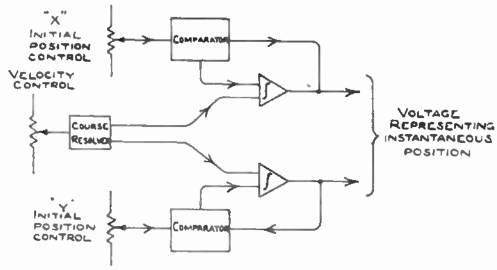

Fig. 1. Block diagram of co-ordinate generator.

has to be chosen so that the integrator can provide a linear output over this range. This scale is determined to a large extent by the h.t. voltages available and for the system to be described was chosen as ½ volt per mile or 1/4000 volts per yard.

The basic integrator circuit is derived from a Miller integrator.* There is one integrator per co-ordinate each integrating a voltage representing the velocity component in that direction. A block diagram of the basic integrator is shown in Fig. 2.

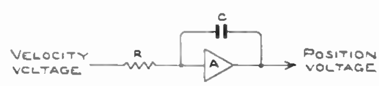

Fig. 2. Basic integrator.

The time constant CR of the Miller integrator is equal to S_v/S_p so that if a velocity scale of 1 volt = 10 knots = (50/9) yd/sec is chosen as the time constant $CR = 4000/(50/9) = 720$ sec ($S_p = 1/4000$ volts/yd).

The required value of CR can be obtained by choice of suitable values for C and R but is usually unsatisfactory due to the need for high values of either R or C.

Attenuation of the velocity scale produces a proportional reduction in the required value of CR but there is a limit to the attenuation possible since the input voltage should be large compared with the grid base of the valve used.

* B. H. Briggs, "The Miller integrator," *Electronic Engineering*, **20**, pp. 243-247, 279-284, 325-330, August/September/October 1948.

This problem has been overcome by an artifice employed to give the effect of a large time constant without the disadvantages of high circuit values.

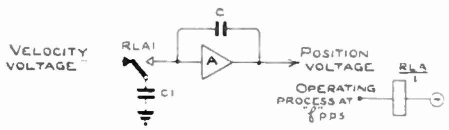

Fig. 3. Modified integrator.

Referring to Fig. 3, the capacitor C1 charges up to the voltage representing the velocity component and when the relay operates it transfers its charge to C.

The output voltage from the integrator thus changes at a rate

$$\frac{dV_o}{dt} = \frac{C_1}{C} \cdot V_i \,.\, f \text{ volts/sec.} \qquad \ldots\ldots\ldots(1)$$

The choice of the relay operating frequency f is not critical but should meet the following requirements:

(*a*) Constancy;

(*b*) High enough for the overall effect on the display to be one of continuous change and not of sudden jumps;

(*c*) Low enough (in conjunction with a suitable mark to space ratio) for the capacitor C1 to charge and discharge fully.

From equation (1) it can be seen that

$$\frac{C_1}{C} = S_p / f\, S_v \qquad \ldots\ldots\ldots(2)$$

from which the velocity scale S_v can be determined.

The choice of the value of capacitor C is limited by the high value of equivalent shunt resistance it must have at the working temperature of the equipment. This shunt resistance produces an error velocity towards the zero position. The leakage resistance of the external circuits, particularly those associated with the grid of the integrator valve, can also have an important effect on this error velocity and great care must be taken to ensure that all leads are kept remote from h.t. supplies. In this circuit, capacitor C was chosen as an 8-μF paper dielectric type with 600 V working at 70°C.

It is worth noting that with the scales quoted earlier a leakage resistance of 2000 megohms from the grid to an h.t. potential of 300 V produces an error velocity towards the zero position of approximately 135 knots. A high degree of compensation for this error velocity can be achieved very simply using a similar circuit to that used for giving the high time constant effect. Referring to Fig. 4, capacitor C2 charges up to the output voltage from the integrator and then discharges into the capacitor C when the relay operates, a 180 deg. phase change being introduced by connecting the relay contacts as shown. By adjustment of the value of C2 under working conditions the error velocity can be reduced to less than ¼ volt per minute at worst, representing an error velocity of less than 30 knots. The relay used in this circuit should, of course, be chosen to have a very high leakage resistance itself or else the circuit will add a greater error than it is capable of compensating.

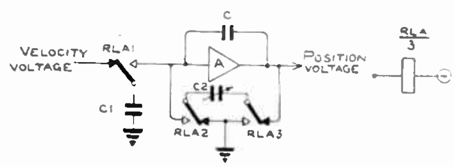

Fig. 4. Modified integrator with compensation for circuit leakage.

The input to each integrator is a voltage representing the resolved component of the velocity along the x and y axes. The method of resolution depends on the type of aircraft the simulator will represent and there are two main types to be considered in this application. These are "fighters" whose rate of change of movement on the display should be realistic, and "miscellaneous" which can be used for a variety of purposes having the characteristic that course and speed need altering at infrequent intervals. The need for realism in the displayed movements of the fighter implies building into the circuit acceleration limiters. This applies to the controls for speed and course setting. A circuit of the controls for the fighter type is given in Fig. 5.

Referring to this figure, the potentiometer on the left is the speed control, the voltage appearing at the slider being chosen to meet the speed range required after choice of S_v.

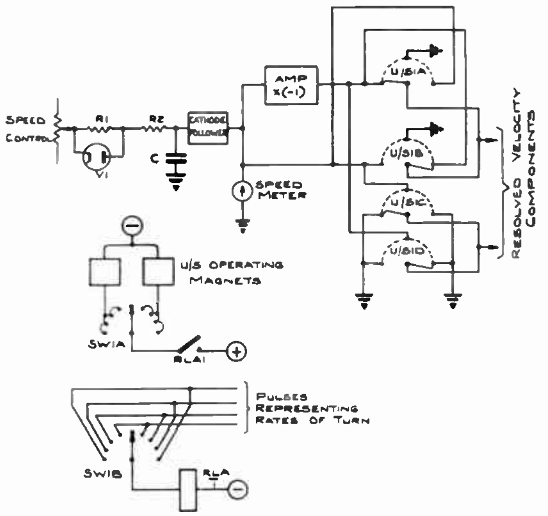

Fig. 5. Block diagram of speed and course control circuit for fighter type aircraft simulator.

The components R1, R2 and C together form an integrating circuit so that sudden changes of velocity caused by movement of the slider do not appear at the output of the cathode follower. The time constant of these components can be chosen to give any desired maximum acceleration. The diode V1 is inserted so that the maximum deceleration is greater than the maximum acceleration as is found in practice. The requirement for a controlled rate of turn can be met by straightforward application of servo techniques, but these would require additional control circuits which are bulky. They are, however, more accurate than the system adopted which, although somewhat less accurate than a real aircraft, has the merits of simplicity, cheapness and small size. In Fig. 5 U/S1 represents a 50-step 2-way uniselector with resistors connected between the banks. The values chosen are such that the output from the pick off wipers are proportional to the sine and cosine of the angle of rotation of these wipers from some position chosen as the zero. This system produces a 50-step approximation to the circle and the course changes in increments of approximately 7½deg. (i.e. course accuracy of $\pm 3\frac{3}{4}°$).

The uniselector is provided with two operating magnets (for turns to port and starboard) and a switch selects one of four pulse trains which provide practical rates of turn when applied to the operating magnets. A further disadvantage of this method is that it takes no account of the velocity, and unreasonable centripetal accelerations can be obtained with high rates of turn and velocities. This effect can of course be counteracted by a proper understanding of the circuit. Figure 6 shows the very much more simple controls used for the "miscellaneous" type. No built-in control of acceleration is introduced but if necessary this effect can be achieved to some degree by careful use of the control knobs.

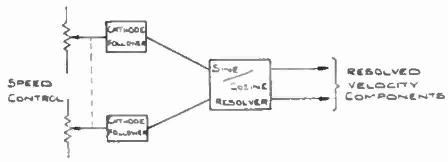

Fig. 6. Block diagram of speed and course control circuits for miscellaneous type aircraft simulator.

The final requirement of this circuit is that the output should be capable of being set anywhere within the exercise area. The output from the basic Miller integrator can be made either positive or negative-going simply by changing the sign of the voltage to which the resistor R is connected. This could be used as a means of performing the positioning, a switch connecting the leak to the appropriate voltage until the required output is reached. The method adopted uses this principle but also provides the facility of enabling any number of the simulators to be grouped together as a formation under the control of a specified leader.

Referring to Fig. 7 the output from the integrator is compared with a voltage representing the desired position. If these are not equal an error voltage of appropriate sign from the comparator is applied to the integrator through the switch SW1 and the resistor R.

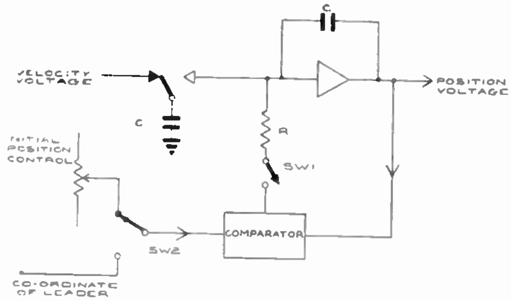

Fig. 7. Method of initial positioning.

This drives the integrator towards equality with the desired position voltage. When equality is reached the error voltage from the comparator becomes zero and the output from the integrator remains equal to the initial position voltage as long as switch SW1 is closed. When this switch is opened the output voltage will change at the rate determined by the voltage applied to the capacitor C. If the time constant C.R is made small it is apparent that the output of the integrator will accurately follow changes in this initial position voltage as long as the switch SW1 is closed. Thus if the switch SW2 is set to select the output from another integrator, the first integrator will accurately follow it. A circuit diagram of the comparator is shown in Fig. 8. The constant current valve V3 is included to ensure that the "equality" voltage output from V4 is constant over the whole range of possible input voltages. The potentiometer RV2 is provided as a means of offsetting the follower from the leader by a variable distance. The follower can be made to break formation and fly off independently merely by opening the switch SW1.

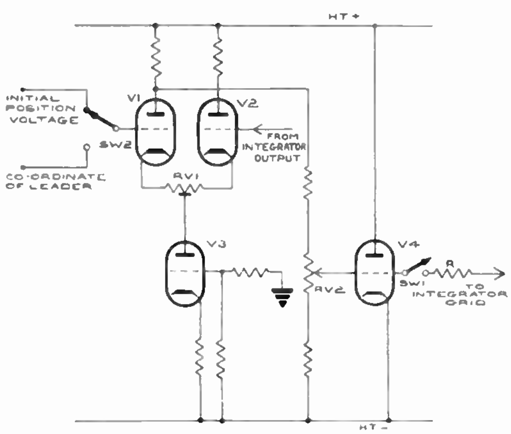

Fig. 8. Circuit diagram of comparator.

3.1. Echo generation

A block schematic of the circuit is shown in Fig. 9, and the complete circuit in Fig. 10. The circuit compares the plan co-ordinates, i.e. the outputs of the integrators, with the instantaneous radar co-ordinates, and produces a pulse at the instant of equality. Two comparators in cascade are used per co-ordinate to give accurate indication of the instant of equality.

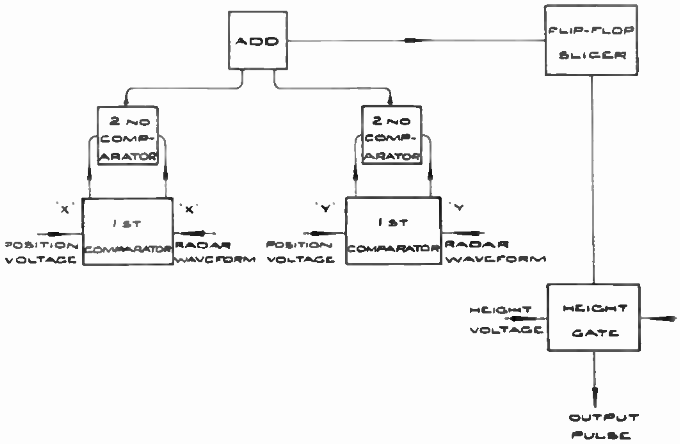

Fig. 9. Block diagram of echo generator.

The first comparators V1 and V2 require a constant current supply provided by V3a and V3b since the voltages at the anodes at equality must be sensibly constant over the whole range of output voltages provided by the integrator. RV1 and RV2 are provided to set this condition.

The second comparator consists of a triode and a pentode with a short base suppressor grid. Consider the situation when the voltage applied to the left-hand grid of V1 is more positive than that applied to the right-hand grid. Under this condition V4a is cut off whilst V5 is conducting via its screen circuit only. When

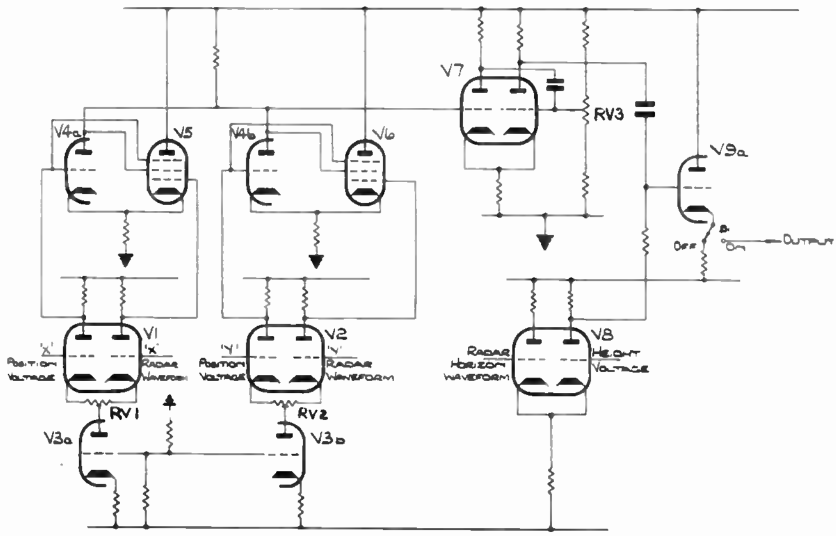

Fig. 10. Circuit diagram of echo generator.

the left-hand grid of V1 is more negative than the right-hand grid, V4a is conducting and V5 is cut off completely. The conducting current through the screen grid of V5 and the anode current of V4a are made equal, and hence the volt drop across the anode load of V4a is the same for both conditions described above. Under the condition when both grids of V1 are at the some potential both V4a and V5 are conducting and passing equal currents via their anode circuits. The screen current of V5 is considerably less than previously, thus producing a positive-going spike across the anode load resistance of V4a. V4a and V4b have a common load resistance and the pulses produced by each of the *x* and *y* comparators add up to produce a pulse of double amplitude when the radar scan passes through the target position.

These pulses are applied to V7 which is a combined "flip-flop slicer." The potentiometer RV3 is set so that the flip-flop fires only when the two pulses add together. The output from this valve is a pulse of constant width and amplitude and is a.c. coupled to an output cathode follower, V9a. The leak of the a.c. coupling circuit is connected to the anode circuit of a comparator valve which compares the target height with a radar horizon waveform. Thus if the target is above the radar horizon the echo pulse appears at the cathode of V9a and if not no pulse appears.

The waveforms representing the instantaneous radar co-ordinates are represented by sawtooth waveforms whose amplitudes are proportional to the sine and cosine of the radar bearing. It is worth noting that a set of these waveforms is required for each separate type of radar simulated, to cater for the varying ranges and data rates of these radars.

The switch S1 of Fig. 10 in the ON position connects the output pulse to a circuit which mixes it with pulses from other echo generators and either a synthetic noise background or, if desired, the video signal from a "live" radar set. This mixed signal is then distributed to the various display systems. A block diagram of the mixing system is shown in Fig. 11 and a circuit diagram of the pulse mixers in Fig. 12. The number of inputs that can be connected to a mixing grid depends on the capacitive loading that can be tolerated on the echo generator output cathode follower and this depends to a large extent on the grouping of these units and the type of capable used.

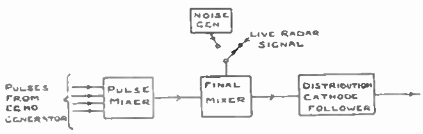

Fig. 11. Block diagram of mixing system.

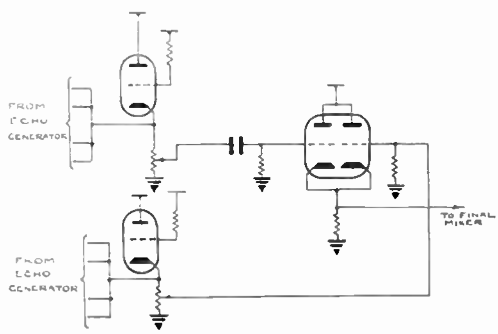

Fig. 12. Circuit diagram of echo-pulse mixer.

3.2. *The Range-Height Picture*

3.2.1. Aircraft co-ordinate generation

As in the case of the plan co-ordinates, the "fighter" type of simulator requires controlled rates of change of the height parameter.

The basic circuit used is the conventional Miller Integrator with a variable time constant to produce the effect of different rates of climb or descent. The choice of a scale for height can be determined in a rather more arbitrary fashion than the scales of position and velocity. If it is to be used directly on a display system as in this application it should be chosen to suit the deflecting system of the display tube.

A block diagram of the height co-ordinate generator for the fighter type of simulator is shown in Fig. 13 and is very similar to the method used for positioning the x and y co-ordinates. The error voltage from the comparator is applied to the leak resistor of a conventional Miller integrator, the value of which is controlled by a switch selecting the desired rate of change of height.

The time constant CR of this circuit is given by $CR = V_i \Big/ \dfrac{dV_o}{dt}$. The use of the comparator in this circuit enables the simulator to follow height changes of the leader as well as course and speed changes. The modified integrator circuit of Fig. 3 could be used in this application particularly if slow rates of change are required. In the miscellaneus type, height control is by a potentiometer covering the required height voltage range.

3.2.2. Echo generation

Probably the most common method of height finding using radar is that using a separate aerial which scans a particular bearing in a vertical plane. Simulation of this method requires:

(i) Selection of the targets lying on the selected bearing.

(ii) Production of a suitable height scanning waveform for comparison with the height co-ordinate of the selected targets.

(iii) Presentation of the echoes produced by (i) gated by (ii).

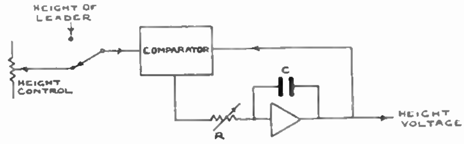

Fig. 13. Block diagram of control of height in fighter type of simulators.

The method evolved to produce the height picture is not considered to be suitable for use in any other application than the one described since it depends upon the presence in the overall system of a 100-way switch required for the presentation of the overall tactical picture together with its associated code information.

The system is shown in block schematic form in Fig. 14. The display shows all targets lying in a lane on a selected bearing as short

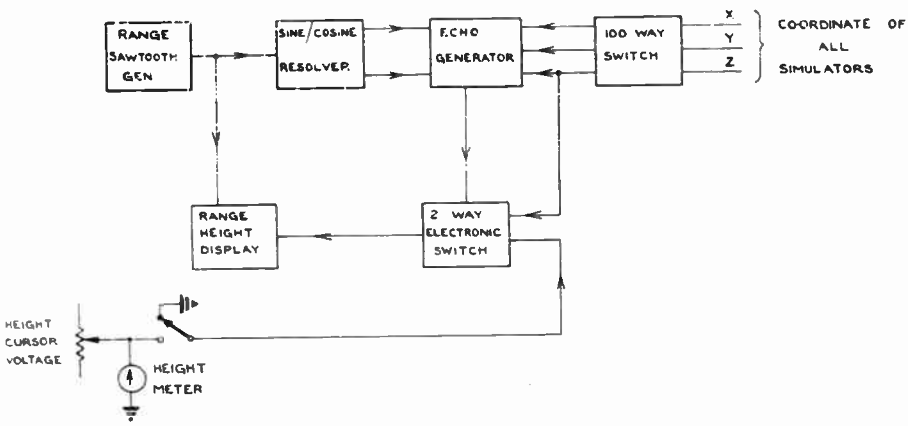

Fig. 14. Block diagram of height presentation system.

horizontal lines at correct range and at distances above the zero which are proportional to their heights. The width of this lane is determined by the method of bearing comparison and in practice is about 5 miles in width at all ranges for the circuit used.

Referring to Fig. 14, a range saw-tooth of amplitude to suit the radar range is resolved into sine and cosine components by a sine/cosine potentiometer set to the selected bearing. The sine and cosine range sawtooth waveforms are applied to an echo generator of the type shown in Fig. 10. The x and y co-ordinates of all the simulators in use are also applied to this echo generator in sequence by the 100-way switch. Any of these that finds equality with both the bearing saw-tooth waveforms produce a pulse which opens an electronic switch and allows the height voltage which is applied simultaneously with the corresponding x and y voltages to produce a vertical deflection of the range sawtooth applied to the display tube. The 100-way switch completes one cycle in approximately 2 seconds, sampling each position for approximately 15 milliseconds. The display tube has a long persistence screen to remove flicker. Measurement of height is made by means of a potentiometer covering the whole height range. The slider of this potentiometer is connected to the electronic switch, so that when no pulses are being provided by the echo generator the voltage at the slider appears as a horizontal line on the display. To avoid confusion on the display the slider of this potentiometer is connected to the electronic switch via a spring loaded toggle switch normally connected to the zero height voltage. Height is read on a meter connected to the potentiometer slider and calibrated directly in feet.

The method of height presentation outlined above is open to objection on many grounds, but principally because it makes the height finding problem far too easy. This objection is recognized and the system is defended on the grounds that other height presentation systems considered by the author for providing some degree of realism were based on one or other of existing height presentation systems. This was considered undesirable in a trainer of this nature as these systems in due course will become obsolete or modified thus necessitating changes to the equipment in the trainer. The system described has been adopted as a means of providing height information from a synthetic radar display to the user.

3.3. *Ship Co-ordinate Generation*

The circuits described so far form the basis of the trainer. There is however one further method of co-ordinate generation required to complete the picture. This is the method used for producing position, course, and speed of the ships used in the exercise.

The inherent drift in the circuit described for the generation of the aircraft co-ordinates can by careful design and layout be reduced to considerably less than 30 knots. However since ships with a speed in excess of about 45 knots are the exception rather than the rule, the residual drift even when as low as 5 knots can amount to a serious error.

For this reason, the electronic method of co-ordinate generation has not been used for ship targets, and a straightforward velodyne-tachogenerator method* of driving a potentiometer has been adopted. A block schematic of the circuit is shown in Fig. 15 and is self-explanatory. The output potentiometer used is a 10 turn helical wire wound type having 18,000 turns, giving a resolution of approximately 50 yd.

4. The Complete System

The basis of the system is the coordinate and echo generation described earlier. These together with a large number of other items not so far described make up the completed system which is shown in schematic form in Fig. 16.

In this section it is proposed to describe some of the special circuits used and their application to this trainer.

4.1. *Individual Radar Pictures*

Each ship is assumed to have the same type of radar and hence only one set of range saw-tooth waveforms need to be generated. The bearing information is provided by a common aerial simulator driving the resolvers necessary for generation of the radar saw-tooth

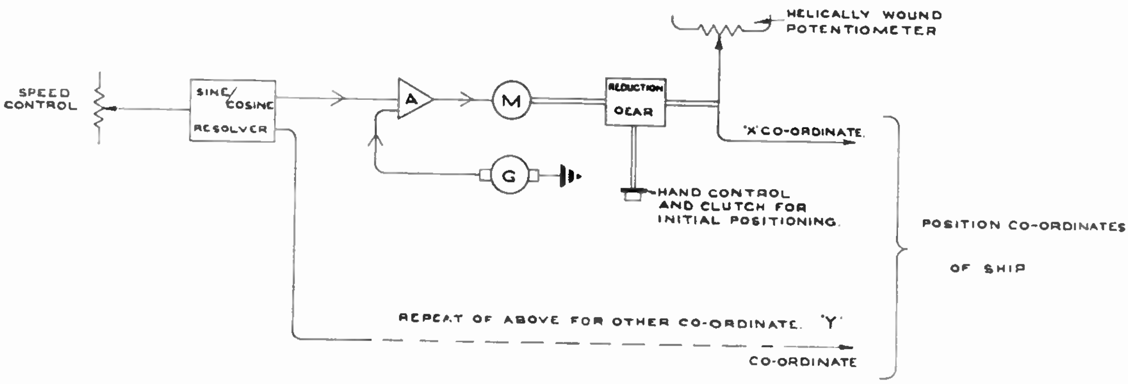

Fig. 15. Method of ship co-ordinate generation.

The reduction gear box must be designed to be free from backlash as far as possible in order to avoid discontinuities in the track whilst the ship is turning. One degree of backlash between the gear box and this potentiometer contact represents some 250 yd which is considered the maximum tolerable. This apparently stringent requirement arises from the fact that the display systems are capable of expansion by a factor of 10. On the fully expanded scale 250 yd represents approximately ¼ in.

* F. C. Williams and A. M. Uttley, "The Velodyne," *J. Instn Elect. Engrs*, **93**, Part IIIA. No. 7, pp. 1256-1274. 1946.

waveforms. Associated with the saw-tooth generating equipment is the generation of switching waveforms and radar brightening waveforms for the display circuits.

The common saw-tooth waveforms are added to each of the voltages representing ship's position to provide an offset sawtooth representing the radar range from the ship's position. It should be noted here that if the scale of the mixed voltages is changed in this adding circuit, as would be the case with straightforward resistance mixing, the scale of the voltages from the integrators must be changed accordingly.

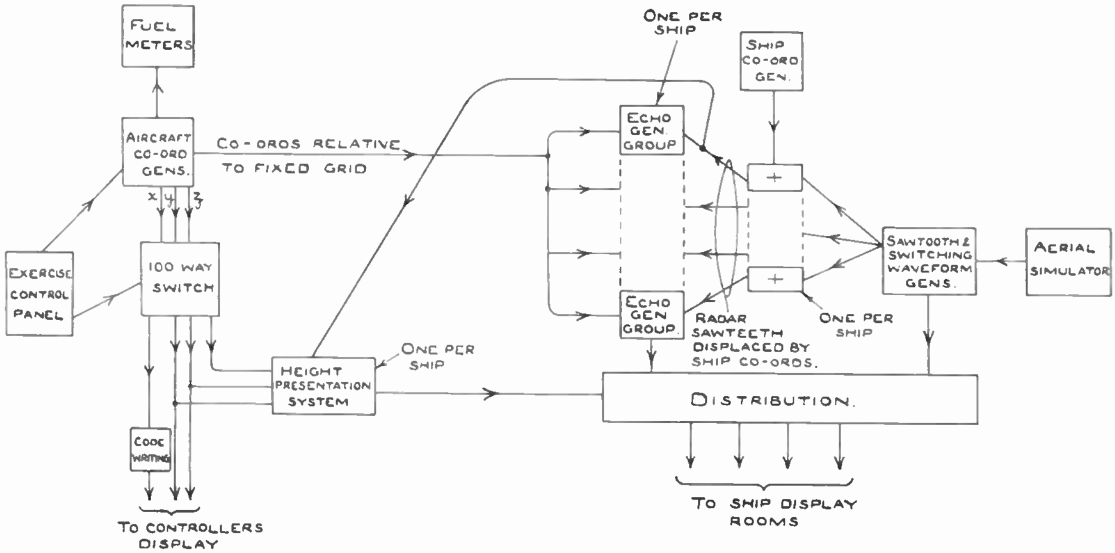

Fig. 16. Block diagram of trainer system.

The waveforms representing radar range from the ships are then applied to the echo generators for production of the radar echoes.

The number of echo generators required in the system for each radar type depends on the number of separate radar pictures required and the maximum number of targets each radar can see. Thus if there are S ships each requiring to see N targets the number of echo generators required is SN per radar type. The control circuits and coordinate generators need be reproduced once only per target. The echoes produced by each group of echo generators are then added to noise and distributed.

4.2. *The Overall Tactical Picture*

This picture is required for the use of the controllers of the exercise. This picture is required to show the instantaneous positions of all targets in the exercise area as dots together with a group of numerals representing the information known about it. This is fixed information determined before the beginning of the exercise, e.g. the ship to which an aircraft belongs, whether friendly or enemy, track reference, etc.

Height information—which is not fixed information—is made available to the controller through a circuit which displays this information on a meter when the track reference number of a simulator is set on a group of switches. The reference number of each simulator is part of the fixed information shown in the numeral group on the tactical picture.

The methods of generating the information required for this picture are considered in more detail below.

4.2.1. The 100-way switch

Generation of the overall picture is based on this switch. It consists essentially of four 25-position uniselectors stepped synchronously and a 4-way electronic switch. The output from each bank of each uniselector is sampled in turn by this switch so there needs to be one 4-way electronic switch per bank of the uniselector. The switching pulses for the electronic switch are arranged so that all the information relating to one target is presented to the appropriate circuits simultaneously.

The uniselectors are stepped by application of a suitable pulse waveform to their operating coils, these pulses being derived from a stable multivibrator circuit synchronized from the code writing equipment. Special precautions are taken to ensure that the uniselectors always remain in step.

The voltages applied to this switch are the *x*, *y* and *z* coordinate voltages of each of the simulators together with the d.c. voltages representing the coded information about each simulator.

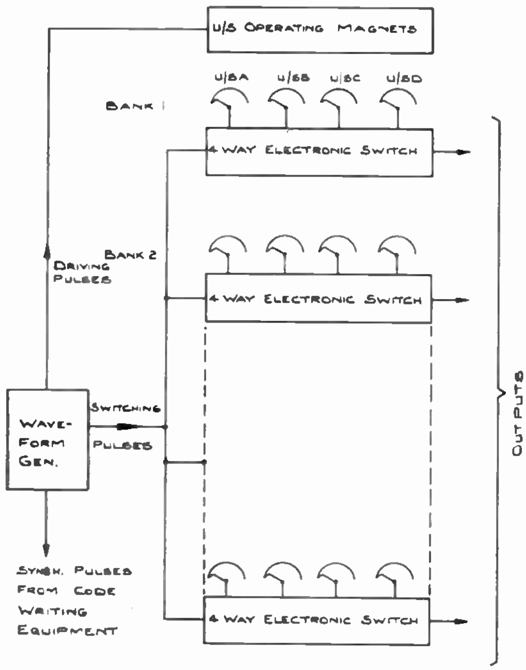

Fig. 17. Block schematic of 100-way switch.

A block schematic of the 100-way switch is shown in Fig. 17.

4.2.2. The code writing equipment

The function of this equipment is to generate the waveforms required to trace the numerals representing information about each target.

These waveforms are cyclograms formed by application of suitable voltages to the deflection system of the display.

The outputs from the 100-way switch applied to this equipment are d.c. voltages representing the desired numeral. These d.c. voltages are compared with reference voltages, the output from the comparators being used to select an appropriate mixture of deflecting waveforms and a suitable brightening pulse or pulses. These are then applied directly to the display system.

The deflecting waveforms consist of sine waves and half sine waves, together with portions of a square waveform. The equipment also provides a synchronizing pulse fed to the waveform generator of the 100-way switch. Code writing is described in a paper which is in course of preparation.

4.2.3. The electronic clock

This circuit makes use of the code writing facility to provide on the overall picture a continuous time record during the exercise.

Referring to Fig. 18, the shaft of an electric motor from a clock carries a cam which closes a pair of contacts and steps a uniselector once per minute. The contacts of the uniselector carry voltages representing the numerals 0–9 arranged sequentially. This uniselector selects a voltage representing the minutes unit of a four digit block indicating time elapsed from the exercise start. It is considered improbable that any exercise will last for longer than 10 hours so that the first digit of the time group can be made zero always. One uniselector is required for each of the other digits, their movements being controlled by the uniselector representing the succeeding digit.

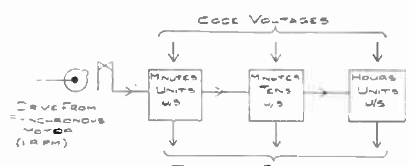

Fig. 18. Block diagram of electronic clock.

4.2.4. Controller's height information

It will have been observed that the tactical picture generated provides no information on the height of the targets in the exercise area.

Methods are available for introducing this information into the numeral group which involve additional complex equipment. The introduction of this equipment is not thought essential to this system and a simpler method which indicates the height of any selected target on a meter has been designed.

The voltages representing the height of the targets are connected to the contacts of a uniselector bank or banks. The controller is provided with a pair of switches on which he sets the reference number, taken from the tactical pictures. These switches determine the state (i.e. whether operated or released), of a group of relays, a change of state causing the uniselector to step over its bank contacts until the new condition is satisfied. The height voltage present on this contact is applied to a meter calibrated directly in feet on the controller's console. Well established methods used by telephone engineers are available for this type of circuit.

4.3. *Target Grouping*

One of the principal features of the method of co-ordinate generation employed is that of formation grouping and subsequent splitting of this formation during the exercise. This same feature can be extended to enable the target to start from a ship's position so as to represent an aircraft taking off from an aircraft carrier.

The preliminary grouping of the targets and ships at the commencement of an exercise when large numbers of ships and aircraft are involved presents a considerable problem.

It has been solved in this trainer by the use of a large multiple plug and socket assembly as used in business accounting machines. This provides a large number of contacts in a small volume.

The plug part is made in two forms, one with provision for permanent wiring and the other for flying leads between contacts.

In use all the co-ordinate voltages from both ship and aircraft simulators are connected to contacts on the socket together with wires from the selector switches SW2 of Fig. 9, the switch being chosen to give the required choice of "leader."

This plug is also used as a convenient point to feed in the voltage representing the code numerals associated with each target.

The use of such a plug and socket has the advantage that certain types of exercises can be wired up as permanent connections and a library of such exercises built up as required, whilst still preserving a large degree of flexibility for exercises of an experimental nature.

5. Additional Features

5.1. *Fuel Meters for the Fighter Type Simulator*

A major factor in the defensive role of fighter aircraft is the endurance of such an aircraft. It is clearly completely unrealistic to make a study of tactics involving the use of such aircraft using a synthetic system such as described in this paper without making an attempt to provide a means of indicating fuel remaining at any instant.

The factors affecting the fuel consumpion of an aircraft are many and varied. They include speed, height, weather conditions, load carried and a host of other factors some of which, for example the human element and aircraft to aircraft variations, are difficult to determine in practice. In addition there must also be considered the number of possible types of aircraft which could be used.

It is no doubt possible but highly impracticable to produce a design which takes into account all possible factors influencing the fuel consumption rate but to make the design feasible and relatively simple a number of assumptions must be made.

The assumptions made are as follows:

(i) The fuel consumption characteristics are considered as a function of speed and height only.

(ii) The fuel consumption characteristics of an aircraft type show no variation from aircraft to aircraft.

(iii) A limited number of aircraft types are considered.

It is most desirable that the circuit designed should be able to make use of the speed and height voltages generated in the aircraft control panels.

The first step in the design procedure is to obtain fuel consumption data on the selected aircraft types. From this data either by trial and error or more elegant mathematical methods it is possible to obtain an approximate equation relating fuel consumption with speed and height. At first sight this would seem to be a formidable task; fortunately the equations which were found to give a satisfactory approximation for the data considered were relatively simple. For the three types of aircraft considered a general equation of the form

$$R = Kv^n - Ah^m$$

was found to give a good approximation to the supplied data.

The problem thus is resolved into appropriately shaping the voltages representing height and velocity obtained from the simulator control circuits, adding and integrating.

A curve having no inflexions, e.g. a power law, can be approximated by a number of straight lines, the accuracy of approximation increasing with the number of lines used. This method of approximation is very suitable for use in electronic circuits using diodes and resistors. If d diodes are used the circuit produces the approximation of $(d+1)$ straight lines. This circuit is also suitable for switching so that the power law can be varied and provides a simple method of selecting one of several aircraft types in one unit.

The method of integration used is that used elsewhere in the system, e.g. coordinate generation. The output is indicated on a meter, full scale deflection representing the tank capacity of the aircraft.

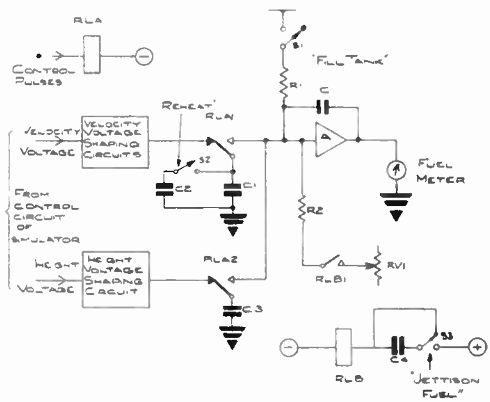

Fig. 19. Block diagram of fuel consumption meter.

A block diagram of the circuit used is shown in Fig. 19, which shows the method used for (i) filling the "tank" to any level, (ii) increasing the fuel consumption when a jet engine is using "reheat," (iii) decreasing the fuel remaining by a predetermined amount as for example when wing tanks are dropped.

It is probably worth mentioning in passing that the integrator used in the above circuit and which is described in Sect. 3.1.1 can be used to perform one of the more difficult mathematical operations used in computing circuits, that of producing an output voltage representing the multiplication of two functions. For example, in certain cases it may be found that a more suitable equation connecting fuel consumption with speed and height takes the form

$$R = \mathrm{f}(v) \,.\, \mathrm{F}(h).$$

In Section 3.1.1 it was shown that

$$\frac{\mathrm{d}V_o}{\mathrm{d}t} = \frac{C_1}{C} \,.\, V_i \,.\, f \text{ volts/sec}$$

and in the applications so far described f, the operating frequency of the relay, has been constant. If f is made proportional to one of the voltages to be multiplied, the output voltage from the circuit will be proportional to the product of two input voltages. Possible circuits for producing pulses at a rate proportional to a voltage include:

(i) Multivibrator methods.

(ii) Control of a d.c. motor fitted with a cam on the motor shaft.

(iii) Relay methods.

5.2. *The Marshalling Unit.*

Some of the ships simulated by the equipment in this trainer will undoubtedly be aircraft carriers. The handling time of the aircraft on such a ship, i.e. their internal movements, taking off, etc. must be considered as part of the overall tactical problem, to be studied. It is thus essential to introduce into the system the realistic time delays associated with such movements.

The unit designed to perform this function is known as a "marshalling unit." Each unit should be capable of handling the maximum number of aircraft that a single aircraft carrier can put in the air at one time. In this instance, for convenience of design, this number was fixed at twenty-five aircraft per aircraft carrier, enabling the circuit design to be based on uniselectors.

The unit must be capable of producing realistic time intervals for the possible movements and provide interlocks to prevent impossible movements, e.g. movement in one step from hangar to airborne. Certain of the permissible movements can take place simultaneously whilst others cannot, e.g. use of the aft lift is not allowable whilst landing on is in progress, and further interlocks are necessary to achieve this. Indication that an impossible operation is being attempted is given by a warning lamp being lit and the whole unit ceasing to function.

Each aircraft is represented by a four position selector switch, and five lamps together with a pair of relays used as information stores. The switch is used to select one of the four possible positions of the aircraft: "Hangar," "Deck Forward," "Deck Aft" or "Airborne." Four of the lamps are used to indicate the present position of the aircraft and the fifth to indicate that a change is about to take place. The action of setting the switch to a different position, firstly lights the change lamp, secondly connects a "marking" voltage to one contact of a uniselector bank. This "marking" voltage is distributed through contacts of the storage relays, arranged so that there is no outlet except to allow other "possible" movements. Several uniselectors are used, one per separate movement in fact. The uniselectors are set in motion to find the "marked" contact by one of a group of keys (again one per movement).

The interlock relays preventing certain movements taking place simultaneously are wired to these keys.

The uniselectors step over their contacts until a marked contact is found and then stop. A circuit is then completed which starts the timing control, a pulsing relay driving a uniselector. This method provides a convenient method of counting down by a factor of up to 50 with a wide and easily adjustable choice of output pulse frequencies. A pulse from this uniselector occurring after the appropriate delay is connected through the first uniselector to the storage relays to change their condition to that representing the new position. This action removes the "marking" voltage from the uniselector contact and if the movement key is still operated the uniselector will search for any other marked contacts.

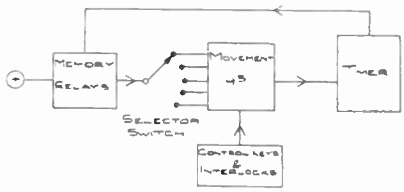

Fig. 20. Block diagram of marshalling unit.

A block schematic of the method is given in Fig. 20. Circuit details are not given as they follow normal Post Office methods.

6. Acknowledgments

This paper is published with the permission of the Admiralty Signal and Radar Establishment.

The author has drawn freely on the unpublished work of Messrs. P. R. Dax, I. A. McCord and H. S. Tomlin of A.S.R.E. in the preparation of this paper and acknowledges this assistance with thanks.

APPLICANTS FOR ELECTION AND TRANSFER

As a result of its December meeting the Membership Committee recommended the following elections and transfers to the Council.

In accordance with a resolution of Council and in the absence of any objections, the election or transfer of the candidates to the class indicated will be confirmed fourteen days after the date of circulation of this list. Any objections or communications concerning these elections should be addressed to the General Secretary for submission to the Council.

Transfer from Associate Member to Member

CASSEY, Rex, B.Sc. *Wellington, New Zealand.*
GODDEN, Alec Williams, B.Sc.(Eng.). *London, W.5.*
MATHEWS, Leonard Frederick. *Dartford.*

Direct Election to Associate Member

CROMPTON, Edward Wyn. *North Harrow.*
MOORE, Flt. Lt. George Robert, M.A., R.A.F. *Harrogate.*
OXLEY, Flt. Lt. John Dawson, B.Sc., R.A.F. *Glasgow.*
SINGER, Wing Cmdr. Herbert Arthur, R.A.F. *Long Crendon.*
SMITH, Arthur Alan. *Bishop Auckland.*
TEMPEL, Julian Andre. *Braunstone.*

Transfer from Associate to Associate Member

NORMAN, John. *Stevenage.*

Transfer from Graduate to Associate Member

CHRISTIAN, Robert Gregory. *Liverpool.*
DUPEE, Sqdn. Ldr. Oswald Arthur, R.A.F. *London, N.W.11.*
GILBERT, Alan Percy. *Stanmore.*
JARRETT, John James. *Higher Denham.*
MACKENZIE, Alasdair. *Edinburgh.*
MARLER, John Robert. *Brentwood.*
MUSGROVE, Stephen Victor. *Orpington.*
REID, Andrew Michael. *Newcastle-on-Tyne.*
SEMPERS, Colin Arthur. *St. Albans.*
SHARMA, Flt. Lt. Dharm Sheef. *Secunderabad.*

Direct Election to Associate

BAILEY, Arthur. *East Molesey, Surrey.*
BRERETON, Walter Leslie. *Salford.*
EMMEL, John William. *Twickenham.*
FERRIS, Major William Victor, B.Sc., Indian E.M.E. *Ahmednagar.*
GILES, Henry. *Dunmow.*
HORNE, Ronald Alfred. *Birmingham.*
JACKSON, John Arthur Cyril. *Chelmsford.*
LAWTON, Arthur Henry. *Nairobi.*
OXBOROUGH, Derek Mervyn. *Bedford.*
PERRY, Thomas Leslie. *Birstall.*
RENTON, Frank Morley. *Edmonton, Alberta.*

Direct Election to Graduate

AVISS, Hugh Wilson. *Tunbridge Wells.*
BREND, Colin James. *Sunbury-on-Thames.*
CRITCHLEY, Arthur Walter. *St. Albans.*
LEWIN, Douglas William. *Ilford.*
POWELL, Alan John. *Chesham.*
SMITH, John Edward. *Uxbridge.*
VENEEAR, Benjamin James, B.Sc.(Eng.)* *Enfield.*
WISE, Frederick Henry. *West Byfleet.*
YORK, Richard Neville. *Pretoria.*

Transfer from Student to Graduate

CHEW BAK KHOON. *Singapore.*
KYI, Maung Aung. *Rangoon.*
McMILLAN, Thomas Lynn. *Glasgow.*

STUDENTSHIP REGISTRATIONS

BALASUBRAMANYAM, T. R., B.Sc.(Hons.). *Madras.*
BHAT, Manohar Ganesh, B.Sc. *Bombay.*
EMERSON MANGALARAJ, Ratnaswamy T., B.Sc. *Madurai.*
FORD, John Bryan Walter. *Shoreham-by-Sea.*
FRISCH, Abraham. *Holon, Israel.*
FROST, Clifford Sylwin. *Ilford.*
GALSTAUN, Garnick C., B.Sc.(Hons.). *London, N.22.*
GOPALAKRISHNAN, Srinivasa, B.Sc. *Madras.*
GRAY, Alexander MacGregor. *Newmains, Lanarkshire.*
HALE, Derek Stanley. *Jersey.*
HANCOCK, Peter Burnell. *New York.*
IGNATIUS, George Joseph L., B.Sc., M.A. *Tiruchirapalli.*
JAGANNADHARAO, Kharidehal, B.Sc., M.Sc. *Vizianagram.*
KELLY, Ronald. *Crawley.*
KESAVAN, Chandrarao, B.Sc. *Kancheepuram.*
KUMAR SHARMA, Sudarshan, B.Sc. *Jullundur City.*
KYAW WIN, Lieut., B.A., Burma Sigs. *Maymyo, Burma.*
LAKSHMI NARAYANA, Bonala, B.Sc. *Khammameth.*
MAHANDHATANAIDU, Hanumantu, B.Sc. *Tallagam.*
MALLORY, Edward Thomas. *Guildford.*
MILLER, Robert George. *London, W.4.*
MURTY, Suryanarayana Yedavilli, M.Sc. *Ganjam, India.*
NAG, Asim, B.c., M.Sc. *Calcutta.*
NANIWADGKAR, Moreshwar Vinayak, B.E. *Barsi.*
NEUMANN, Shimon Siegfried. *Ramat Gan, Israel.*
NIRODI, Suresh Bhavanishanker, B.Sc. *Bombay.*
NOHAR SINGH, Rama, M.Sc. *Dharwa.*
NOOR, Bad Shah. *London, S.W.1.*
OKEAHIALAM, Heron Onyekwere Donatus. *Enugu, Nigeria.*
OKONKWO, Godwin Chukunwike Okoye. *Yaba, Nigeria.*
ONWUEGBUZIA, Stephen Onuorah. *London, N.6.*
PAILOOR, Gopalakrishna, B.Sc. *Welwyn Garden City.*
PALEKER, Shashishekhar Ramkrishna. *Bombay.*
PANSHIKAR, Manohar Atmaram, B.Sc. *Dadar, Bombay.*
PEETT, Edward John. *Sittingbourne.*
RADHAKRISHNAN, Gopalaswamy, B.Sc. *Kumbakonam.*
RAJANGAM, K. B., M.A. *East Tambaram.*
RAJIAH, Mallapuram, B.Sc. *Secunderabad.*
RAMA RAYA KINI, P., B.Sc. *Bangalore.*
RANGASWAMY, Harihar, B.Sc., M.Sc. *Bombay.*
SANTOS, Stephen Verissimo, Akinola. *London, S.E.27.*
SESHAGIRI RAO, Rangarao. *Bangalore.*
SHARMA, Vishnu Deo, B.Sc. *Raghbirpuri.*
SHINDE, Manohar Bhimrao. *Shahapur.*
SRINIVASAN, Subramania, B.Sc. *New Delhi.*
SRIVASTAVA, Mathura Prasad, B.Sc. *Basti.*
STOBART, Peter Cresswell. *Stevenage.*
SUBRAHMANYAN, P. *Kollengode.*
TAYLOR, Raymond Brian. *Brierley Hill.*
TEWARI, Shivasewak, B.Sc., M.Sc. *Etawah.*
TIMBRELL, Leslie Alan. *Wellingborough.**
WIGWE, Shyngle Iwogbo Joseph. *Lagos.*

* Reinstatement.

THE CALCULATION OF CHARACTERISTIC IMPEDANCE BY CONFORMAL TRANSFORMATION*

by

J. C. Anderson, M.Sc. (Associate Member)†

Recommended by the Papers Committee as an example of the application of a specialized mathematical method to the solution of problems in radio and electronic engineering.

SUMMARY

The basic theory is reviewed and the method is applied to the particular case of a coaxial transmission line with a cylindrical outer and a strip inner. It is demonstrated that higher characteristic impedance can be achieved with this configuration than with the normal coaxial construction employing cylindrical wire as an inner.

1. Introduction

In many electrode configurations, the transmission line amongst them, the electric and magnetic fields have variations in two dimensions only, the fields in the direction of the third dimension being uniform. It is this sort of problem which can readily be dealt with by the theory of a complex variable.

If we consider a complex variable, Z, whose value is given by

$$Z = x + jy \qquad \text{.........(1)}$$

we can show its value diagrammatically by plotting it as a point on a graph having the cartesian axes x and y. As x and y vary, the point representing Z will trace out a curve in the xy plane, which we may call the Z plane.

If now we have a second complex variable, W, which is related to the first one by some equation of the form

$$W = f(Z) \qquad \text{.........(2)}$$

then clearly as Z varies, so also will W. If we define W as having components such that

$$W = u + jv \qquad \text{.........(3)}$$

then we may represent W on a graph of v against u. As Z traces out a path in the Z plane, W will trace out some other path in the W plane, i.e. in the v against u graph.

Now we shall show that if the relationship between W and Z is of a certain kind—known mathematically as analytic—we may take u and v as representing electrostatic potential and field respectively. Furthermore, by choosing the right sort of analytic function we may transform a uniform field in the W plane to fit any particular electrode structure in the Z plane. In this way we are enabled to plot out the electric lines of force and the equipotential lines within the electrode structure being considered.

Once the right function has been found the functional relationship between W and Z enables calculation of capacitance.

2. Analytic Functions

Consider a small change in δZ in Z, and the corresponding change δW in W. The differential coefficient of the function will be defined by

$$\frac{\delta W}{\delta Z} = \lim_{\delta Z \to 0} \frac{\delta W}{\delta Z} = \lim_{\delta Z \to 0} \frac{f(Z+\delta Z) - f(Z)}{\delta Z} \qquad \text{.........(4)}$$

A complex function is said to be analytic whenever this derivative exists and is unique.

By the statement that it must be unique we mean that the result must be the same for a change of Z in the x or in the y direction.

Now for a change in the x direction alone $\delta Z = \delta x$, and

$$\frac{dW}{dZ} = \frac{\partial W}{\partial x} = \frac{\partial}{\partial x}(u + jv) = \frac{\partial u}{\partial x} + j\frac{\partial v}{\partial x} \qquad \text{.........(5)}$$

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Similarly for a change in the y direction alone, $\delta Z = \delta y$ and

$$\frac{\mathrm{d}W}{\mathrm{d}Z} = \frac{\partial W}{\partial(\mathrm{j}y)} = \frac{1}{\mathrm{j}}\frac{\partial}{\partial y}(u+\mathrm{j}v) = \frac{\partial v}{\partial y} - \frac{\mathrm{j}\,\partial u}{\partial y} \quad \ldots.(6)$$

If these two equations are to yield the same results, as they must do if the function is to be analytic, we have

$$\left.\begin{aligned}\frac{\partial u}{\partial x} &= \frac{\partial v}{\partial y}\\ \frac{\partial v}{\partial x} &= -\frac{\partial u}{\partial y}\end{aligned}\right\} \quad \ldots\ldots\ldots(7)$$

These are known as the Cauchy-Riemann equations, and represent a necessary condition for the derivative $\delta W/\delta Z$ to be unique at a point and for the function f(Z) to be analytic at that point. It can be shown that, if eqns. (7) are satisfied, then $\delta W/\delta Z$ is the same for any arbitrary direction. The equations therefore also represent sufficient conditions.

3. The Conformal Transformation

Now returning to the Z and W planes, we have pointed out that to each point in the Z plane will correspond some point in the W plane. Furthermore if, for a given function $W = \mathrm{f}(Z)$, any point (x, y) in the Z plane traces out some curve $x = \mathrm{F}(y)$ in the Z plane, the corresponding point (u, v) in the W plane will trace out some curve $u = \mathrm{F}_1(v)$. Thus a point in the Z plane is transformed to a point in the W plane, and the function which accomplishes this is called a transformation.

The above applies to any function, but when the function is also analytic, then $\mathrm{d}W/\mathrm{d}Z$ is independent of the direction of the change dZ, and the entire infinitesimal region in the vicinity of the point W is similar to the infinitesimal region in the vicinity of Z. A transformation having these properties is called a conformal transformation.

3.1. *Representation of Electrostatic Fields*

A most important point to note is that because $Z = x + \mathrm{j}y$, then in the Z plane the lines of constant x will always be perpendicular to the lines of constant y. Similarly in the W plane, the constant u lines will always be perpendicular to the constant v lines. This is illustrated in Fig. 1. Now these are precisely the conditions in an electrostatic field, where the potential lines are always everywhere perpendicular to the lines of electric force. Thus Z and W are eminently suited to represent an electrostatic field.

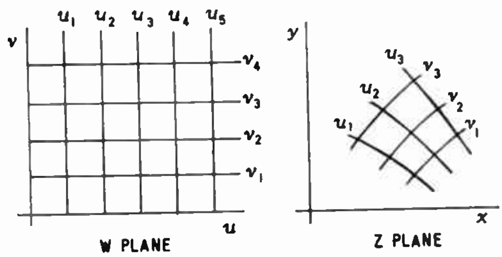


Fig. 1. Illustrating the orthogonal properties of a transformation

To demonstrate this, suppose we take u as representing the potential function in a particular field in volts. The electric field strength will be given by

$$E_x = -\frac{\partial u}{\partial x} \text{ and } E_y = -\frac{\partial u}{\partial y} \quad \ldots\ldots\ldots(8)$$

Now the change dv in v corresponding to the changes dx and dy in x and y respectively will be given by

$$\mathrm{d}v = \frac{\partial v}{\partial x}\mathrm{d}x + \frac{\partial v}{\partial y}\mathrm{d}y \quad \ldots\ldots\ldots(9)$$

But from the equations (7) we have

$$-\mathrm{d}v = \frac{\partial u}{\partial y}\mathrm{d}x - \frac{\partial u}{\partial x}\mathrm{d}y = -E_y\,\mathrm{d}x + E_x\,\mathrm{d}y \quad \ldots\ldots\ldots(10)$$

Now eqn. (10) clearly represents the total change in the E field, so that the total change in electric flux corresponding to the change in field dv will simply be given by

$$-\mathrm{d}\psi = \varepsilon\,\mathrm{d}v$$

where ε is the permittivity of the medium. Thus, if u is the potential function, providing we choose our reference for zero flux at $v = 0$, the flux will be given by

$$-\psi = \varepsilon v \text{ coulombs/metre} \quad \ldots\ldots\ldots(11)$$

Similarly if v were chosen as the potential function, then u would represent the flux.

The way in which the conformal transformation is used in practice is to take a uniform field in the W plane, represented by equally spaced u and v lines, and choose a function which transforms it to a configuration in the Z plane which fits with the electrode arrangement being considered.

4. The Inverse Cosine Transformation

We will now consider the functional relationship between W and Z to be

$$W=\cos^{-1}Z \qquad \text{.........(12)}$$

i.e. $x+jy=\cos(u+jv)=\cos u\cosh v-j\sin u\sinh v$. Equating real and imaginary parts leads to

$$x=\cos u\cosh v \quad \text{and} \quad y=-\sin u\sinh v,$$

from which we get

$$\frac{x^2}{\cosh^2 v}+\frac{y^2}{\sinh^2 v}=1 \qquad \text{.........(13)}$$

and

$$\frac{x^2}{\cos^2 u}-\frac{y^2}{\sin^2 u}=1 \qquad \text{.........(14)}$$

Eqn. (13) will be recognized as the equation of a family of ellipses, whilst eqn. (14) is that of a family of hyperbolae. These are sketched in Fig. 2.

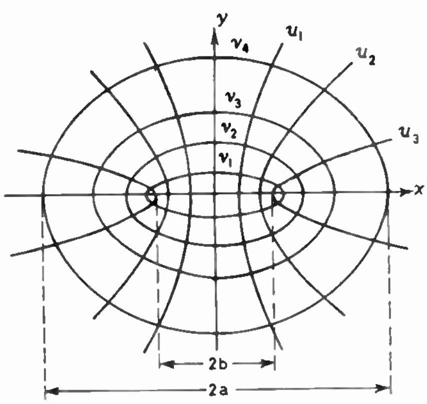


Fig. 2. Inverse cosine transformation.

This transformation is obviously suitable for the case of confocal elliptic cylinders. A limiting case of these is where we have a transmission line in which the central conductor is a strip stretching between the foci of an elliptic outer conductor. If the latter is of relatively large radius, little error will be introduced by treating a cylindrical outer as an ellipse. We will now consider an example of such a case.

4.1. *A particular case*

Let us take a strip inner conductor of width $2a$, in a cylindrical outer of radius $6a$ (say). Let the potential of the inner be V_0 and that of the outer zero.

It is necessary to introduce a scale factor so that when we transform to the Z plane, the contour spacings fit with the particular case. Referring to the diagram of the cross-section through the line shown in Fig. 2, we clearly require that when $y=0$ the minimum possible value for x', and therefore for Z', should be a (primed symbols apply to the particular problem). Now, in general the minimum value of Z is obtained when $W=0$ and will be 1. Thus we require that when $Z=1$, $Z'=a$, i.e. $Z=Z'/a$.

Now we will let v be the potential so u is the field between the conductors. We must make these fit with the particular problem. This can be done by inserting arbitrary constants in the functional relationship and then evaluating them for the particular problem under discussion.

$$\text{Let} \quad W'=A\cos^{-1}\left(\frac{Z'}{a}\right)+jB=u'+jv' \qquad \text{........(15)}$$

Now we require $v'=0$ when v is such as to make $x=6a$ when $y=0$. Using eqn. (13) if we put $y=0$ we have $x=\cosh v$ so that when $x=6$, $v=2{\cdot}48$. From eqn. (15) we have

$$u'+jv'=AW+jB$$
$$=Au+Ajv+jB$$

When $v'=0$
$$0=Av+B$$
$$=2{\cdot}48A+B$$

therefore $B=-2{\cdot}48A$

When $v'=V_0$ we have $v=0$ as this corresponds to the surface of the inner conductor.

Then $V_0=B$ and so $A=-V_0/2{\cdot}48$.

Substituting for A and B we now have

$$u'+jv'=V_0\left(\frac{-W}{2{\cdot}48}+j\right)$$

$$u'+j(v'-V_0)=\frac{-V_0}{2{\cdot}48}\cos^{-1}\frac{(Z')}{(a)}$$

$$\frac{x'+jy'}{a}=\cos\left[\frac{-2{\cdot}48}{V_0}(u'+j(v'-V_0))\right]$$

This leads to

$$\left.\begin{aligned}\frac{x'^2/a^2}{\cosh^2 f(v')} + \frac{y'^2/a^2}{\sinh^2 f(v')} &= 1 \\ \text{and} \quad \frac{x^2/a^2}{\cos^2 f(u')} - \frac{y'^2/a^2}{\sin^2 f(u')} &= 1\end{aligned}\right\} \quad \ldots\ldots(16)$$

$$\text{where } f(v') = \frac{2{\cdot}48v'}{V_0} - 2{\cdot}48 \quad \ldots\ldots\ldots(17)$$

$$\text{and} \quad f(u') = \frac{2{\cdot}48}{V_0}\, u' \quad \ldots\ldots\ldots(18)$$

From eqns. (16), (17) and (18), we may plot the values of x'/a and y'/a for various values of u' and v'. This is done in Fig. 3, which illustrates the extent of the error introduced by the approximation of the outer to a cylinder.

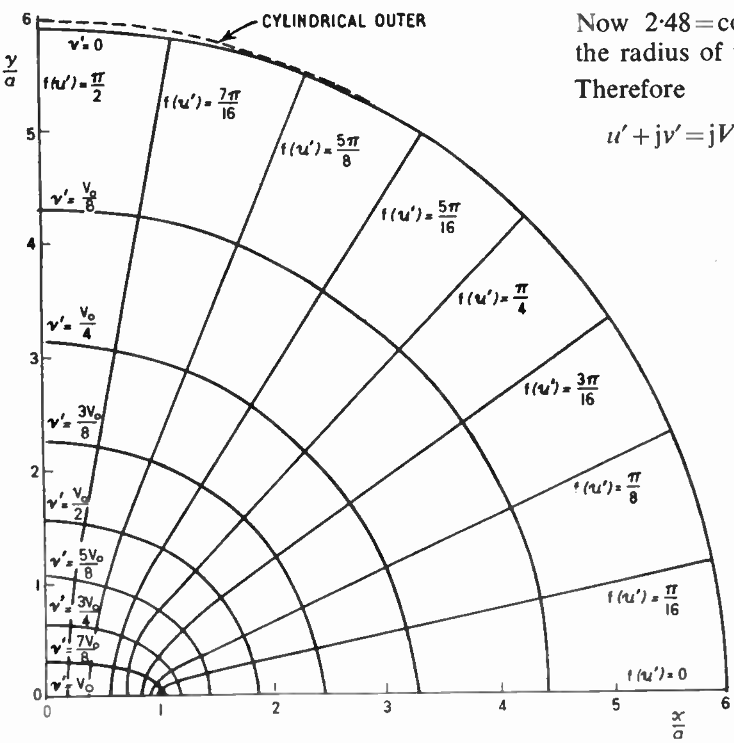
5. Calculation of Capacitance

Now in general we have $W = \cos^{-1} Z$ so we may write

$$Z = \cos W = \frac{\exp(jW) + \exp(-jW)}{2}$$

from which $\quad \exp(2jW) - 2Z\exp(jW) + 1 = 0$

whence $\quad \exp(jW) = Z \pm (Z^2 - 1)^{\frac{1}{2}}$



i.e. $$W = \frac{1}{j}\log(Z \pm (Z^2 - 1)^{\frac{1}{2}}) \quad \ldots\ldots\ldots(19)$$

Now if R is a complex number,

$$\log R = \log|R|\underline{/\theta} = a + jb$$

so that

$$|R|\underline{/\theta} = \exp(a + jb) = \exp(a)\exp(jb) = \exp(a)\underline{/b},$$

therefore $b = \theta$ and $\exp(a) = |R|$ or $a = \log|R|$.

Hence $\quad \log R = \log|R| + j\theta,$

which may be written

$$\log R = \log|R| + j \arg R \quad \ldots\ldots\ldots(20)$$

Applying this to (19)

$$W = \frac{1}{j}\left\{\log|Z \pm (Z^2 - 1)^{\frac{1}{2}}| + j\arg[Z \pm (Z^2 - 1)^{\frac{1}{2}}]\right\}$$

For our particular case

$$u + jv' = jV_0 - \frac{V_0}{2{\cdot}48}\cos^{-1}\frac{Z'}{a}$$

$$= jV_0 + \frac{jV_0}{2{\cdot}48}\left[\log\left|\left(\frac{Z'}{a}\right) \pm \sqrt{\left(\frac{Z'}{a}\right)^2 - 1}\right| + \right.$$

$$\left. + j\arg\frac{Z'}{a} \pm \frac{\sqrt{Z'^2}}{a} + 1\right]$$

$$= jV_0 + j\,\frac{V_0 \log|\text{mod}|}{2{\cdot}48} - \frac{V_0}{2{\cdot}48}\arg$$

Now $2{\cdot}48 = \cosh^{-1}6 = \cosh^{-1}(b/a)$ where b is the radius of the outer.

Therefore

$$u' + jv' = jV_0 + \frac{jV_0 \log \text{mod}}{\cosh^{-1}(b/a)} - \frac{V_0}{\cosh^{-1}(b/a)}\arg$$

Fig. 3. Inverse cosine transformation for strip and cylinder,

$$\frac{\text{diameter of cylinder}}{\text{width of strip}} = 6.$$

Equating real and imaginary parts

$$v' = V_0 \frac{\left((1 + \frac{\log Z \pm (Z^2 - 1)^{\frac{1}{2}}}{a}\right)}{\cosh^{-1}(b/a)} \quad \ldots\ldots(21)$$

and $$u' = \frac{-V_0 \arg\left(\frac{Z \pm (Z^2 - 1)^{\frac{1}{2}}}{a}\right)}{\cosh^{-1}(b/a)} \quad \ldots\ldots\ldots(22)$$

Now the flux $\psi = \varepsilon u'$ and, by Gauss' Law, the total charge q is given by the number of flux lines ending on the conductor. This will be found from the change in ψ in going once round the perimeter.

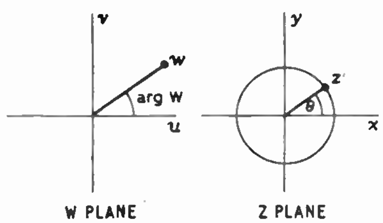


Fig. 4. Polar notation in the Z and W planes.

Now referring to Fig. 4, if we represent Z by the polar notation $|Z|\underline{/\theta}$, to go once round the perimeter we must change θ by 2π. We require to know what change in the argument of W this will imply.

Now $\arg W$

$$= \arg [Z \pm (Z^2 - 1)^{\frac{1}{2}}]$$

$$= \arg [|Z|\underline{/\theta} \pm \sqrt{(|Z|^2 \underline{/2\theta} - 1)}]$$

$$= \arg [|Z|\underline{/\theta} \pm \sqrt{(|Z|^2 \cos 2\theta + j|Z|^2 \sin 2\theta - 1)}]$$

Now when θ changes by 2π the square-rooted term in the above expression does not change, so the total change in argument is equal to the change in $|Z|\underline{/\theta}$, which is 2π.

$$\text{Thus } q = \frac{2\pi\, \varepsilon\, V_0}{\cosh^{-1}(b/a)}$$

from which it follows that the capacitance of the line is given by

$$C = \frac{2\pi\varepsilon}{\cosh^{-1}(b/a)} \text{ farads per metre} \quad \ldots\ldots\ldots(23)$$

5.1. *Particular case*

For the case of $b = 6a$ quoted above, with air dielectric, we get

$$C = \frac{10^{-9}}{18 \times 2{\cdot}48} = 22{\cdot}4 \text{ pF/metre.}$$

6. Characteristic Impedance of Strip/Cylindrical Lines

From the well-known symmetry between inductance and capacitance for a line, we may write down the expression for inductance as

$$L = \frac{\mu}{2\pi} \cosh^{-1}\left(\frac{b}{a}\right) \text{ henry/metre} \quad \ldots\ldots\ldots(24)$$

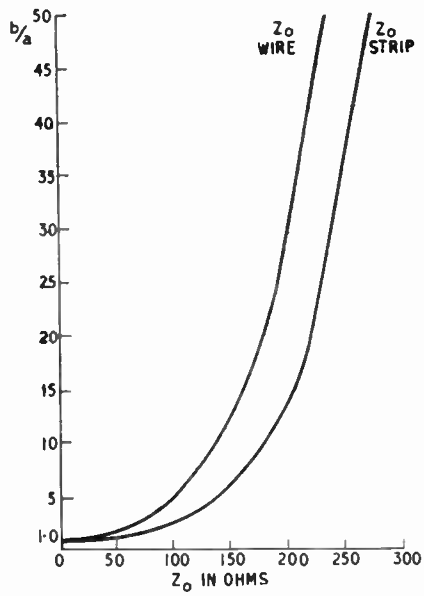


Fig. 5. Characteristic impedance of strip/cylinder and cylinder/cylinder transmission lines.

Thus, neglecting resistance and leakance—permissible at high frequency—we get the expression for characteristic impedance as

$$Z_0 = \frac{1}{2\pi}\sqrt{\frac{\mu}{\varepsilon}} \cosh^{-1}\left(\frac{b}{a}\right) \text{ ohms} \quad \ldots\ldots(25)$$

For air dielectric this gives

$$Z_0 = 60 \cosh^{-1}\left(\frac{b}{a}\right) \text{ ohms} \quad \ldots\ldots.(26)$$

Referring to the earlier definitions, b/a is the ratio of the radius of the outer to half the width of the strip, i.e. it is the ratio of the diameter of the outer to strip width. It is emphasized that the assumption throughout the above has been that the thickness of the strip is negligible compared with its width. In general, the width should be ten or more times the thickness.

For the normal coaxial line the characteristic impedance is given by $Z_0 = 138 \log_{10}(b/r)$ ohms for air dielectric, where r is the radius of the inner conductor.

Taking $b/r = b/a$, Fig. 5 shows the values of Z_0 for variation of b/a. It will be seen that, for a strip of width equal to the diameter of the inner wire, the same characteristic impedance may be attained with less than half the diameter of outer conductor. This could be particularly useful for higher values of characteristic impedance, where a coaxial construction is required.

7. Conclusions

A single case of the use of the conformal transformation has been illustrated here; there are a number of transformations suitable for a wide range of electrode structures. The inverse cosine one itself may be applied to the field between perpendicular planes with a finite gap, the field between two confocal hyperbolic cylinders,* between a hyperbolic cylinder and a plane conductor extending from the focus to infinity and the field between two infinitely long strips.

Ramo and Whinnery† give other functions covering the field at a corner between two plane electrodes and that between two parallel conducting cylinders; they also give the Schwarz transformation covering general polygons, which enables calculation of the fields in the vicinity of stepped electrodes. Titchmarsh‡ gives a useful treatment from the pure mathematics standpoint, and quotes a number of other useful functions.

* L. A. Pipes, "Applied Mathematics for Engineers and Physicists," pp. 489-492 (McGraw-Hill, New York, 1946). (A derivation of the example given in this paper is included.)

† S. Ramo and J. R. Whinnery, "Fields and Waves in Modern Radio," 2nd Edn. (Wiley, New York, 1953).

‡ E. C. Titchmarsh, "Theory of Functions" (Oxford University Press, 1932).

DEVELOPMENTS IN COMPONENT DESIGN

An Interim Report on the International Symposium held in Malvern in September, 1957.

The International Symposium on Electronic Components was held under the sponsorship of the Ministry of Supply at the Royal Radar Establishment, Malvern, between September 24th-26th. Its purpose was to review and discuss component developments, new materials for components, application techniques and the problems of severe environmental conditions in component design. Although the emphasis was on components for use in military equipment, it is felt that such applications lead to improvements in other fields, and that the material presented will be of general interest to members of the Institution. The Institution has therefore obtained permission from the Ministry of Supply to publish in the *Journal* the extended summaries which make up the Interim Report of the Symposium, and the first two papers appear below. Subsequent issues will contain the summaries of the remaining twenty-five papers. It is understood that the complete proceedings of the Symposium, including the full text of discussions, will be published in the near future, and an announcement about this will be made in due course.

Problems Affecting the Design of Service Components and a Survey of Present and Future United Kingdom Developments

G. W. A. Dummer, M.B.E.*

After reviewing briefly the two organizations in Great Britain concerned with components for military use (Radio Component Research and Development Committee and the Radio Component Standardization Committee), Mr. Dummer went on to say that one of the problems facing both research workers and industry in components was the increasing number and types which were needed for particular requirements. Apart from temperature and humidity considerations, the design and production of components were now required to meet certain standards regarding vibration and shock, miniaturization for transistor circuits, guaranteed life under severe environmental conditions, nuclear radiation hazards and even resistance to acoustic noise from jet engines.

A new technical classification system which can cover present and future requirements was now required. Four main classifications were envisaged for ground, ship, airborne and guided missile components, with sub-classifications covering temperature, vibration, life, nuclear radiation effects, etc.

The reliability problem for complex military electronic equipment was stressed and, based on an analysis of fault rates from a number of sources, the following order of component faults was given† (1) valves, (2) resistors, (3) capacitors, (4) transformers, (5) switches, (6) relays, (7) meters, (8) plugs and sockets, (9) cables and connectors. It was emphasized that with the lower failure rates required by the Services (of the order of 0·1% or even 0·01% under given conditions), it was essential to test much larger quantities of components than previously. Equipment for the automatic testing of 1,000 components (initially resistors) was described briefly.

In an attempt to bring practical design into what is admittedly the extremely complicated field of reliability, an airborne radar system was being built at R.R.E. in order to determine the maximum reliability which can be obtained by modern design. It has been constructed in order to examine in detail techniques such as circuit and system simplification, vigorous environmental pre-testing of all components during construction, the use of the best tried components and valves, including wired-in valves and cathode ray tubes, adequate derating of all valves and components and the use of recently developed components to replace valves wherever possible, e.g. magnetic amplifiers replacing valve servo amplifiers, magnetic modulators replacing hydrogen thyratron modulators, and the protection of all components by equipment sealing or potting. It is intended that this equipment will be flown and subjected to continuous environmental testing in order to provide a standard of reliability for a typical complex airborne radar system.

* Superintendent, Technical Services, Royal Radar Establishment, Malvern.

U.D.C. No. 621.396.69

† See also: G. W. A. Dummer, "The maintainability of services equipment, Part 1," *J.Brit.I.R.E.*, 15, p. 283, June 1955.

The influence of assembly techniques on components was discussed in some detail. The effect of plastic, potted circuits on components due to shrinkage after gelation and the effect of pressure on strain-sensitive components in potted circuits was instanced. The introduction of printed wiring and its effect on the shapes, the leads, the ease of soldering and "snap-in" methods of insertion prior to dip soldering were also discussed.

The effect of transistors on components has been considerable and the reduction in size for low voltage operation was mentioned. Voltages of 1·5, 3, 6, 9, 15, 30 and 50 V had been standardized by the British Radio Components Standardization Committee.

High temperature components were described and equipment operating in an ambient temperature of 150°C was demonstrated in the laboratory. Experimental work on transformers at very much higher temperatures was described briefly and two transformers operating at 500°C were demonstrated.

A programme of work was initiated some three years ago on the relationship between the life and ambient temperature of components and short life rating figures have now been obtained for many types of components.

Finally, in an attempt to look into the future, it was suggested that the reliability of components in guided missiles was probably the greatest single factor requiring effort within the next five years. Future electronic constructions would probably use printed wiring with miniature components and transistors either hand or automatically assembled and dip soldered.

Work on smaller and smaller components may lead to film techniques in which thin, resistive, capacitive, and even thin magnetic films now being made experimentally may lead to solid circuit assemblies. A solid circuit multivibrator was shown in which a piece of silicon had been doped and shaped to form the équivalent of four normal transistors with the resistors and capacitors deposited in film form directly on the silicon block with interleaving insulating films. The size of this four transistor multivibrator was 3/10 in. square by 1/10 in. thick.

New Concepts, Techniques and Components for Transistor Circuits

A. W. Rogers*

Mr. Rogers opened his lecture by saying that the impact of the transistor on American Service equipment was considerable. The U.S. is at present producing over 30,000,000 transistors a year and the Army alone had a projected utilization of about 100,000,000 transistors in ultimate production. He then reviewed some of the components which had been developed in America to operate in transistor circuits and showed a number of dip soldered printed wiring constructions in which these components were used. He then drew attention to the parts density or packing of components and gave instances of volumetric efficiency within systems and components. In a helmet radio the average packing of the components was 216,000 components per cubic foot. Small resistors ranged as high as 3,500,000 per cubic foot whilst transformers and i.f. coils ranged between 33,000 and 63,000 per cubic foot. Transistors ranged between 50,000 and 150,000 per cubic foot. Even with this density of packing the overall volume used was only 27% of the total volume available in the box. He was of the opinion that the efficient packaging of components in equipment was one of the most important requirements of the future.

He then discussed reliability and gave a maximum permissible failure rate for components as one in 10,000 per 1,000 operating hours. This means that a 1,200 component system will have a reliability goal of one equipment failure per 8,000 hours of operating time; a larger system such as Nike ground guidance equipment with 10,000 components will have a reliability goal of one system failure per 1,000 hours of operating time.

Mr. Rogers then described some of the tests being carried out on resistors and capacitors. For resistors a total of 3,000 carbon composition and deposited carbon film resistors had been placed on continuous life test at rated wattage and 70°C. Failure rates for deposited carbon resistors were about 0·3% per thousand hours and for carbon composition resistors the figure was 10% per 1,000 hours (for 5% resistance change in both cases). Some 12,000 capacitors were now being similarly evaluated.

He concluded by emphasizing the need for improved volumetric efficiency and for new concepts to pack large numbers of components in a small space with high efficiency and reliability.

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U.D.C. No. 621.396.69:621.382

APPROXIMATE RELATIONS BETWEEN TRANSIENT AND FREQUENCY RESPONSE*

by

H. H. Rosenbrock, B.Sc., Ph.D.†

A paper presented at the Convention on "Electronics in Automation" in Cambridge on 28th June 1957. In the Chair: Dr. Denis Taylor.

SUMMARY

The paper is based upon an existing graphical method for obtaining the transient response of a stable, linear device from its frequency response, and vice versa. This method is extended to cover certain difficult cases, and examples in electrical and in chemical engineering are given.

1. Introduction

It is well known[1] that the dynamic behaviour of a linear device can be described in terms of its response either to a sudden change of input (usually a step or an impulse) or to steady sinusoidal inputs of different frequencies. The two descriptions are equivalent, so that the transient response can be obtained from the frequency response or vice versa. Moreover the frequency response is usually easier to calculate than the transient response, whereas the transient response often gives more useful information about the practical behaviour of the device. Thus there are advantages in being able to make the conversion from one type of response to the other without too much labour.

The mathematical relationship between transient response and frequency response can be written[2]

$$\mathrm{f}(t)=H(0)+\frac{2}{\pi}\int_{\omega=0}^{\infty}\mathcal{J}H(\mathrm{j}\omega)\,.\cos\omega t\,.\mathrm{d}(\log\omega), \qquad \ldots\ldots\ldots(1)$$

where $H(\mathrm{j}\omega)$ is the complex function expressing the steady-state response of a stable, linear device at frequency ω, and $\mathrm{f}(t)$ is the response of the same device to a unit step-change of input at $t=0$. The symbol $\mathcal{J}$ implies that only the imaginary part of $H(\mathrm{j}\omega)$ is to be taken. The converse relationship can be written

$$H(\mathrm{j}\omega)=\mathrm{f}(\infty)+\omega\int_{t=0}^{\infty}t[\mathrm{f}(t)-\mathrm{f}(\infty)]\sin\omega t\,\mathrm{d}(\log t)+ +\mathrm{j}\omega\int_{t=0}^{\infty}t[\mathrm{f}(t)-\mathrm{f}(\infty)]\cos\omega t\;\mathrm{d}(\log t) \qquad \ldots\ldots\ldots(2)$$

Even when the data are given in an analytical form, equations (1) and (2) are usually difficult to evaluate, and in practice the data are often obtained numerically by experiment. For this reason a numerical or graphical method for evaluating the equations is desirable. Such a graphical method has been described in an earlier paper[2], and it makes use of a pair of transparent cursors. To evaluate eqn. (1), the graph of $\mathcal{J}\,H(\mathrm{j}\omega)$ is plotted on special graph paper against $\log\omega$. Then the first (cosine) cursor, shown in Fig. 1, is placed over this graph and slid along to left or right until the appropriate value of t is shown on the horizontal scale. The sum of the intercepts of the graph on the vertical scales of the cursor then gives the value of the integral in eqn. (1) to within 1 or 2 per cent. By moving the cursor to left or right the integral can be evaluated again for different values of t, and the behaviour of $\mathrm{f}(t)$ can in this way be found. A similar (sine) cursor[3] is used to evaluate the first integral in eqn. (2).

The result of applying the cosine cursor to the graph of a function $\mathrm{F}(\omega)$ is written $\mathcal{C}\{\mathrm{F}(\omega)\}$, and is given by the formula

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U.D.C. No. 517.512:661

$$\mathcal{C}\left\{F(\omega)\right\} = \frac{\pi}{200}\int_0^\infty \frac{F(\omega)}{\omega}\cos \omega t \, d\omega \quad .....(3)$$

Similarly the result of using the sine cursor is

$$\mathcal{S}\left\{F(\omega)\right\} = \frac{200}{\pi}\int_0^\infty \frac{F(\omega)}{\omega}\sin \omega t \, d\omega \quad ....(4)$$

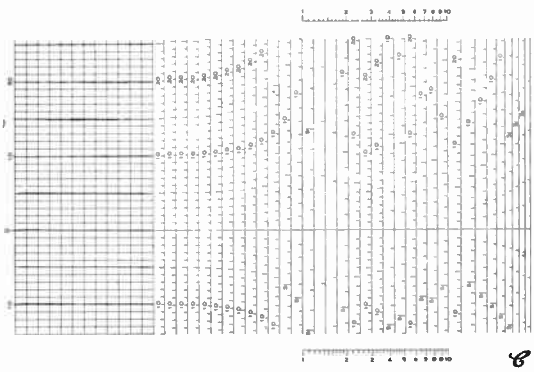

Fig. 1. The cosine cursor.

With this notation the graphical method for evaluating eqns. (1) and (2) can be written

$$f(t) = H(0) + \frac{1}{100}\,\mathcal{C}\left\{\mathcal{I}H(j\omega)\right\} \quad ....(5)$$

and

$$H(j\omega) = f(\infty) + \frac{\pi\omega}{200}\,\mathcal{S}\left\{t[f(t) - f(\infty)]\right\}$$
$$+\, j\frac{\pi\omega}{200}\,\mathcal{C}\left\{t[f(t) - f(\infty)]\right\} \quad ....(6)$$

Equations (5) and (6) describe the methods given earlier[2] and for many purposes they are the most convenient. In certain problems, however, their use may be limited in the following ways:

(i) When the frequency-response of the device has a wide pass-band over which the phase changes by a large amount, $\mathcal{I}\,H(j\omega)$ will contain much detail and will be tedious to compute. The method indicated by eqn. (5) then becomes inconvenient and its accuracy is impaired. This situation may arise in amplifiers for television channels, or when a large number of similar circuits are connected in cascade[4].

(ii) Equation (5) is easily generalized to allow for inputs other than a unit step: if the Laplace transform of the input is $\theta(p)$ it is merely necessary to replace $\mathcal{I}H(j\omega)$ by $\mathcal{I}\,[j\omega\,\theta(j\omega)\,H(j\omega)]$. It is, however, necessary that this last expression should remain finite for all values of ω, and for this reason eqn. (5), as generalized, cannot be used when the input is sinusoidal. This difficulty arises in finding the transient response of an electrical circuit when a sinusoidal input is suddenly applied to it.

(iii) Equation (6) is convenient for finding the lower-frequency end of the response, which is often the part of most interest. At higher frequencies, however, the results become inaccurate because $\{\mathcal{C}\,t[f(t) - f(\infty)]\}$ and $\mathcal{S}\{\,t[(t) - f(\infty)]\,\}$ become small, while the errors in these quantities remain roughly constant in magnitude. After multiplication by ω the magnitude of the errors may be large. This difficulty has arisen in finding the frequency response of a distillation column from the transient response as calculated on an electronic computer.

In the following Sections alternative formulae are presented by which these difficulties can be avoided, and examples of their use are given.

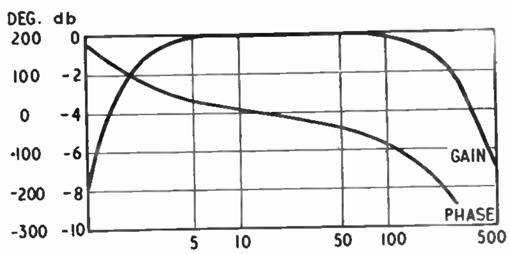

Fig. 2. Gain and phase characteristics of a device having a wide pass-band.

2. Devices having a Wide Pass-band

Figure 2 shows the type of gain and phase characteristic in question. Over the region, between $\omega = 1$ and $\omega = 500$, where the gain is high, the phase changes by about 600°. The graph of $\mathcal{I}\;H(j\omega)$ will therefore show a number of successive positive and negative loops, as in Fig. 3. For wider pass-bands the number of loops of $\mathcal{I}\;H(j\omega)$ would be greater, and the labour of drawing the graph would become excessive. Moreover, the loops of $\mathcal{I}\,H(j\omega)$ tend to become narrow at the higher frequencies, and this reduces the accuracy of the results obtained from eqn. (5).

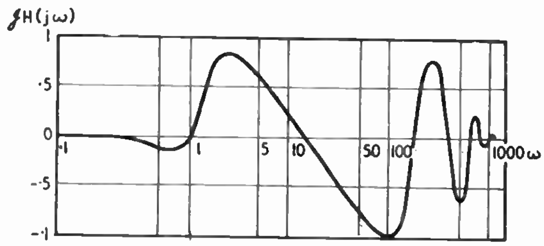

Fig. 3. Sketch of $\mathcal{I}H(j\omega)$ corresponding to Fig. 2.

To avoid this difficulty the origin of time is changed so that the input is a unit step at time $t=-\tau$. The output at time t is then given (Section 8) by the following formulae:

$$\left.\begin{aligned} f(t)&=0, && t<-\tau\\ &=\tfrac{1}{2}\{A(|t|)-B(|t|)+H(j\infty)\}, && -\tau\leqslant t\leqslant 0\\ &=\tfrac{1}{2}\{A(|t|)+B(|t|)+H(j\infty)\}, && 0\leqslant t\leqslant\tau\\ &=A(|t|),\ [=B(|t|)+H(j\infty)], && t\geqslant\tau \end{aligned}\right\} \quad\ldots\ldots\ldots(7)$$

where

$$A(|t|)=H(0)+\frac{1}{100}\,C\left\{\mathcal{I}[(H(j\omega)-H(j\infty))e^{j\omega\tau}]\right\} \quad\ldots\ldots\ldots(8)$$

$$B(|t|)=\frac{1}{100}\,S\left\{\mathcal{R}[(H(j\omega)-H(j\infty))e^{j\omega\tau}]\right\} \quad\ldots\ldots\ldots(9)$$

If $H(j\omega)-H(j\infty)$ is written in the form $re^{j\theta}$ the functions to which the cursors are applied are

$$\mathcal{I}[(H(j\omega)-H(j\infty))e^{j\omega\tau}]=r\sin(\omega\tau+\theta)\ldots\ldots(10)$$

$$\mathcal{R}[(H(j\omega)-H(j\infty))e^{j\omega\tau}]=r\cos(\omega\tau+\theta)\ldots\ldots(11)$$

Equations (7), (8) and (9) hold for any value of τ, but they are most convenient to use when τ is chosen so that $\omega\tau+\theta$ is as nearly constant as possible over the pass-band. The equations are appropriate for devices of a low-pass or band-pass nature. For high-pass devices eqn. (5) is usually more appropriate.

Equations (7), (8), and (9) incidentally give an insight into the transient behaviour of the devices to which they are applicable. For example, if $\omega\tau+\theta$ is negligibly small over the whole pass-band, $A(|t|)$ will be constant, and $f(t)$ will be symmetrical about $t=0$. If the graph of θ against ω is slightly convex upwards and $\omega\tau+\theta$ can be made zero at the ends of the pass-band it will then be small and positive in the middle of this band. The variable part of $A(|t|)$ will be positive,* and the curve of transient response will be lifted in the region near $t=0$. These results are illustrated in Fig. 4, and many others of the same kind can be derived.

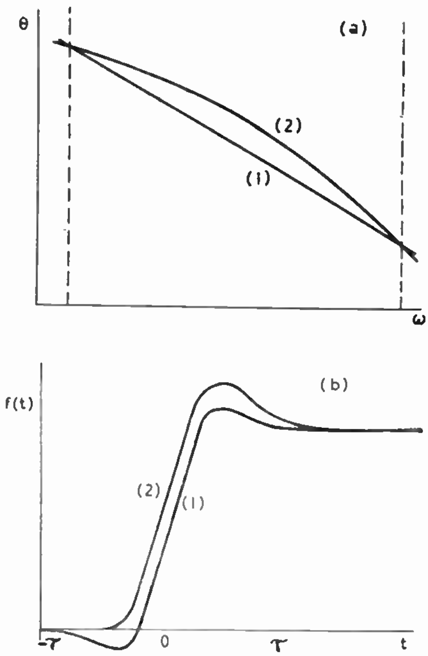

Fig. 4. (*a*) Sketch of two phase characteristics. (*b*) Sketch of the corresponding responses to a unit step-function.
The scale of ω in (*a*) is linear, and the broken lines define the pass-band. The slope of curve (1) in (*a*) gives τ.

2.1. *Example*

The transfer function from which Fig. 2 was calculated is

$$H(p)=\frac{p^4}{[p^2+1\cdot 6p+1]^2}\cdot\frac{1}{[(p/1000)^2+1\cdot 6(p/1000)+1]^{10}} \quad\ldots\ldots\ldots(12)$$

and this will be used as an example. The most convenient value of τ was found to be 0·016 sec,

* The time derivative of $A(|t|)$ will in these circumstances correspond to a pair of echoes.[5]

and the resulting values of $r\sin(\omega\tau+\theta)$ and $r\cos(\tau\omega+\theta)$ are shown in Fig. 5. Fig. 6 shows a number of points of the transient response obtained by use of the cursors according to eqn. (7). It will be noticed that although both cursors must be used when $-\tau<t<\tau$, a pair of values of $A(|t|)$ and $B(|t|)$ gives two values of $f(t)$.

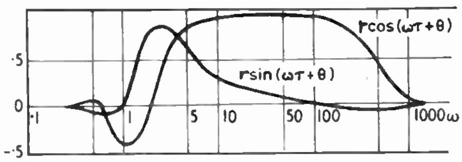

Fig. 5. Graphs of $r\cos(\omega\tau+\theta)$ and $r\sin(\omega\tau+\theta)$ for the system of Fig. 2.

A direct analytical solution for the transient response corresponding to the transfer function given in eqn. (12) would be tedious to find. The responses corresponding to its two factors can, however, be easily found. They are respectively

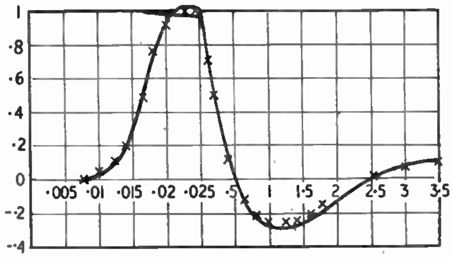

Fig. 6. Transient response of the system of Fig. 2. The curves are based on an analytical solution and the crosses show points obtained from Fig. 5 by use of the cursors. The scale of time is contracted after $t=0{\cdot}025$ sec.

$$f_1(t)=e^{-0\cdot8t}\Big\{(-3{\cdot}2778+1{\cdot}3t)\sin 0{\cdot}6t+ +(1-0{\cdot}4333t)\cos 0{\cdot}6t\Big\} \quad\ldots\ldots\ldots(13)$$

and[6]

$$f_2(t)=\frac{1}{0{\cdot}6^{20}\,2^9\,9!}\int_0^u u^{10}e^{-4u/3}\left[\sqrt{(\pi/2u)}\,J_{19/2}(u)\right]du \quad\ldots\ldots\ldots(14)$$

where in eqn. (14), $u=600t$ and $J_{19/2}$ is the Bessel function of the first kind and order 19/2. The required response is then given[7] by the convolution integral

$$f(t)=\int_0^t f_1(t-x)\,f_2'(x)\,dx \quad\ldots\ldots\ldots(15)$$

Now for small values of t, $f_1(t-x)\cong 1$, while for large values of t, $f_2'(x)$ behaves in eqn. (15) very much like a unit impulse at $x=0{\cdot}016$ sec. It follows that for small values of time $f(t)\cong f_2(t)$, while for large values of time $f(t)\cong f_1(t-0{\cdot}016)$. The two functions $f_2(t)$ and $f_1(t-0{\cdot}016)$ are plotted in Fig. 6 for comparison with the results from the cursors.

3. Suddenly-applied Sinusoidal Input

A sinusoidal input can be represented in the usual way by a complex function $Ae^{j\Omega t}$. The physical input is equal to the real part of this function, and the complex number A determines the amplitude and phase of the sinusoid. If the input is zero up to time $t=0$, and thereafter is equal to the real part of $Ae^{j\Omega t}$ then the output will be the real part of $f(t)$, where $f(t)$ is given (Section 8) by

$$f(t)=Ae^{j\Omega t}\Big[H(j\Omega)+ +\frac{1}{200}\mathcal{C}\Big\{\mathcal{I}[H(j\Omega+j\omega)-H(j\Omega-j\omega)] -j\frac{1}{200}\mathcal{C}\Big\{\mathcal{R}[H(j\Omega+j\omega)-H(j\Omega-j\omega)]\Big\}\Big] \ldots(16)$$

Equation (16) is exact, but when $H(j\omega)$ is nearly symmetrical about Ω the following simpler approximation may be used

$$f(t)=Ae^{j\Omega t}\Big[H(j\Omega)+\frac{1}{100}\mathcal{C}\Big\{\mathcal{I}\,H(j\Omega+j\omega)\Big\}\Big] \quad\ldots\ldots\ldots(17)$$

This is the usual band-pass to low-pass transformation, as will be seen by comparison with eqn. (5).

3.1. *Example*

As an example, the transfer function

$$H(p)=\frac{1}{p^2+p+1} \quad\ldots\ldots\ldots(18)$$

will be considered. The input will be zero up to time $t=0$ and thereafter will be equal to $\cos t$. Thus in eqn. (16), $A=1$, $\Omega=1$, and $H(j\Omega)=-j$. Fig. 7 shows the functions

$-\mathcal{I}[H(j\Omega+j\omega)-H(j\Omega-j\omega)]$

and

$-\mathcal{R}\,[H(j\Omega+j\omega)-H(j\Omega-j\omega)]$

and in Fig. 8 are shown a number of points obtained by use of the cursors according to eqn. (16).

The Laplace transform of the output is

$$\frac{p}{p^2+1} \cdot \frac{1}{p^2+p+1}$$

from which the output is found to be

$$f(t)=\sin t - 1{\cdot}155e^{-0{\cdot}5t}\sin 0{\cdot}866t \quad \ldots\ldots(19)$$

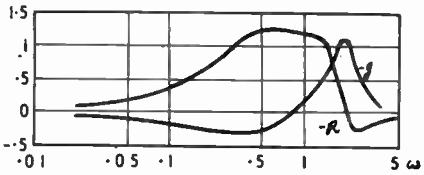

Fig. 7. Graphs of $-\mathscr{J}[H(j\Omega+j\omega)-H(j\Omega-j\omega)]$ and $-\mathscr{R}[H(j\Omega+j\omega)-H(j\Omega-j\omega)]$ for a particular system.

A graph of this function is shown in Fig. 8 for comparison with the points obtained by the approximate method.

In this example $H(p)$ represents a low-pass characteristic, highly unsymmetrical about Ω, so that eqn. (17) cannot be used.

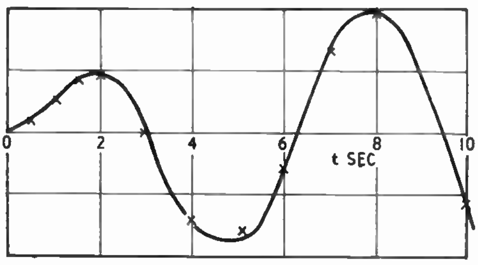

Fig. 8. Transient response of the system of Fig. 7. The curve is based on an analytical solution and the crosses show points obtained from Fig. 7 by means of the cursors.

4. Frequency Response from Transient Response

An alternative formula to eqn. (6) is (Section 8):

$$H(j\omega)=f(0)+ + \frac{\pi}{200}\mathscr{C}\left\{t\frac{d}{dt}f(t)\right\} - j\frac{\pi}{200}\mathscr{S}\left\{t\frac{d}{dt}f(t)\right\} \quad \ldots\ldots\ldots(20)$$

In this formula, if $f(t)$ is discontinuous at the origin, $f(0)$ is to be evaluated just after $t=0$ and $t\frac{d}{dt}f(t)$ is to be put equal to zero when $t=0$.

The quantities on the right-hand side of equation (20) are not multiplied by ω, and the magnitude of the error will therefore be roughly the same for all frequencies. On the other hand the differentiation cannot be done graphically with any high degree of accuracy. The formula may nevertheless be useful in some circumstances. For example:

(i) The response to a unit impulse, for $t>0$ is $h(t)=d/dt\, f(t)$. If $h(t)$ is available, eqn. (20) can easily be evaluated.

(ii) If $f(t)$ is obtained by calculation rather than experiment, the differentiation can be carried out numerically with good accuracy. This situation is illustrated by the example.

4.1. *Example*

The behaviour of a distillation column is governed[8] by a set of equations of the form

$$\frac{dx_n}{dt} = V_{n-1}\,y_{n-1} - V_n\,y_n + L_{n+1}\,x_{n+1} - L_n x_n \quad \ldots\ldots\ldots(21)$$

where x_n defines the liquid composition on the nth plate, and y_n is a non-linear function of the x. The simplest way to solve these equations in the general case is[9] by a step-by-step integration using an electronic digital computer. Then in order to investigate the automatic control of such a column it is desirable to have the frequency response to certain types of small disturbance. The electronic computer can most easily give the response to small step-disturbances. The response of the column to these, if they are small enough, will be linear, and the corresponding frequency responses can therefore be obtained from the computed step-responses by eqn. (20).

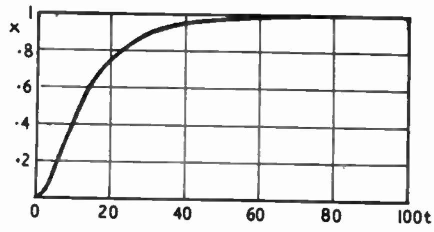

Fig. 9. Transient response of the top plate of a 5-plate enriching section to a step-change of feed composition. The response was calculated by means of an electronic computer.

Figure 9 shows the computed response of the composition on the top plate of a 5-plate enriching section to a step-change of vapour feed composition. The differentiation in eqn.

(20) was carried out by differencing the compositions at two successive times and so obtaining the derivative at the average of these times. The resulting graph of $t\,dx/dt$ against log t is shown in Fig. 10.

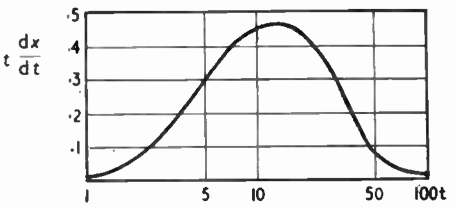

Fig. 10. Graph of $t\,dx/dt$ obtained from the computed response shown in Fig. 9.

The relation between y and x in this problem was artificially linearized, making possible an analytical solution[10] for the frequency response:

$$H(j\omega)=\frac{1}{(1-35\omega^2+9\omega^4)+j\omega(15-28\omega^2+\omega^4)}$$

This is shown in Fig. 11, together with a number of points obtained from Fig. 10 by means of the cursors.

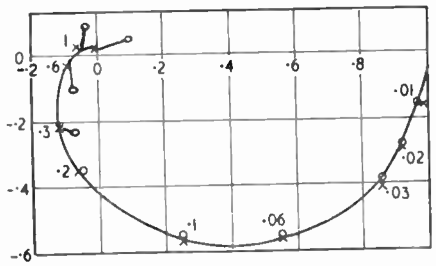

Fig. 11. Frequency response corresponding to the transient response of Fig. 9.

The line is drawn through points obtained from an analytical solution. The crosses show points obtained from Fig. 10 by means of the cursors according to eqn. (20). Points obtained by eqn. (6) are shown by circles.

Figure 11 also shows points obtained by use of the cursors according to eqn. (6). The error in these is less at the lower frequencies, but becomes excessive for ω greater than about 0·2. With the scales used for the graph of $t[f(t)-f(\infty)]$ an error of one unit in the results from the cursors gives in fact an error of $0{\cdot}08\omega$ in $\mathcal{I}\,H(j\omega)$ or $\mathcal{R}\,H(j\omega)$. For the method of eqn. (20) on the other hand a similar error of one unit gives an error at all frequencies of 0·016 in $\mathcal{I}H(j\omega)$ or $\mathcal{R}\,H(j\omega)$.

5. Conclusions

By the methods given in Sections 2, 3 and 4, the difficulties described in Section 1 may be avoided. The errors in the approximate results depend upon the problem and upon the care taken with the graphical work. They can usually be kept without difficulty below about 2 per cent. of the greatest value among the results.

The examples given have been chosen so that an analytical solution is possible. The benefits of the approximate method become more evident as the complexity of the problem increases, since the amount of work required increases much less rapidly than for an analytical solution.

6. Acknowledgments

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7. References

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8. Appendix: Derivation of formulae

It is assumed throughout that $H(p)$ has no poles on or to the right of the imaginary axis. Then if

$$H(j\omega)=F(j\omega)+H(j\infty), \qquad \ldots\ldots\ldots(22)$$

$f(t)$ can be expressed as the sum of two parts:

(i) A part $f_1(t)$ corresponding to $H(j\infty)$ and given by

$$f_1(t)=H(j\infty)\,U(t) \qquad \ldots\ldots\ldots(23)$$

where $U(t)$ is a unit step-funtcion at $t=0$.

(ii) A part $f_2(t)$ corresponding to $F(j\omega)$ and given by

$$f_2(t)=\frac{1}{2\pi}\int_{-\infty}^{\infty}\frac{F(j\omega)}{j\omega}\,e^{j\omega t}\,d\omega+\tfrac{1}{2}F(0) \qquad \ldots\ldots\ldots(24)$$

Hence if $t=\tau+\theta$

$$f_2(\tau+\theta)=\frac{1}{2\pi}\int_{-\infty}^{\infty}\frac{F(j\omega)}{j\omega}\,e^{j\omega(\tau+\theta)}d\omega+\tfrac{1}{2}F(0) \qquad \ldots\ldots\ldots(25)$$

$$=\frac{1}{\pi}\int_{0}^{\infty}\mathcal{R}\left[\frac{F(j\omega)}{j\omega}\,e^{j\omega\tau}\,e^{j\omega\theta}\right]d\omega+\tfrac{1}{2}F(0) \qquad \ldots\ldots\ldots(26)$$

$$=\frac{1}{\pi}\int_{0}^{\infty}\left\{\mathcal{J}\left[F(j\omega)\,e^{j\omega\tau}\right]\frac{\cos\omega\theta}{\omega}+\mathcal{R}\left[F(j\omega)\,e^{j\omega\tau}\right]\frac{\sin\omega\theta}{\omega}\right\}d\omega+\tfrac{1}{2}F(0) \qquad \ldots\ldots\ldots(27)$$

$$=\frac{1}{\pi}\int_{0}^{\infty}\left\{\mathcal{J}\left[F(j\omega)\,e^{j\omega\tau}\right]\frac{\cos\omega|\theta|}{\omega}\mp\mathcal{R}\left[F(j\omega)\,e^{j\omega\tau}\right]\frac{\sin\omega|\theta|}{\omega}\right\}d\omega+\tfrac{1}{2}F(0) \qquad \ldots\ldots\ldots(28)$$

where the negative sign must be taken for the second term if $\theta<0$ and the positive sign if $\theta>0$. Then since

$f_2(\tau+\theta)=0$, $\theta\leqslant-\tau$, and $F(0)=H(0)-H(j\infty)$,

$$\varphi(\theta)=f(\tau+\theta)=0, \qquad \theta<-\tau$$

$$=H(j\infty)+\tfrac{1}{2}\left\{\frac{2}{\pi}\int_{0}^{\infty}\mathcal{J}\left[\Big(H(j\omega)-H(j\infty)\Big)e^{j\omega\tau}\right]\frac{\cos\omega|\theta|}{\omega}\,d\omega-\frac{2}{\pi}\int_{0}^{\infty}\mathcal{R}\left[\Big(H(j\omega)-H(j\infty)\Big)e^{j\omega\tau}\right]\frac{\sin\omega|\theta|}{\omega}\,d\omega+H(0)-H(j\infty)\right\}, \qquad -\tau\leqslant\theta\leqslant0$$

$$=H(j\infty)+\tfrac{1}{2}\left\{\frac{2}{\pi}\int_{0}^{\infty}\mathcal{J}\left[\Big(H(j\omega)-H(j\infty)\Big)e^{j\omega\tau}\right]\frac{\cos\omega|\theta|}{\omega}\,d\omega+\frac{2}{\pi}\int_{0}^{\infty}\mathcal{R}\left[\Big(H(j\omega)-H(j\infty)\Big)e^{j\omega\tau}\right]\frac{\sin\omega|\theta|}{\omega}\,d\omega+H(0)-H(j\infty)\right\}, \qquad 0\leqslant\theta\leqslant\tau$$

$$=H(j\infty)+\frac{2}{\pi}\int_{0}^{\infty}\mathcal{J}\left[\Big(H(j\omega)-H(j\infty)\Big)e^{j\omega\tau}\right]\frac{\cos\omega|\theta|}{\omega}\,d\omega+H(0)-H(j\infty), \qquad \tau\leqslant\theta, \text{ or}$$

$$=H(j\infty)+\frac{2}{\pi}\int_{0}^{\infty}\mathcal{R}\left[\Big(H(j\omega)-H(j\infty)\Big)e^{j\omega\tau}\right]\frac{\sin\omega|\theta|}{\omega}\,d\omega, \qquad \tau\leqslant\theta \qquad \ldots\ldots\ldots(29)$$

With a change of notation (t for θ and $f(t)$ for $\varphi(\theta)$) eqns. (29) are equivalent to eqns. (7), (8) and (9).

If instead of a unit step-function the applied function is $Ae^{j\Omega t}\,U(t)$ then

$$K(p)=\frac{AH(p)}{p-\mathrm{j}\Omega}=\int_0^\infty \mathrm{f}(t)\,\mathrm{e}^{-pt}\,\mathrm{d}t \qquad \text{.........(30)}$$

Hence

$$K(p+\mathrm{j}\Omega)=\int_0^\infty \mathrm{f}(t)\,\mathrm{e}^{-(p+\mathrm{j}\Omega)t}\,\mathrm{d}t \qquad \text{.........(31)}$$

$$=\int_0^\infty \left[\mathrm{f}(t)\,\mathrm{e}^{-\mathrm{j}\Omega t}\right]\mathrm{e}^{-pt}\,\mathrm{d}t \qquad \text{.........(32)}$$

This can be inverted to give

$$\mathrm{f}(t)\,\mathrm{e}^{-\mathrm{j}\Omega t}=\frac{1}{2\pi\mathrm{j}}\int_{-\mathrm{j}\infty}^{\mathrm{j}\infty} K(p+\mathrm{j}\Omega)\,\mathrm{e}^{pt}\,\mathrm{d}p \qquad \text{.........(33)}$$

Let $K(p+\mathrm{j}\Omega)$ have a simple pole at $p=0$, with residue a. Then

$$\mathrm{f}(t)\,\mathrm{e}^{-\mathrm{j}\Omega t}=\frac{1}{2\pi}\int_{-\infty}^{\infty} K[\mathrm{j}(\Omega+\omega)]\,\mathrm{e}^{\mathrm{j}\omega t}\,\mathrm{d}\omega+\tfrac{1}{2}a \qquad \text{......(34)}$$

$$=\frac{1}{2\pi}\int_0^\infty \left\{K_+\mathrm{e}^{\mathrm{j}\omega t}+K_-\mathrm{e}^{-\mathrm{j}\omega t}\right\}\mathrm{d}\omega+\tfrac{1}{2}a \quad (35)$$

where

$$K_+=K[\mathrm{j}(\Omega+\omega)] \quad \text{and} \quad K_-=K[\mathrm{j}(\Omega-\omega)]$$

Then

$$\mathrm{f}(t)\,\mathrm{e}^{-\mathrm{j}\Omega t}=$$

$$\frac{1}{2\pi}\int_0^\infty \left\{(K_++K_-)\cos\omega t+\mathrm{j}\,(K_+-K_-)\sin\omega t\right\}\mathrm{d}\omega,$$

$$+\tfrac{1}{2}a \qquad \text{.........(36)}$$

Now $\mathrm{f}(t)=0$ when $t<0$ so from the odd and even symmetry of $\sin\omega t$ and $\cos\omega t$ respectively,

$$\frac{1}{2\pi}\int_0^\infty (K_++K_-)\cos\omega t\,\mathrm{d}\omega+\tfrac{1}{2}a$$

$$=\frac{1}{2\pi}\int_0^\infty (K_+-K_-)\sin\omega t\,\mathrm{d}\omega,\ t>0$$

$$\text{.........(37)}$$

Hence

$$\mathrm{f}(t)\,\mathrm{e}^{-\mathrm{j}\Omega t}=\frac{1}{\pi}\int_0^\infty (K_++K_-)\cos\omega t\,\mathrm{d}\omega+a,$$

$$\text{.........(38)}$$

or

$$=\mathrm{j}\frac{1}{\pi}\int_0^\infty (K_+-K_-)\sin\omega t\,\mathrm{d}\omega \quad \text{......(39)}$$

Now

$$K_+=\frac{AH[\mathrm{j}(\Omega+\omega)]}{\mathrm{j}\omega},\ K_-=\frac{AH[\mathrm{j}\Omega-\omega)]}{-\mathrm{j}\omega}$$

and

$$a=AH(\mathrm{j}\Omega)$$

Thus

$$\mathrm{f}(t)=A\,\mathrm{e}^{\mathrm{j}\Omega t}$$

$$\times\left[H(\mathrm{j}\Omega)+\frac{1}{\pi}\int_0^\infty \left\{H(\mathrm{j}\Omega+\mathrm{j}\omega)-H(\mathrm{j}\Omega-\mathrm{j}\omega)\right\}\frac{\cos\omega t}{\mathrm{j}\omega}\mathrm{d}\omega\right]$$

$$\text{.........(40)}$$

or

$$=A\,\mathrm{e}^{\mathrm{j}\Omega t}\times$$

$$\times\left[\mathrm{j}\frac{1}{\pi}\int_0^\infty \left\{H(\mathrm{j}\Omega+\mathrm{j}\omega)+H(\mathrm{j}\Omega-\mathrm{j}\omega)\right\}\frac{\sin\omega t}{\mathrm{j}\omega}\mathrm{d}\omega\right]$$

$$\text{.........(41)}$$

Equation (16) follows at once from eqn. (40).

With the notation of eqns. (22), (23) and (24):

$$F(p)=\int_0^\infty \left[\frac{\mathrm{d}}{\mathrm{d}t}\,\mathrm{f}_2(t)\right]\mathrm{e}^{-pt}\,\mathrm{d}t \quad \text{....(42)}$$

and since $\mathrm{d}/\mathrm{d}t\,\mathrm{f}_2(t)$ will tend to zero as t tends to infinity in all cases of interest,

$$F(\mathrm{j}\omega)=\int_0^\infty \left[\frac{\mathrm{d}}{\mathrm{d}t}\,\mathrm{f}_2(t)\right]\mathrm{e}^{-\mathrm{j}\omega t}\,\mathrm{d}t \quad \text{.........(43)}$$

Thus

$$H(\mathrm{j}\omega)=H(\mathrm{j}\infty)+$$

$$+\int_0^\infty \left[t\,\frac{\mathrm{d}}{\mathrm{d}t}\,\mathrm{f}_2(t)\right]\frac{\cos\omega t-\mathrm{j}\sin\omega t}{t}\,\mathrm{d}t \quad \text{......(44)}$$

and since

$$\lim_{t\to 0}\ \mathrm{f}(t)=\mathrm{f}_1(+0)=H(\mathrm{j}\infty)$$

and

$$\frac{\mathrm{d}}{\mathrm{d}t}\,\mathrm{f}_1(t)=0,\ t>0,$$

eqn. (20) follows at once.

DIPLOMAS IN TECHNOLOGY

THE National Council for Technological Awards has just published its first Report, for the period December 1955 to July 1957.

The Council was set up by the Minister of Education, on the recommendation of the National Advisory Council on Education for Industry and Commerce. Its purpose is to create and administer technological awards to students in technical colleges. The permanent secretariat is now established at 9 Cavendish Square, London, W.1.

The Council has already issued a Memorandum on the recognition of courses in technical colleges leading to the Diploma in Technology. On the question of eligibility to take approved courses, the present report states that there is advantage in flexibility in the standard of admission. In general, the requirement is either five subjects in the General Certificate of Education of which at least two appropriate to the course to be followed must be at Advanced level, or a good Ordinary National Certificate. The Council has not attempted to establish a standard definition of a "good" Ordinary National Certificate, but has considered individually the proposals put forward by each college. The report also emphasizes the importance of the sandwich course and integrating practical training in industry with academic study.

So far the Council has considered 104 applications for recognition of courses covering Chemistry, Biology, Physics, Building, Metallurgy, Instrument Technology and most branches of Engineering. Fifty courses have been approved, one of which—at the Northern Polytechnic—is in "Physics and Technology of Electronics." Otherwise, recognition has not so far been given to other courses specifically in radio, communication or electronic engineering. Eight courses in Electrical Engineering have approved and some of these permit students to take electronics and communications engineering—but only as optional subjects.

It must, however, be emphasized that the National Council for Technological Awards does *not* initiate courses. The planning of courses and application to the Council for recognition is the responsibility of individual colleges. A high standard is expected of the academic content of approved courses, and the report stresses that it is the function of colleges to prepare syllabuses in collaboration with the industries which the courses will serve.

The Chairman of the Council, Lord Hives, has stated that whilst a number of large industrial organizations support courses leading to the Diploma in Technology, there are many firms who have yet to make their contribution. Lord Hives urged industry to examine future developments in terms of new processes and new materials and the corresponding demand for more technologists. It is to be hoped, therefore, that more positive action may be taken in securing recognition by the National Council of courses in the communications and electronics engineering field.

In urging more radio manufacturers to use their influence with local colleges, an editorial in "Wireless World" has stated* "The growing importance of electronics in the national economy is sufficient justification for the strongest possible representation on the Boards of Studies appointed by the Council."

Indeed, the collaboration of all industry is essential if the Council is to obtain its objective of securing, within a few years, an annual output of 9,000 technologists trained by technical colleges, against the present output of some 5,000.† Not the least of the Council's proposals is the recommendation that teachers should not be given an excessive "teaching load," and that senior members of industrial organizations take an active part in the activities of local colleges.

So far nearly 1,000 students are now studying courses leading to the Dip.Tech. and in order to increase this intake, the report of the Council makes reference to the need for a substantial amount of new building required by existing and new technical colleges. The Council also considers it desirable for colleges to have residential facilities. Whilst the absence of such facilities has not prevented courses being recognized up to the present, it is intended to make their provision a condition, possibly by the time that the renewal of the recognition of courses is considered in five years' time.

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* "Training technologists," *Wireless World*, **64**, page 1, January 1958.

† This is in addition to the present output of some 5,000 scientists a year of whom about 1,000 come from the technical colleges. The Universities also supply about 2,000 technologists a year.

REPORT OF THE ANNUAL GENERAL MEETING OF SUBSCRIBERS TO THE Brit.I.R.E. BENEVOLENT FUND

The Meeting was held on 27th November, 1957, and the Chair was taken by Rear-Admiral Sir Philip Clarke, K.B.E., C.B., D.S.O.

1. To confirm the Minutes of the Annual General Meeting of subscribers held on 31st October, 1956

The Minutes of the Annual General Meeting dated 31st October, 1956, had been published in the October 1956 *Journal*. The Chairman's proposal that those Minutes be taken as a correct record was approved unanimously.

2. To receive the Annual Report of the Trustees

Admiral Clarke stated that the Annual Report had been published in the October 1956 *Journal*. The President, Mr. Marriott, who regretted missing the meeting of the subscribers, had recently attended the 144th Annual Festival of Reed's School and was convinced that the association with Reed's School was of mutual advantage. The Trustees were grateful to the School for the way in which they had helped in looking after the children of incapacitated or deceased members. Although the Fund contributed towards the fees of the school, this was only a very small proportion of the total cost of the child attending the school.

The disappointing part of the Benevolent Report was that 25 fewer members had supported the fund in the last year; only about one in six of the members of the Institution gave any subscription. The Trustees made a special appeal to every member to contribute something, however small, to the Fund. Admiral Clarke especially expressed appreciation of the substantial contributions made by the various Radio Industry Clubs and Electric & Musical Industries Ltd.

The Report was adopted unanimously.

3. To receive the Income and Expenditure Account and the Balance Sheet for the year ended 31st March, 1957

Admiral Clarke called on the Honorary Secretary of the Fund, Mr. G. D. Clifford, to present the Accounts.

Mr. Clifford first referred to the existence of several Bursaries which had been purchased. Not only were these endowments of immediate help to the School, but they enabled the Trustees to give valuable help to a boy who would benefit from a Grammar School education.

He then went on to point out that the effect of 16 years' work by successive Trustees could now be seen in the Balance Sheet. Every year the Trustees had endeavoured to have a surplus for investment and, as a result, the income from this source was now nearly half that received in contributions.

Investment income had not been built up at the expense of those needing assistance, and not only had the Trustees been able to help in the education of the children of deceased members, but had also distributed a considerable sum to members in distress, or to dependants.

The Accounts were unanimously approved.

4. To elect Trustees for the year 1957-58

The Secretary, Mr. G. D. Clifford, stated that the retiring Trustees, Mr. G. A. Marriott, Rear-Admiral Sir Philip Clarke, Mr. G. A. Taylor and Mr. A. H. Whiteley had offered themselves for re-election, and had recommended that Mr. A. A. Dyson be elected to fill the remaining vacancy caused by the death of Mr. E. J. Emery.

The nominations were approved unanimously.

5. To appoint the Honorary Solicitors, and
6. To appoint the Honorary Accountant

Admiral Clarke paid tribute to the work carried out in honorary capacities by Messrs. Braund and Hill and Mr. R. H. Jenkins, and he moved their re-appointment as Honorary Solicitors and Honorary Accountant respectively.

These proposals were carried unanimously.

7. Any other business

The Secretary confirmed that he had not received notice of any other business. The Chairman again thanked all subscribers for their support and closed the meeting.

. . . Radio Engineering Overseas

A selection of abstracts from European and Commonwealth journals received in the Library of the Institution. Members may borrow these journals under the usual conditions. All papers are in the language of the country of origin of the journal unless otherwise stated. The Institution regrets that translations cannot be supplied.

621.3.029.66

Dielectric walls with a small reflection coefficient at microwaves. H. MEINKE. *Nachrichtentechnische Zeitschrift*, **10**, pp. 551-558, November 1957.

A summary is given relating to the various methods for manufacturing walls with a small reflection coefficient over a narrow band as well as a wide band. The properties of suitable synthetic dielectric, particularly in a mixture with graphite dust and iron dust, are given.

621.317.32

Noise voltage measurements on low impedance circuit elements with the aid of a valve voltmeter with a preceding transformer. W. NONNENMACHER. *Nachrichtentechnische Zeitschrift*, **10**, pp. 559-563, November 1957.

It can be shown that equivalent noise impedance of less than 1 ohm can be obtained in the low and medium frequency ranges on a valve voltmeter when an input transformer is employed. In this way it is possible to measure the thermal noise in circuit elements with a very small impedance. Additional noise from the transformer can be avoided.

621.317.332.6:621.397

The measurement of the reflection coefficient in television transmission lines and equipment. E. THINIUS. *Nachrichtentechnische Zeitschrift*, **10**, pp. 548-550, November 1957.

A method is described which is suitable for measuring the magnitude and the phase of the reflection coefficient of equipment on a cable or of mismatch points in cables at operational frequencies. The whole frequency band is recorded by an oscilloscope so that points of major reflections can be detected immediately.

621.317.755

Low-voltage oscilloscope tubes. F. DE BOER and W. F. NIENHUIS. *Philips' Technical Review*, **19**, pp. 158-164, No. 5, November 1957.

An account is given of the development of two oscilloscope tubes requiring an anode voltage of only 400 V. One tube has been designed for symmetrical, the other for asymmetrical application of the sweep voltage. The fluorescent layer is applied on top of a conductive tin oxide coating (to prevent the screen charging up) and is not covered by a binder layer, as such a layer would decelerate the beams electrons excessively. The conductive coating makes it possible to touch the tube while measurements are being made; furthermore, the cathode can be earthed, so that the heater can be fed from a normal heater winding in the transformer. The electrode system has been designed in such a way as to make the adjustment of brightness and focus independent of each other. The design of a small light-weight oscilloscope is discussed.

621.373.4.011

The generalization of the Barkhausen formula for electronic oscillators. M. DRAGANESCU. *Automatica si electronica, Bucharest*, **1**, No. 3, May-June 1957, pp. 112-115.

The Barkhausen formula is generalized with a view to obtaining the condition for the setting up of the oscillations, the amplitude of the oscillations, the dynamic stability of the oscillations, and the oscillation frequency. This is possible by taking into account the non-linearities of the oscillating system and using the operational calculus, linearized elements being substituted for non-linear ones. By way of application a classical oscillator circuit is studied taking grid currents into consideration. Original results are obtained.

621.373.52

Transistor oscillators and their sensitivity to load impedance. W. HERZOG. *Nachrichtentechnische Zeitschrift*, **10**, pp. 564-569, November 1957.

The second four-terminal network equation for transistors is used for the determination of the "internal" feedback in an oscillator circuit and the equivalence of this feedback to the normal external feedback is shown. Two types of transistor oscillators are given which are insensitive to load impedances and conditions are given under which two load impedances can be connected simultaneously and without interfering with each other or with the circuit.

621.373.52

Experimental and theoretical investigation into frequency stabilized transistors for 8 Mc/s. H. SCHAFFHAUSER and M. J. O. STRUTT. *Archiv der Elektrischen Übertragung*, **11**, November 1957, pp. 455-460.

The amplifier, its load and the feedback network are treated as fourpoles, which are cascaded to a single four-pole for which the condition of oscillation obtains. A definite circuit is then considered, yielding two equations for the transistor oscillator. Taking into account the variations of transistor parameters, caused by temperature variations, the oscillator circuit, containing a quartz element, may be calculated. Frequency variation and output voltage as influenced by variations of temperature, load and supply tension have been measured. The input parameters of the transistor are discussed as dependent on temperature and on emitter current. Attempts were made to adjust the series resistance of the emitter so that the input parameters remain practically constant upon temperature variations. No general rule could be found for this adjustment.

621.375.2.029.6:621.385.3.029.6

A 4,000 Mc/s wide-band amplifier using a disc-seal triode. J. P. M. GIELES. *Philips' Technical Review*, **19**, pp. 145-156, No. 5, November 1957.

Description of an amplifier for frequencies from 3,800 to 4,200 Mc/s of 8 db, the amplifier can deliver a power of 0·5 W and 1·5 W respectively. The input and output can be connected to waveguides. For connecting several of these amplifiers in cascade, coupling is preferably effected by means of ferrite isolators. The variation of group delay in the frequency band covered is much smaller than with the i.f. amplifiers normally used in beam transmitters.

621.375.4

An analysis of transient response of junction transistor amplifiers. J. C. BHATTACHARYYA. *Journal of the Institution of Telecommunication Engineers, New Delhi*, **3**, pp. 297-303, September 1957.

An exact solution of the one-dimensional diffusion equation as is applicable to a junction type of transistor has been obtained by the method of Laplace's transform. The solution has been utilized to derive an expression for the short-circuited output collector current with a step input forcing function. The time-independent part of this expression is found to be identical with the relation for steady state collector current as given earlier by Shockley. Experimental results on the transient response of a OC70 junction transistor have been shown to agree closely with the response obtained theoretically. It is concluded that under ordinary conditions of operation the major physical process underlying transistor action must be the diffusion of minority carrier across the base region.

621.396.677.53

Polarization d/f system. J. GROBKOPF and K. VOGT. *Nachrichtentechnische Zeitschrift*, **10**, pp. 572-579, November 1957.

Theoretical and experimental investigations concerning a new direction-finding system for short waves free from night error are presented. The advantage of the new system, when compared with the conventional Adcock system of comparable qualities, is given by its smaller dimensions.

621.396.677.75

Some investigations of dielectric aerials. (Mrs.) R. CHATTERJEE, and S. K. CHATTERJEE. *Journal of the Institution of Telecommunication Engineers, New Delhi*, **3**, pp. 280-284, September 1957.

A comparative study of the expressions for radiation pattern of a circular dielectric rod aerial excited in the HE_{11} mode as obtained by the Schelkunoff's equivalence principle and by the application of Huyghen's principle over the whole rod shows that in the $\varphi = 0^\circ$ and $\varphi = 90^\circ$ planes, the beamwidth of the major lobe and the structure of the radiation pattern differ in the former case but are the same in the latter case. It is also shown that Halliday and Kiely's theory, also based on Huyghen's ray theory, does not indicate any variation in these factors in the two planes. This is justified due to the vector nature of the Schelkunoff's principle and the scalar nature of the original Huyghen's principle.

621.396.677.8:621.396.11

Microwave antenna characteristics in the presence of of an intervening ridge. R. VIKRAMSINGH, M. N. RAO and S. UDA. *Journal of the Institution of Telecommunication Engineers, New Delhi*, **3**, pp. 274-279, September 1957.

Using two transmitters and two receivers operating in the 2,000 Mc/s region, the following experiments were carried out with an intervening ridge between terminals 14 km apart: (i) using horizontal as well as vertical polarization, beam shapes were determined in the azimuth plane and vertical plane; (ii) dependence of signal level on the angles of elevation of the axes of the parabolic mirrors was also studied; (iii) polarization patterns were recorded by rotating one dipole and keeping the other fixed; (iv) the effect of defocusing of the dipole with respect to the parabola on the received signal intensity was studied.

621.396.826

Observations on back scatter echoes for short-wave transmission. HANS-ULRICH WIDDEL. *Archiv der Elektrischen Übertragung*, **11**, November 1957, pp. 429-439.

With back scatter observations by means of pulsed shortwave broadcasting transmitters, emphasis is placed in particular on an investigation of the influence of aerial radiation patterns on the observability and interpretability of the echoes. The maximum observable back scatter range depends on the vertical radiation pattern of the aerial system. Insufficient horizontal directivity and poor front-to-back ratio makes the echo pattern complex and its interpretation more difficult, particularly when the perimeter of the skip zone fails to be concentric with respect to the transmitter site. By means of a simple mathematical conversion method the propagation paths and the origin of most echo signals could be determined. The vast majority comes from the ground and undergoes intermediate reflection in the F2 layer in both go and return paths. Scatter processes in the ionosphere itself, reflections at geographic objects and polar aurora structures, were also detected.

621.397.335

Assessment of the use of horizontal synchronization for television receivers. HELMUT LUTZ. *Archiv der Elektrischen Übertragung*, **11**, November 1957, pp. 461-470.

The paper derives the parameters that are important for judging the horizontal synchronization of television sets, in particular with the use of regulating circuits. The noise characteristics and the response to a sine or pulse disturbance are calculated under the assumption of linear relationships in the regulator. An arrangement is devised for measuring these characteristics. For judging the pulling-in and holding properties the non-linear regulator characteristic must be known completely. A measuring arrangement is described for plotting the unstable portion of the regulating characteristic, and a geometrical construction method for determining the phenomena in the pulling-in process.

621.397.62

A luminous frame around the television screen. J. J. BALDER. *Philips' Technical Review*, **19**, pp. 156-158, No. 5, November 1957.

Observers viewing a picture screen were asked to adjust the brightness and width of a uniformly illuminated frame around the picture so as to give most agreeable viewing conditions. This experiment was made with 20-25 observers each under conditions of a number of luminance values of screen and surroundings. The preferred frame width was found to be independent of screen and ambient luminance and averaged about 0·3 times the half screen dimensions for all observers. The preferred frame luminance was found to increase with the screen luminance as well as with the luminance of the surroundings. It appears that the frame luminance should be adjustable between 0 and 40 cd/m².