l

BBC

ENGINEERING DIVISION

MONOGRAPH

NUMBER 76: DECEMBER 1968

The Variable Inductance Frequency Modulator

by

J. E. PACKMAN, B.Sc., C.Eng., M.I.E.E.
(Designs Department, BBC Engineering Division)

BRITISH BROADCASTING CORPORATION

PRICE FIVE SHILLINGS

BBC ENGINEERING MONOGRAPH

No. 76

THE VARIABLE INDUCTANCE FREQUENCY MODULATOR

by

J. E. Packman, B.Sc., C.Eng., M.I.E.E.

(Designs Department, BBC Engineering Division)

DECEMBER 1968

BRITISH BROADCASTING CORPORATION

FOREWORD

THIS is one of a series of Engineering Monographs published by the British Broadcasting Corporation. About six are produced every year, each dealing with a technical subject within the field of television and sound broadcasting. Each Monograph describes work that has been done by the Engineering Division of the BBC and includes, where appropriate, a survey of earlier work on the same subject. From time to time the series may include selected reprints of articles by BBC authors that have appeared in technical journals. Papers dealing with general engineering developments in broadcasting may also be included occasionally.

This series should be of interest and value to engineers engaged in the fields of broadcasting and of telecommunications generally.

Individual copies cost 5s. or 5s. 9d. by post, while the annual subscription is £1 post free. Orders can be placed with newsagents and booksellers, or BBC PUBLICATIONS, 35 MARYLEBONE HIGH STREET, LONDON, W.1.

CONTENTS

PREVIOUS ISSUES IN THIS SERIES

No.	Title	Date
1.	*The Suppressed Frame System of Telerecording*	JUNE 1955
2.	*Absolute Measurements in Magnetic Recording*	SEPTEMBER 1955
3.	*The Visibility of Noise in Television*	OCTOBER 1955
4.	*The Design of a Ribbon Type Pressure-gradient Microphone for Broadcast Transmission*	DECEMBER 1955
5.	*Reproducing Equipment for Fine-groove Records*	FEBRUARY 1956
6.	*A V.H.F./U.H.F. Field-strength Recording Receiver using Post-detector Selectivity*	APRIL 1956
7.	*The Design of a High Quality Commentator's Microphone Insensitive to Ambient Noise*	JUNE 1956
8.	*An Automatic Integrator for Determining the Mean Spherical Response of Loudspeakers and Microphones*	AUGUST 1956
9.	*The Application of Phase-coherent Detection and Correlation Methods to Room Acoustics*	NOVEMBER 1956
10.	*An Automatic System for Synchronizing Sound on Quarter-inch Magnetic Tape with Action on 35-mm Cinematograph Film*	JANUARY 1957
11.	*Engineering Training in the BBC*	MARCH 1957
12.	*An Improved 'Roving Eye'*	APRIL 1957
13.	*The BBC Riverside Television Studios: The Architectural Aspects*	JULY 1957
14.	*The BBC Riverside Television Studios: Some Aspects of Technical Planning and Equipment*	OCTOBER 1957
15.	*New Equipment and Methods for the Evaluation of the Performance of Lenses for Television*	DECEMBER 1957
16.	*Analysis and Measurement of Programme Levels*	MARCH 1958
17.	*The Design of a Linear Phase-shift Low-pass Filter*	APRIL 1958
18.	*The BBC Colour Television Tests: An Appraisal of Results*	MAY 1958
19.	*A U.H.F. Television Link for Outside Broadcasts*	JUNE 1958
20.	*The BBC's Mark II Mobile Studio and Control Room for the Sound Broadcasting Service*	AUGUST 1958
21.	*Two New BBC Transparencies for Testing Television Camera Channels (Out of Print)*	NOVEMBER 1958
22.	*The Engineering Facilities of the BBC Monitoring Service*	JANUARY 1959
23.	*The Crystal Palace Band I Television Transmitting Aerial*	FEBRUARY 1959
24.	*The Measurement of Random Noise in the presence of a Television Signal*	MARCH 1959
25.	*A Quality-checking Receiver for V.H.F. F.M. Sound Broadcasting*	JUNE 1959
26.	*Transistor Amplifiers for Sound Broadcasting*	AUGUST 1959
27.	*The Equipment of the BBC Television Film Studios at Ealing*	JANUARY 1960
28.	*Programme Switching, Control, and Monitoring in Sound Broadcasting*	FEBRUARY 1960
29.	*A Summary of the Present Position of Stereophonic Broadcasting*	APRIL 1960
30.	*Film Processing and After-processing Treatment of 16-mm Films*	MAY 1960
31.	*The Power Gain of Multi-tiered V.H.F. Transmitting Aerials*	JULY 1960
32.	*A New Survey of the BBC Experimental Colour Transmissions*	OCTOBER 1960
33.	*Sensitometric Control in Film Making*	DECEMBER 1960
34.	*A Mobile Laboratory for UHF and VHF Television Surveys*	FEBRUARY 1961
35.	*Tables of Horizontal Radiation Patterns of Dipoles Mounted on Cylinders*	FEBRUARY 1961
36.	*Some Aspects of Optical Lens Performance*	APRIL 1961
37.	*An Instrument for Measuring Television Signal-to-noise Ratio*	JUNE 1961
38.	*Operational Research on Microphone and Studio Techniques in Stereophony*	SEPTEMBER 1961
39.	*Twenty-five Years of BBC Television*	OCTOBER 1961
40.	*The Broadcasting of Music in Television*	FEBRUARY 1962
41.	*The Design of a Group of Plug-in Television Studio Amplifiers*	APRIL 1962
42.	*Apparatus for Television and Sound Relay Stations*	JULY 1962
43.	*Propagational Factors in Short-wave Broadcasting*	AUGUST 1962
44.	*A Band V Signal-frequency Unit and a Correlation Detector for a VHF/UHF Field-strength Recording Receiver*	OCTOBER 1962
45.	*Vertical Resolution and Line Broadening*	DECEMBER 1962
46.	*The Application of Transistors to Sound Broadcasting*	FEBRUARY 1963
47.	*Vertical Aperture Correction using Continuously Variable Ultrasonic Delay Lines*	MAY 1963
48.	*The Development of BBC Internal Telecommunications*	MAY 1963
49.	*Apparatus for Measurement of Non-linear Distortion as a Continuous Function of Frequency*	JULY 1963
50.	*New Methods of Lens Testing and Measurement*	SEPTEMBER 1963
51.	*Radiophonics in the BBC*	NOVEMBER 1963
52.	*Stereophony: the effect of cross-talk between left and right channels*	MARCH 1964
53.	*Aerial distribution systems for receiving stations in the l.f., m.f., and h.f. bands*	JULY 1964
54.	*An Analysis of Film Granularity in Television Reproduction*	AUGUST 1964
55.	*A Review of Television Standards Conversion*	DECEMBER 1964
56.	*Stereophony: the effect of interchannel differences in the phase/frequency and amplitude/frequency characteristics*	DECEMBER 1964
57.	*Drop-out in Video-tape Recording*	JUNE 1965
58.	*Sine-squared pulse and bar testing in colour television*	AUGUST 1965
59.	*The acoustic design and performance of a new free-field sound measurement room*	SEPTEMBER 1965
60.	*Colorimetric Analysis of Interference in Colour Television*	FEBRUARY 1966
61.	*Sporadic E Ionization and Television Interference*	MARCH 1966
62.	*Automatic monitoring*	APRIL 1966
63.	*The design of transmission lines and single-stage switching circuits for a line-store standards converter*	AUGUST 1966
64.	*Data for the acoustic design of studios*	NOVEMBER 1966
65.	*Tristimulus spot colorimeter*	DECEMBER 1966
66.	*VHF aerial gain calculation using tables of mutual resistance between the radiating elements*	FEBRUARY 1967
67.	*Pulse sound: A System of Television Sound Broadcasting Using Pulses in the Video Waveform*	MARCH 1967
68.	*Recent research on studio sound problems*	JULY 1967
69.	*A Survey of the Development of Television Test Cards Used in the BBC*	SEPTEMBER 1967
70.	*The dynamic characteristics of limiters for sound programme circuits*	OCTOBER 1967
71.	*The Programme Effects Generator*	NOVEMBER 1967
72.	*Colour Sensitometric Parameters in Colour Film Telerecording*	MARCH 1968
73.	*Sonic booms and other aircraft noise in studios*	APRIL 1968
74.	*The non-linear characteristics of klystron amplifiers*	JUNE 1968
75.	*Pulse-code modulation for high-quality sound-signal distribution*	DECEMBER 1968

THE VARIABLE INDUCTANCE FREQUENCY MODULATOR

SUMMARY

The monograph describes an electronic method of varying the effective value of an inductance and derives the theoretical value of this inductance. This method is exploited in an oscillator whose frequency is determined by a control voltage. An expression is obtained for the voltage required to cause a certain deviation in frequency, and in the course of this analysis it is shown that the centre frequency should be reasonably stable over a period of time. An evaluation is made of the distortion which may be introduced by this circuit when driven by a sinusoidal waveform, and of the conditions to be fulfilled for minimum distortion. It is further shown that it is possible to produce a pre-emphasized frequency response without the use of an external network.

A practical modulator design is described and the performance analysed theoretically. The results are compared with those obtained in practice. It is thereby shown that the apparatus would be of adequate performance to be used in broadcast transmitters for a pilot tone stereophonic transmission system.

SYMBOLS

A	ratio of deviation to carrier frequency
a	ratio of carrier currents in the total secondary to primary
C	tuning capacitance (F)
C_c	coupling capacitance (F)
e	instantaneous e.m.f. induced in the primary winding (volts)
$\hat{e}$	amplitude of carrier voltage at TR1 collector (volts)
F_D	peak deviation (Hz)
f_D	instantaneous deviation (Hz)
f_c	carrier frequency (Hz)
h	ratio $\frac{\hat{V}_2}{\hat{V}_3}$ for minimum distortion
I	amplitude of carrier current in primary winding (amps)
$\hat{I}_2$	amplitude of carrier current in TR2 collector (amps)
$\hat{I}_3$	amplitude of carrier current in TR3 collector (amps)
I_2	total current excluding carrier (d.c. +l.f.) in TR2 collector (amps)
I_3	total current excluding carrier (d.c. +l.f.) in TR3 collector (amps)
I_s	steady d.c. current in TR2 or TR3 (amps)
j	operator equal to $\sqrt{-1}$
K	quotient of transistor transconductance (mhos) divided by collector current (amps) $= 37 \times (\text{volts})^{-1}$
k	coupling coefficient of transformer
L	primary inductance of transformer (Henrys)
L_s	half secondary inductance of transformer (Henrys)
M	mutual inductance of total primary to half-secondary (Henrys)
m	a factor proportional to the unbalance of carrier current in the two halves of the secondary
n	turns ratio of half-secondary to total primary
P	product of kna
Q	product of $2I_sR_e$
R_e	value of the resistors in the emitter circuits of TR2 and TR3 (ohms)
T	pre-emphasis time constant (seconds)
t	time (seconds)
V	bias voltage for TR2 and TR3 (volts)
$\hat{V}_2$	amplitude of modulating voltage at TR2 base (volts)
$\hat{V}_3$	amplitude of modulating voltage at TR3 base (volts)
V_{eb}	emitter-base voltage of TR2 and TR3 (volts)
v_2	instantaneous value of $\hat{V}_2$ (volts)
v_3	instantaneous value of $\hat{V}_3$ (volts)
ω	angular velocity of the modulating voltage vector (rads/sec)
ω_c	angular velocity of the carrier voltage vector (rads/sec)
ω_D	angular velocity of the deviation (rads/sec)
Y_{21}	forward transfer admittance of TR2 and TR3 in common-base configuration (mhos)
Z_2	common base forward transfer impedance of TR2 (ohms)

1. Introduction

1.1 *General*

The performance specification of a frequency-modulated transmitter for broadcast use includes two requirements which are difficult to satisfy simultaneously. One is that its frequency can be varied linearly by the programme signal and the other is that the mean frequency shall remain constant over a period of time within a limit that is much smaller than the maximum deviation frequency.

1.2 *Possible Systems for Frequency Modulation*

There are at least three ways of overcoming this problem. One is to have a modulator which does not inherently have good frequency stability, compare its mean output frequency with a crystal oscillator, and adjust the modulator frequency by means of a closed loop servo system. A second way is to deviate a crystal oscillator in frequency. The third way is to have a modulator operating at a much lower frequency than that broadcast but producing the full deviation and mixing to the final frequency with a stable crystal oscillator. It is usually possible to achieve a better percentage stability at lower oscillator frequencies since stray capacitances etc. are a smaller proportion of the whole and hence the first two systems often use a low frequency modulator and reach the final frequency by multiplication.

The first system has the disadvantage of complexity and the possibility of 'drop-outs' if there is an interruption of the supply. It does, however, have the advantage that it can be locked to a multiple of line frequency in colour television transmitters and make intermodulation products have a frequency corresponding to minimum visibility on the picture.

The second system which deviates a crystal oscillator obviously does not take full advantage of the inherent stability of such an oscillator. Hence the frequency stability is degraded compared with that of a normal crystal oscillator but is still considerably better than with a non-crystal oscillator. Moreover, small deviation necessitates multiplication and this also multiplies the frequency error.

The third system will have an increasing stability the more the modulator centre frequency is lowered but a correspondingly worse linearity since the fractional deviation becomes higher. If the centre frequency is carefully chosen a compromise can be reached which gives adequate performance on both requirements. Care has to be taken in mixing to the final frequency to avoid spurious products, but otherwise the system introduces no serious complications. This system has been used in the VRFM modulator described in an earlier BBC Engineering Monograph.

2. Essential Design Considerations

This monograph deals with the design of a modulator using the third system, and its application to a high quality broadcast transmitter suitable for stereophonic or monophonic signals. Basic design considerations are those of centre frequency and allowable distortion.

2.1 *Choice of Centre Frequency*

It is possible to build an LC oscillator that has a centre frequency stability of 1 part in 10^4 over a period of several weeks. If 250 Hz were considered to be the limit of mean frequency excursion this would imply a centre frequency of 2·5 HMz. This would be well within the limits laid down in the BBC which, in turn, are well within the international limit of $\pm$3,750 Hz. The rated system deviation in use in the BBC is $\pm$75 kHz, which represents a 3 per cent deviation of 2·5 MHz. This is not high enough to cause undue distortion, as will be seen later.

It is not advisable to translate the frequency of the modulator to Band II (88–100 MHz) in one step as this would make it difficult to reduce the level of other mixer products in the output to a reasonable level and would involve designing a special filter for each output frequency. It has therefore been found convenient initially to raise the frequency to a standard intermediate frequency of 10·7 MHz and to use a single design of filter to remove spurious mixer products. This standard intermediate frequency could be higher if used in a television transmitter where 30–40 MHz is more common. The disadvantage of the higher frequency is offset by only requiring one design of filter. Alternatively, the modulator could have a 6 MHz centre frequency and this could be mixed with the vision I.F. oscillator output.

2.2 *Performance Requirements*

A frequency modulator should have sufficient frequency response to enable coded stereophonic signals to be passed without undue degradation. The pilot tone system adopted by the BBC transmits the monophonic (left plus right) information as a normal compatible signal and the stereophonic information (left minus right) in a double sideband amplitude modulation of a suppressed sub-carrier of 38 kHz. In addition, a low level pilot tone of 19 kHz is transmitted for receiver decoding purposes. Unequal transmission of sum and difference signals will cause crosstalk between left and right channels. The frequency response must therefore be very flat between 30 Hz and 53 kHz, assuming an audio range on each channel of 30 Hz to 15 kHz. Typical requirements are that the gain difference shall not exceed 0·5 dB and the phase difference 4°.

Non-linearity in the modulator naturally gives rise to harmonic distortion, but with a stereophonic system it also causes crosstalk between the decoded left and right hand signals. When the modulating frequency is high these components may not be harmonically related. As an example, assume that the modulating frequency is 12 kHz, thus producing normal multiplex components of 12, 26, and 50 kHz. Second order distortion will produce in-band components at 14, 24, 38, and 52 kHz, giving rise to a decoded signal at 14 kHz in both left and right output channels. Third order distortion will produce in-band components at 2, 36, 38, and 40 kHz, giving rise to a decoded signal at 2 kHz in both channels. As these 2 kHz components are derived from a high frequency signal with considerable pre-emphasis they will be at a relatively high level. Therefore, to achieve a reasonable performance the linearity requirements are stringent. A suitable specification limit for harmonic distortion of a single tone, fed directly into the modulator, would be 0·25 per cent (− 52 dB), maintained at all frequencies up to 53 kHz.

3. The Principle of Variable Inductance

The value of an inductance is determined by the magnitude of the back e.m.f. produced when a change of current is made. If this change is produced by an alternating current, and an e.m.f. of the same frequency and phase can be induced into the inductance from a mutually coupled circuit, there is an apparent change of inductance. If the amplitude of the signal in the secondary circuit can be controlled then such an inductance can be used in the tuned circuit of a frequency-modulated oscillator.

Consider the transformer in Fig. 1, in which currents are fed into the primary and each half of the secondary winding by three current transformers. The phase relationship of the currents in the two halves of the secondary is such that their magnetic fields are in opposition.

If L is the primary inductance
- M is the mutual inductance of total primary to half secondary
- a is the ratio of carrier currents in the total secondary to primary
- m is a factor proportional to the unbalance of carrier current in the two halves of the secondary, being zero when the currents are equal

$$e = j\omega_c LI \sin \omega_c t + j\omega_c \frac{MaI}{2}(1-m)\sin \omega_c t - j\omega_c \frac{MaI}{2}(1+m)\sin \omega_c t$$

$$= j\omega_c I \sin w_c t\left[L + \frac{Ma}{2}(1-m) - \frac{Ma}{2}(1+m)\right]$$

$$= j\omega_c I \sin \omega_c t[L - Mam]$$

In an a.c. circuit

Instantaneous Voltage $= j\omega L \times$ Instantaneous Current

$$\therefore \text{Apparent Primary Inductance} = \frac{e}{j\omega_c I \sin \omega_c t}$$

$$= L - Mam \qquad (1)$$

If L_S = Actual inductance of one half-secondary

and k = Coupling coefficient of primary to secondary

$$M = k\sqrt{LL_s}$$

If n = turns ratio of half-secondary to primary

$L_S = n^2 L$ assuming a high degree of inter-turn coupling

$$\therefore \quad M = k\sqrt{Ln^2L} = knL \qquad (2)$$

Substituting (2) in (1)

Apparent Primary Inductance $= L(1 - knam)$ (3)

4. Deviation Produced by an Inductance Change

An expression can be obtained for the deviation produced by an inductance change which, by expansion, will also give the amplitude of the distortion components.

Equation (3) shows that as k, n, and a are constants, the inductance varies linearly with m, the unbalance of the secondary currents. If the primary inductance is used as part of the tuned circuit in an LC oscillator, then the frequency will vary with the unbalance, but not according to a linear law.

Suppose that, if $m = 0$, i.e. there is no unbalance in the secondary currents, such an oscillator has a frequency given by:

$$\omega_c = \frac{1}{\sqrt{LC}} \text{ i.e. } f_c = \frac{1}{2\pi\sqrt{LC}} \qquad (4)$$

If $m \neq 0$, then the oscillator will deviate to a frequency $(f_c + f_D)$, where f_D is the amount of deviation

$$\text{and } 2\pi(f_c + f_D) = \frac{1}{\sqrt{L(1-knam)C}}$$

$$= \frac{1}{\sqrt{LC}\sqrt{1-knam}}$$

$$\text{but, from (4) above } \frac{1}{\sqrt{LC}} = 2\pi f_c,$$

$$\therefore 2\pi(f_c + f_D). = \frac{2\pi f_c}{\sqrt{1-knam}} = 2\pi f_c(1 - knam)^{-\frac{1}{2}}$$

$$\text{whence } f_D = f_c[(1 - knam)^{-\frac{1}{2}} - 1]$$

By binomial expansion this equals:

$$f_D = f_c\left[\frac{knam}{2} + \frac{3}{8}(knam)^2 + \frac{5}{16}(knam)^3 + \dots\right] \qquad (5)$$

The first term represents a linear variation of f_D with m; the remaining terms represent non-linear elements. Thus if m were made directly proportional to the amplitude of a modulating signal a certain degree of non-linearity would ensue. However, a simple method for reducing this non-linearity will be outlined later.

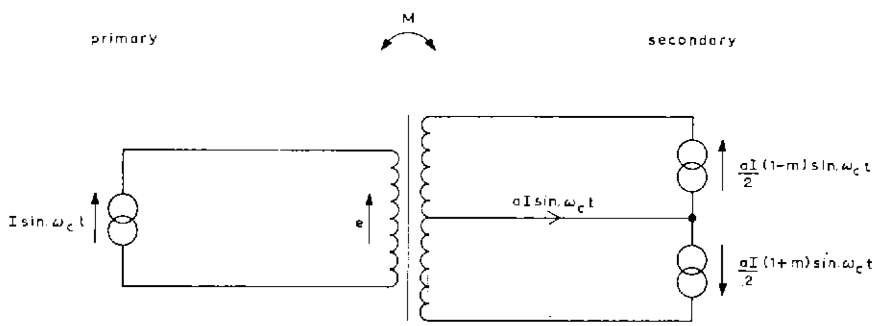

Fig. 1 — Transformer circuit illustrating the principle of variable inductance

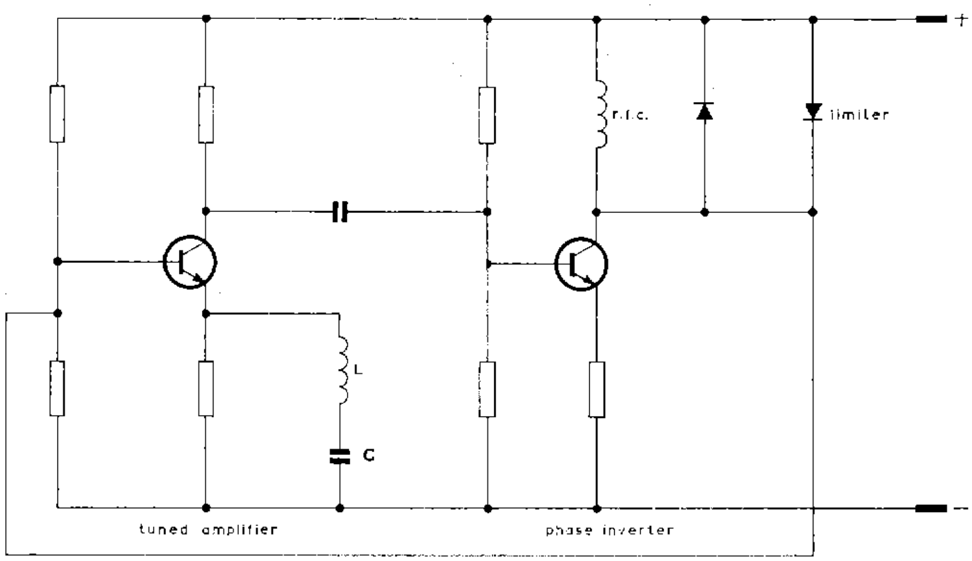

Fig. 2 — Basic oscillator circuit

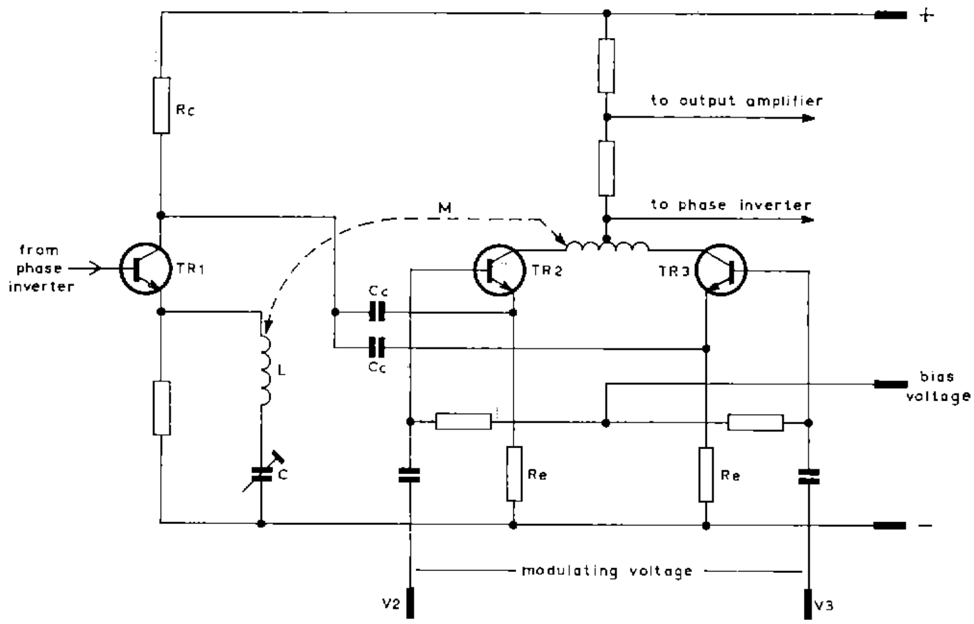

Fig. 3 — Oscillator coupled to variable inductance circuit

5. Application to a Practical Circuit

5.1 *Circuit Derivation*

The next step is to design an oscillator which will satisfy the condition of equation (4). A suitable basic oscillator is shown in Fig. 2.

At the resonant frequency of L and C, the left-hand transistor has maximum gain and a phase shift of 180°, from base to collector. The right-hand transistor always has a phase shift of 180° from base to collector. No other components produce phase shift provided that they are correctly chosen. Thus the circuit oscillates at the resonant frequency of L and C at an amplitude governed by the limiter diodes. To accommodate the transformer and secondary current generators the circuit can be modified as shown in Fig. 3, in which TR1 corresponds to the left-hand (tuned amplifier) transistor of Fig. 2, and the phase inverter transistor is not shown.

If the impedance of the coupling capacitors, C_c, is low at the r.f. oscillation frequency (although high at l.f.) and if the resistance of R_c is large compared with the emitter impedance of TR2 and TR3, the flow of oscillator current is principally from the tuned circuit, through TR1, the coupling capacitors, the transistors TR2 and TR3, and the two halves of the transformer secondary. It should be noted that these currents flow into opposite ends of the complete secondary and emerge together at the centre point. The magnetic fields which they generate are therefore in opposition.

Transistors TR2 and TR3 are effectively current generators and hence the currents in the half-secondary windings are not affected by any back e.m.f. from the primary winding. The sharing of the currents in TR2 and TR3 is governed by their relative emitter input impedances. These input impedances can be altered by varying the l.f. currents through these transistors by means of the voltage on their respective bases. Hence the current balance, and thus the frequency deviation is controlled by the base-to-base voltage on TR2 and TR3. This modulation effect exploits one of the most stable characteristics of a transistor, namely the forward transfer admittance Y_{21}, which for all transistors is about 37 I_c mhos where I_c is the collector current in amperes.

In Fig. 3, the modulation input instantaneous voltages v_2 and v_3 are not assumed equal for reasons that will become apparent when distortion is being evaluated.

5.2 *Sensitivity of the Practical Circuit*

Let $\hat{I}_2$ = Amplitude of r.f. collector current in TR2

$\hat{I}_3$ = Amplitude of r.f. collector current in TR3

These are equivalent to the currents in the secondary of the transformer of Fig. 1.

$$\frac{aI}{2}(1+m)\sin\omega_c t=\hat{I}_2\sin\omega_c t$$

$$\frac{aI}{2}(1-m)\sin\omega_c t=\hat{I}_3\sin\omega_c t$$

$$\therefore 1+m=\frac{2}{aI}\hat{I}_2 \qquad (6)$$

$$1-m=\frac{2}{aI}\hat{I}_3 \qquad (7)$$

Subtracting (7) from (6)

$$2m=\frac{2}{aI}(\hat{I}_2-\hat{I}_3)$$

But from adding (6) to (7)

$$aI=\hat{I}_2+\hat{I}_3$$

$$\therefore m=\frac{\hat{I}_2-\hat{I}_3}{\hat{I}_2+\hat{I}_3} \qquad (8)$$

But $$\hat{I}_2\sin\omega_c t=\frac{\hat{e}\sin\omega_c t}{Z_2} \qquad (9)$$

where $\hat{e}$ = amplitude of input carrier voltage (the same for TR2 and TR3)

and Z_2 = common base forward transfer impedance of TR2

$$=\frac{1}{Y_{21}}$$

$$=\frac{1}{KI_2}$$

where K is a constant ($=37$ volts^{-1})

and I_2 is the collector current of TR2 (as defined)

Substituting in (9)

$$\hat{I}_2\sin\omega_c t=KI_2\hat{e}\sin\omega_c t$$

$$\therefore \hat{I}_2=KI_2\hat{e}$$

$$\text{similarly } \hat{I}_3=KI_3\hat{e}$$

Thus (8) becomes

$$m=\frac{I_2-I_3}{I_2+I_3} \qquad (10)$$

If the modulation input is sinusoidal and of angular velocity ω the voltage on the emitter of TR2 is:

$$V-V_{eb}+\hat{V}_2\sin\omega t$$

where V is the bias voltage (see Fig. 3)

V_{eb} is the base emitter voltage of TR2 and TR3

$\hat{V}_2$ is the amplitude of v_2 (see Fig. 3)

Similarly, the voltage on the emitter of TR3 is:

$$V-V_{eb}-\hat{V}_3\sin\omega t$$

where $\hat{V}_3$ is the amplitude of v_3 (see Fig. 3)

$\hat{V}_3$ is assumed negative since it will normally be in anti-phase to $\hat{V}_2$.

As the collector currents are practically equal to the emitter currents

$$I_2=\frac{1}{R_e}(V-V_{eb}+\hat{V}_2\sin\omega t)$$

and $$I_3=\frac{1}{R_e}(V-V_{eb}-\hat{V}_3\sin\omega t)$$

Substituting in (10)

$$m=\frac{(\hat{V}_2+\hat{V}_3)\sin\omega t}{2(V-V_{eb})+(\hat{V}_2-\hat{V}_3)\sin\omega t}$$

The steady-state emitter current of TR2 and TR3 is given by:

$$I_S=\frac{V-V_{eb}}{R_e}$$

$$\therefore m=\frac{(\hat{V}_2+\hat{V}_3)\sin\omega t}{2I_SR_e+(\hat{V}_2-\hat{V}_3)\sin\omega t} \qquad (11)$$

Total instantaneous modulating input voltage $=v_2+v_3$, i.e. v_2 and v_3 are assumed to be in phase opposition.

To evaluate a simple, approximate formula for the sensitivity, all unbalance will be ignored and $\hat{V}_2$ assumed equal to $\hat{V}_3$.

Then total inst. mod. volt. $=(\hat{V}_2+\hat{V}_3)\sin\omega t=2\hat{V}_2\sin\omega t$ and also, Equation (11) becomes:

$$m=\frac{2\hat{V}_2\sin\omega t}{2I_SR_e}$$

substituting in the first term of Equation (5)

$$f_D=f_c\frac{kna(2\hat{V}_2\sin\omega t)}{4I_SR_e}=f_c\frac{kna}{4I_SR_e}\times\text{(inst. mod. volt.)}$$

or, in peak values

$$F_D=\frac{kna}{4I_SR_e}.f_c\times\text{(peak modulating voltage)} \qquad (12)$$

5.3 *Stability of the Centre Frequency*

The basic oscillator shown in Fig. 2 operates at a frequency governed only by L and C provided that all other components have a negligible, or at least stable, phase shift. If L and C can be kept stable, then the basic oscillator will have a stable centre frequency.

When considering the effect of the deviation producing circuit on the stability it will be appreciated that only oppositely sensed drifts in the two halves of the circuit will cause a centre frequency shift. Thus the frequency drift will be proportional to the difference in parameters in the sensitivity Equation (12) likely to arise between the two halves and any terms that are assumed equal during the course of this derivation. Consider in turn each of the parameters in (12):

The product of $I_s R_e$. This is base-emitter V_{eb} subtracted from the bias voltage. The bias voltage is common to the two halves and will therefore have no effect. V_{eb} is a basic property of the semi-conductor material used in the transistors and is not liable to significant change. In any event it would generally be small in relation to the bias voltage. Since R_e is an external component it may be formed of a high stability resistor. I_s is therefore also fixed.

The transformer coupling coefficient k. If an enclosed dust iron assembly is used for the transformer core, the coupling will not be varied by external effects. Considering that it is only the difference of the coupling coefficients of each half-secondary to the primary that can cause a frequency shift, there is unlikely to be a drift on this account.

The transformer turns ratio n. This clearly will not change.

The current ratio a. Since all other parts of the circuit are common, only a relative change in the common-base current gains of the modulating pair of transistors will cause a frequency shift. These gains very closely approach unity and do not normally significantly change during the life of a transistor.

D.C. voltage from modulating source. This would obviously cause a frequency shift but could be completely avoided by coupling to the source via capacitors. Since, in most practical cases, these will have to be of such a large value as to require electrolytic types, they should be of tantalum and not of aluminium construction to avoid the effects of leakage current. Other low leakage types will also be suitable.

The forward transfer admittance Y_{21}. This is one of the most stable of transistor characteristics and a relative change between transistors would therefore be most unlikely.

The coupling capacitors C_c. If these are large enough they will not cause a frequency shift if they change in value. However, as will be seen later when discussing pre-emphasis they cannot be very large. Indeed, they may even be series resonant circuits, in which case a relative change of L or C would cause an unbalance of oscillator current

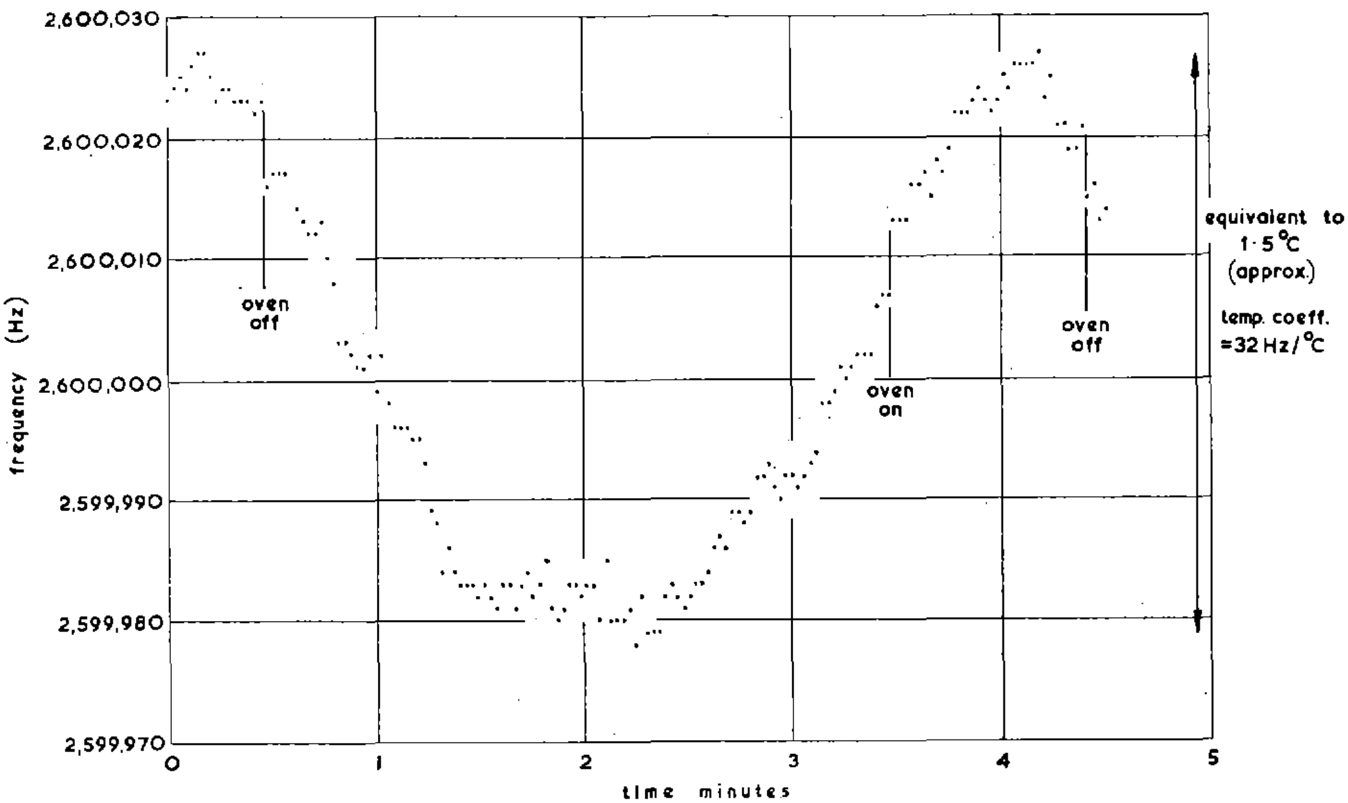

Fig 4 — Short-term frequency stability of VIFM, showing effects of oven temperature cycling and hence a frequency shift. But as both elements of these resonant circuits are fixed they can be made stable.

It would appear that the most likely source of drift would be the main tuning circuit. The capacitor of this must be variable in order to set the centre frequency, and therefore will be large and more liable to proximity effects, unless adequate precautions are taken. The very fact that it is variable may also give rise to frequency changes unless caution is exercised. The mechanical clamping of the enclosed magnetic circuit of the transformer must also be arranged with care. In order to avoid changes due to thermal variations, the modulator should be enclosed in a temperature controlled oven. The unit designed for broadcast use has been housed in a small commercial oven with a simple mercury contact thermometer operating a heater via a transistor switch. The operating temperature was chosen to be 45°C, so as to be high enough to accommodate most ambient temperatures. The differential was $\pm 0{\cdot}7$°C, and the effect of this differential is shown in Fig. 4. The overall frequency swing of about 50 Hz is quite satisfactory to meet any usual service requirement.

The long-term stability is shown in Fig. 5, which shows only a typical case and relates in particular to a new unit. Once 'run-in' the initial drift is much lower. Some units have been known to drift down in frequency, but they usually tend to a small positive drift such that a 1 kHz change takes more than three months to build up. The reasons for this drift are not clear but would appear to be associated with the core material of the main tuning inductance.

The mechanical construction chosen is shown in Fig. 6. The modulator, but not the phase splitter, is mounted on a printed circuit and housed in an oven whose dimensions just fit a BBC standard chassis. The main tuning capacitor is split into large-swing and small-swing types, the former being accessible for adjustment only by means of a screwdriver, thus avoiding any mechanical stress on its shaft.

5.4 *Harmonic Distortion*

Examination of Equations (3) and (5) shows that there is some non-linearity because it is the inductance and not the frequency which is linearly related to the unbalance factor m. But when Equation (11) is examined it is seen that, provided $\hat{V}_2$ does not equal $\hat{V}_3$, m does not vary linearly with the input voltage $(\hat{V}_2 + \hat{V}_3) \sin \omega t$. Clearly there must be a relative value of $\hat{V}_2$ and $\hat{V}_3$ which will give optimum linearity.

The condition for this can be determined by expanding the expressions for m, m^2, and m^3 obtained from Equation (11) and substituting these in Equation (5), ignoring all ultimate terms greater than third order.

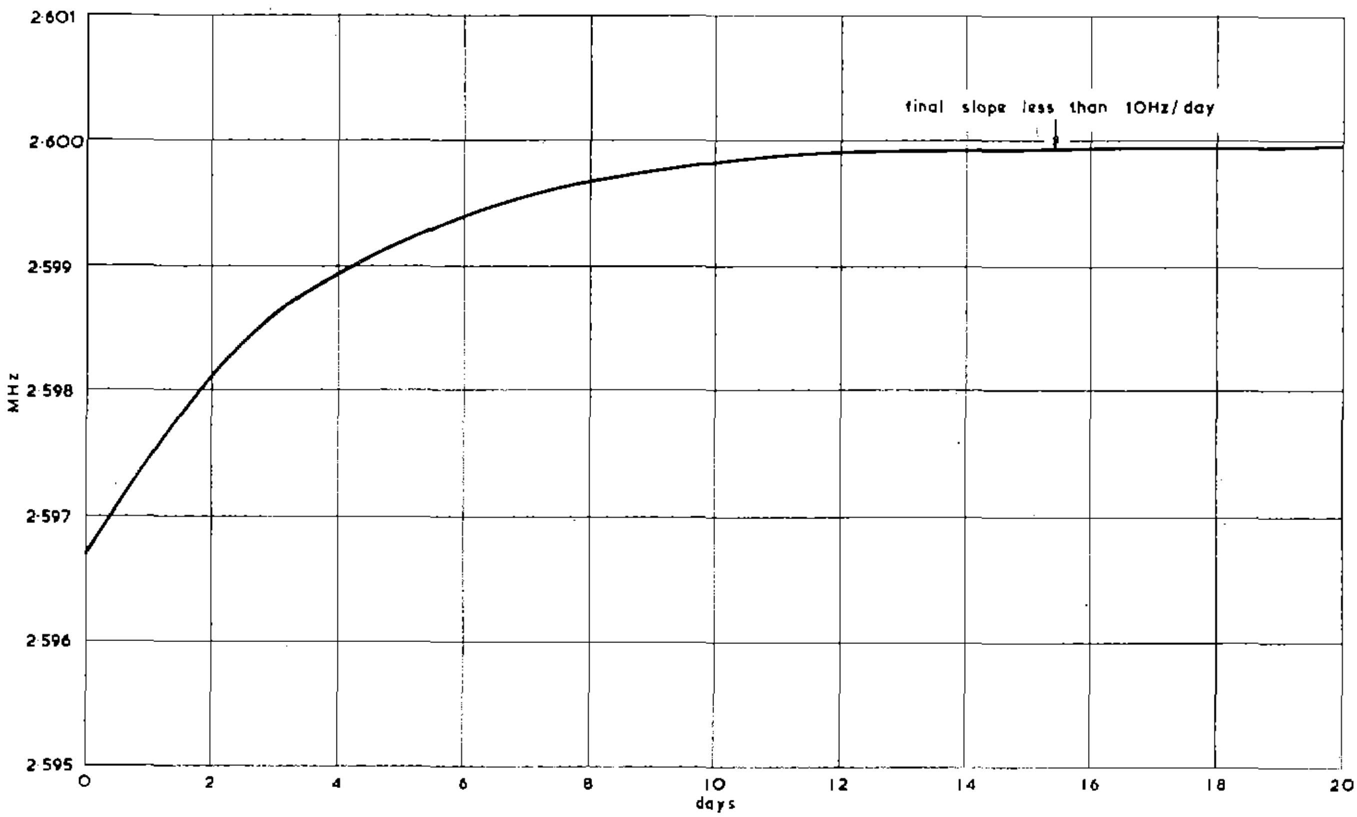

Fig. 5 — Long-term frequency stability of a typical VIFM when new

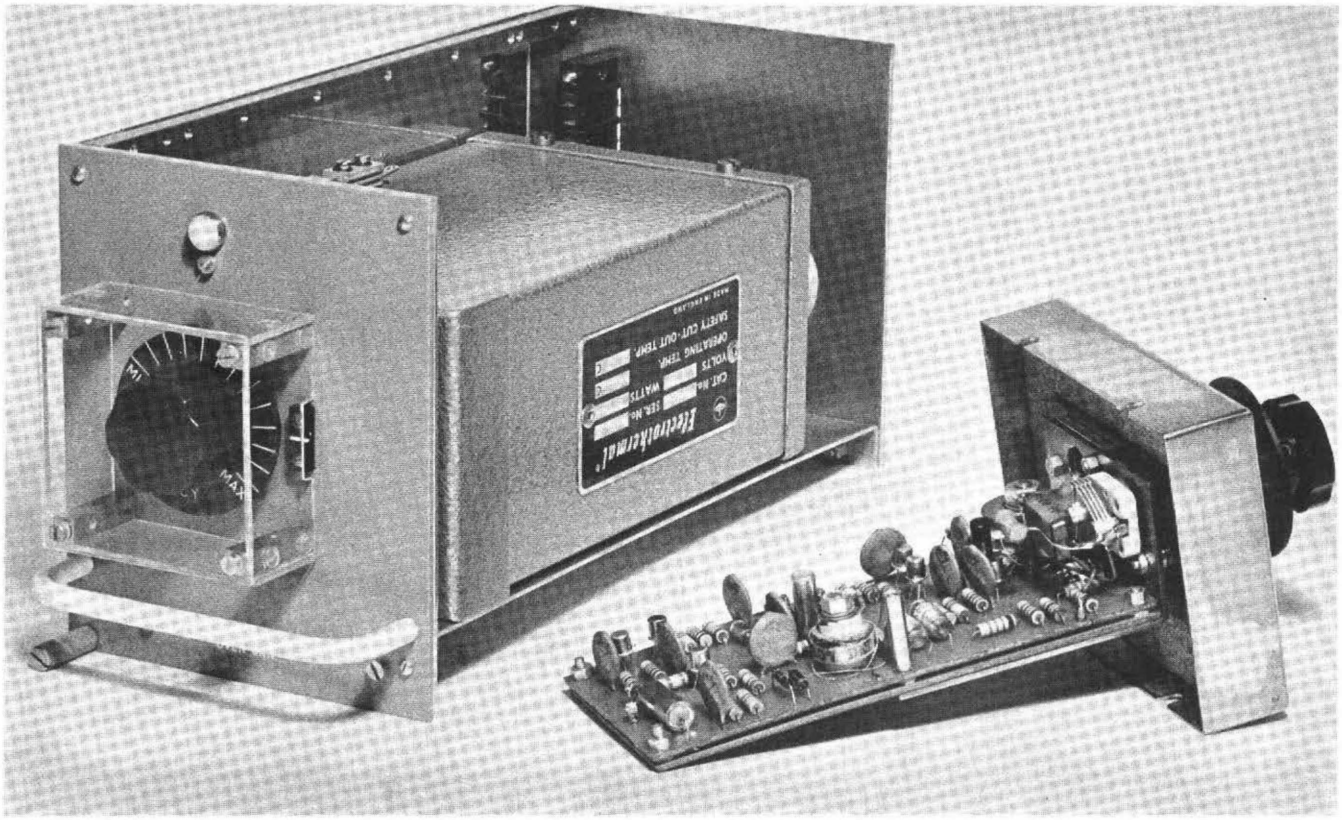

Fig. 6 — General view of the modulator and of the unit housed in an oven

For convenience let $2I_sR_e = Q$ and using this in Equation (11)

$$m = \frac{\hat{V}_2+\hat{V}_3}{Q}\sin\omega t\left[1+\frac{\hat{V}_2-\hat{V}_3}{Q}\sin\omega t\right]^{-1} = \frac{\hat{V}_2+\hat{V}_3}{Q}\sin\omega t\left[1-\frac{\hat{V}_2-\hat{V}_3}{Q}\sin\omega t+\frac{(\hat{V}_2-\hat{V}_3)^2}{Q^2}\sin^2\omega t\dots\right] \quad (13)$$

$$m^2 = \frac{(\hat{V}_2+\hat{V}_3)^2}{Q^2}\sin^2\omega t\left[1+\frac{\hat{V}_2-\hat{V}_3}{Q}\sin\omega t\right]^{-2} = \frac{(\hat{V}_2+\hat{V}_3)^2}{Q^2}\sin^2\omega t\left[1-\frac{2(\hat{V}_2-\hat{V}_3)}{Q}\sin\omega t+\dots\right] \quad (14)$$

$$m^3 = \frac{(\hat{V}_2+\hat{V}_3)^3}{Q^3}\sin^3\omega t\left[1+\frac{\hat{V}_2-\hat{V}_3}{Q}\sin\omega t\right]^{-3} = \frac{(\hat{V}_2+\hat{V}_3)^3}{Q^3}\sin^3\omega t\left[1\dots\right] \quad (15)$$

Now let $P = kna$ and substitute this and the values of m, m^2, and m^3 given by (13), (14), and (15) in Equation (5),

$$\frac{f_D}{f_c} = \left[\frac{Pm}{2}+\frac{3P^2m^2}{8}+\frac{5P^3m^3}{16}\dots\right] = \sin\omega t\left[\frac{P(\hat{V}_2+\hat{V}_3)}{2Q}\right]+\sin^2\omega t\left[-\frac{P(\hat{V}_2+\hat{V}_3)(\hat{V}_2-\hat{V}_3)}{2Q^2}+\frac{3P^2(\hat{V}_2+\hat{V}_3)^2}{8Q^2}\right]$$

$$+\sin^3\omega t\left[\frac{P(\hat{V}_2+\hat{V}_3)(\hat{V}_2-\hat{V}_3)^2}{2Q^3}-\frac{3P^2(\hat{V}_2+\hat{V}_3)^2(\hat{V}_2-\hat{V}_3)}{4Q^3}+\frac{5P^3(\hat{V}_2+\hat{V}_3)^3}{16Q^3}\right]+\dots$$

$$\frac{f_D}{f_c} = \frac{P(\hat{V}_2+\hat{V}_3)}{2Q}\left\{\sin\omega t+\frac{1}{Q}\left[\frac{3P(\hat{V}_2+\hat{V}_3)}{4}-(\hat{V}_2-\hat{V}_3)\right]\sin^2\omega t\right.$$

$$\left.+\frac{1}{Q^2}\left[\frac{5P^2(\hat{V}_2+\hat{V}_3)^2}{8}-\frac{3P(\hat{V}_2+\hat{V}_3)(\hat{V}_2-\hat{V}_3)}{2}+(\hat{V}_2-\hat{V}_3)^2\right]\sin^3\omega t\dots\right\} \quad (16)$$

Now $\sin^2 \omega t = \frac{1}{2}(1 - \cos 2\omega t)$ which represents a centre frequency shift together with second harmonic distortion. Neither of these is desirable and both should be eliminated. This condition would be achieved if:

$$\frac{3P(\hat{V}_2+\hat{V}_3)}{4} = (\hat{V}_2 - \hat{V}_3) \tag{17}$$

But as $\hat{V}_2 \propto \hat{V}_3$ let $\hat{V}_2 = h\hat{V}_3$ where h is a constant, whence

$$3P(h+1) = 4(h-1)$$

$$h(4-3P) = 4+3P$$

$$\therefore h = \frac{4+3P}{4-3P} \tag{18}$$

If h is made equal to this, then Equation (17) is satisfied and both the centre frequency shift and second harmonic distortion will be avoided. As all the quantities involved are constants the condition will hold for all modulating frequencies and all deviations.

If Equation (17) is then used to simplify Equation (16) the $\sin^2$ terms disappear and the equation becomes:

$$\frac{f_D}{f_c} = \frac{P(\hat{V}_2+\hat{V}_3)}{2Q} \sin \omega t + \frac{P^3(\hat{V}_2+\hat{V}_3)^3}{32Q^3} \sin^3 \omega t \tag{19}$$

Now let $A = \frac{P(\hat{V}_2+\hat{V}_3)}{2Q}$

$$\therefore \frac{f_D}{f_c} = A \sin \omega t + \frac{A^3}{4} \sin^3 \omega t$$

$$= A\left[\sin \omega t + \frac{A^2}{16}(3 \sin \omega t - \sin 3\,\omega t)\right]$$

$$= A\left[\left(1+3\frac{A^2}{16}\right) \sin \omega t - \frac{A^2}{16} \sin 3\,\omega t\right] \tag{20}$$

From which it follows that the percentage of third harmonic is:

$$\frac{A^2}{16+3A^2} \times 100 \tag{21}$$

Now from (12)

$$\frac{F_D}{f_c} = \frac{kna(\hat{V}_2+\hat{V}_3)}{4I_sR_e} = A$$

$\therefore$ for a centre frequency of 2·5 MHz and a deviation of 75 kHz

$A = \frac{75 \times 10^3}{2{\cdot}5 \times 10^6} =$ 0·03 and, of course, smaller for lesser deviation

Whenever A is of this order $3A^2$ is negligible compared with 16 and the expression for the third harmonic becomes:

$$\frac{A^2}{16} \times 100 \text{ per cent} \tag{22}$$

Substituting this value for A in Equation (20) gives

$$\frac{f_D}{f_c} = 0{\cdot}03(1{\cdot}00017 \sin \omega t - 0{\cdot}00056 \sin 3\omega t)$$

$$\simeq A \sin \omega t \tag{23}$$

thus confirming that, for this order of practical deviation and centre frequency, the approximation of Equation (12) is justified. From Equation (22) the percentage of third harmonic in this case is:

$\frac{0{\cdot}09}{16} = 0{\cdot}0056$ per cent corresponding to -85 dB

Even if $A = 0{\cdot}1$ the third harmonic would only amount

to $\frac{1}{16} = 0{\cdot}062$ per cent corresponding to -64 dB

5.5 *Pre-emphasis*

Examination of Fig. 3 will show that for signals applied to the bases of TR2 and TR3 in anti-phase, the emitter resistors R_e have the coupling capacitors C_c in parallel, the junction of the two coupling capacitors being a virtual earth. Thus, if the modulating frequency is high enough, the value of R_e is effectively decreased. Equation (12) shows that this will produce an increase in sensitivity.

If such a frequency modulation system requires a pre-emphasis characteristic of T seconds then it can be simply obtained by making

$$C_cR_e = T \tag{24}$$

This will avoid the use of external networks and the associated amplification to compensate for their loss.

If no pre-emphasis is required a circuit modification will usually be required. At the r.f. oscillator frequency the reactance of the coupling capacitors must be small compared with the common-base input impedance of TR2 and TR3 in parallel and this is a low impedance. At the highest modulation frequency the reactance of these capacitors must be large compared with R_e which, in turn, is typically large compared with the transistor input impedance. To meet this condition the carrier frequency should be very much greater than the highest modulation frequency. In practice, the coupling capacitors are reduced to a value acceptable to the modulation frequency and then series resonated with an inductance at the carrier frequency to produce a low impedance. The circuit must be designed so that these tuned circuits do not become a controlling influence on the frequency. Often a compromise will be necessary such that any remaining pre-emphasis is removed by an external de-emphasis network.

6. Practical Design of a Broadcast Modulator

6.1 *Band I FM Service*

Since this modulator is capable of a wide deviation with very little distortion it is advisable to keep the centre frequency as low as possible so that any drift off centre frequency is correspondingly as low as possible. For this reason, and for compatibility with existing designs of BBC transmitting equipment, a centre frequency of 2·6 MHz was chosen, at which the full standard deviation of 75 kHz was to be produced. In order to be suitable for a multiplex signal of a pilot-tone stereo system the modulation frequency must be able to vary between 30 Hz and 53 kHz with very little phase or amplitude variation. The actual noise must be at least 70 dB below the peak modulation and distortion components must be at least 50 dB below the fundamental.

From Equation (20), ignoring the third harmonic terms

$$\frac{f_D}{f_c}=A\left(1+\frac{3A^2}{16}\right)\sin\omega t$$

hence
$$\frac{F_D}{f_c}=A\left(1+\frac{3A^2}{16}\right)$$

but
$$\frac{F_D}{f_c}=\frac{75\times10^3}{2\cdot6\times10^6}=2\cdot88\times10^{-2}$$

and from this it is obvious that, as A is much less than 1, $\frac{3A^2}{16}$ will be negligible compared with 1

$$\therefore A=2\cdot88\times10^{-2}$$

The third harmonic distortion, from (22) is

$$\frac{A^2}{16}\times 100 \text{ per cent}$$

$$=0\cdot0052 \text{ per cent corresponding to } -85\cdot7 \text{ dB.}$$

In practice the conditions of Equation (18) may be difficult to meet exactly and a residue of second harmonic may remain. The correction is virtually applied by making the gain of one input of the modulating signal exactly h times that of the other. It is instructive to determine the amount of distortion which would result from making the amplitudes of each input the same, i.e. $\hat{V}_2=\hat{V}_3$. Equation (16) then becomes:

$$\frac{f_D}{f_c}=A\sin\omega t+\frac{3}{2}A^2\sin^2\omega t+\frac{5}{2}A^3\sin^3\omega t \qquad (25)$$

$$=A\left[\sin\omega t+\frac{3A}{4}(1-\cos2\omega t)+\frac{5A^2}{8}(3\sin\omega t-\sin3\omega t)\right]$$

$$=A\left[\frac{3A}{4}+\left(1+\frac{15A^2}{8}\right)\sin\omega t-\frac{3A}{4}\cos2\omega t-\frac{5A^2}{8}\sin3\omega t\right]$$

which represents:

(i) a centre frequency shift of $\dfrac{\frac{3}{4}A}{1+\frac{15A^2}{8}}\times 75$ kHz at

maximum modulation deviation of 75 kHz

$$=\frac{450A}{8+15A^2}\text{kHz}$$

(ii) Second harmonic of
$$-20[\log(8+15A^2)-\log 6A]\text{ dB}$$

(iii) Third harmonic of
$$-20[\log(8+15A^2)-\log 5A^2]\text{ dB}$$

For the design under consideration $A=2\cdot88\times10^{-2}$ whence

centre frequency shift at full modulation $=1\cdot62$ kHz

second harmonic $=-33\cdot3$ dB

third harmonic $=-65\cdot7$ dB

If it is feasible to obtain, and hold, 20 dB of cancellation of the $\sin^2$ term in Equation (25) this can be rewritten:

$$\frac{f_D}{f_c}=A\sin\omega t+\frac{3}{20}A^2\sin^2\omega t$$

(It has been shown that the effect of the $\sin^3$ term can be ignored in these practical cases.)

$$\therefore\frac{f_D}{f_c}=A\left(\frac{3A}{40}+\sin\omega t-\frac{3A}{40}\cos2\omega t\right)$$

which represents a centre frequency shift of only 162 Hz and second harmonic of $-53\cdot3$ dB.

These values of centre frequency shift and second harmonic are quite acceptable in practice and appropriate values have been assigned to the basic components of Fig. 3 to satisfy these conditions.

The resultant practical circuit is shown in Fig. 7.

In this, TR1, 2, 3, and 4 form the modulator as previously described. Each transistor is biased such that a steady current of 1 mA flows and the attenuator across the limiter diodes in the collector of TR4 sets the oscillator current in TR1 emitter to 1 mA d.a.p. The oscillator current is passed to TR2 and TR3 by a series resonant circuit in order to avoid excessive increase in sensitivity at the higher modulation frequencies. The 4 μH inductor is resonant with the 910 pF capacitor in series and this with $R_c=5$ kΩ gives a time constant of 4·55 μsec. The current then passes through TR2 and TR3 to the secondary of the transformer.

Since slight phase shifts occur between the primary and secondary of the transformer in practice, they are corrected by a 510 pF capacitor at TR1 collector. If this is not done a variable damping appears across the transformer primary resulting in amplitude modulation. A choke which introduces loss above 2·6 MHz is placed in the centre tap of the secondary to reduce high frequency gain which might otherwise result in instability. The signal is then phase inverted and limited by TR4, the limiter consisting of two silicon diodes back to back. To compensate for phase shifts at this stage there is an undecoupled emitter impedance of 220 Ω and 220 pF in parallel. At frequencies high compared with 2·6 MHz 220 pF will fully decouple the 220 Ω emitter resistance, and this will give TR4 a high gain. A 68 Ω resistor is therefore placed in series to avoid this. The attenuated output is fed back to

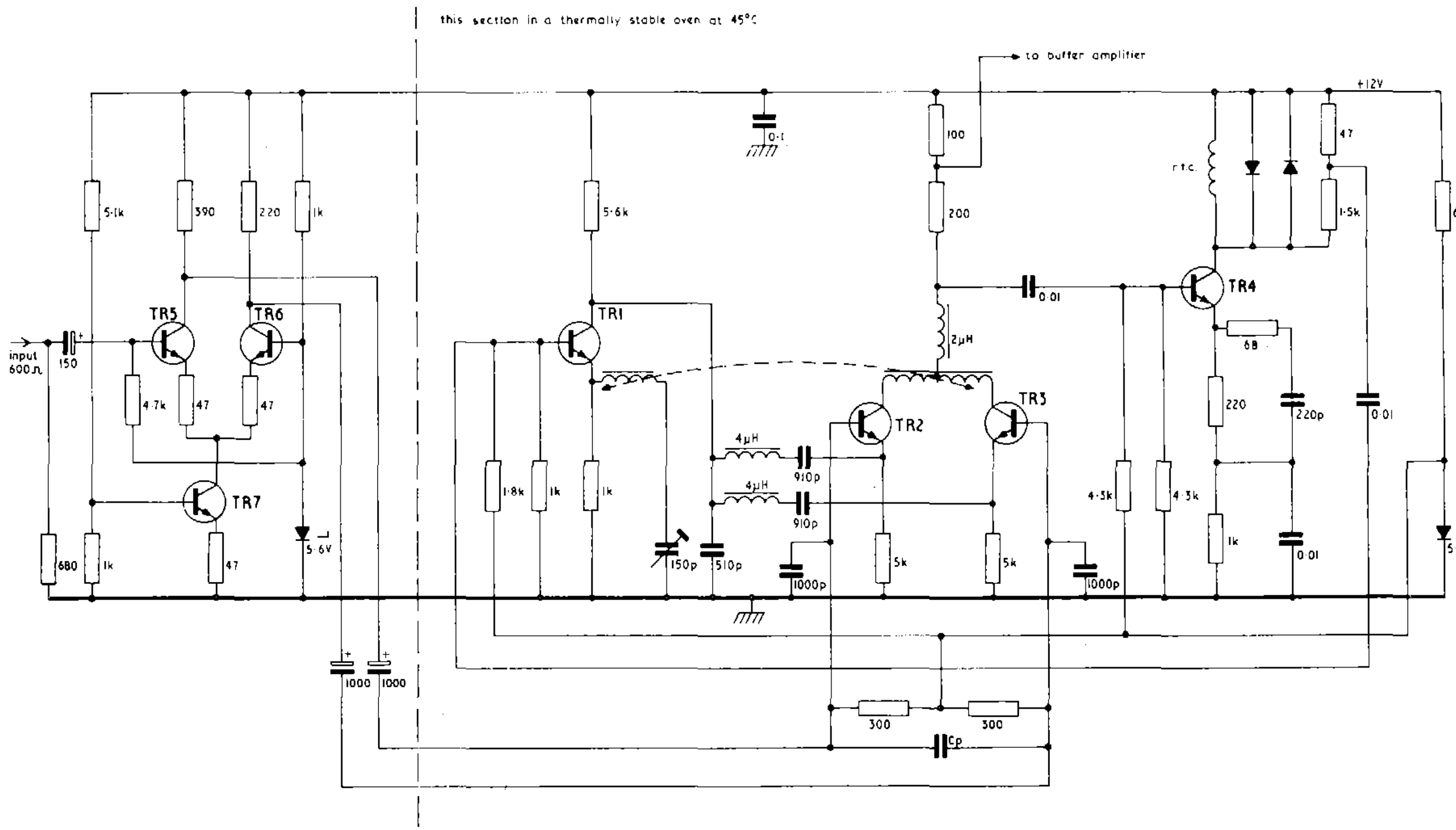

Fig. 7 — Circuit of variable inductance frequency modulator

TR1 where the tuned circuit across TR1 emitter selects the fundamental.

The unbalanced input from the modulation source is phase split by a long-tailed pair TR5 and TR6, the long tail being formed by TR7. Unequal collector loads cause the unbalance required for distortion correction and the output is taken to a 600 Ω load formed by the bias resistors of TR2 and TR3. The capacitor Cp corrects the frequency response of the modulator.

Values are given on this circuit for most components. Other parameters not shown are given below:

$L = 23{\cdot}2\ \mu H$	$a = 0{\cdot}6$ (estimated)
$k = 0{\cdot}84$	$I_2 = I_3 = I_s = 1\ mA$
$n = 0{\cdot}65$	$V = 5{\cdot}6$ volts

From Equation (12)

$$\hat{V}_2 + \hat{V}_3 = \frac{4F_D I_s R_e}{f_c kna}$$

By substituting known values $\hat{V}_2 + \hat{V}_3 = 1{\cdot}762$ volts.

From Equation (18), for minimum distortion

$$h = \frac{4+3P}{4-3P} \text{ where } h = \frac{\hat{V}_2}{\hat{V}_3} \text{ and } P = kna = 0{\cdot}3276$$

$$\therefore h = \frac{4{\cdot}983}{3{\cdot}017} = 1{\cdot}651$$

But $\hat{V}_2 + \hat{V}_3 = (h+1)\hat{V}_3$

$$\therefore \hat{V}_3 = \frac{\hat{V}_2 + \hat{V}_3}{h+1} = \frac{1{\cdot}762}{2{\cdot}655} = 0{\cdot}665 \text{ volts}$$

$$\hat{V}_2 = h\hat{V}_3 = 1{\cdot}655 \times 0{\cdot}663 = 1{\cdot}097 \text{ volts}$$

In the practical circuit shown in Fig. 7, $\hat{V}_2 + \hat{V}_3$ was found to be 2·04 volts for 75 kHz deviation, making the modulator 1·3 dB less sensitive than predicted. The separate components of the input were 1·23 volts and 0·81 volts, i.e. a ratio of 1·52, whereas theory predicts a ratio of 1·651.

The distortion characteristics of this modulator are shown in Fig. 8, with a modulating frequency of 1 kHz applied as a monophonic signal. The drive voltage asymetry had been adjusted to give a second harmonic separation of 54 dB, which was about the limit of the test apparatus available. A roughly comparable centre frequency shift was produced. The third harmonic shown was almost entirely produced by the phase splitter, whose performance was somewhat restricted by the practical necessity of having only a 12-volt supply.

The demodulated, but not decoded, output of the modu-

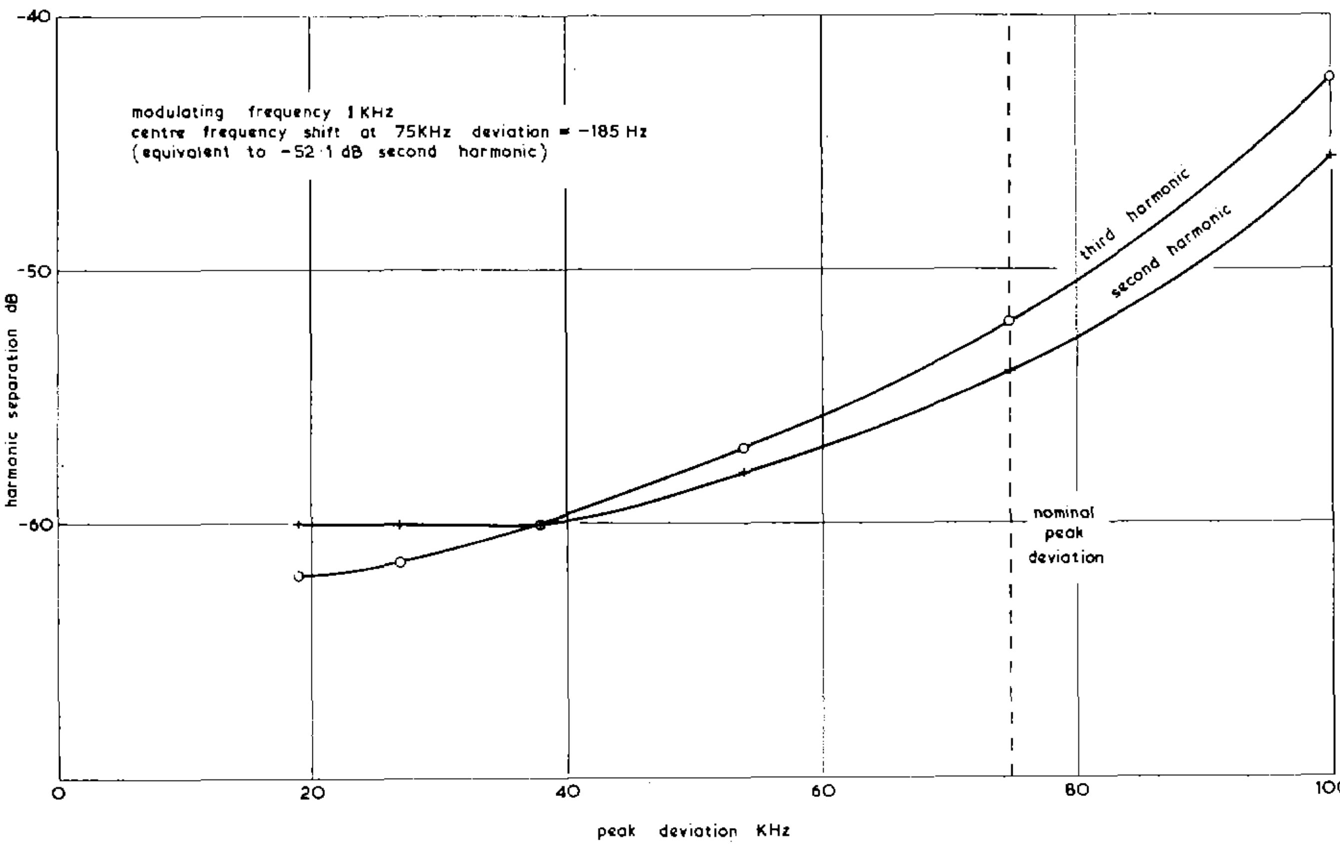

Fig. 8 — Harmonic distortion of a typical VIFM with asymmetrical drive voltages

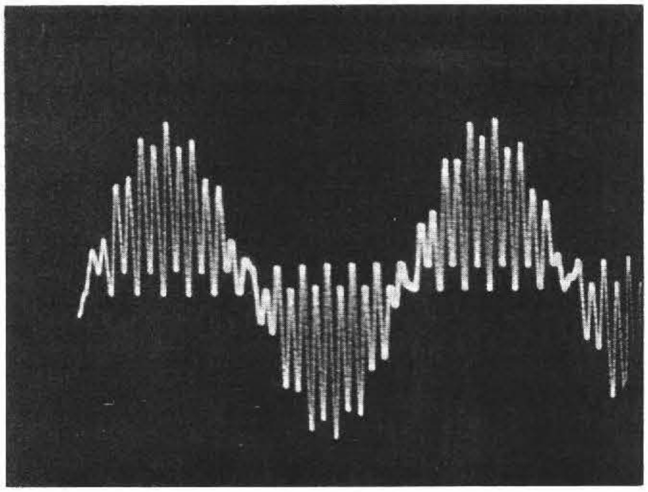

Fig. 9 — *Demodulated output with pilot tone stereophonic transmission and* 1·5 *kHz in one channel only*

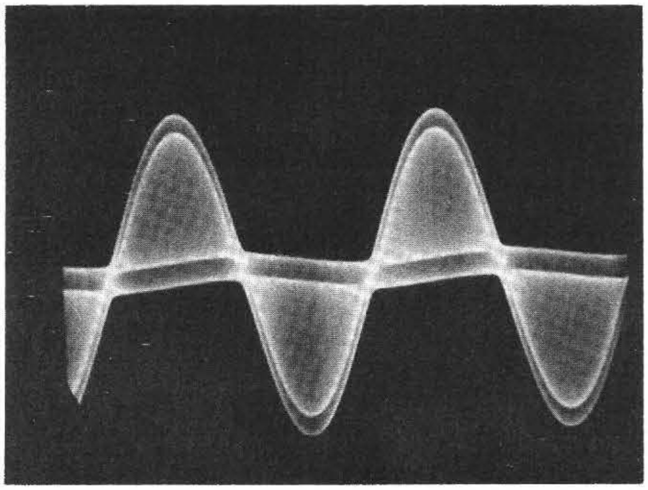

Fig. 10 — *Demodulated output with pilot tone stereophonic transmission and* 30 *Hz in one channel only*

lator is shown in Figs. 9 and 10. The former shows a 1·5 kHz signal present on either the left or the right channel so that the sum and difference signals are equal. Fig. 10 shows the effect of reducing this frequency to 30 Hz. The base line is no longer straight because of phase shifts, caused by coupling capacitors, between the components at 30 Hz and 38 kHz. If this base line is not straight it signifies the presence of linear crosstalk. Fig. 11 shows the variation of crosstalk with frequency. Except at high frequencies the linear component is predominant, and tends towards a 6 dB per octave slope at low frequencies. Mid-band figures of about 50 dB can usually be attained by careful adjustment of the compensating capacitor C_p in Fig. 7. At high frequencies the non-linear components become predominant for the reasons explained in Section 2.2.

6.1.1 *Possible Simplifications for Monophonic Operation*

The phase splitter formed by TR5, 6, and 7 may be replaced by a transformer if it is not required to operate above about 20 kHz. The secondary of this transformer must be tapped and returned to the bias zener. The unbalance can be achieved either by having an unequal number of turns each side of the centre tap or by introducing a resistance into one leg. This will usually require a higher input level since the phase splitter has been designed to have gain. The transformer will allow balanced or unbalanced input.

The coupling capacitors C_c need not be series resonated if a built-in pre-emphasis characteristic is required. For BBC use, a 50 μs time constant is standard. Thus from Equation (21)

$$C_c R_e = 50\ \mu s$$
$$\therefore C_c = 10{,}000\ pF$$

6.2 625-*line Television Sound*

If the modulator runs at 6 MHz centre frequency its output may be mixed with a vision carrier to produce the required r.f. channel for television sound. In this case the importance of the other mixer products is reduced since the alternative sideband of the complete transmission would be rejected by existing vestigial-sideband shaping filters; also the transfer oscillator signal is the vision carrier and hence a breakthrough of this, even only 30 dB below sound carrier, is not an embarrassment. The peak deviation requirement is only 50 kHz and not 75 kHz.

$$\text{Now} \qquad \frac{F_D}{f_c} = \frac{5 \times 10^4}{6 \times 10^6} = 0{\cdot}833 \times 10^{-2}$$

As has been shown, it can be assumed that this is the value of A. (Page 14.) To evaluate the basic distortion of the modulator under balanced drive conditions, this value is inserted into Equation (25) as before.

Then centre frequency shift at full modulation = 469 Hz

second harmonic	= −44·8dB
third harmonic	= −87·2 dB

These are already within acceptable limits and it is therefore unnecessary to resort to assymetrical input voltages for further correction. Indeed, since stereophony is not required, the pre-emphasis can be made integral to the unit and the input isolated by a transformer. This gives a very simple and cheap modulator. The one performance parameter that will be degraded is the centre frequency stability since temperature and possibly other effects are worse at the higher centre frequency. This is somewhat alleviated by the lower percentage deviation requiring less sensitive circuits.

There is no reason why assymetrical input voltages should not be used if it is wished to improve the performance beyond 0·6 per cent distortion. The pre-emphasis can be retained since Equation (18) does not contain the magnitude of Re and so the correction will still be maintained at all operating frequencies.

If it is desired to keep the centre frequency at an exact

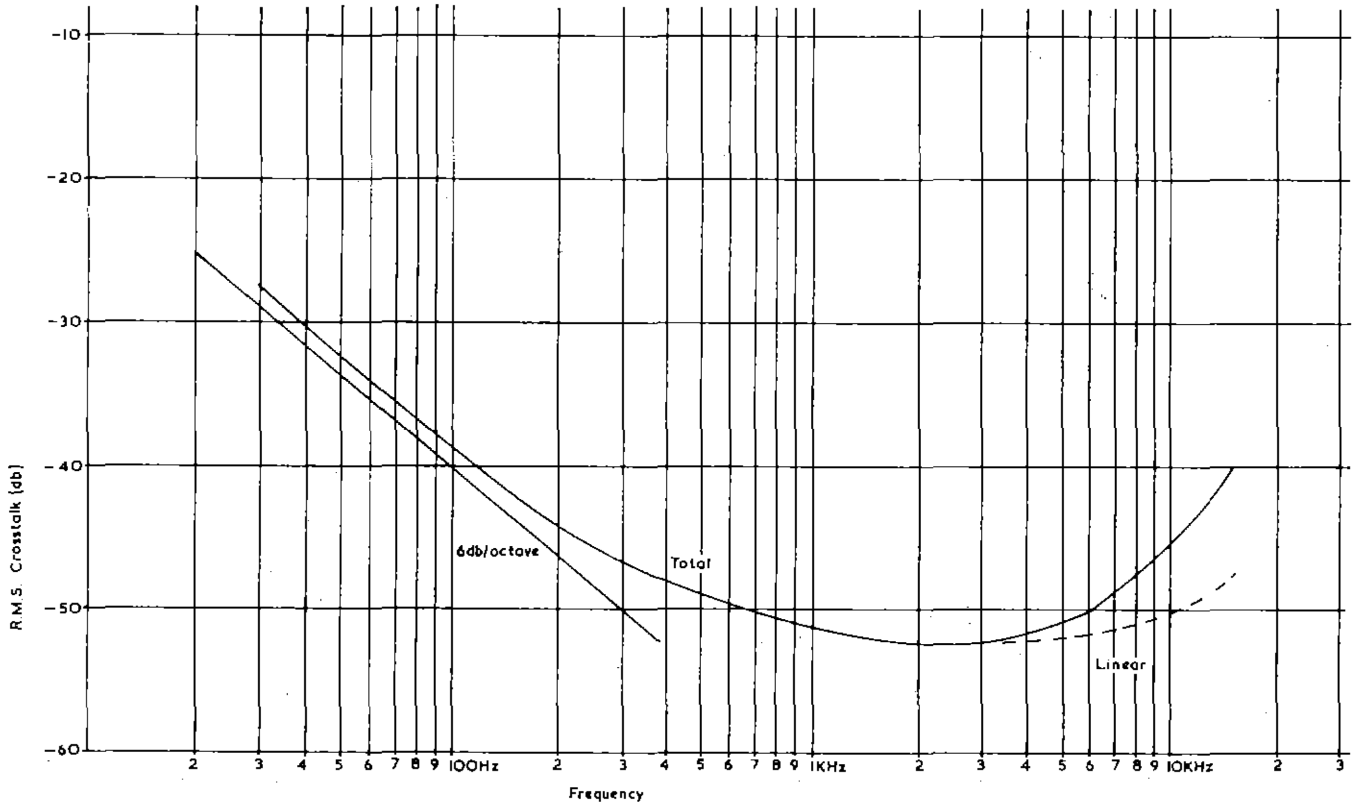

Fig. 11 — *Interchannel crosstalk characteristics with de-emphasis*

figure, the modulator centre frequency may be automatically controlled by a d.c. input added to the programme. To obtain this control signal a standard frequency and that of the modulator are compared in a discriminator and the resultant programme output integrated to obtain the d.c. signal which can be used for correction.

7. Application to a Band II Transmitter

The block diagram of a Band II drive or low power transmitter is shown in Fig. 12. The modulator output is limited to remove any spurious amplitude modulation and mixed with the output of a 13·3 MHz crystal oscillator. The desired output band centred on 10·7 MHz is selected by a filter that has sufficient selectivity to reject the fourth harmonic of the modulator at 10·4 MHz. This 10·7 MHz signal is then mixed with the output of a second crystal oscillator to produce the desired Band II signal which is then increased in level by a tuned amplifier until an output of some 15 watts is obtained. If the unit is being used as a transmitter in its own right a filter to stop harmonics is added to the output circuit. This drive unit is completely solid-state and is already in service at a number of BBC transmitting stations including Wrotham and Holme Moss, which radiate stereophonic programmes. Numerous monophonic versions are in use in remote unattended stations.

The complete drive unit is shown in Fig. 13. From left

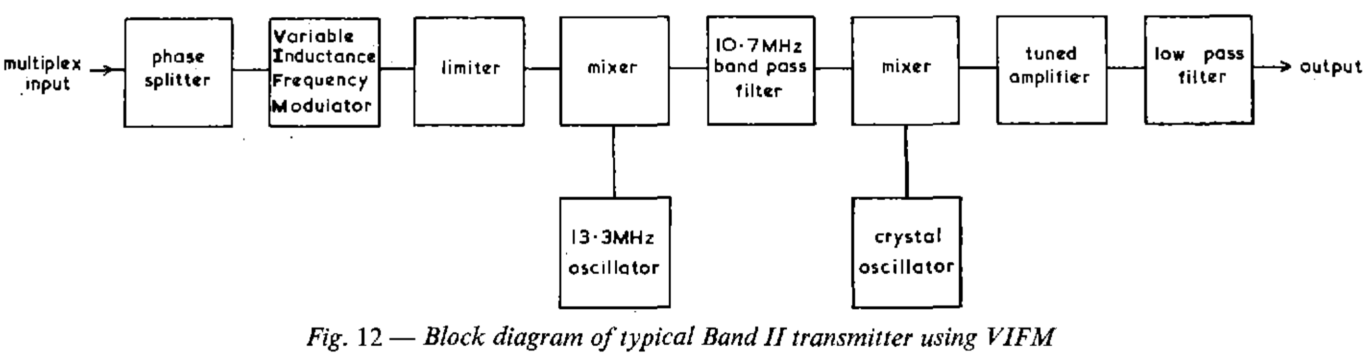

Fig. 12 — *Block diagram of typical Band II transmitter using VIFM*

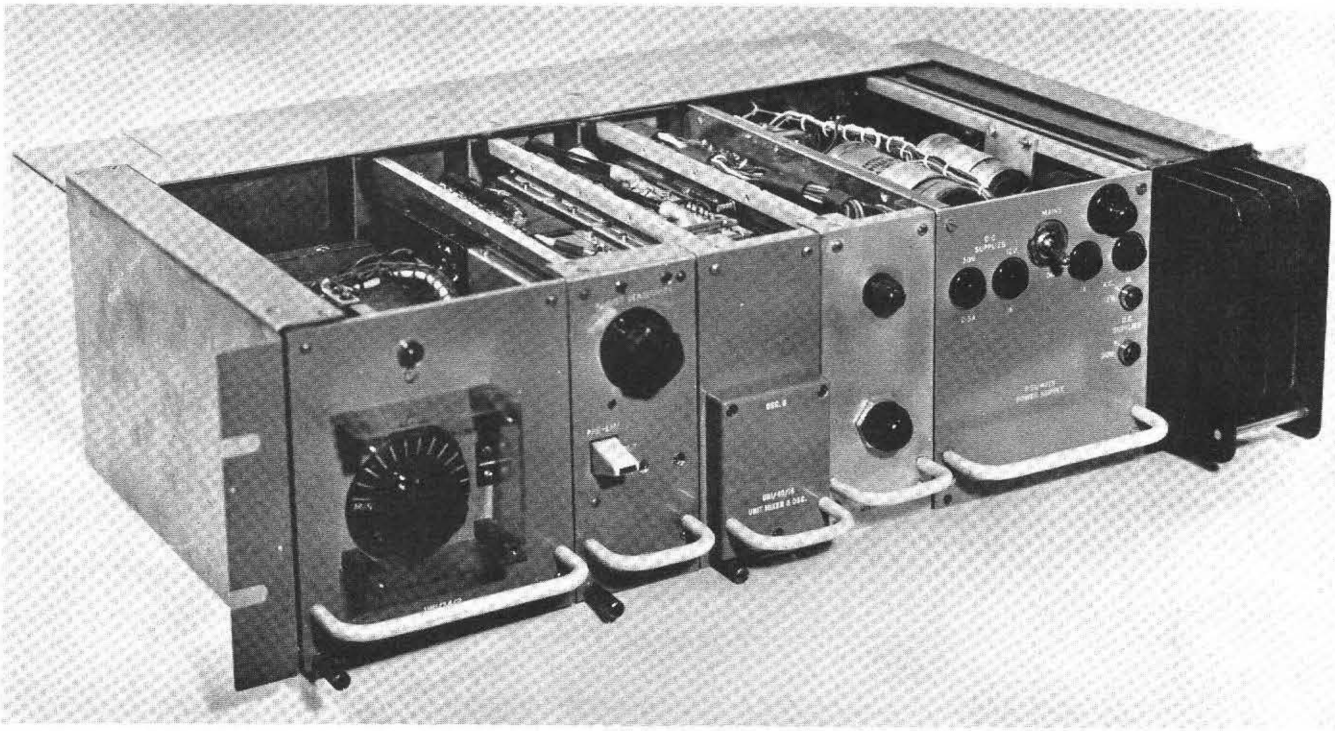

Fig. 13 — General view of a complete Band II drive

to right the units of Fig. 12 are accommodated as follows:

(*a*) The modulator and oven assemblies.

(*b*) Phase splitter, limiter, and mixer.

(*c*) Second oscillator, second mixer, and first stages of output amplifier.

(*d*) Oven power supplies and first oscillator.

(*e*) D.C. power supply.

(*f*) Final stages of output amplifier.

All the plug-in units carrying r.f. signals contain individually screened boxes and use a multi-way co-axial type of back connector.

8. Conclusions

It has been shown that a simple and inexpensive linear frequency modulator can be constructed on the principle described. It accepts a wide band of modulating frequencies and introduces very little phase or amplitude distortion. The noise performance of the practical unit is adequate even for multiplex stereo systems, producing a signal-to-noise ratio of about 60 dB in either the left or right output channel.

Two versions have been designed for BBC use, the MD3/1A, with integral pre-emphasis for monophonic systems and the MD3/2 without pre-emphasis, primarily for stereophony. The former requires a UN1/33 input unit, and the latter a UN1/78. When provided with an external pre-emphasis unit, such as the NE1/6, this latter unit is suitable for monophony.

The design of this modulator is covered by BBC Patent No. 1076831.

9. Acknowledgements

Grateful acknowledgements are made to many members of the BBC Designs Department, in particular Mr G. G. Johnstone, who have contributed to the design of this equipment, and also to members of Research Department.

10. Bibliography

Tibbs, C. E., and Johnstone, G. G., **Frequency Modulation Engineering,** Chapman and Hall, pp. 244–6, 1956.

Peachey, F. A., Toombs, R., and Smart, D. L., **Apparatus for Television and Sound Relay Stations,** BBC Engineering Monograph No. 42, July 1962.

BBC ENGINEERING TRAINING MANUALS

The following manuals by members of the Engineering Division of the BBC have been prepared primarily for the Corporation's operating and maintenance staff. They have been made available to a wider public so that the specialized knowledge and experience contained in them may be open to all interested in the engineering side of sound and television broadcasting.

Microphones—A. E. Robertson, B.Sc.(Eng), M.I.E.E. 75s. net, by post 76s. 3d. 357 pp.

Sound and Television Broadcasting: General Principles—K. R. Sturley, Ph.D., B.Sc., M.I.E.E. 45s. net, by post 46s. 5d., 382 pp.

This manual explains the basic principles of sound and television broadcast engineering and operations.

Television Engineering: Principles and Practice—S. W. Amos, B.Sc.(Hons.), A.M.I.E.E., and D. C. Birkinshaw, M.B.E., M.A., M.I.E.E.

Vol. I: Fundamentals, camera tubes, television optics, electron optics. 45s. net in U.K. only, by post 46s. 3d. 301 pp.
(Written in collaboration with J. L. Bliss, A.M.I.E.E.)

Vol. II: Video-frequency amplification. 45s. net, by post 46s. 3d., 270 pp.

Vol. III: Waveform Generation. 45s. net, by post 46s. 3d., 224 pp.

Vol. IV: General Circuit Techniques. 35s. net, by post 36s. 3d., 277 pp.

These manuals are published by ILIFFE BOOKS LTD, DORSET HOUSE, STAMFORD STREET, LONDON, S.E.1, for the British Broadcasting Corporation, and are available from the publishers or from BBC PUBLICATIONS, 35 MARYLEBONE HIGH STREET, LONDON, W.1.

ENGINEERING TRAINING SUPPLEMENTS

		s.	*d.*
No. 5	An Introduction to Wideband Aerials	3	0
No. 6	Programme Meters	3	0
No. 9	Frequency Modulation	6	6
No. 11	Lighting for Television Outside Broadcasts	6	0
No. 14	Colorimetry	4	6

These Engineering Training Supplements are available from BBC TECHNICAL PUBLICATIONS SECTION, HAREWOOD HOUSE, HANOVER SQUARE, LONDON, W.1.

Published by the British Broadcasting Corporation, 35 Marylebone High Street, London, W.1. Printed in England by The Broadwater Press Ltd, Welwyn Garden City, Herts. *SBN: 563 08408 1*