Founded 1925
Incorporated
by Royal Charter 1961
To promote the advancement of radio, electronics and kindred subjects by the exchange of information in these branches of engineering

Volume 47 No. 5 May 1977

The Radio and Electronic Engineer

The Journal of the Institution of Electronic and Radio Engineers

Technology and the Annan Committee

A Synopsis

WITH terms of reference as wide as those given to the Annan Committee on the Future of Broadcasting, it is easy to be overcome by the heady opportunities offered by the newer technologies—the 'unlimited' channel capacity of future glass fibre systems, the attraction of return-path television, the possibilities of international satellite ventures, Teletext, and so on. Indeed we considered all of these things—but at the risk of being thought pedestrian attached rather more importance to ensuring a greater equality of opportunity for all to receive the radio and television programmes currently transmitted within the United Kingdom.

The question of the public's right to receive is a tempting one to debate—but clearly it would be impractical to recommend that everyone in the country should have the right to receive a signal if they wanted it. The costs of getting the signal to the most isolated inhabitant would be prohibitive. The coverage question is however of vital importance, for the life of the existing 405-line transmitters is rapidly moving to a close and unless full priority is given to the planned extension of u.h.f. transmissions to all populations exceeding 500 by the early 1980s, several hundred thousand people, currently able to watch television, will be unable to receive any signal at all.

The question of radio coverage is no less important for with ever increasing demands on the spectrum the present situation is manifestly unsatisfactory. The completion of v.h.f. radio coverage, particularly in the deprived areas of Scotland and Wales, must thus be improved with all possible speed, for night reception on m.f. will never be adequate, however present services are redistributed. The question of v.h.f. radio presents other problems too, for the reluctance of many to use v.h.f. sets has been long evident. Difficulty of set tuning is cited as the main reason for not preferring v.h.f. (though 75% of the public own v.h.f. sets)—but inadequate set aerial provision is undoubtedly a prime cause of poor reception in many areas. Current technology offers the possibility of producing easily tunable sets and it would be in everyone's interest if they didn't have to be imported!

There was frequent criticism of the sound quality of television receivers and although industry stated that there was little demand for better quality sound, it is noteworthy that many imported sets sell well on this ticket. Representations were made also about the transmission of stereo sound for television—mainly by music lovers—and this too needs further examination, for, though there are undoubted difficulties in achieving a realistic synchronism between the picture transmitted and the stereo sound, the problem should be minimal where music is concerned. The fact that there are now several techniques for the transmission of dual-channel sound within the normal video bandwidth encouraged us to recommend experiments in this area—particularly with dual language transmissions. Although clearly the ability to select either of two languages means that one will effectively be dubbed, it would be for many, e.g. Welsh speakers, preferable to receive a dubbed programme in their own tongue rather than always be faced with listening to a foreign language.

So what of the future? Even with the modest 15-year horizon to which the Committee worked, the fact that economic conditions rather than technology dominate the pace of development made clear recommendations difficult. One conclusion which became evident at an early stage, however, was that even with effectively unlimited technological potential, there is no point in investing in it unless it can be used well. In the broadcasting context this means that there must be adequate programme material to fill the available capacity without dilution of programme standards. We did not feel that broadcasting alone at present could realistically justify more channel space than is currently available. True, wide-band cable could serve many needs, but operated for pay-TV on a wide scale it could work against the public interest if it resulted in exclusive rights being acquired to programme material which would otherwise be freely available through existing outlets. At this stage therefore we favoured continuation of experiments in the local use of cable—but on a far more carefully monitored and realistically funded scale than those which have failed in the past.

And yet, in the end, it is difficult to avoid the conclusion that a broad-band cable system is the ultimate, logical, way of distributing all forms of fixed communication signals, leaving 'the air' free for mobile users. The time for such a system is nevertheless not yet, for present technology is expensive and with cheaper alternatives around the corner it would be prudent to wait. It was interesting to us to find that Germany, facing the same problem, had decided that investment in cable at the moment could best be devoted to improving its telex capability—unexciting—maybe, but realistic!

Then there were other questions:

Would video cassettes make a large impact on the kinds of programme transmitted, or encourage the use of the air for recording in the 'silent hours'? As technology stands today we thought not—but technology has a habit of advancing rapidly and a vigilant eye is needed to make sure that its potential is not wasted.

Where should Teletext stand—most of us were keen for it to go ahead—but Viewdata was outside our terms of reference and could not Viewdata in time involve the same sort of editorial decisions as broadcasting?

What about the re-use of Bands I and III when v.h.f. ceases, or the use of a clearance of Band II up to 108 MHz? Here again broadcasting is only one claimant and though we recommended that v.h.f. local radio and a fourth national radio channel should command the highest priority for frequencies in Band II and that space in Bands III and I should be reserved for 'local' television, at some future date, we were acutely aware of the rapidly increasing needs of other users.

These latter are but a few of the problems which involve the use of developing technology. The rate of development and the scarcity of spectral resources, as well as the fickleness of economic uncertainty, left us in no doubt that the time had come to establish a standing advisory body, a 'Telecommunications Advisory Committee', to keep such matters under continuous review and advise Government accordingly in relation to general technological investment policy. We were clear that there is a need for Government to consider all telecommunication technology as a whole and casual committees with specialized terms of reference akin to the old T.A.C. are not appropriate to a future where common technology will serve a variety of hitherto independent areas of application.

GEOFFREY SIMS

Professor G. D. Sims, O.B.E., Ph.D. (Fellow 1966), Vice-Chancellor of the University of Sheffield since 1974, was a member of the Committee on the Future of Broadcasting under the chairmanship of Lord Annan which reported in March of this year, two and a half years after its setting up by the present Government. Professor Sims serves on or has served on many other committees, government, educational or scientific, for example the Economic Development Committee for the Electronics Industry and its Manpower Working Group, the British Library Organizing Committee, the Government Delegation to the OECD Inter-Governmental Conference on Highly Qualified Manpower, the Naval Education Advisory Committee, the Electronics Research Council, the Electrical Engineering Board of the CNAA, and, since last year, as first Chairman of the British Council Engineering and Technology Advisory Committee. A graduate of Imperial College, London, he has worked both in industrial and government research laboratories and after seven years on the teaching staff of the Department of Electrical Engineering at University College London, he was appointed in 1963 to the Chair of Electronics and Head of the Department at the University of Southampton. Professor Sims has published numerous papers on electronics and on education and he has written two books and edits a monograph series on electrical subjects. Papers in *The Radio and Electronic Engineer* have been on 'Microwave semiconductor devices' and (as a co-author) on 'The future education of electronic engineers'. In 1974 he was chairman of the organizing committee of the IERE Conference on The Electronics Industry and Higher Education.

UDC 621.372.55: 621.391.822.4: 621.396.62.029.55

Noise equalization in h.f. receiving systems

D. C. BUNDAY,
A.M.B.I.M., C.Eng., F.I.E.E., F.I.E.R.E.*

SUMMARY

The paper analyses a concept of noise equalization that can be applied in order to reduce the noise contribution due to intermodulation effects without affecting the basic signal to noise performance. Some practical applications of the noise equalization technique are outlined and the resulting operational benefits demonstrated.

* *Engineering Operations Division, Government Communications Headquarters, Oakley, Priors Road, Cheltenham GL52 5AJ.*

1 Introduction

The 2 to 30 MHz spectrum continues to enjoy its long established popularity as a long-range communications medium with all its well-known advantages and disadvantages. Indeed all the indications are of a steady increase in the number of transmissions with an even greater increase in the total power transmitted as equipment is superseded and improved communication standards brought into effect. At the same time the application of modern design principles to major h.f. stations has resulted in the use of wideband aerial arrays feeding multiple numbers of remotely tunable receivers in more flexible systems.

The wideband flexible approach tends to produce its own problems particularly if active devices in the system like distribution amplifiers are exposed to the full output from a wideband aerial prior to any filtering. The wide range of signal levels almost inevitably leads to the generation of intermodulation products (i.p.s) at levels in excess of the noise floor of the system. These i.p.s then inhibit system performance. The design of equipment to accept high signal levels without detectable i.p.s tends to be both complex and expensive.[1] At the same time the relatively high external noise levels have led to the opinion that low noise factors are neither necessary nor desirable in h.f. receiving systems.[2]

These two references present useful information on noise, signal levels, i.p.s etc. that has general application in this field. The one factor that appears not to have been exploited is the excess of noise at the lower frequencies due to increasing external noise levels with reduction in frequency. At the same time the noise due to the active devices remains essentially constant with the frequency. Noise equalization seeks therefore to exploit this excess noise aspect by reducing input noise levels to provide a more constant noise over-ride across the frequency range at the input to these active devices. It has the direct effect of reducing signal levels at the active devices to the lowest acceptable values, thereby reducing i.p.s levels by a proportionally greater amount depending on the order of the i.p.s, whilst at the same time retaining virtually the same overall noise performance.

2 Noise Sources and System Performance

The main sources of noise in an h.f. receiving system are those external to the system and the ideal system should not in itself add noise to the unavoidable external noise. In practice the system also produces internal noise and it is the balance between external and internal noise that has to be controlled in order to produce optimum performance. When performing noise calculations it is most convenient to work in terms of noise temperature ratio expressed in logarithmic form where appropriate. The investigation of other aspects of system performance often requires that noise temperature be converted to noise power. In these investigations it is essential to consider the noise levels at the input and output of each active device as well as taking the more conventional approach of referring all noise to the system input as a measure of performance. The comparison of noise and signal levels is assisted by converting noise levels in a 3·1 kHz band to dB relative to 1 μV in 75 Ω where at reference temperature of 290 K the equivalent noise voltage is −30·3 dBμV.

2.1 External Noise

The noise external to the system comprises atmospheric, man-made and galactic noise and is described in CCIR Report No. 322.[3] The level of this noise can vary widely, particularly at the lower frequencies, but the CCIR conclude that man-made noise tends to set the lower limit and give man-made noise levels that are representative of a low noise area at quiet times. These levels are generally accepted as the minimum external noise levels for system planning purposes and they are expressed in noise temperature ratio terms and shown for practical purposes as curve A in Fig. 1.

It is then necessary to consider how the external noise is coupled into the system through the aerial. The curve A in Fig. 1 represents the noise temperature, and hence the noise power in a given bandwidth, available from a loss-free short vertical aerial over a perfectly conducting ground plane. In practice there will be aerial inefficiency, ground and mismatch losses that reduce the external noise available to the system and these must be taken into account although they can be minimized by the use of balanced and matched aerial arrays. The spatial distribution of the external noise will also affect the external noise available to the system. It has been shown[4] that atmospheric noise due to distant thunderstorms is likely to show directive properties. Propagation of man-made noise on the other hand is principally over power lines

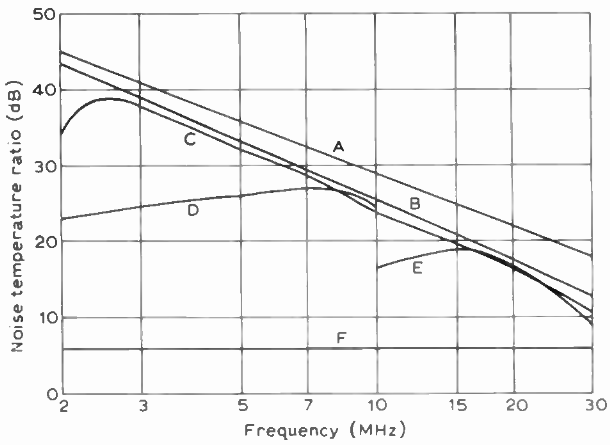


Fig. 1. Aerial feeder output noise under quiet noise conditions.
A CCIR low noise area at quiet times.
B Log periodic array feeder output at quiet times.
C Combined high and low band rhombic feeder output at quiet times.
D 12 m elevated feed monopole feeder output at quiet times.
E 6 m elevated feed monopole feeder output at quiet times.
F Amplifier/receiver input noise level (7 dB noise factor).

and by ground wave. Hence in an inland location remote from a built-up area in the UK or Western Europe man-made noise may reasonably be assumed to be omnidirectional. Thus for quiet noise conditions the external noise available to the system will tend to remain constant regardless of aerial gain or directivity. However if the external noise is noticeably directional then the external noise available from a directional aerial will depend on the relative directions of aerial and noise.

It is important to note that at h.f., and to a first approximation, aerial gain and directivity do not normally affect the minimum level of external noise available to the system to any significant degree.

2.2 Internal Noise

The unavoidable internal noise sources comprise the noise contributions from all system losses including ground losses, mismatch losses, transformer losses and feeder losses etc., normally assumed to be at reference temperature of 290 K, together with the thermal noise contribution resulting from the noise factor of the active devices in the system. This noise adds to the net external noise at any particular point in the system to produce the overall basic noise or noise floor for the system.

Additionally there is the noise contribution due to intermodulation where given a relatively small number of signals contributing i.p.s the net result, stemming mainly from the third and higher order products, can appear like a noise spectrum. This latter noise contribution is only of significance if its level exceeds the basic noise floor of the system and hence knowledge of the basic noise at each point in the system is a prime requirement.

Typical CCIR quiet noise performance curves for a number of common aerial configurations feeding a receiver or distribution amplifier through a transformer and 400 metres of 75 ohm RG164/U coaxial cable are shown in Curves B to E of Fig. 1. The noise levels shown represent the noise coupled to the device input and account for external noise plus system loss contributions. To these must be added the device noise, shown as Curve F of Fig. 1, and the effect of this noise addition on the noise levels shown can be determined using Fig. 2 which is derived as follows.

$$T_T = T_{IN} + T_R = \frac{T_A}{L} + 290\left(1 - \frac{1}{L}\right) + T_R$$

where

T_T = total noise temperature at device input in kelvins
T_{IN} = noise temperature coupled to device input (K)
T_R = noise temperature of device input itself (K)
T_A = noise temperature of aerial (K)
L = loss between aerial and device input

and

$$10 \log_{10} \frac{T_{IN}}{T_R} = \text{noise temperature over-ride (dB)}$$

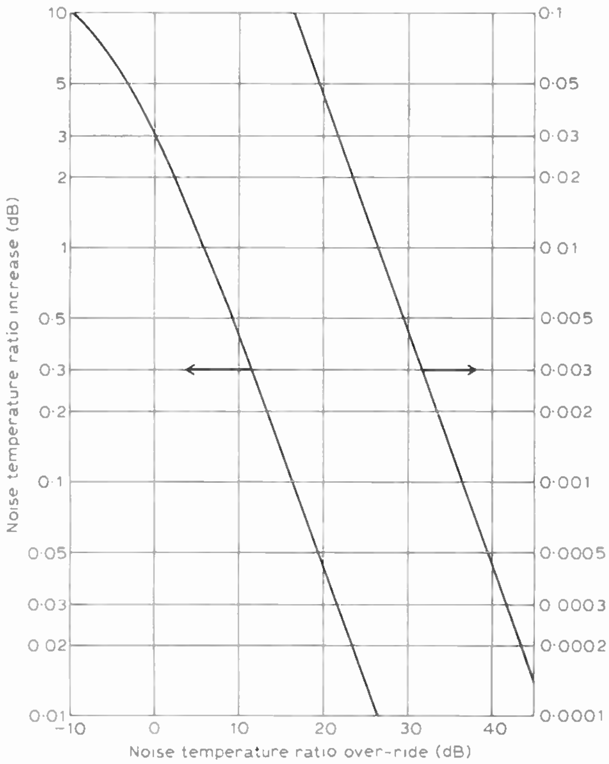


Fig. 2. Noise increase due to amplifier input noise temperature.

while

$$10 \log_{10} \frac{T_T}{T_{IN}} = \text{noise temperature increase (dB)}$$

It will be noted that a 30 dB noise over-ride produces a noise increase of less than 0·005 dB while anything in excess of a 16 dB noise over-ride gives a noise increase of less than 0·1 dB and that a 10 dB noise over-ride yields a noise increase of less than 0·5 dB.

Inspection of Fig. 1 shows that the highest noise output is obtained from a log periodic array (Curve B) since aerial efficiency is maximized while ground and mismatch losses are minimized. This is closely followed by the combined high and low band rhombics (Curve C) where the losses due to the typical combining filter account for much of the difference. The monopole performance of Curves D and E is significantly lower, particularly at the lower ends of the bands, due to the increasing magnitude of aerial inefficiency, ground and mismatch losses. The difference between Curves A and B is essentially due to the transformer and feeder losses which range from less than 2 dB at 2 MHz to nearly 6 dB at 30 MHz.

2.3 Intermodulation Levels

The levels of intermodulation produced in an active device are a function of the linearity of the device and are normally dependent on the performance of the output stage of the device since this is where signal levels are usually at a maximum. It is traditional to measure intermodulation by applying two equal level signals of similar frequency and to specify the intermodulation performance in terms of the output intercept point. This is the output level at which, by linear extension of the fundamental and intermodulation product output curves, the intermodulation output level is theoretically equal to the fundamental output level. Typical values for the output intercept point of earlier generation devices currently in use are +60 dBm second order and +36 dBm third order. Comparable values of output intercept point for modern state of the art devices are typically +96 dBm second order and +60 dBm third order. The practice of referring all levels to the output removes the gain variable from the device although gain is an important factor in system design as will be seen later.

Intermodulation products only become a problem if their levels are significant when compared to the basic noise levels existing in any system. A simple measure can be made when intermodulation levels are equal to basic noise levels and Fig. 3 shows, for each of the standards of output intercept point performance quoted above, the fundamental output levels to produce intermodulation equals noise across 75 ohm in a 3·1 kHz band. When considering system performance it is convenient to adopt this measure and introduce the concept

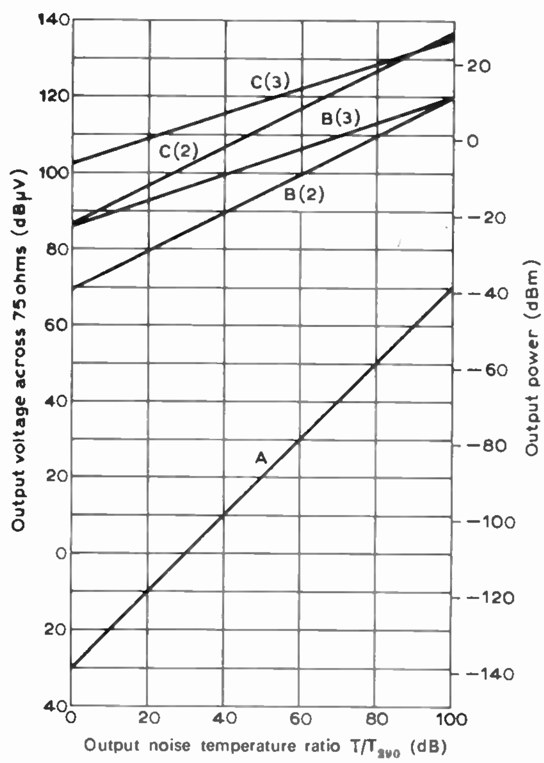


Fig. 3. Amplifier output signal levels for intermodulation equals noise.

- A Equivalent noise voltage in 3·1 kHz band.
- B (2) 2nd-order i.p.s equal noise in 3·1 kHz for +60 dBm output intercept point.
- B (3) 3rd-order i.p.s equal noise in 3·1 kHz for +36 dBm output intercept point.
- C (2) 2nd-order i.p.s equal noise in 3·1 kHz for +96 dBm output intercept point.
- C (3) 3rd-order i.p.s equal noise in 3·1 kHz for +60 dBm output intercept point.

of the signal handling capability of a system where this is defined as being the output level at any frequency which produces intermodulation levels equal to CCIR quiet noise levels at that or a harmonic frequency. This signal handling capability level may be referred to other points in the system through knowledge of the various gains and losses. Hence at system levels less than the signal handling capability level the effects of intermodulation may generally be neglected and the signal handling capability levels set an upper limit on intermodulation-free system performance.

2.4 System Signal Handling Capability

The system signal handling capability levels, as defined in Section 2.3, give a useful measure of system performance with respect to noise. It is of interest therefore to examine the factors which affect the signal handling capability of a system. The basic situation is shown graphically in Fig. 4 and it is important to note that a 36 dB improvement in second-order output intercept point produces only an 18 dB improvement in signal handling capability whilst a 24 dB improvement in third-order output intercept point produces only a 16 dB improvement in signal handling capability, i.e. an improvement of half and two-thirds respectively in accordance with the $(n-1)/n$ law deduced from Fig. 4. Thus, if desired, the signal handling capability of a system may be improved by substituting devices of higher output intercept point.

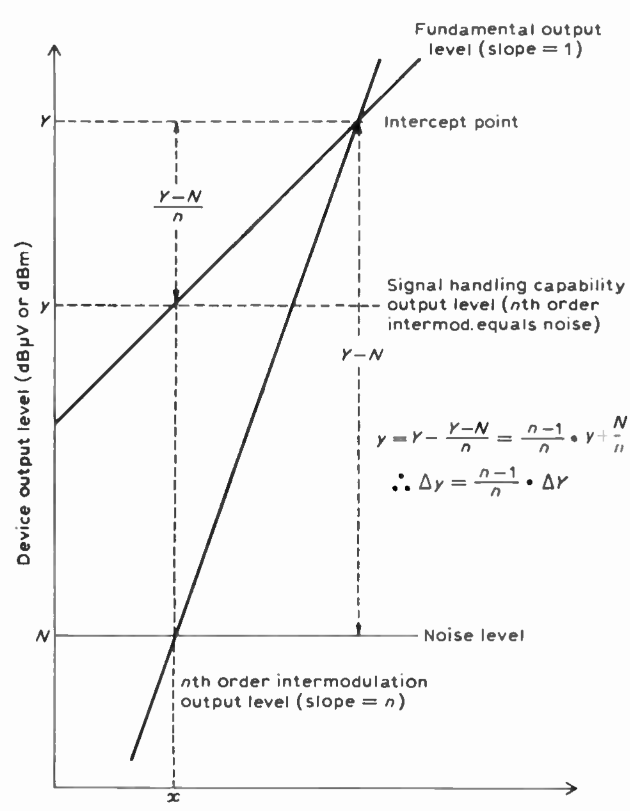


Fig. 4. Signal handling capability and the effect of output intercept point.

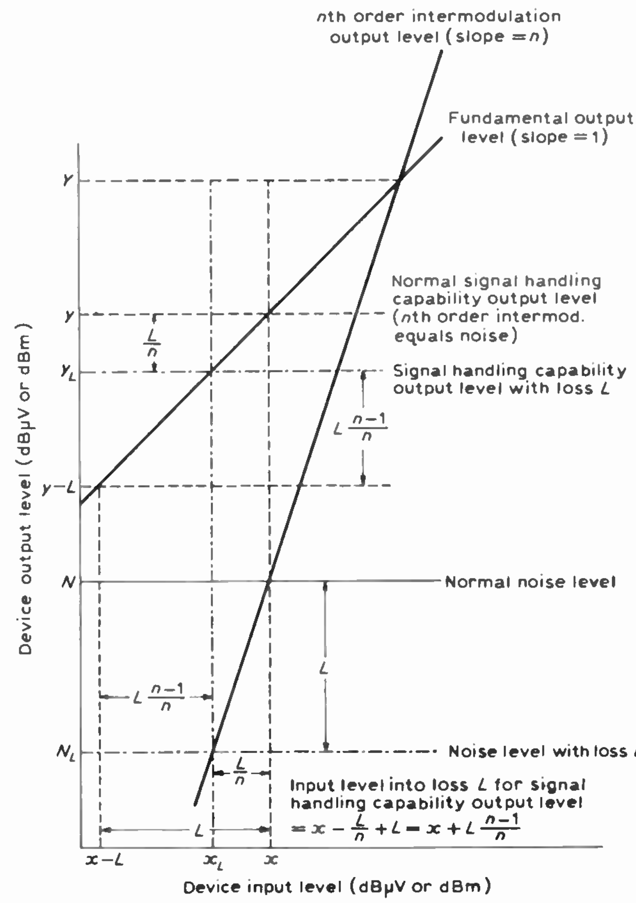


Fig. 5. Signal handling capability and the effect of inserting loss.

This is likely to prove expensive in practice and only a part of the improvement in output intercept point is realized in system performance terms.

A similar improvement in system signal handling capability can be achieved by reducing gain, or alternatively introducing loss, provided that an effective external noise over-ride can be maintained. The situation is shown graphically on Fig. 5 where it can be seen that introducing loss L, or reducing gain by L, reduces the noise level by L and hence the signal handling capability of the output is apparently reduced by L/n. However the input signal level to produce the original output conditions will also be affected by the reduction in gain or introduction of loss. The net effect is that the system input level may be increased by $L(n-1)/n$ before intermodulation equals noise under the new system conditions, i.e., the system signal handling capability is actually improved by $L(n-1)/n$. This method of improving system performance by introducing loss is clearly much less expensive than the alternative of replacing existing devices with new ones of higher performance, yet has similar effects. It must be emphasized that introducing loss is normally only acceptable in terms of system performance if there is an adequate noise over-ride, i.e. the introduction of loss reduces both signal and noise levels by much the same amount and does not produce an unacceptable degradation in signal to noise. However, the lower the noise factor of the active devices the larger the loss that can be tolerated and hence, everything else being equal, the better the signal handling capability of the system.

2.5 System Signal Handling Performance

The determination of actual intermodulation levels requires knowledge of the input levels and hence, by application of gains and losses, the output levels as well as the output intercept points of system devices. In an aerial system the terminated output signal levels may be determined relative to the incident field strength. The relationship can be considered to be dependent on an aerial factor (AF) such that

$$20 \log_{10} V_A = 20 \log_{10} E_I + AF \text{(dB)}$$

where V_A is the terminated aerial output and E_I is the incident field strength.

Now where:

P_A = power received from aerial,
G_A = aerial power gain relative to isotropic radiator,
R = matched terminating impedance,

it can be shown[5] that

$$P_A = \frac{V_A^2}{R} = \frac{E_I^2}{120\pi} \cdot \frac{\lambda^2}{4\pi} \cdot G_A$$

Therefore

$$20 \log_{10} V_A = 20 \log_{10} E_I + 12{\cdot}8 + 10 \log_{10} R + \\ + 10 \log_{10} G_A - 20 \log_{10} f \text{ (MHz)}$$

and

$$AF \text{(dB)} = 12{\cdot}8 + 10 \log_{10} R + \\ + 10 \log_{10} G_A - 20 \log_{10} f \text{ (MHz)}$$

The output levels for an ideal zero-gain, loss-free aerial terminated in 75 ohms are shown as Curve A of Fig. 6. The terminated feeder output levels for the same aerial configurations as considered in the preceding Section are also shown on Fig. 6 as Curves B to E. Curve B assumes a log periodic array of constant 10 dB gain, whilst Curve C assumes rhombics with an average 15 dB gain but with the actual gain varying between 10 and 20 dB depending on frequency. Curves D and E assume a constant 4·8 dB gain for each type of monopole. The effect of aerial gain is to increase the signal levels, and hence and to a greater extent any i.p.s levels, for those signals in the forward direction whilst reducing signal levels and i.p.s for those signals arriving from other directions. The net effect is likely to be that whilst the i.p.s noise spectrum effects from a high-gain, highly directive aerial system may well be comparable with those from a low-gain aerial due to the multiplicative action, there are likely to be rather more discrete high-level i.p.s in the higher gain case.

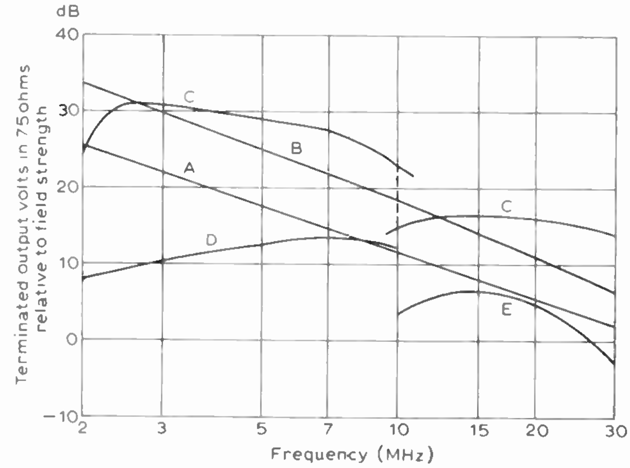


Fig. 6. Aerial feeder output signal levels relative to field strength.
A Zero gain lossless aerial.
B Log periodic array feeder output.
C Combined high and low band rhombic feeder output.
D 12 m monopole feeder output.
E 6 m monopole feeder output.

The system signal handling capability of the typical combined high and low band rhombic aerial system, referred back to the level of the incident field strength to match CCIR quiet noise conditions in a 3·1 kHz band, feeding a 30 dB gain distribution amplifier of 7 dB noise factor and +60 and +36 dBm second- and third-order intercept points respectively is shown in Fig. 7. Curve A represents the system CCIR quiet noise level in terms of equivalent incident field strength with Curve B showing the system signal handling capability, again in terms of incident field strength, for second-order i.p.s equal to CCIR quiet noise. Curves C and D show the third-order system signal handling capability with Curve C representing the in-band i.p.s case. It will be appreciated that in arriving at these signal handling capability figures account had to be taken of signal levels at frequency f producing intermodulation that equalled noise at frequency nf thus adding a frequency dimension to this property. It will also be noted that, under CCIR quiet noise conditions, incident field strengths of significantly less than 1 mV/m will exceed the system signal handling capabilities and produce intermodulation levels in excess of noise thus inhibiting system performance. In practice such a system, which is typical of many in use today, can be expected to suffer from significant intermodulation problems. The lower noise levels at the high frequencies together with the relatively higher signal levels at the lower frequencies tend to compound the problem in that the system is more sensitive to the $f+f$ and $2f+f$ i.p.s with the system signal handling capability correspondingly reduced as can be seen from Fig. 7. Changing the amplifier for one of improved output intercept point performance will help the situation generally but is unlikely to improve the lower frequency signal handling capability to significantly better than 1 mV/m.

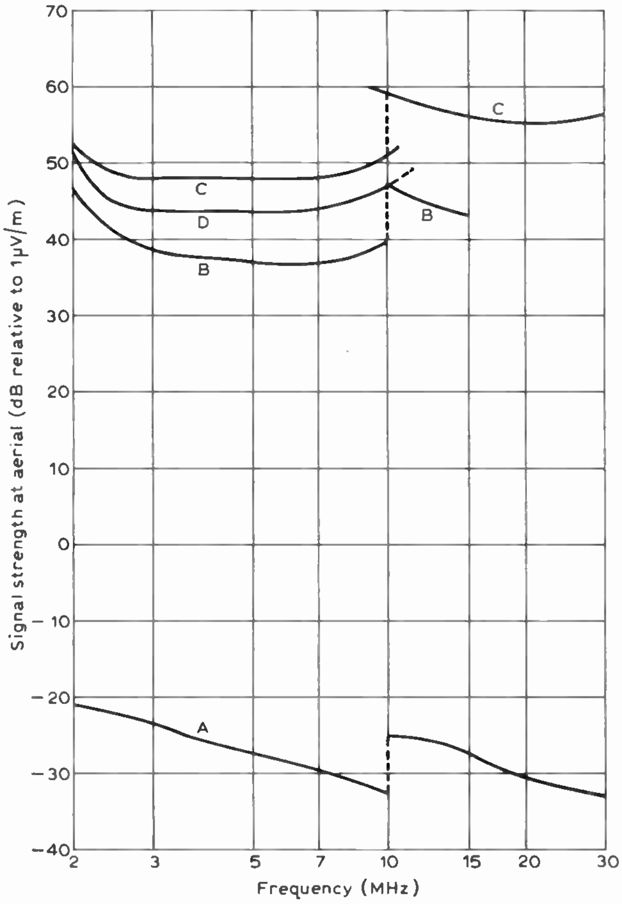


Fig. 7. Rhombic aerial system signal handling capability levels. Typical combined high and low band rhombic aerials feeding into distribution amplifier of 30 dB gain and 7 dB noise factor.

- A Equivalent signal strength to produce system CCIR quiet noise level.
- B Signal strength to produce 2nd-order $(f+f)$ intermod. equals CCIR quiet noise at $2f$ (+60 dBm intercept point).
- C Signal strength to produce 3rd-order $(2f-f)$ intermod. equals CCIR quiet noise at f (+36 dBm intercept point).
- D Signal strength to produce 3rd-order $(2f+f)$ intermod. equals CCIR quiet noise at $3f$ (+36 dBm intercept point).

3 Noise Equalization

3.1 The Concept

The objective of noise equalization is to reduce basic noise levels to the minimum necessary to provide effective noise over-ride at each stage, i.e. external noise limiting throughout the system. At one and the same time signal levels are reduced to the minimum levels compatible with retaining basic signal to noise performance. This in turn optimizes the signal handling capability of the system. A suggested overall design objective is that noise levels in the system based on CCIR quiet noise at the input should not increase by more than 0·5 dB relative to the signal levels, i.e. basic signal to CCIR quiet noise at the aerial should not be deteriorated by more than 0·5 dB. From Fig. 2 it can be seen that an overall CCIR quiet noise over-ride of slightly more than 9 dB would meet the case. When more than one stage is involved the noise over-ride at each stage will probably vary between 12 and 15 dB in order to produce an overall 9 to 10 dB over-ride effect.

3.2 The Ideal Case

It can be seen from Fig. 1 that to equalize the CCIR quiet noise of Curve A down to a uniform noise temperature ratio of 16 dB in order to achieve a 10 dB noise over-ride at the input to a 7 dB noise factor amplifier, Curve F, would require an equalization loss of 29 dB at 2 MHz falling to 2 dB at 30 MHz. The introduction of such an equalizer would reduce signal to CCIR quiet noise by less than 0·5 dB but would have a profound effect on the signal handling capability of the system as shown in Fig. 8. The improvement in signal handling capability for the in-band $2f-f$ third-order i.p.s equals noise follows the established $L(n-1)/n$ law and is shown as Curve C. The improvement in signal handling capability for the out-of-band $f+f$ second-order and $2f+f$ third-order i.p.s is more pronounced due to the reduction in equalization loss, and hence a lesser effect on noise level, with increasing frequency. The improvement in these out-of-band cases is shown in Curves B and D of Fig. 8. It can be expressed in terms of the loss at frequencies f and nf such that the improvement in signal handling capability

$$\Delta SH_f = L_f - \frac{L_{nf}}{n} = L_f\frac{n-1}{n} + \frac{\Delta L}{n}$$

where

$$\Delta L = L_f - L_{nf}$$

Atmospheric noise levels are lower at the higher frequencies and in the practical situation signal levels tend to be higher at the lower frequencies[1,2] and in these circumstances the enhanced improvement in signal handling capability for the out-of-band i.p.s has considerable benefit. This benefit is best appreciated in terms of the effective reduction in existing intermodula-

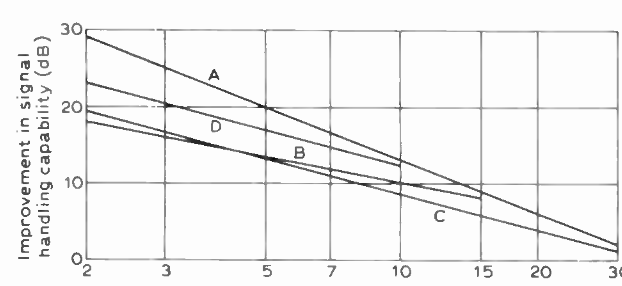


Fig. 8. Theoretical improvement in signal handling capability levels with noise equalization.

- A Equalization loss CCIR quiet noise to 16 dB noise temperature ratio.
- B Improvement in signal handling capability for 2nd-order $(f+f)$ intermod. equals noise at $2f$.
- C Improvement in signal handling capability for 3rd-order $(2f-f)$ intermod. equals noise at f.
- D Improvement in signal handling capability for 3rd-order $(2f+f)$ intermod. equals noise at $3f$.

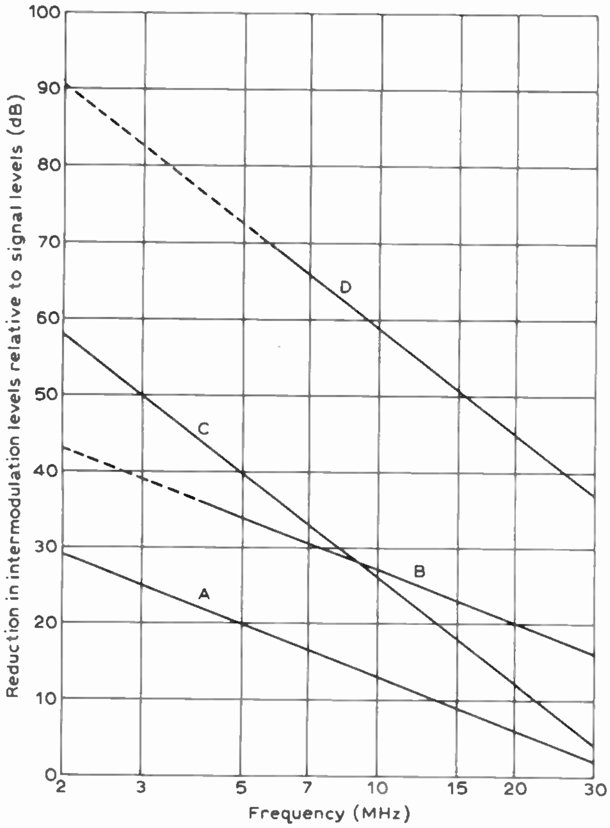


Fig. 9. Theoretical reduction in intermodulation levels with noise equalization.

- A Equalization loss CCIR quiet noise to 16 dB noise temperature ratio.
- B Reduction in 2nd-order $(\frac{1}{2}f+\frac{1}{2}f)$ intermodulation levels relative to f signal levels.
- C Reduction in 3rd-order $(2f-f)$ intermodulation levels relative to f signal levels.
- D Reduction in 3rd-order $(2\cdot\frac{1}{3}f+\frac{1}{3}f)$ intermodulation levels relative to f signal levels.

tion levels, relative to signal and noise levels at frequency f, that results from the noise equalization of a system. The value of this effective reduction in existing nth order intermodulation levels is n times the improvement in signal handling capability at frequency f/n for out-of-band i.p.s and n times the improvement in signal handling capability at frequency f for the in-band case. The effects are shown in Fig. 9 where Curve A again represents the theoretical noise equalizer loss characteristic for the CCIR quiet noise condition, Curve B the effective reduction in the level of the $\frac{1}{2}f+\frac{1}{2}f$ second-order i.p.s, Curve C the effective reduction in the levels of the $2f-f$ third-order i.p.s and Curve D the effective reduction in the levels of the $2\cdot\frac{1}{3}f+\frac{1}{3}f$ i.p.s. Whether or not the full benefit of these reductions in intermodulation levels can be observed in practice will depend on their level relative to the system noise floor.

The interesting and important aspect of the noise equalization concept is that at the higher frequencies, i.e. where out-of-band i.p.s are significant, intermodulation levels are reduced to a greater extent than would result from comparable increases in output intercept point or reduction in amplifier gain. Also since noise is constant with frequency, after equalization the signal handling capability for both in-band and out-of-band i.p.s of the same order will be the same for both cases. Given that the same output performance can be maintained, i.e. output intercept point constant, then the lower the noise figure of the receiver or amplifier the larger the noise equalization loss that can be introduced and the greater the benefit to the system in reducing intermodulation levels and improving signal handling capabilities.

The effects of noise equalization on the value of system signal handling capability under CCIR quiet noise conditions can be deduced in more general terms by assuming an approximate value for the aerial noise temperature T_A, relative to 290 K, such that

$$T_A\,(\text{dB}) = 50-20\log_{10} f\,(\text{MHz}).$$

The total loss L_E (dB) between aerial and amplifier or receiver input under noise equalized conditions is then determined by the amplifier or receiver noise temperature T_R, relative to 290 K, and the required noise over-ride N_R (dB) such that

$$\begin{aligned} L_E\,(\text{dB}) &= T_A\,(\text{dB})-T_R\,(\text{dB})-N_R\,(\text{dB}) \\ &= 50-20\log_{10} f\,(\text{MHz})-T_R\,(\text{dB})-N_R\,(\text{dB}) \end{aligned}$$

Under these conditions the noise temperature input level, relative to 290 K, at the amplifier or receiver is constant at $T_R\,(\text{dB})+N_R\,(\text{dB})$ and this expression can be converted to noise voltage terms by applying Boltzmann's constant, the bandwidth B (Hz) and the matched terminating impedance R. Thus the equalized noise voltage V_{NE} at the input to the amplifier or receiver is given by

$$\begin{aligned} 20\log_{10} V_{NE} &= T_R\,(\text{dB})+N_R\,(\text{dB})+ \\ &\quad +10\log_{10} B+10\log_{10} R-204. \end{aligned}$$

Under these same conditions the input signal level V_{AE} at the amplifier or receiver is given by

$$\begin{aligned} 20\log_{10} V_{AE} &= 20\log_{10} V_A-L_E\,(\text{dB}) \\ &= 20\log_{10} E_1+AF\,(\text{dB})-L_E\,(\text{dB}) \end{aligned}$$

Now where

- P_{IP} = dBm value of amplifier or receiver output intercept point for nth order products,
- V_{IP} = amplifier or receiver output voltage level corresponding to P_{IP},
- G_R = gain of amplifier or receiver,
- V_{SH} = amplifier or receiver output signal handling capability level for nth order products under quiet conditions,
- E_{SH} = incident field strength corresponding to signal handling capability level under quiet conditions.

Then from Fig. 4 and for a matched system of constant impedance R

$$20\log_{10}V_{SH}=20\log_{10}V_{IP}-\frac{1}{n}[20\log_{10}V_{IP}-$$

$$-10\log_{10}G_R-20\log_{10}V_{NE}]$$

$$=\frac{n-1}{n}[20\log_{10}V_{IP}]+\frac{1}{n}\times$$

$$\times[10\log_{10}G_R+20\log_{10}V_{NE}]$$

$$=\frac{n-1}{n}[P_{IP}(\text{dBm})+10\log_{10}R-30]+$$

$$+\frac{1}{n}[10\log_{10}G_R+20\log_{10}V_{NE}]$$

$$=10\log_{10}G_R+20\log_{10}V_{AE}$$

when

$$E_I=E_{SH}.$$

Hence substituting for $20\log_{10}V_{NE}$ and $20\log_{10}V_{AE}$ and simplifying

$$20\log_{10}E_{SH}=\frac{n-1}{n}[P_{IP}(\text{dBm})-T_R(\text{dB})-$$

$$-N_R(\text{dB})-10\log_{10}G_R]+$$

$$+\frac{1}{n}[10\log_{10}B-174]-$$

$$-10\log_{10}G_A+7{\cdot}2.$$

This analysis demonstrates very clearly that improvements in signal handling capability require high output intercept point, low device noise temperature and low device gain although the effect of changes in all these factors are subject to the $(n-1)/n$ law. Increasing the noise levels or noise bandwidth also improves the signal handling capability but these aspects are subject to a $1/n$ law. Hence an n dB increase in aerial noise temperature above CCIR quiet noise conditions produces only a 1 dB increase in actual signal handling capability. At the same time the signal handling capability is inversely related to the aerial gain and otherwise independent of frequency.

The approximate values of signal handling capability, referred to the incident field strength, for a noise equalized zero-gain lossless aerial feeding the 30 dB gain, 7 dB noise factor devices assumed in Fig. 3 are shown in Fig. 10.

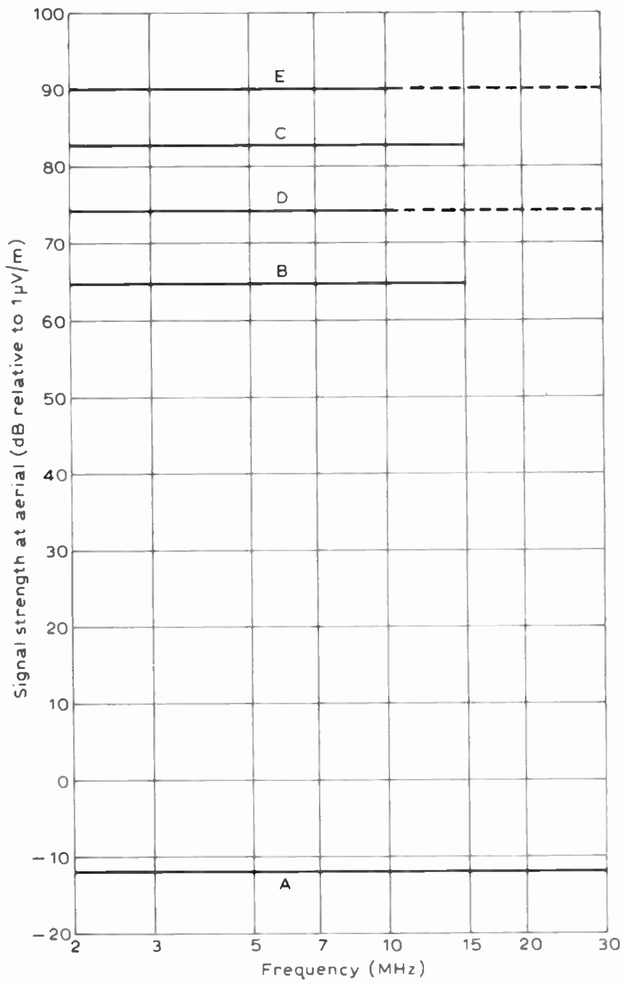


Fig. 10. Theoretical signal handling capability levels for noise equalized zero gain aerial system.

Noise equalized zero gain loss free aerial feeding into distribution amplifier of 30 dB gain and 7 dB noise factor with 10 dB noise over-ride.

A Equivalent signal strength to produce system CCIR quiet noise level.

B Signal strength to produce 2nd-order $(f+f)$ intermod. equals CCIR quiet noise at $2f$ (+60 dBm intercept point).

C Signal strength to produce 2nd-order $(f+f)$ intermod. equals CCIR quiet noise at $2f$ (+96 dBm intercept point).

D Signal strength to produce 3rd-order $(2f-f$ and $2f+f)$ intermod. equals CCIR quiet noise at f and $3f$ (+36 dBm intercept point).

E Signal strength to produce 3rd-order $(2f-f$ and $2f+f)$ intermod. equals CCIR quiet noise at f and $3f$ (+60 dBm intercept point).

3.3 Practical Implementation

The equalization loss proposed in Curve A of Figs. 8 and 9 is only applicable in certain cases if the noise equalization is introduced directly at the aerial output and this implies the use of a remote pre-amplifier preceding the aerial feeder. Such an arrangement tends to be unusual in the h.f. band, although it is often normal practice at higher frequencies, but as will be seen later it does enable maximum performance to be realized particularly at the high frequency end of the band. In practice the use of active devices is normally constrained to the output of the aerial feeder in order to avoid the problems of remote siting, protection and maintenance. This means that when noise equalization is applied to the input of the first active device the value of equalization used has to take account of the line losses etc. These losses being low at low frequencies and higher at the high frequencies, they tend to steepen the slope of the noise output curve, thus enhancing the benefits to be gained from noise equalization, although these benefits tend to be marginally reduced at the higher frequencies by some loss in overall noise performance due to inadequate noise override.

When introducing noise equalization it is important to keep in perspective the fundamentally imprecise and variable noise conditions that are encountered. Thus whilst the values of noise equalization that precisely meet a given situation may be difficult, if not impossible, to realize exactly, values within a dB or so of the ideal will probably suffice. The use of standard constant-resistance bridged-T equalizer design can be expected to satisfy most cases given some adjustment of the available variables. The author has found the assistance of a small programmable calculator to be invaluable in this respect but a comprehensive set of design curves is also available.[6]

To illustrate the case and assess the practical effects on system performance noise equalization has been applied to the combined high and low band rhombics of Figs. 1, 6 and 7. Curve E of Fig. 11 shows the value of equalization required to give a 12 dB CCIR quiet noise over-ride at the input to a distribution amplifier of 7 dB noise factor. It will be noted that the loss falls below zero from 17·5 MHz upwards so that at 30 MHz a gain of some 7·5 dB is required. However, even at 30 MHz the net CCIR quiet noise over-ride is still some 4·5 dB although from Fig. 2 it can be seen that this

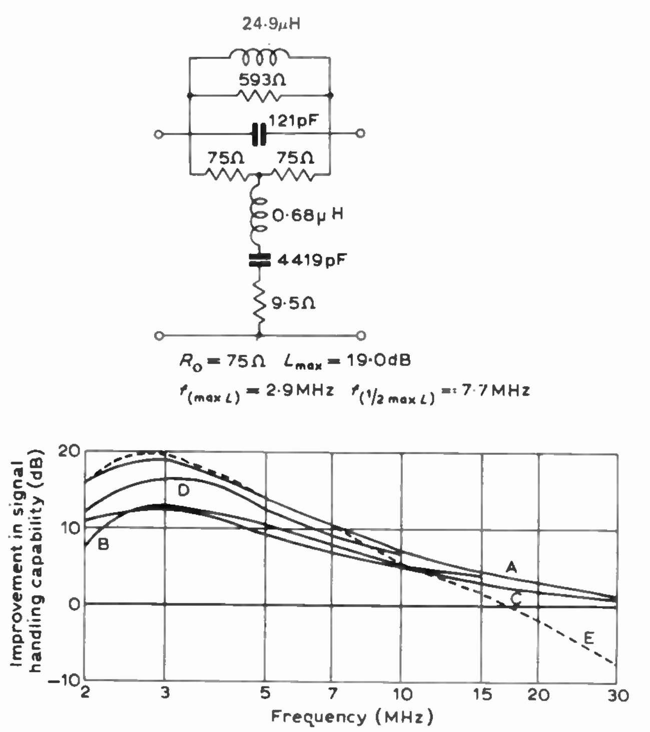


Fig. 11. Improvement in signal handling capability levels for a noise equalized rhombic aerial system.
A Actual equalization loss (circuit shown).
B Improvement in signal handling capability for 2nd-order $(f+f)$ intermod. equals noise at $2f$.
C Improvement in signal handling capability for 3rd-order $(2f-f)$ intermod. equals noise at f.
D Improvement in signal handling capability for 3rd-order $(2f+f)$ intermod. equals noise at $3f$.
E Equalization loss required to yield 12 dB noise over-ride (CCIR quiet noise to 7 dB noise factor).

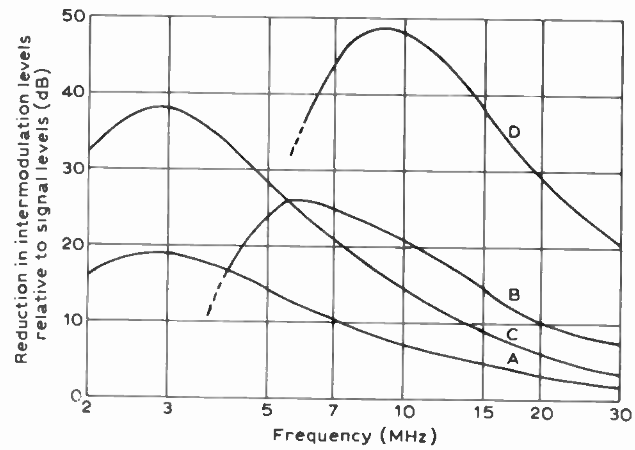


Fig. 12. Reduction in intermodulation levels for a noise equalized rhombic aerial system.
A Actual equalization loss in dB.
B Reduction in 2nd-order $(\frac{1}{2}f+\frac{1}{2}f)$ intermodulation levels relative to f signal levels.
C Reduction in 3rd-order $(2f-f)$ intermodulation levels relative to f signal levels.
D Reduction in 3rd-order $(2\cdot\frac{1}{2}f+\frac{1}{2}f)$ intermodulation levels relative to f signal levels.

equates to a loss in system signal to CCIR quiet noise of some 1·3 dB. The practical noise equalizer loss is shown as Curve A of Fig. 11. Also shown is the noise equalizer circuit and basic design parameters together with the improvements in signal handling capability to be expected —Curves B, C and D. It will be seen that Curves A and E of Fig. 11 are almost identical from 2 MHz to 7·5 MHz and that at 30 MHz Curve A shows 1·5 dB loss rather than the desired 7·5 dB gain. This means that the CCIR quiet noise over-ride at 30 MHz is reduced to only 3 dB with the result, from Fig. 2, that signal to CCIR quiet noise falls by 1·75 dB, i.e. the introduction of noise equalization has degraded the actual system signal to CCIR quiet noise by less than 0·5 dB. This can be claimed to meet the noise equalization design objective although it is still possible to improve the higher frequency performance by a dB or so by the use of a suitable pre-amplifier should this be important.

The effective reduction in existing intermodulation levels due to the noise equalization of Curve A is shown in Curves B to D of Fig. 12. Again the very pronounced reduction in third-order out-of-band i.p.s, Curve D, is worthy of comment and attention. The signal handling capabilities, in terms of incident field strength, of the noise equalized rhombic systems assumed are shown in Fig. 13 and this should be compared with Fig. 7. Firstly, it will be noted from Curves A that the system CCIR quiet noise level is essentially unchanged although a marginally higher field strength is required in the noise equalized case to compensate for the 0·5 dB increase in noise level. Curves B, C and D all reflect the higher field strength values to produce i.p.s equal to CCIR quiet noise in the noise equalized case since the intermodulation free dynamic range of the system has in-

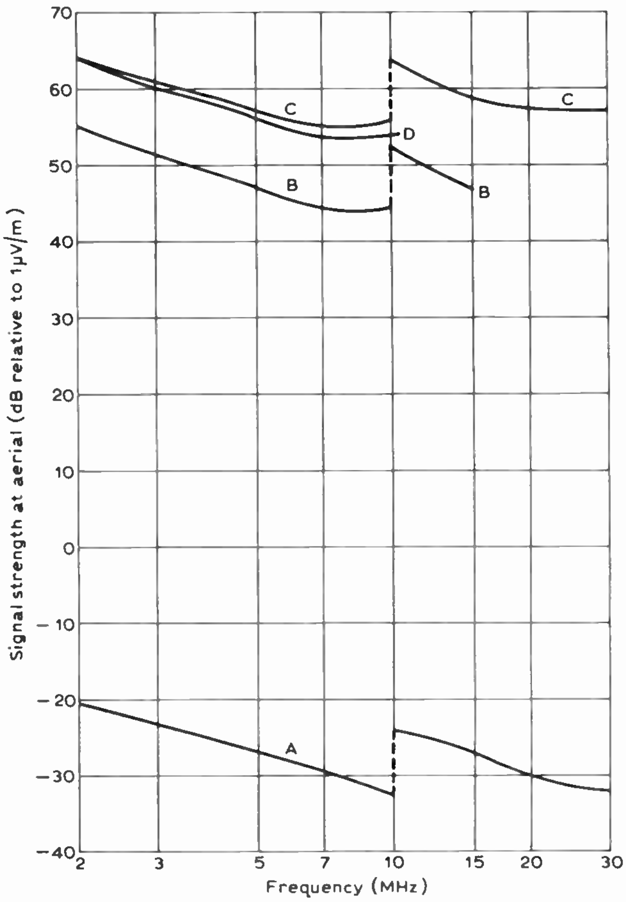


Fig. 13. Noise equalized rhombic aerial system signal handling capability levels.

Typical noise equalized combined high and low band rhombic aerials feeding into distribution amplifier of 30 dB gain and 7 dB noise factor.

A Equivalent signal strength to produce system CCIR quiet noise level.

B Signal strength to produce 2nd-order $(f+f)$ intermod. equals CCIR quiet noise at $2f$ (+60 dBm intercept point).

C Signal strength to produce 3rd-order $(2f-f)$ intermod. equals CCIR quiet noise at f (+36 dBm intercept point).

D Signal strength to produce 3rd-order $(2f+f)$ intermod. equals CCIR quiet noise at $3f$ (+36 dBm intercept point).

creased by the improvement in the signal handling capability. It is particularly interesting to note that, as forecast, Curves C and D virtually coincide in the noise equalized case showing that the signal handling capability for in-band and out-of-band third-order i.p.s is now essentially the same thus removing the frequency dimension. Further, this third-order signal handling capability has now been increased to approximately the 1 mV/m level and although intermodulation problems may remain they will assume similar significance with input signal levels regardless of frequency. Similarly changing the amplifier for one of higher ouput intercept point is now likely to have more beneficial effects and could virtually eliminate intermodulation problems. It is important to note that the essential difference between the curves of Fig. 13 and those of Fig. 10 is due to the gain of the rhombic aerials.

3.4 Observed Results

The author has now designed and fitted noise equalization in a considerable number of existing aerial systems including typical rhombic systems of the type outlined and various log periodic arrays as well as circularly disposed arrays of monopoles.[7] Assessing the results of fitting the noise equalizers is very difficult on a quantitative basis due to the intrinsically variable nature of both noise and signal levels although there is no doubt room here for longer term statistical studies.

The first attempt to make quantitative assessments of the effects of noise equalization on a typical rhombic aerial system involved measuring the apparent changes in system noise level due to equalization. The equalization loss was then subtracted from these apparent noise level changes in order that the results could be presented as potential improvements in available signal to noise ratio. These results are shown for various times of the day in Fig. 14 and generally confirm the expectations particularly insofar as the 10 to 20 MHz range is concerned. Based on this evidence the decision was taken to introduce noise equalization into practical use. The emphasis was placed on allowing the operator to convince himself of the desirability of the technique rather than to present it as some instant palliative thus inevitably provoking

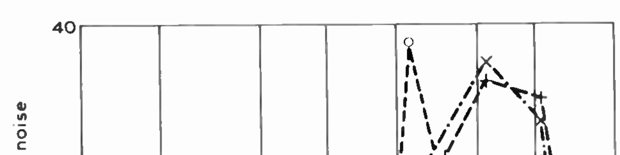


Fig. 14. Improvement in rhombic aerial system signal-to-noise ratio with noise equalization.

Typical combined high and low band rhombic feeding 7 dB noise factor amplifier of 30 dB gain and +60 and +36 dBm 2nd- and 3rd-order output intercept points.

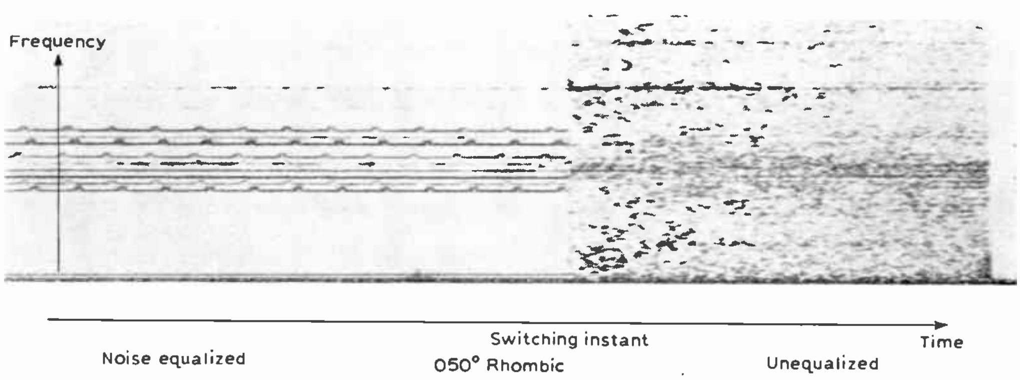


Fig. 15. Sample sonograph of noise equalizing a rhombic aerial system. 12·436 MHz at 20.00 hours on 23rd March 1976.

immediate cynicism. The method adopted has been to provide the operator with a means of remotely switching the noise equalizer in and out of his aerial circuit leaving him to decide which mode of operation is preferable.

The first reaction to providing a noise equalization capability tends to be a suspicion that introducing loss into the system may produce loss of signals. There is no evidence to date to suggest that this loss of signal has occurred although of course it is theoretically possible on those rare occasions when external noise falls significantly below CCIR quiet noise levels. In satisfying himself on this score the operator inevitably finds some apparent signals that are themselves intermodulation products and the effect here is often dramatic in that quite high level signals are lost completely when the noise equalizer is switched into circuit. An equally dramatic improvement often results when observing low-level signals, particularly in the 8 to 20 MHz band. What was barely readable, or indeed unreadable, is found to be comfortably readable under noise equalized conditions. Naturally insofar as high level signals are concerned the effects are less noticeable but even here the operators tend to comment on an improvement in signal quality and crispness under noise equalized conditions. To some extent the effect of noise equalization can in this respect be likened to the difference between listening to high and low fidelity audio systems. Certainly the operators appear to find the noise equalized signal environment more comfortable to work with and end up by keeping the noise equalizer permanently in circuit.

Further attempts to assess the practical results of noise equalization were made by the operating staff using the Sonograph. This machine enables a short sample, just over 2 seconds in duration, of an audio signal to be analysed into its various frequency components over the time period. A typical effect of noise equalization on the receiver audio output from a rhombic aerial system can be seen from Fig. 15. This shows a Sonograph of receiver output around the noise equalizer switching instant for a received signal at approximately 12·5 MHz. It will be noted that in the absence of noise equalization a wide spectrum of noise components virtually obscures the wanted signal and that most of this noise is effectively removed by noise equalization.

4 Conclusions

The noise equalization concept as described, using CCIR quiet noise or any other suitable noise levels as a reference, can be applied to existing h.f. receiving systems by performing a few relatively simple calculations allied with noise and loss measurements as appropriate. The cost of implementation is low with minimal disturbance to the system and the benefits obtainable are particularly marked in the case of wideband systems using aerial amplifiers and distribution arrangements providing multiple feeds to the receivers themselves. The effect of reducing signal levels to the minimum compatible with retaining basic signal to noise performance enhances the system performance by improving its signal handling capability, as defined earlier, thereby reducing intermodulation problems very considerably. The operator still has full use of his receiver aerial attenuator which is likely to be more effective in reducing problems due to strong signals as a result of the noise equalization prior to the receiver.

Improvements to existing h.f. receiving systems apart, the concept is seen as a basic design tool in new system design which, given some knowledge of the signal environment, will permit more realistic specification of the required gain, noise factor and output intercept point of the various active devices with some hope that the required performance can be achieved at an acceptable cost. In this respect noise equalization has the additional benefit that it reduces the noise performance of all aerial configurations to a common base and enables true comparison of the practical performance of different aerial configurations to be made on a system basis.

The application of the noise equalization and signal handling capability concepts in new system design causes greater emphasis to be placed on features such as noise factor, gain per stage, intervening losses and output intercept point than has been traditional at h.f. although these same features have long been regarded as crucial in the v.h.f. and microwave fields.

5 Acknowledgments

The author is particularly grateful to the many colleagues whose efforts in realizing the hardware and evaluating the results have made this paper possible. The paper is published with the consent of the Director GCHQ and with the permission of the Controller of Her Majesty's Stationery Office.

6 References

7 Winn, R. F. E., 'Operational aspects of h.f. receiver design', *Electronics and Power*, **20**, pp. 440–4, June 1974.

2 Sosin, B. M., 'H.f. communications receiver performance requirements and realization', *The Radio and Electronic Engineer*, **41**, pp. 321–9, July 1971.

3 CCIR Report No. 322, 'World Distribution and Characteristics of Atmospheric Radio Noise', International Telecommunications Union, 1964.

4 Bradley, P. A. and Clark, C., 'Atmospheric radio noise and signals received on directional aerials at high frequencies', *Proc. Instn Elect. Engrs*, **111**, pp. 1534–40, September 1964.

5 Glazier, E. V. D. and Lamont, H. R. L., 'Transmission and Propagation', 'The Services Text Book for Radio, Vo. 5, Chap. 9 (HMSO, London, 1957).

6 Academy of Motion Picture Arts and Sciences Research Council, 'Motion Picture Sound Engineering' (Van Nostrand, New York, 1938), Chaps. XVI and XVII.

7 Starbuck, J. T., 'A multiple-beam high-frequency receiving aerial system', *The Radio and Electronic Engineer*, **37**, pp. 229–35, April 1969.

Manuscript first received by the Institution on 18th October 1976 and in final form on 6th January 1977. (Paper No. 1764/Com. 146.)

The Author

Derek Bunday (Fellow 1970) joined the Post Office Telephones as a Youth in Training in 1946 and subsequently moved to the Research Station at Dollis Hill in 1953. Part-time studies at Southampton University and Willesden Technical College led to the award of a City and Guilds Full Technological Certificate in telecommunications in 1956 and a Higher National Certificate plus endorsements in 1957, as well as an IEE prize and a City and Guilds Silver Medal. Whilst at Dollis Hill he was engaged on projects as diverse as the development of branching filter networks for 4 GHz television links and the testing of transatlantic submarine cables. Moving to Government Communications Headquarters in 1960, the next six years were spent working on a wide variety of microwave system projects. Then came a spell of two years with the European Space Research Organization's Control Centre Department where he became Head of the Equipment Division. From November 1967 he carried additional responsibilities as Operations Coordinator for the first ESRO satellite which was successfully launched and operated from May 1968. Returning to GCHQ later in 1968 he has filled a number of management posts and is currently Head of the Special Projects Branch of the Engineering Division with a wide range of interests ranging from h.f. d.f. to satellite communications.

UDC 621.391.883.2: 621.376: 621.396.96

Radar signal loss resulting from sub-optimal phase quadrature processing

G. C. BAGLEY, C.Eng., M.I.E.R.E.*

SUMMARY

The effective difference in sensitivity between two types of detection threshold is derived. They are a single threshold placed on the envelope of a complex signal, and a pair of thresholds placed on the modulus of each of the resolved quadrature components. It is shown that the loss which results for the sub-optimal case is acceptably small.

Notation

$Q(\alpha, \beta)$	Marcum's Q function
α	normalized signal-to-noise voltage ratio
β	normalized threshold voltage
α'	required increased value of α
β'	effective increased value of β
t	threshold modulus for each quadrature component
$A(t)$	normal probability integral
$I_0(x)$ and $I_1(x)$	modified Bessel functions

1 Introduction

A two-path receiver[1] is one in which signals are resolved into two quadrature components and processed at base-band. In its optimal form, the two components are eventually combined as the square-root of the sum of the squares in order to obtain the signal envelope. The magnitude of this is independent of signal phase angle and may be compared to a chosen detection threshold.

An alternative and perhaps simpler method is to place a separate threshold on the modulus of each quadrature component. It has been shown[2] that this method is sub-optimal since it favours some signal phase-angles more than others.

This paper is concerned with the value of the equivalent loss of signal-to-noise ratio incurred.

* *Royal Signals and Radar Establishment, St. Andrew's Road, Great Malvern, Worcestershire WR14 3PS.*

2 False Alarm Probability

For fairness of comparison, the false-alarm probability of the two types of detector must be made equal.

The receiver noise in both cases may be resolved into two uncorrelated quadrature components each having a Gaussian distribution and together they form a bi-variate Gaussian distribution given by:

$$P_n(x, y) = \frac{1}{2\pi} \exp\left(-\frac{x^2+y^2}{2}\right) \tag{1}$$

This is equivalent to eqn. (1) of Ref. 2 but with the variance set to unity.

The modulus of the noise has a Rayleigh distribution given by:

$$P_n(r) = r \exp\left(-\frac{r^2}{2}\right) \tag{2}$$

The probability of this exceeding a threshold β is given by:

$$P_{fa} = \exp\left(-\frac{\beta^2}{2}\right) \tag{3}$$

In the case of the separate thresholds applied to the modulus of each separate component, we obtain what we will call a 'square' threshold rather than the 'circular' one of constant radius. (See Fig. 1.)

The probability of a false alarm is now the probability that the tip of the noise vector does not lie within the square defined by $-t < x < t$ and $-t < y < t$. That is

$$P_{fa} = 1 - (A(t))^2 \tag{4}$$

where $A(x)$ is the normal probability integral given by:

$$A(x) = \frac{1}{\sqrt{(2\pi)}} \int_{-x}^{x} \exp\left(-\frac{t^2}{2}\right) dt$$

The notation is that of Ref. 3 (page 931). Tables are available in Ref. 5. For $P_{fa} = 0{\cdot}1$

we have $t = 1{\cdot}948$ and $\beta = 2{\cdot}146$

When both types of threshold have been set to equal

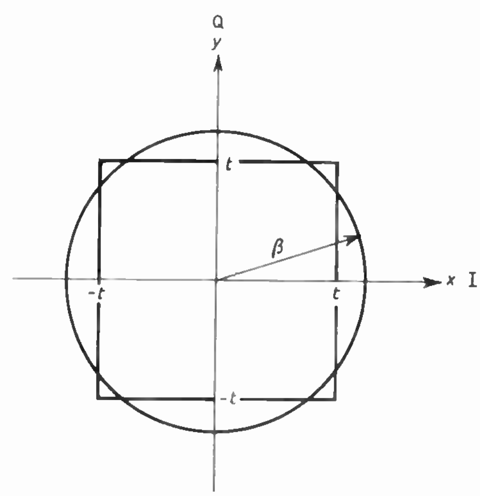


Fig. 1. Square and circular detection thresholds.

probability of false alarm we may now compare the different signal levels necessary to give the same detection probability.

3 Signal Detection

The reason for the difference of performance is that although the circular threshold treats all signals of the same amplitude with equal favour, the square threshold is more lenient to signals whose phase lies near to the axes than those lying in the middle of each quadrant. The latter have the corners of the square to overcome.

The probability of a signal of amplitude α overcoming the threshold β in the case of the circular threshold is given by Marcum's Q function[4] namely:

$$Q(\alpha, \beta) = \int_{\beta}^{\infty} r \exp\left(-\frac{r^2+\alpha^2}{2}\right) I_0(\alpha r)\, dr \qquad (5)$$

Here α is $\sqrt{(2E/N)}$, the normalized signal-to-noise ratio.

In the case of the square threshold we consider the bivariate Gaussian form used in (1) but with non-zero mean values:

$$P_{s+n}(x, y) = \frac{1}{2\pi} \exp\left(-\frac{(x-\alpha\cos\theta)^2+(y-\alpha\sin\theta)^2}{2}\right) \qquad (6)$$

This should be convolved with $p(\theta) = 1/2\pi$ to allow for the unknown, equiprobable signal phase and then compared to the square threshold.

The detection probability P_d would be the sum of all the probability lying outside the square. Both the convolution and the integral would be tedious to perform, so we make an approximation and replace the square threshold by its average circular value for all phase angles, and continue to use (5). We are replacing $\overline{Q(\alpha, r(\theta))}$ by $Q(\alpha, \bar{r})$ where r is the 'radius' of the square threshold.

The validity of this approximation has not been studied over a wide range of signal-to-noise ratios, but we note that it is exact for $\alpha = 0$ and that it will therefore be good for weak signals.

The equivalent circular threshold β' is given by

$$\beta' = \frac{4}{\pi}\int_0^{\pi/4} t \sec\theta\, d\theta = \frac{4}{\pi} t \log\tan\frac{3\pi}{8} = 1{\cdot}1222\, t \qquad (7)$$

(The above integral is evaluated over one octant. By symmetry the result is true for all θ.)

Thus we have, for equal $P_{fa} = 0{\cdot}1$,

$$\beta = 2{\cdot}146 \quad \text{and} \quad \beta' = 1{\cdot}1222\, t$$

but

$$t = 1{\cdot}948 \quad \text{giving} \quad \beta' = 2{\cdot}1860$$

It can be seen that, although from a noise-only point of view the thresholds are the same, a signal of uniformly distributed phase will have a larger equivalent radial threshold to overcome in the case of the square threshold, and the signal voltage would have to be scaled up to obtain the same P_d.

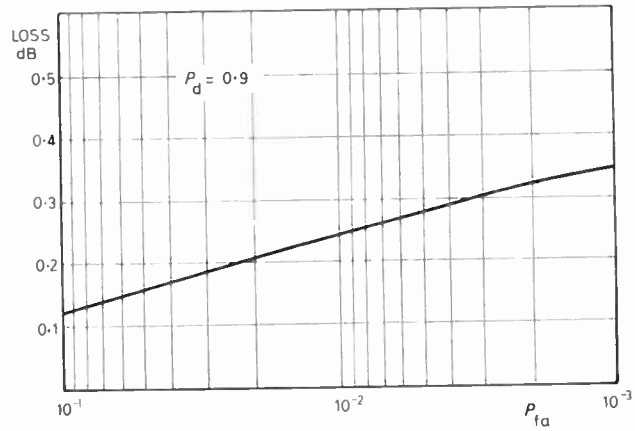


Fig. 2. Single strike probability of false alarm.

Referring to Appendix F of Ref. 4 we find expressions for the partial derivatives of the Q function with respect to α and β.

These may be used with Table 9.8 of Ref. 3 to obtain the correct scaling factor for α to maintain P_d when using the square threshold:

$$\alpha' = \alpha + (\beta' - \beta)\frac{\partial\alpha}{\partial\beta} \quad \text{where} \quad \frac{\partial\alpha}{\partial\beta} \quad \text{is} \quad -\frac{I_0(\alpha\beta)}{I_1(\alpha\beta)} \qquad (8)$$

The ratio of α' to α is the required loss.

A graph of this result for P_{fa} ranging from 0·1 to 0·001 and $P_d = 0{\cdot}9$ is presented in Fig. 2 where the necessary increase of α has been expressed as a ratio (in dB).

4 Conclusions

Reference 2 has shown that 'it is better to detect a noisy signal of unknown phase by setting a threshold on the envelope than by setting thresholds on its two quadrature components'.

It is now shown that the loss of effective signal-to-noise ratio incurred by making the wrong choice of threshold type may be acceptably small for many practical applications.

5 References

1 Skolnik, M. I., 'Introduction to Radar Systems', p. 438 (McGraw-Hill, New York, 1962).
Tucker, D. G. and Griffiths, J. W. R., 'Detection of pulse signals in noise', *Wireless Engineer*, **30**, pp. 264–73, November 1953.

2 Wetherall, P. R., Whitfield, G. R. and Woodward, P. M., 'A comparison of two methods of detecting a signal of unknown phase in the presence of noise', *RRE Journal*, No. 52, pp. 141–6, April 1965.

3 Abramowitz, M. and Stegun, I. A., 'Handbook of Mathematical Functions' (Dover, New York, 1965).

4 Helstrom, C. W., 'Statistical Theory of Signal Detection', Appendix F (Pergamon, Oxford, 1968).

5 Dwight, H. B., 'Mathematical Tables' (Dover, New York, 1961).

Manuscript received by the Institution on 31st August 1976. (Short Contribution No. 184/AMMS80.)

UDC 621.373.9

Mobile radio frequency synthesizers based on surface acoustic wave oscillators

P. M. GRANT, B.Sc., Ph.D., C. Eng., M.I.E.E.,*
K. M. HUSSAIN, B.Sc.,*
R. C. CORNER*
and
Professor J. H. COLLINS,
B.Sc., M.Sc., C.Eng., F.I.E.E., F.I.E.R.E.*

SUMMARY

Delay-stabilized surface acoustic wave oscillators are attractive for use as stable, low-noise sources in digital frequency synthesizers for mobile radio. Here the design and operating principles of the surface acoustic wave oscillator are briefly reviewed. Its application is subsequently detailed in three experimental frequency synthesizer modules, which phase lock the surface acoustic wave oscillator to a reference crystal oscillator to achieve the superior stabilization required in mobile radio applications. First a single channel u.h.f. synthesizer is described which possesses superior efficiency when compared with conventional frequency multiplier based transmitters. Later digital dividers are added extending this module into multichannel synthesizers which ultimately have potential applications in h.f., v.h.f. and u.h.f. mobile radio systems.

* *Department of Electrical Engineering, University of Edinburgh, King's Buildings, Mayfield Road, Edinburgh EH9 3JL.*

1 Introduction

The-delay stabilized surface wave oscillator (s.a.w.o.)[1–4] possesses several features which could prove attractive for mobile radio applications. When compared with conventional crystal oscillators (based on bulk acoustic wave resonators) the advantages of s.a.w.o.s are rugged construction, high fundamental operating frequency (10 MHz–2 GHz), output powers up to 1 W, and low single sideband (s.s.b.) f.m. noise performance.[1] A significant attraction is the ability during design to trade off the oscillator Q to give either high stability ($Q \sim 10000$) or a frequency deviation capability up to 1% of the operating frequency ($Q \sim 100$). This latter feature ultimately permits a single s.a.w.o. to cover several v.h.f./u.h.f. radio channels. Therefore, the s.a.w.o. characteristics are intermediate between the bulk crystal oscillator and LC voltage-controlled oscillator (v.c.o). (See Table 1.) However, s.a.w.o.s do not possess either the ultimate long-term stability of the former or the electronically-controllable octave bandwidth of the latter. For completeness the resonator stabilized s.a.w.o.[5] has been included in Table 1. However, its restricted deviation precludes its use in multichannel synthesizers and hence it is not considered further in this paper.

Delay-stabilized s.a.w.o.s have many potential uses in communication, radar and telemetry equipments. This paper describes the design and operation of three experimental mobile radio frequency synthesizer modules incorporating these s.a.w.o.s. First a single channel synthesizer (Fig. 1) is reported which replaces the frequency multipliers in existing u.h.f. radiotelephone transmitters with a s.a.w.o. operating directly at 460 MHz carrier frequency. The requirement for accurate long and medium-term frequency stability, ± 6 parts in 10^6, is met by controlling the s.a.w. oscillator directly from a stable bulk crystal oscillator with a phase locked loop. The module maintains lock over the $-10°C$ to $+40°C$ temperature range with a $\pm 10\%$ adjustment of the supply voltage. By careful choice of loop time constants, the s.a.w.o. can also be frequency modulated. Thus, the module can be used to retrofit v.h.f./u.h.f. personal radiotelephones. The module is later extended, with variable divider logic, into multichannel frequency synthesizers, e.g. Fig. 2. Preliminary results are presented for both v.h.f. and u.h.f. s.a.w.o.-based synthesizers. Alternative designs of synthesizers incorporating s.a.w.o.s are included and their relative advantages discussed. The experimental modules reported here were designed primarily to demonstrate operating principles, and hence were not optimized either in terms of efficiency or power output.

2 S.A.W. Oscillator Principles

The delay stabilized s.a.w.o., shown schematically in Fig. 1, comprises a delay line whose output is amplified and fed back to the input. S.a.w.o. delay lines (Fig. 3) are fabricated on polished quartz substrates by depositing

Table 1. Comparison of properties of various oscillators.

Oscillator type	Operating mode	Frequency range (Hz)	Effective loaded Q	Deviation	Temperature coefficient (-30 to $+70$°C) (parts in 10^6/deg C)	S.s.b. f.m. noise at 10 kHz offset from 500 MHz source dB/Hz
Conventional quartz crystal	resonator stabilized	10^4–2×10^7 (fundamental)	5×10^3–2×10^6	up to 500 parts in 10^6	<1	-140 (multiplied 10 MHz)
LC-based voltage controlled	resonator stabilized	10^3–10^{10}	10^1–10^3	up to octave bandwidth	typically 10	-100
Surface acoustic wave	resonator stabilized	10^7–2×10^9	10^4–10^5	<100 parts in 10^6	~1	no measured results
	delay stabilized		10^2–10^4	10^4–10^2 parts in 10^6		
			dependent on delay line length			-105 to -140 (Q dependent)

and subsequently etching a 500–2000 Å aluminium film to yield the desired interdigital transducer (i.d.t.) patterns. The s.a.w. velocity of 3157 m/s introduces a delay of ~ 3 μs/cm, resulting in a wavelength, λ, of 6·8 μm (i.d.t. finger width 1·7 μm) at 460 MHz. When the amplifier gain exceeds the delay line loss oscillation occurs according to the phase condition:

$$\phi_e + \frac{2\pi f_n L}{v} = 2\pi n \qquad (1)$$

where L is the acoustic path length, f_n the frequency of the nth mode, v the acoustic velocity and ϕ_e the phase shift through the feedback loop. This oscillator multimodes[2] at a comb of frequencies spaced by $1/T_D$ where T_D is the loop transit delay, which is approximately L/v. However, the s.a.w. delay line is a frequency filter whose response is controlled by the i.d.t. geometries. Oscillator delay lines are usually designed with a $(\sin x)/x$ frequency response where the main lobe peak coincides with the desired operating frequency, f_0. When the nulls are arranged to suppress the unwanted frequencies stable (single mode) operation is achieved.

This oscillator can now be deviated by altering the feedback delay with an electronically controlled phase shifter. Selection of a $\pm\pi/2$ phase shift provides a frequency deviation,

$$\Delta f = \mp \frac{f_0}{4L} \qquad (2)$$

without upsetting the single-mode operation. Thus, the selection of L which controls directly the oscillator Q (Ref. 3)

$$Q = \pi L \qquad (3)$$

and hence the spectral linewidth and s.s.b. f.m. noise performance[1,2] also sets the available deviation. Analysis[4]

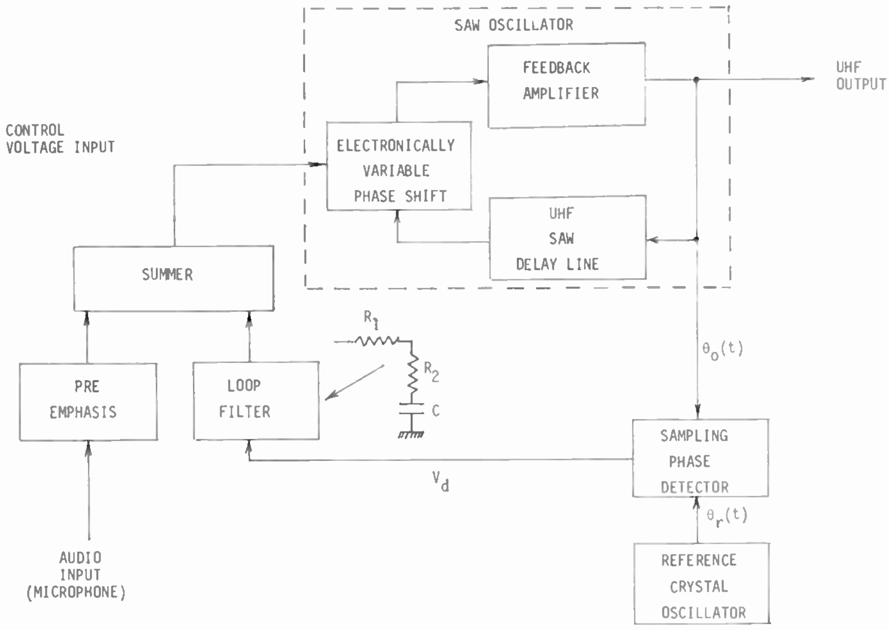


Fig. 1. Single channel synthesizer module.

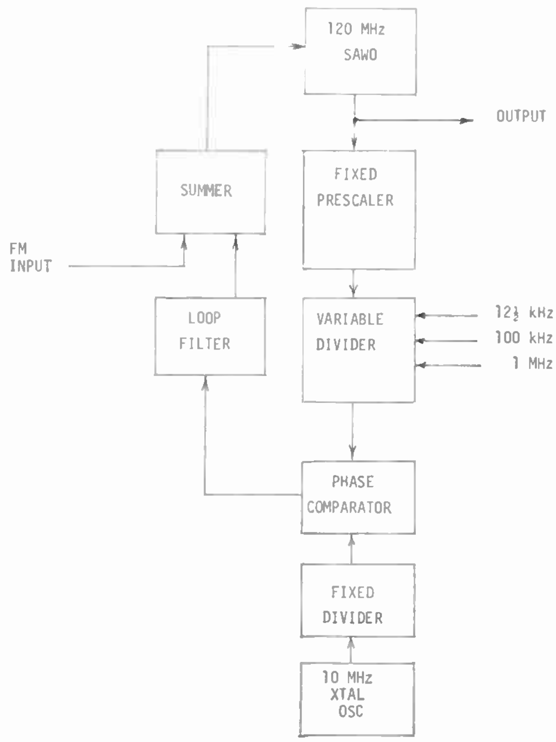


Fig. 2. V.h.f. s.a.w.o.-based synthesizer.

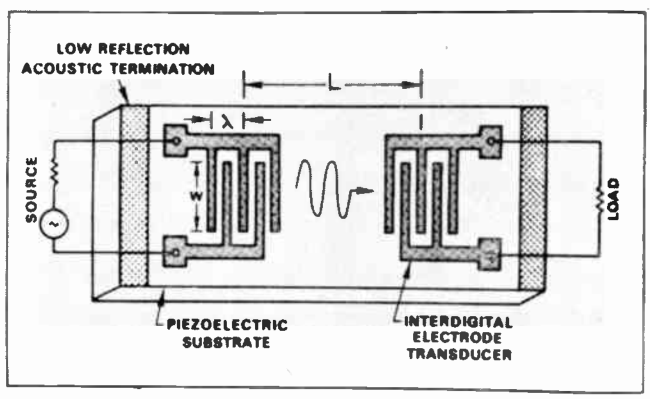
is currently aimed at establishing the maximum modulation rate for s.a.w.o.s. Modulation indices of unity have been achieved in s.a.w.o.s employing modulation rates of 10 T_D. Measurements of short-term stability on high-power, 5 W, s.a.w.o.s have shown results which are comparable with, if not superior to, existing crystal oscillator performance (Table 1).

Several problems hinder the direct incorporation of s.a.w.o.s in radiotelephones. First, the oscillator medium-term (temperature) stability must meet the ± 3 kHz u.h.f. radiotelephone requirements[7] over the -10°C to $+40$°C operating temperature range. This can be achieved with crystal oscillators based on AT cut bulk acoustic wave resonators possessing a crystal cut angle tolerance of $\pm 1'$ of arc. The most stable s.a.w.o.s are fabricated on ST cut quartz substrates, which can be less highly toleranced without degrading the oscillator stability. However s.a.w.o.s only possess a stability of ~ 1 part in 10^6 per deg C over the required temperature range,[6] hence their medium-term stability is always inferior. Secondly, the s.a.w.o. is sensitive to instabilities in the phase response of the feedback amplifier. It is for this reason that wideband feedback amplifiers have been predominantly used in s.a.w.o.s. Here we have explored the feasibility of incorporating a low-power tuned loop amplifier with a class C power amplifier to achieve the superior d.c. to r.f. efficiency which is required for battery powered equipments. Thirdly, s.a.w.o. current long-term ageing rates[8] (~ 1 part in 10^6/month) necessitate regular equipment realignment. These three deficiencies have been overcome in the following synthesizer[9] modules by phase locking the s.a.w.o. to a reference crystal oscillator. Although the ultimate deviation of a s.a.w.o. (1%) is small compared to a v.c.o., s.a.w.o. based synthesizers are well matched to the bandwidth of existing u.h.f. pretuned radiotelephone equipments.

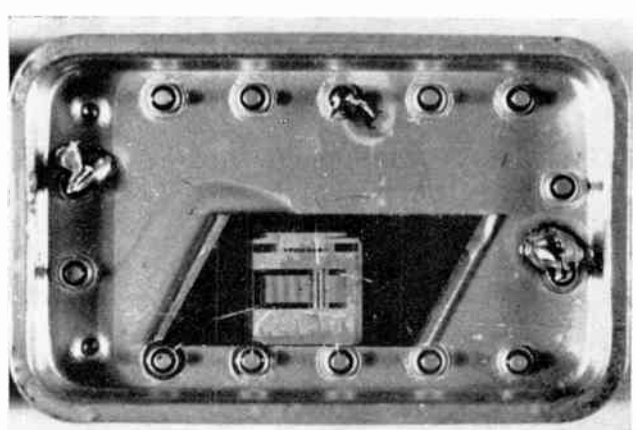
3 Single-channel Frequency Synthesizer

3.1 460 MHz S.A.W. Oscillator

The 460 MHz $L = 400\ \lambda$ delay line, which was fabricated on a 7·6 mm × 1·8 mm (0·3 in × 0·7 in) rhombic shaped ST, X quartz substrate (Fig. 3(b)), was characterized by exciting it with a shortburst of r.f. signal.[10] The insertion loss was 21·5 dB at the centre frequency of 459·73 MHz and the delay was 870 ns. The s.a.w.o. was constructed with a narrowband feedback amplifier, comprising three tuned MMT 2857 transistor stages, with 30 dB open loop gain. Output power of 0·2 mW was obtained for 64 mW d.c. input. Both the efficiency and output power level can be improved with class C tuned power amplifiers. The electronically variable phase shift was obtained with a CR network. Variation of the BA 141 varactor diode reverse bias between 1 volt and 7 volt altered its capacitance from 17 pF to 4 pF giving a 50° change in the phase shift through the network. This deviated the oscillator by 160 kHz. When temperature cycled from -10°C to $+40$°C the u.h.f. output from the



(a) Schematic.



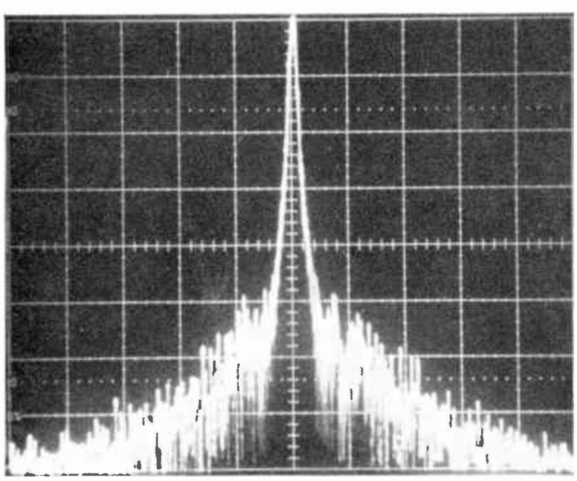
(b) 460 MHz delay line in prototype 3·8 × 2·5 cm package.

Fig. 3. S.a.w. delay line

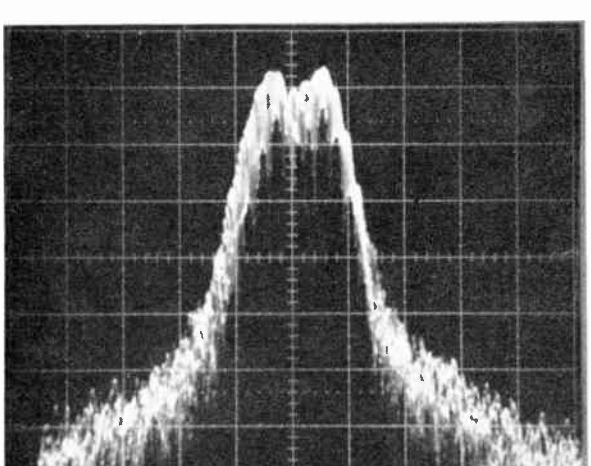
s.a.w.o. altered by 30 kHz. This temperature performance can be improved by almost an order of magnitude when a wideband feedback amplifier is used.[10] However, in principle this oscillator possesses sufficient electronically-controllable frequency adjustment to compensate for all temperature effects, power supply tolerances ageing mechanisms, *and* permit f.m. deviation of 5 kHz to be achieved.

3.2 Phase-locked Sampler

The sampling phase detector used a step recovery diode to convert the reference input from the crystal oscillator into a train of short (500 ps) pulses. These pulses drove a diode bridge which sampled the s.a.w.o. output and subsequently held it in a f.e.t. buffer amplifier. When combined with the 460 MHz s.a.w.o. the open-loop gain was approximately $10^5\ s^{-1}$. The phase detector power consumption was 400 mW. Our s.a.w.o. synthesizer was designed for a low-pass loop with a low audio cut off frequency, 40 Hz. This permits the loop to track slow transients, e.g. temperature and ageing effects, without responding to the oscillator modulation. In addition the −6 dB/octave audio pre-emphasis severely attenuates any s.a.w.o. deviation at modulation rates below 100 Hz.



(a) Without deviation.



(b) Deviated ±2·5 kHz at 1 kHz rate (analyser bandwidth 100 Hz).

Fig. 4. Spectrum analysis of synthesizer output.

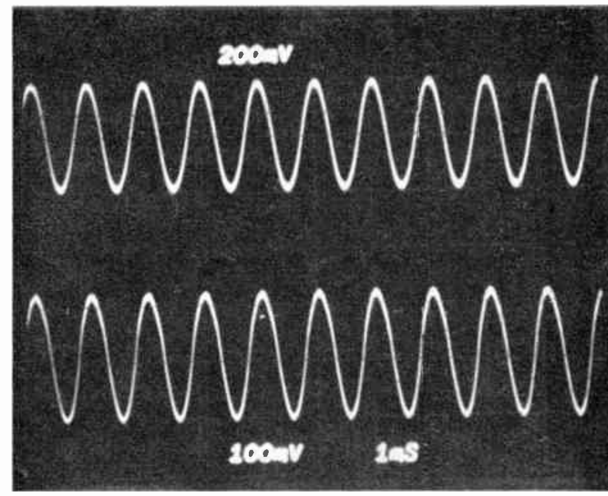


Fig. 5. Synthesizer modulation.

Upper trace: Input modulation (1 kHz tone).
Lower trace: Demodulated 460 MHz output.

3.3 Synthesizer Performance

The 460 MHz synthesizer module was constructed with a 10 MHz laboratory standard supplying the reference frequency. Figure 4(a) shows a spectrum analysis of the s.a.w.o. module output with no modulation input. In comparison Fig. 4(b) shows the output when deviated 5 kHz by the 1 kHz audio tone shown in the upper trace of Fig. 5. The lower trace of Fig. 5 shows the demodulated 460 MHz output. The similarity between traces of Fig. 5 gives no evidence of any distortion from the phase control loop. Phase lock was maintained, with no change in the output frequency, while the module was temperature cycled from −10°C to +40°C and the power supply adjusted ±10%.

The significance of this new approach to u.h.f. radiotelephone transmitter design can be assessed by comparing the synthesizer performance against commercially available products. A comparison based on component complexity is not valid as the prototype module design has not been optimized. With the recent launch of commercially available s.a.w. i.f. filters[11] for domestic colour television receivers, which cost only 80p in quantity production, it is our opinion that any doubts as to the reliability and availability of s.a.w. delay lines should now be unfounded. Contrary to the television requirement the s.a.w.o. has a much higher tolerance of delay line fabrication errors as the precise filter characteristic is not a prime factor in maintaining stable oscillation.

It is suggested that the most relevant assessment of our module is based on a comparison of its power consumption against existing radiotelephone transmitters. Although the basic s.a.w.o. is inefficient, when the feedback amplifier is operated at low loop power (1–10 mW), to minimize d.c. power consumption, and class C amplifiers are employed to obtain the required (100 mW–1 W) output, then the overall oscillator efficiency will improve with higher output powers as shown in Fig. 6. Efficiencies of 50% have already been achieved[7] at 5 W output power. Adding the power drain of the sampling phase detector and reference crystal oscillator gives the projected synthesizer module efficiency shown in Fig. 6. When compared against existing UK Pye Pocketphone equipments the efficiency of our module should be comparable with these equipments up to ~ 100 mW output power. Above this level we predict that the s.a.w.o. based design of u.h.f. transmitters has significant advantages, hence it now deserves further serious consideration by equipment manufacturers. These single channel synthesizers, employing narrow oscillator deviation, can in principle be implemented with either delay or resonator stabilized s.a.w.o.s. The lower insertion loss of the s.a.w. resonator[5] is a major attraction in battery powered equipments.

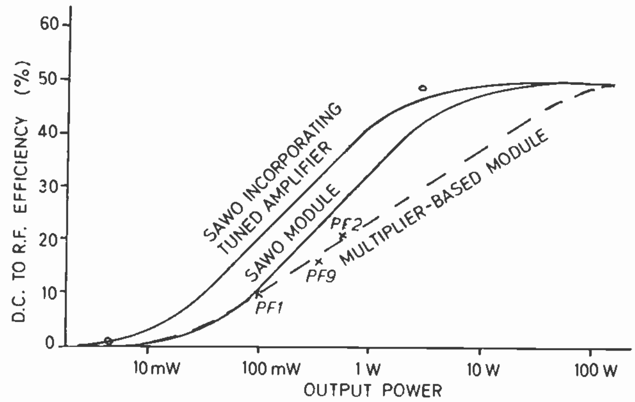


Fig. 6. Comparative performance prediction for Pocketphone transmitters based on conventional frequency multiplication and s.a.w phase-locked oscillator modules.

4 Multichannel Frequency Synthesizers

4.1 V.h.f. Synthesizer Design

Our first demonstration multichannel synthesizer, which operated at v.h.f. (Fig. 2), followed standard indirect design techniques[9] where the output from the voltage controlled s.a.w.o. is divided and subsequently compared against a reference crystal controlled frequency in a phase comparator. If the variable divider is programmed to count in unity increments then the comparison frequency equals the channel spacing, 12·5 kHz. The d.c. control voltage from the comparator is filtered and fed back to control the s.a.w.o. The loop filter can again be designed with a cut-off at low audio frequencies permitting direct modulation of the s.a.w.o. with the voice or data input.

As in the earlier module the prime goal is to minimize the power consumption. This dictates that maximum use must be made of complementary metal oxide silicon (c.m.o.s.) logic. Unfortunately, the maximum operating frequency of c.m.o.s. is 5 MHz, hence the output of the s.a.w.o. must be prescaled in a fixed counter before the c.m.o.s. variable divider (Fig. 2).

4.2 V.h.f. Synthesizer Performance

The module used a 120 MHz v.h.f. s.a.w.o. incorporating a readily available $L = 350\,\lambda$ delay line. The long delay (~ 3 μs) restricts the number of available channels in this experimental synthesizer. Prescaling was achieved in a Plessey SP 8655 emitter coupled logic (e.c.l.) fixed divider. A c.m.o.s. 4059 counter, which was programmed from external thumbwheel decade switches, was used for the variable divide-by-9600. C.m.o.s. phase comparison and fixed reference division was also used. With a passive RC loop filter the power consumption, 50 mW, in the digital control loop arose predominantly from the e.c.l. prescaler. The oscillator tuned from 119·90 MHz to 120·05 MHz, permitting the synthesis of 12 channels with 12·5 kHz spacing. The prescaling by 32 limits the overall division ratio of the variable divider, and reduces the comparison frequency by the same factor of 32, from 12·5 kHz to 390·625 Hz. This restricts both the loop gain and the tuning time. These performance deficiencies have resulted in more sophisticated approaches in the design of the following u.h.f. synthesizers.

4.3 U.h.f. Synthesizer Designs

In a multichannel frequency synthesizer which retains a low loop cut-off frequency to permit direct modulation of the s.a.w.o. with the voice input there is a limit to the minimum oscillator Q. The Q must be sufficiently high to maintain a narrow oscillator linewidth stopping the transmitter radiating excessive energy (−90 dB) into adjacent radio channels. This sets a minimum Q value of ~ 500. Equation (3) subsequently yields the minimum delay line length as $L = 160\,\lambda$, and equation (2) gives the maximum deviation for a 480 MHz s.a.w.o. as 1·5 MHz. Although this oscillator deviation is low in comparison with the octave bandwidth of an LC v.c.o., synthesis of 120 u.h.f. 12·5 kHz radio channels is adequate for many mobile radio applications. In fact, the current bandwidth of Pye PF9 u.h.f. pretuned Pocketphones is only 1·5 MHz.

Here two distinct design approaches are reported which overcome the restricted comparison frequency when a fixed prescaler is inserted before the variable divider. Either the new variable dual modulus prescalers can be incorporated (Fig. 7) or the v.c.o. output can be mixed with a separate stable oscillator down-converting (Fig. 8) to frequencies which can be accommodated in the variable dividers.

The variable prescalers[9] are based on e.c.l. counters which normally divide by a fixed radix, e.g. 10, but in response to an asynchronous input from the programming counter they will count one extra pulse and divide by 11. Variable prescalers can be cascaded or combined

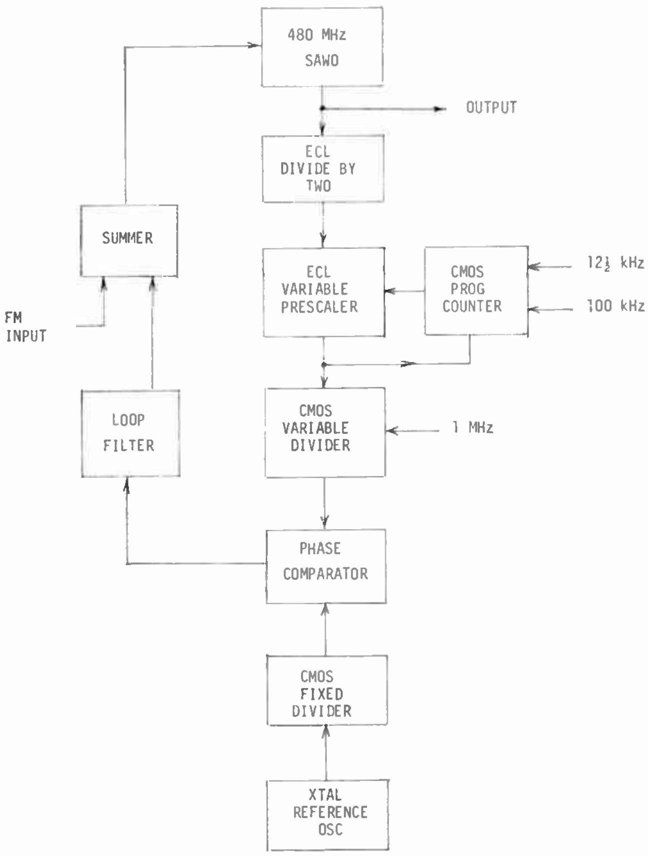


Fig. 7. U.h.f. s.a.w.o.-based synthesizer.

with fixed dividers to form high speed variable radix counters. When followed by conventional c.m.o.s. variable dividers the synthesizer comparison frequency does not require to be reduced as in the 120 MHz synthesizer. The new system (Fig. 7) offers high divider speed and complexity for moderate power consumption. Normally, the loop filter would have difficulty in suppressing the control line ripples introduced by the variable prescaler at subharmonics of the comparison frequency. This is not a problem in the s.a.w.o. synthesizer module as the loop cut-off is purposely kept to low audio frequencies.

Preliminary experiments are now being performed on a u.h.f. synthesizer based on this design concept. The synthesizer incorporates an e.c.l. fixed divide-by-two followed by a variable divide-by-80/81 with the remainder of the control electronics realized entirely in c.m.o.s. circuitry. Loop power consumption of 150 mW is contributed primarily by the e.c.l. prescalers. The edge triggered phase detector provides a lock range equal to the capture range. With a $L = 200\ \lambda$ 480 MHz delay line the synthesizer initially covered 30 channels between 479·8 MHz and 480·175 MHz at 12·5 kHz spacings. The feedback amplifier and phase shifter are now being optimized to obtain the full $\pm\pi/2$ phase shift for 96 channel coverage. Detailed measurements will subsequently be performed on synthesizer spectral purity and tuning time.

This synthesizer is limited by the s.a.w.o. deviation to operation in either the transmitter or receiver. With two oscillator delay lines, which are offset in frequency by the radio i.f., and sidestep logic[9] in the variable divider, one synthesizer can operate in both the transmitter and receiver. Alternatively it may be possible with ladder i.d.t. design[12] to operate one multi-moded s.a.w.o.[13] stably at either the transmitter or receiver frequencies.

In the alternative u.h.f. synthesizer design (Fig. 8) the requirement for the high speed variable prescaler is removed by mixing the u.h.f. v.c.o. output with a separate u.h.f. fixed stable oscillator operating slightly below the minimum v.c.o. output frequency. Low-pass filtering selects the difference frequency which is subsequently divided in the digital control loop. Moderate prescaling may be required dependent on the desired number of channels. This approach for synthesizer design is not presently favoured as any drift or noise on the high stability fixed u.h.f. oscillator is transferred to the v.c.o. Thus, there is an excellent application here for the single-channel phase-locked s.a.w.o. module reported earlier.

In principle, this u.h.f. synthesizer could be implemented with two s.a.w.o.s. One would be used as an intermediate Q u.h.f. v.c.o., while the other would provide the high Q u.h.f. fixed output. However, the enhanced electronic tuning range of the conventional LC-based v.c.o. is attractive as it permits the synthesizer to be offset by the radio i.f. for use in both the transmitter and receiver. In Fig. 8 the s.a.w.o. operates on a harmonic of the radio i.f., e.g. 23 MHz. On switching from transmit to receive the reference frequency is added to the s.a.w.o. output in a second mixer to introduce an i.f. offset. The v.c.o. subsequently retunes without increasing the operating frequency or incorporating sidestep into the variable divider.

The use of a v.c.o. rather than a s.a.w.o. as the controlled oscillator in this synthesizer necessitates a wideband control loop to improve the low Q v.c.o. short term stability. In addition the large change in loop gain introduced by the variable divider will require careful design of the loop filter if capture is to be maintained at all frequencies. This may require the frequency selection information also to be used to control the loop filter time constants. The selection of an LC v.c.o. or s.a.w.o as the controlled oscillator involves a delicate compromise between tuning range, noise performance and efficiency. The s.a.w.o. offers improved wideband s.s.b. f.m. noise with increased loop power. In contrast the LC v.c.o. is constrained by the varactor diode breakdown voltage which sacrifices tuning range in high power low noise sources.

The synthesizers reported here have predominantly used low cut-off frequencies in the loop filter to permit simultaneous f.m. of the module with audio information. This has the disadvantage that it increases the tuning time to approximately one second. In applications such as the new urban cellular systems,[14] where channel allocations

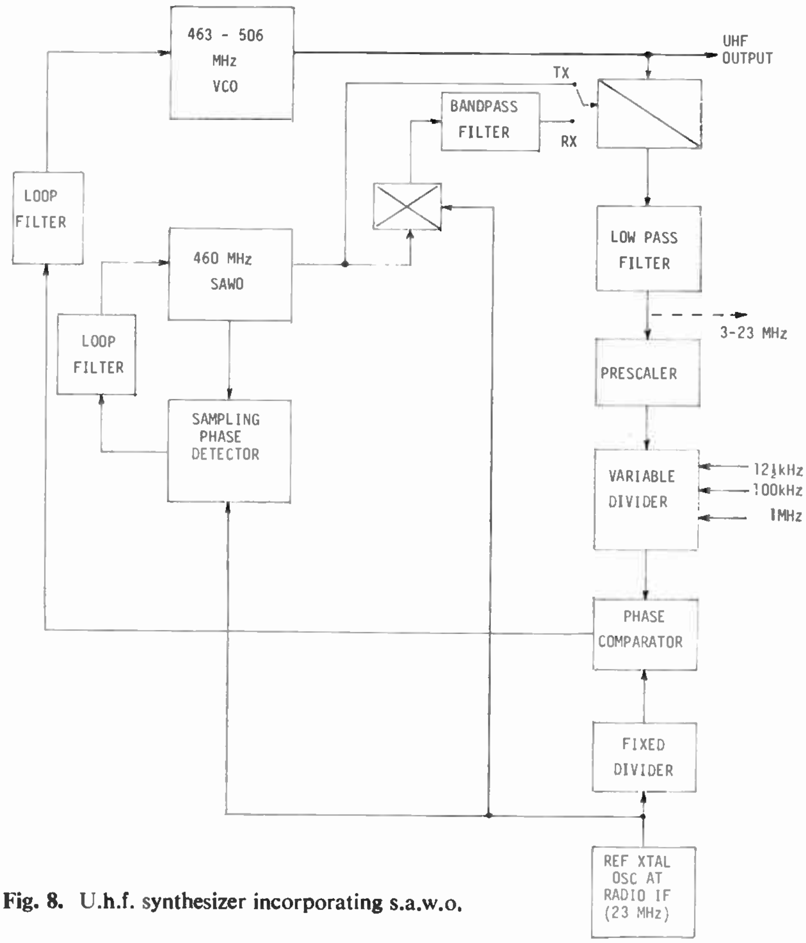


Fig. 8. U.h.f. synthesizer incorporating s.a.w.o.

are reassigned as the mobile alters it location, the long tuning times will result in loss of information. Here sophisticated design techniques, which increase the comparison frequency above the channel spacing to give a wideband control loop, are required to permit fast synthesizer switching. This will also permit the use of low Q (e.g. 100) s.a.w.o.s with a consequent fivefold increase in the number of synthesizer channels, over that reported here. Normally modulation of this synthesizer with information would have to be performed externally. However, if the audio input is suitably pre-emphasized then it can be modulated within the wideband control loop.

5 Conclusions

The paper has outlined the design of delay stabilized single moded s.a.w.o.s and compared them briefly against existing crystal oscillators. Table 1 highlights the fact that although the s.a.w.o. Q is lower than a crystal oscillator, when operated at high loop power the higher s.a.w.o. operating frequency gives it a noise performance comparable with a multiplied crystal source. The current performance of s.a.w.o.s, which utilize separately packaged delay line and maintaining amplifier, is predicted to improve significantly with the recent development of thin film hybrid s.a.w.o.s (Fig. 9) which minimize interconnection and packaging parasitics. Direct incorporation of a u.h.f. s.a.w.o. in personal radiotelephone equipments, which must cover a 50°C temperature range with $< \pm 3$ kHz frequency drift, is currently not possible. Therefore phase-locking techniques were employed to control the s.a.w.o. from a reference crystal oscillator. The demonstrated u.h.f. s.a.w.o. single channel synthesizer module was capable of frequency modulation

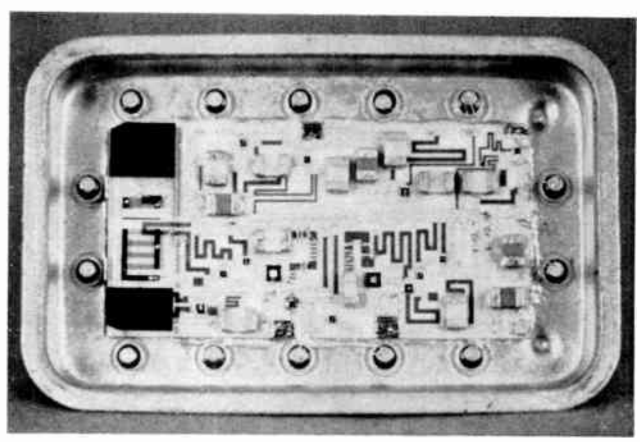


Fig. 9. 458·75 MHz hybrid s.a.w.o. incorporating custom designed thin-film amplifier.
(Courtesy Hewlett Packard Ltd., South Queensferry)

with greater than 5 kHz deviation without unlocking the loop. Its superior d.c. to r.f. efficiency (Fig. 6) is most significant in battery-powered mobile radio equipments.

The 1–2 MHz frequency deviation of intermediate Q u.h.f. s.a.w.o.s also permits the design of multichannel frequency synthesizers based on these oscillators, which cover approximately a hundred u.h.f. radio channels. The design and operation of prototype v.h.f. and u.h.f. synthesizers has been briefly reported.

The u.h.f. s.a.w.o. is also applicable in h.f. synthesizers. This can be accomplished by outputting the h.f. difference frequency after the low-pass filter (see dotted output Fig. 8) in place of the u.h.f. output. This immediately overcomes the requirement for controlling the absolute accuracy of the high Q u.h.f. s.a.w.o. Although a u.h.f. v.c.o. is capable of deviation over the complete 3–30 MHz h.f. band, a miniature, rugged, high power low Q u.h.f. s.a.w.o. is capable of deviation over $>\frac{1}{3}$ of the required bandwidth. Coverage of the entire h.f. band could then be obtained by switching the high Q s.a.w.o. between three separate delay lines, to achieve a low noise, s.a.w.o. based, h.f. synthesizer.

The recent development of multi-moded s.a.w.o.s[13, 15] will extend considerably the number of radio channels which can be covered by s.a.w.o. based frequency synthesizers. Stable, single-moded operation[2] can be obtained by exciting these oscillators with a short r.f. pulse[13] or by combining two s.a.w. oscillators.[15] Multi-mode synthesizers do require more hardware than those reported here and the synthesis of 12·5 kHz channels requires an 80 μs (30 cm long) delay line. However, they do promise a considerable extension to the tuning range of s.a.w.o.-based synthesizers in the near future.

A most significant development is a new oscillator[16] employing conventional s.a.w. i.d.t.s to generate and detect a surface skimming acoustic wave. When compared to the s.a.w.o.s reported here it promises increased operating frequency plus vastly improved temperature performance and the insensitivity of bulk wave propagation to surface contamination should give oscillator ageing rates comparable with conventional crystal controlled oscillators.

6 Acknowledgments

The authors wish to acknowledge the loan of 120 and 460 MHz delay lines from Hewlett Packard Limited, and 480 MHz delay lines from the Marconi Research Laboratories. The assistance of colleagues here in the department and helpful discussions and suggestions with various industrial companies is also gratefully acknowledged. These studies were sponsored by the British Science Research Council.

7 References

1 Lewis, M. F., 'The design and performance limitations of s.a.w. oscillators', IEE Conference Publication, No. 109, pp. 63–72, September 1973.

2 Burnsweig, J., 'Design considerations and performance capabilities of surface acoustic wave delay line oscillators', Proceedings D.R.M.E., Microwave Acoustic Applications Meeting Paris, pp. 258–88, April 1976.

3 Voller, H. G. and Claiborne, L., 'R.f. oscillator control utilizing surface acoustic wave delay lines', Proceedings of 28th Frequency Control Symposium, Fort Monmouth, pp. 256–9, May 1974.

4 Lawrence, M. W., 'Surface acoustic wave oscillators', *Wave Electronics*, **2**, pp. 199–218, July 1976.

5 Bell, D. T. and Li, R. C. M., 'Surface acoustic wave resonators', *Proc. IEEE*, **64**, No. 5, pp. 711–21, May 1976.

6 Dias, J. F., Karrer, H. E., Kusters, J. A., Matsinger, J. H. and Schultz, M. B., 'The temperature coefficient of delay-time for X-propagating acoustic surface-waves on rotated Y-cuts of alpha quartz', *IEEE Trans. on Sonics and Ultrasonics*, **SU-22**, No. 1, pp. 46–50, January 1975.

7 U.K. Ministry of Post and Telecommunications, UHF Private Land Mobile Radio Specification MPT 101 and MPT 102.

8 Gratz, S. C., 'Surface acoustic wave oscillators', *Electronic Industries*, **1**, No. 3, pp. 20–2, November 1975.

9 Stokes, V. O., 'Techniques for frequency synthesis', *Proc. Instn Elect. Engrs*, **120**, No. 10R, pp. 1057–77, October 1973.

10 Grant, P. M., Corner, R. C. and Collins, J. H., 'Surface acoustic wave oscillator modules for mobile radio', IEE Conference Publication No. 139, pp. 86–9, June 1976.

11 Deacon, J. M. and Heighway, J., 'S.a.w. filters for t.v. receivers', *IEEE Trans. on Consumer Electronics*, **CE-21**, No. 4, pp. 390–5, 1975.

12 Lewis, M. F., 'Surface wave acoustics—applications to safety at sea', Proc. Symp. on Marine Electronics, Southampton University, pp. 15–21, July 1976.

13 Browning, I., Crabb, J. and Lewis, M. F., 'A s.a.w. frequency synthesizer', Proc. 1975 IEEE Ultrasonics Symposium, pp. 245–7 (IEEE Cat. No. 75 CHO 994-4SU).

14 Bearse, S. V., 'The year of giant growth arrives for land mobile communications', *Microwaves*, **13**, No. 1, pp. 38–47, January 1974.

15 Parker, T. E. and Sage, J. P., 'A s.a.w. oscillator using two acoustic paths', Proc. 1976 IEEE Ultrasonics Symposium, pp. 243–7 (IEEE Cat. No. 76 CH 1120 5SU).

16 Browning, T. I. and Lewis, M. F., 'New family of bulk-acoustic-wave devices employing interdigital transducers', *Electronics Letters*, **13**, No. 5, pp. 128–30, 3rd March 1977.

Manuscript first received by the Institution on 21st December 1976 and in final form on 7th February 1977 (Paper No. 1765/CC270).

UDC 621.373.52.029.42/5

A low distortion tri-wave to sine converter

W. A. EVANS, B.Sc., M.Sc., C.Eng., M.I.E.E.*

and

V. SCHIFFER, Dipl. Ing., M.Sc.†

SUMMARY

A new triangle-to-sine shaper circuit is described based upon the modified transfer function of a differential pair of transistors. Computer-aided design techniques have been used to establish an optimized set of design criteria. Theoretical figures as low as six parts per million total harmonic distortion (t.h.d.) are possible with the circuit. Measurements taken on a prototype have fulfilled the promise of the circuit with t.h.d. figures of better than 0·05% with operation to 10^5 hertz.

* *Department of Electrical and Electronic Engineering, University College of Swansea, Singleton Park, Swansea SA2 8PP.*

† *Wandel and Goltermann, Reutlingen, West Germany; formerly a Visiting Research Fellow at Swansea.*

Principal Symbols

$D(t)$	deviation for a sinewave
$H(u_i)$	transfer function
h_{FE}	d.c. transfer ratio
I_1	collector current
I_s	saturation current
I_0	source current
k	Boltzmann's constant
q	electron charge
R_s	source load resistor
R_c	channel resistance when $V_{GD} = 0$
R_d	drain load resistor
T	absolute temperature
t.h.d.	total harmonic distortion
u_i	differential input voltage
$\hat{u}_i$, V_m, $\hat{U}_{in}$	peak voltages
u_T	temperature voltage
u_{be}	base emitter voltage
V_p	pinch-off voltage

1 Introduction

In recent years the utility of the conventional linear oscillator has been progressively challenged by the function generator signal source. The principal distinguishing feature of nearly all these instruments lies in the use of a triangular waveform as the basis from which many other waveforms are derived. In particular, a square wave occurs naturally at some point within the circuitry of all triangle generators. Ramp and pulse waveforms result when the slopes of the sides of the triangle are made unequal and the sinewave is invariably derived using a non-reactive filter or shaping network. Apart from the variety of waveshapes available, function generators provide frequency coverage of typically $10^{-4} < f < 10^7$ hertz. In addition, it is no problem to sweep the frequency over any region within this range. The sweeping can be accomplished with great linearity at relatively high sweep rates over several decades with careful design.

As a source of the ubiquitous sinewave, however, the function generator, despite rapid strides in design, compares unfavourably in purity of output when compared with oscillators based on RC networks of the Wien, phase-shift or twin-tee variety. Function generators seldom exhibit total harmonic distortion (t.h.d.) figures much below 0·5% whilst a t.h.d. figure of 0·05% is relatively easy to obtain with a simple linear oscillator. The main cause of the high distortion figure lies in the generation of a non-ideal triangle wave as well as non-ideal performance of shapers. This paper examines the latter problem leading to the conception and design of a new non-linear shaper of promising performance. A prototype version of the circuit has yielded distortion figures less than 0·05% up to 10^5 hertz.

Shaping Techniques

Methods for shaping triangles into sinusoids fall into two categories:

(a) Linear segment approximation techniques.

(b) Non-linear methods based on the transfer characteristics of electronic devices.

The former is by far the most popular in use in present-day equipment. The number of breakpoints and their positions determine the quality of the sinewave. Strategies for the selection of suitable breakpoints can vary:

(1) Select the breakpoints to minimize the difference between the approximation function and the sinewave in absolute or relative terms. Barnes has examined both tangential and chordal approximations concluding that the latter yields the best result.[1] No indication was given, however, whether the chordal approximation led to an optimum solution to breakpoint selection in terms of an established criterion such as minimizing t.h.d.

(2) Select the breakpoints to eliminate as many harmonics as possible. The choice of breakpoints according to this constraint has been investigated by Schiffer and Evans.[2]

(3) Select the breakpoints to minimize t.h.d. Breakpoint selection according to this constraint is also examined by Schiffer and Evans.

Practical circuits using piecewise linear approximation techniques are many and varied. Ritchie and Young[3] and others have described the now familiar diode resistor array having separate halves for the positive and negative parts of the shaper. Limitations with this circuit include temperature effects coupled with the inability to obtain a 'flat' top. The condition implies infinite conductance of the extreme branches of the array, a condition not met even with high performance diodes. Lack of symmetry between positive and negative halves of the shaper is a further deleterious effect. Klein and Hagenbeuk have developed a rather ingenious circuit in which antiphase triangle inputs are injected at the bases of a differential pair of switching transistors.[4] The current source common to both transistors comprises a combination of a conventional transistor current source and a resistor diode array similar to one half of the previous shaper with the diodes reversed. The combination functions as a conductance which varies as a cosine function so that the currents flowing in each transistor are sine functions. Half sinusoids are generated in anti-phase at the respective collectors so that the voltage appearing between the collectors is a complete sinusoid. Apart from improvements in symmetry, because the same resistor diode network is used for both halves of the sinewave, the advantage of this system lies in its potential for truly zero slope at the peak of the sinewaves. This is a practical possibility since reverse biased silicon devices have virtually zero conductance. The big disadvantage, however, lies in the susceptibility of the circuit to crossover distortion which is difficult to eliminate. Both peak and crossover distortion are problems overcome in a circuit due to Paull and Evans which uses a series of trapezoidal generators to synthesize the sinewave output.[5] Distortion figures as low as 0·1% have been achieved with this approach using a nine and a half segment shaper. Circuit complexity and a number of high tolerance resistors remains a basic problem to be overcome.

The use of non-linear transfer characteristics of electronic devices has the potential of reduction in circuit complexity and the number of adjustments required in setting up the circuit. Middlebrook and Richer have described the use of a field effect transistor as a non-linear element.[6] The basis of the shaping action lies in exploiting the drain characteristic below pinch-off. From a series of numerical calculations optimum values were found for the normalized parameters $V_m:V_p$, $R_s:R_c$ and $R_d:R_c$. The calculations demonstrate that a minimum value of distortion of 0·35% occurs in a region very close to $R_s = R_d = R_c$. The normalized amplitude producing this figure occurs when $V_m:V_p = 1{\cdot}33$. The optimum distortion figure as the temperature varies indicates that devices with the higher pinch-off are least affected but the dependence still leaves something to be desired when incorporation into a precision instrument is contemplated. The final practical circuit required two critical adjustments first to trim R_D for minimum second harmonic distortion and second to vary V_m for minimum total harmonic distortion.

The potential of the differential pair as a triangle-sinewave converter was recognized by Grebene in 1972.[7] The circuit contained as essential elements a pair of matched transistors having equal current sources in the respective emitters. A single emitter degeneration resistor spans the two emitters. In recent correspondence, Meyer and co-workers examine this configuration in some detail through experimental and computer simulation studies.[8] A minimum value of t.h.d. is quoted at 0·2% for a $\hat{u}_i:u_T$ ratio of 6·6, corresponding to a value of $\hat{u}_i$ of 175 mV for $u_T = 26$ mV. Operation to 1 MHz is claimed before rapid increase in t.h.d. occurs, due to irregularities in the peaks of the output sinusoid.

3 Circuit Model of Transconductance Shaper

Figure 1 shows the schematic of the proposed new shaper for analysis purposes. The collector current I_1 as a function of base-emitter voltage u_{be} can be approximated by:

$$I_1 \approx I_s \exp\left(\frac{q}{kT} u_{be}\right) \tag{1}$$

This relation is accurate over many decades of current: deviation is apparent only at low levels of collector current where leakages become significant and high levels where access resistances cause additional voltage drops.

If $R_1 = R_2 = 0$ and $R_5 = \infty$ the voltage at the output is given by:

$$U_0(u_i) = \frac{-R_6 I_0}{1+\exp\left(\dfrac{u_i}{u_T}\right)} + U_{cc} \qquad (2)$$

where

$$u_T = kT/q$$

$$U_0(0) = \frac{-R_6 I_0}{2} \qquad (3)$$

The transfer function $H[u_i/u_T]$ is given by:

$$H\left[\frac{u_i}{u_T}\right] = U_0(0) - U_0(u_i)$$

$$= R_6 I_0 \left[-\tfrac{1}{2} + \frac{1}{1+\exp\left[\dfrac{u_i}{u_T}\right]}\right] \qquad (4)$$

When used on its own this characteristic yields a t.h.d. of 1·3% at best. By inserting emitter resistors $R_1 = R_2 = R$ it may be shown that this distortion figure reduces to 0·3% comparing favourably with the results obtained by Middlebrook and Richer.

An essential feature of a good sine approximation is one in which the first derivative is zero at the peak of the sinewave output. This can never be accomplished with the above characteristic alone since:

$$\frac{d\left[H\left(\dfrac{u_i}{u_T}\right)\right]}{d\left[\dfrac{u_i}{u_T}\right]} = -\frac{R_6 I_0 \exp\left(\dfrac{u_i}{u_T}\right)}{\left[1+\exp\left(\dfrac{u_i}{u_T}\right)\right]^2} < 0 \qquad (5)$$

for all u_i.

It is possible to replace the zero slope condition by the inclusion of an additional stage by allowing R_5 to take a finite value:

$$U_0(u_i) = \frac{-R_5 R_6}{(R_5+R_6)} I_1(u_i) + \frac{R_6(R_3+R_4)}{R_4(R_5+R_6)} u_i + \frac{R_5}{(R_5+R_6)} U_{cc} \qquad (6)$$

where

$$I_1(u_i) = \frac{I_0}{1+\exp\left(\dfrac{u_i}{u_T}\right)} \qquad (7)$$

and

$$U_0(0) = \frac{-R_5 R_6}{(R_5+R_6)} I_1(0) + \frac{R_5}{(R_5+R_6)} U_{cc} \qquad (8)$$

The new transfer function is given by:

$$H\left[\frac{u_i}{u_T}\right] = \frac{R_5 R_6}{(R_5+R_6)} \times [-I_1(0) + I_1(u_i)] - \frac{R_6(R_3+R_4)}{R_4(R_5+R_6)} u_i \qquad (9)$$

where

$$I_1(0) = \frac{I_0}{2}$$

Equation (9) reduces to:

$$H\left[\frac{u_i}{u_T}\right] = \frac{R_5 R_6 I_0}{(R_5+R_6)}\left[-\tfrac{1}{2} + \frac{1}{1+\exp\left(\dfrac{u_i}{u_T}\right)}\right] - \frac{R_6(R_3+R_4)}{R_4(R_5+R_6)} u_i \qquad (10)$$

For the slope to be zero at the peak value $\hat{u}_i$:

$$\left.\frac{d\left[H\left(\dfrac{u_i}{u_T}\right)\right]}{d\left(\dfrac{u_i}{u_T}\right)}\right|_{u_i = \pm\hat{u}_i} = -\frac{R_5 R_6}{(R_5+R_6)} \times \frac{I_0 \exp\left(\dfrac{\hat{u}_i}{u_T}\right)}{\left[1+\exp\left(\dfrac{\hat{u}_i}{u_T}\right)\right]^2} - \frac{R_6(R_3+R_4) u_T}{R_4(R_5+R_6)} = 0 \qquad (11)$$

and from equations (10) and (11)

$$H\left(\frac{u_i}{u_T}\right) = \frac{R_5 R_6}{(R_5+R_6)} I_0 \left\{-\tfrac{1}{2} + \frac{1}{1+\exp\left(\dfrac{u_i}{u_T}\right)} + \frac{\exp\left(\dfrac{\hat{u}_i}{u_T}\right)}{\left[1+\exp\left(\dfrac{\hat{u}_i}{u_T}\right)\right]^2}\frac{u_i}{u_T}\right\} \qquad (12)$$

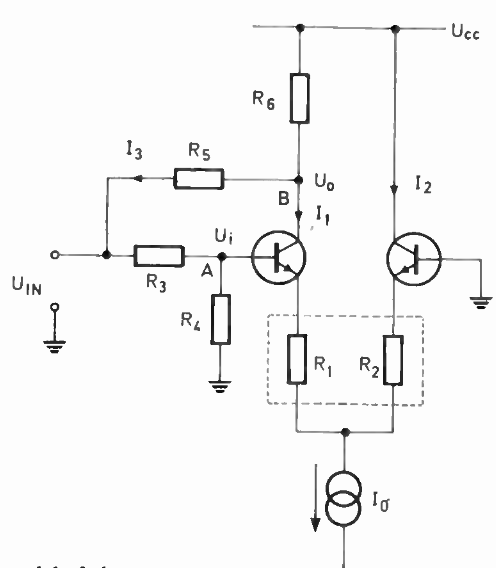


Fig. 1. Circuit model of shaper.

Only one quantity, the relative peak input voltage, is left to be adjusted. The best way to adjust this parameter is by minimizing the distortion of the resulting sinewave. This can be accomplished using the theory developed in Appendix 1. A BASIC program yielded the following results:

$$\text{t.h.d.} = 0{\cdot}0169\% \quad \text{when} \quad \hat{u}_i = 1{\cdot}197u_T \tag{13}$$

and

$$\frac{R_3}{R_4} = \frac{\hat{U}_{in}}{1{\cdot}197u_T} - 1$$

where $\hat{U}_{in}$ is the peak value of the triangle wave in Fig. 1.

This distortion figure means a great improvement compared with the best figures possible with the simple linear segment approximation techniques.[1,2] The resulting deviation of the transfer function from an ideal sine function incorporating the above value is shown in curve A of Fig. 2. Close examination reveals the curvature to be too high in a region roughly approximated by $-\hat{u}_i < 2u_i < \hat{u}_i$. This can be reduced by inserting small resistances R_1 and R_2 into the emitters to TR1 and TR2 respectively. The new expression for collector current is contained in the equation:

$$\frac{u_i}{u_T} = -\ln\left\{\frac{1}{\left[\frac{I_1(u_i)}{I_0}\right]} - 1\right\} - \frac{I_0 R_2}{u_T} + \frac{I_1(u_i)}{I_0}(R_1+R_2)\frac{I_0}{u_T} \tag{14}$$

The values for collector current have to be calculated by approximation for every value of the differential voltage u_i because no explicit relationship $I_1(u_i)$ exists. The first derivative, however, can easily be found:

$$\frac{d\left[\frac{I_1(u_i)}{I_0}\right]}{d\left(\frac{u_i}{u_T}\right)} = \frac{1}{\dfrac{1}{\frac{I_1(u_i)}{I_0}\left[1 - \frac{I_1(u_i)}{I_0}\right]} + \dfrac{I_0}{u_T}(R_1+R_2)} \tag{15}$$

Differentiating equation (9)

$$\frac{d\left[H\left(\frac{u_i}{u_T}\right)\right]}{d\left[\frac{u_i}{u_T}\right]} = \frac{R_5 R_6 I_0}{R_5+R_6}\,\frac{d\left[\frac{I_1(u_i)}{I_0}\right]}{d\left[\frac{u_i}{u_T}\right]} - \frac{R_6(R_3+R_4)}{R_4(R_5+R_6)}u_T \tag{16}$$

Substituting for

$$\frac{d\left[\frac{I_1(u_i)}{I_0}\right]}{d\left[\frac{u_i}{u_T}\right]}$$

from equation (15) and equating to zero to satisfy the peak condition:

$$\left.\frac{d\left[H\left(\frac{u_i}{u_T}\right)\right]}{d\left[\frac{u_i}{u_T}\right]}\right|_{u=\pm\hat{u}} = \frac{R_5 R_6 I_0}{R_5+R_6} \times \frac{1}{\dfrac{1}{\frac{I_1(\hat{u}_i)}{I_0}\left[1-\frac{I_1(\hat{u}_i)}{I_0}\right]} + \dfrac{I_0(R_1+R_2)}{u_T}} - \frac{R_6(R_3+R_4)u_T}{R_4\,(R_5+R_6)} = 0 \tag{17}$$

Using equation (17), equation (9) reduces to:

$$H\left(\frac{u_i}{u_T}\right) = \frac{R_5 R_6 I_0}{(R_5+R_6)}\left\{-\frac{I_1(0)}{I_0} + \frac{I_1(u_i)}{I_0} - \frac{\frac{u_i}{u_T}}{\frac{I_1(\hat{u}_i)}{I_0}\left[1-\frac{I_1(\hat{u}_i)}{I_0}\right] + \frac{I_0}{u_T}(R_1+R_2)}\right\}$$

The values $I_1(0)$, $I_1(u_i)$ and $I_1(\hat{u}_i)$ are calculated according to equation (14).

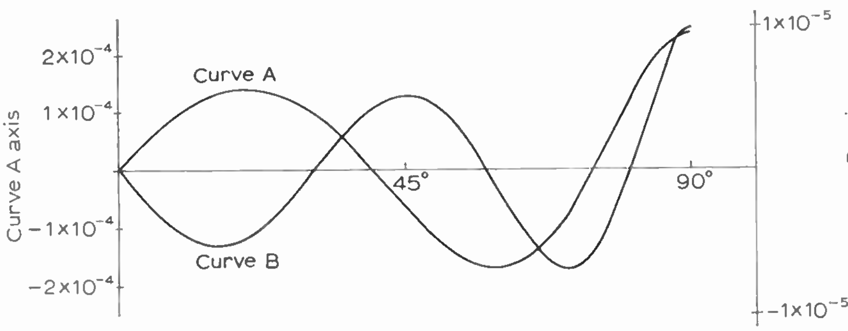


Fig. 2. Deviation from an ideal sinewave.

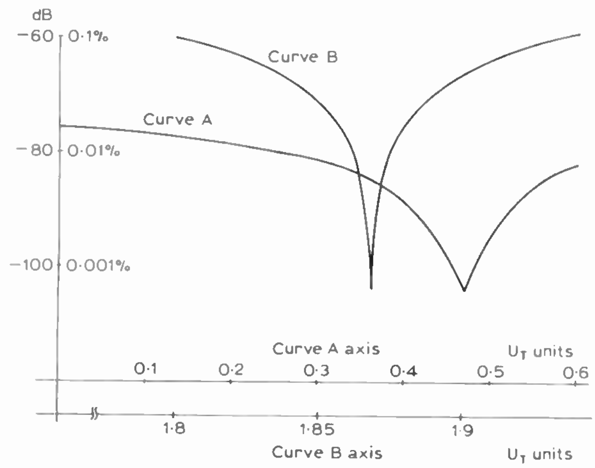


Fig. 3. Total harmonic distortion versus voltage drop across R_1 or R_2 (Curve A) and $\hat{U}_i$ (Curve B).

Thus a further parameter 'the relative voltage drop'

$$\frac{R_1+R_2}{2}\,\frac{I_0}{u_T}$$

is required. Figure 3 demonstrates the theoretical performance of the network for non-ideal settings of (R_1+R_2) and $\hat{u}_i$. Again a computer analysis established an optimum setting when:

$$\frac{R_1+R_2}{2}=0{\cdot}47\,\frac{u_T}{I_0} \quad \text{and} \quad \hat{u}_i=1{\cdot}8685u_T \qquad (17)$$

and

$$\frac{R_3}{R_4}=\frac{\hat{U}_{in}}{1{\cdot}8685u_T}-1$$

Using these values the t.h.d. is computed as 6×10^{-6}. The deviation of the transfer function from the ideal sine characteristic incorporating the above values is shown in curve B of Fig. 2.

4 Practical Circuit Considerations

The practical circuit on which measurements were taken is shown in Fig. 4. As with similar circuits of this nature there are three main aspects deserving of attention:

(1) Transfer characteristic stability

(2) D.c. stability

(3) Frequency performance.

The transfer characteristic of the shaper will deviate from its original characteristic if any of the design parameters that have influence on the transfer characteristic change their values. These are the temperature voltage u_T, and the current I_0.

The change in u_T can either be compensated, using a temperature-dependent potential divider at the input, or reduced by placing the differential transistor pair in a constant-temperature enclosure. The former approach is popular with designers of logarithmic modules. From the theoretical calculations of the shaper performance, it is known that the triangular wave present at the collector is 2·57 times the sinewave. About 12% of the triangular

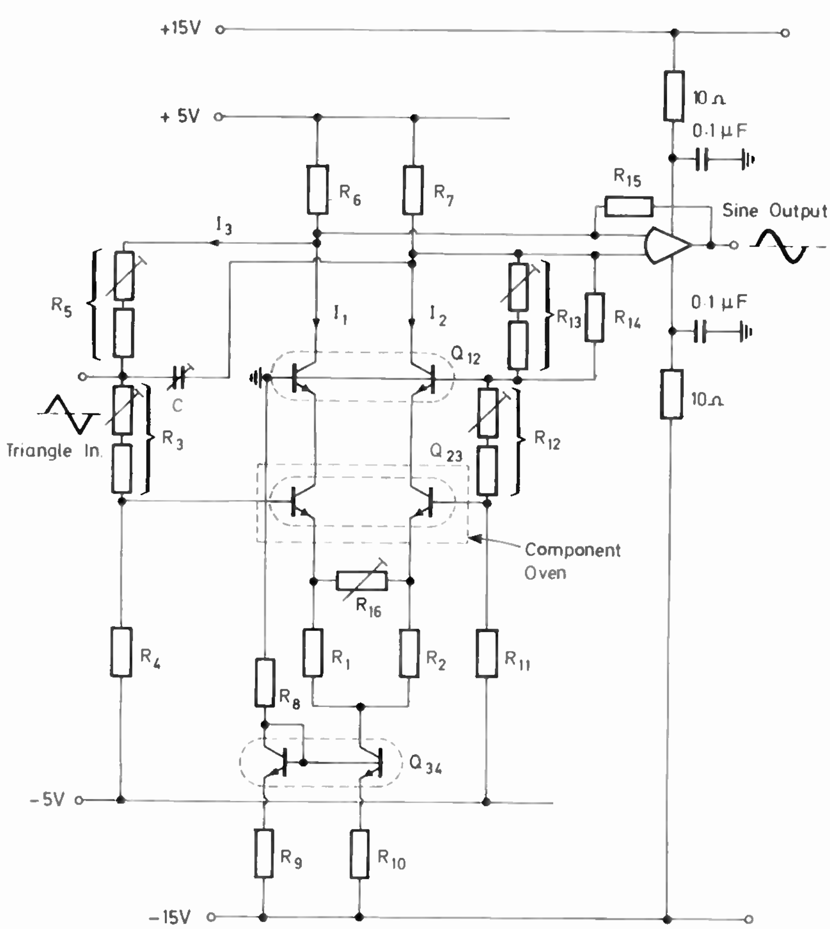
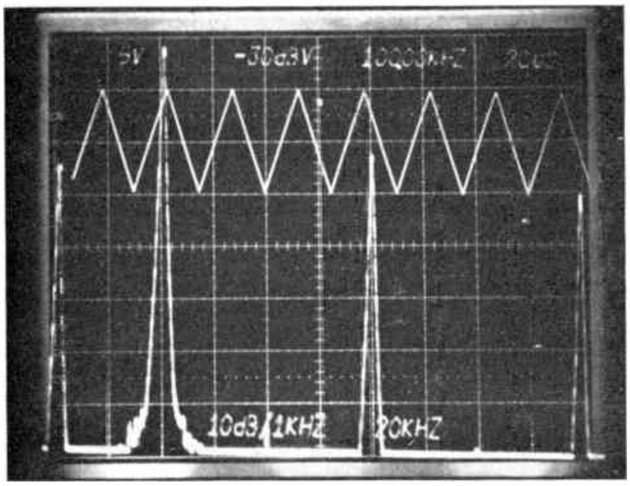


Fig. 4. Practical circuit of transconductance sine shaper.

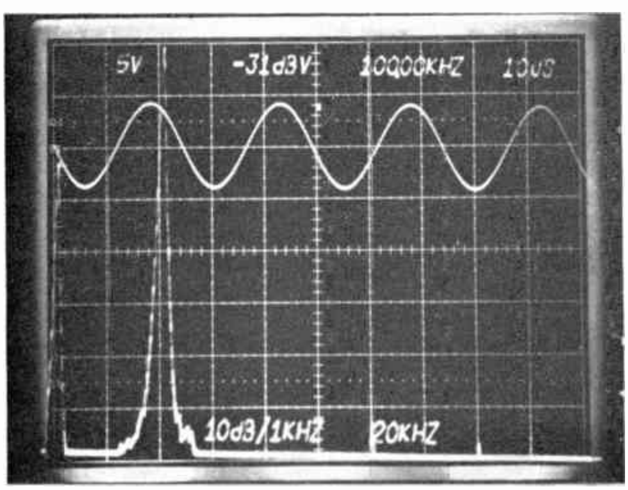
A B C D



E F

(a) Test signal;

A B C D



E F

(b) Shaper output.

Fig. 5. Time and frequency domain performance of shaper.

Scales:

TIME DOMAIN
- A volts per div.
- D time per div.

FREQUENCY DOMAIN
- B reference level (graticule top)
- C dot frequency
- E resolution bandwidth
- F frequency span per div.

wave is harmonic distortion; thus the distortion due to changes in I_0 is 0·31 times the relative changes. This means that the current I_0 has to be controlled very exactly. A current mirror was used to accurately define I_0.

Any changes in d.c. levels will not influence the final distortion figures provided the transfer characteristic remains unchanged. Temperature effects were minimized in all the experimental work by mounting the shaper pair of transistors in a miniature temperature-controlled oven. Changes in output d.c. level, though not contributing to distortion itself, should be kept as low as possible in a precision sine shaper. The obvious and most effective solution is the provision of differential outputs followed by a differential post amplifier. The use of a cascode configuration reduces voltage changes at the collectors of the shaper pair to a minimum, maintaining the integrity of the exponential conformance characteristic as well as improving overall frequency response. A comprehensive discussion of suitable devices, together with anticipated performance when operating in a non-linear mode, is given by Gilbert.[9] To alleviate the low-pass characteristic of the non-linear stage, feed-forward compensation was used.

5 Results and Conclusions

One of the problems not mentioned during the theoretical discussion, but encountered during the building and testing of the circuit, was the effect of noise. Transistor noise is always a problem in circuits where low level signals are to be amplified; so it was in this circuit with the non-linear transistor pair. In the initial test set-up, 'popcorn' noise, in particular, was found to be very damaging to the performance of the circuit because this type of noise cannot be easily filtered.

Obtaining a suitable test triangle presented its own inevitable problems, because commercial equipment—particularly the specification on second harmonic distortion—was found unequal to the task. A triangle/square generator of high purity has been designed to provide the required test signal. Figure 5(a) shows the purity of the test signal in both time and frequency domains. The second harmonic component in the triangle wave is some 70 dB down on the fundamental. Figure 5(b) shows the shaper output characteristics. Second harmonic remains low, 80 dB down on the

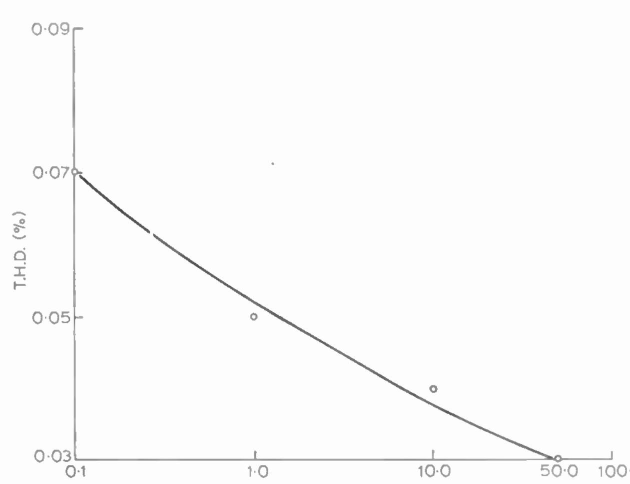


Fig. 6. Graph of t.h.d. versus frequency.
○ measured t.h.d.

fundamental and third harmonic is 73 dB down on the fundamental. Fourth harmonic has increased by 2 dB, but is 76 dB down on the fundamental. Measurements taken using a t.h.d. meter, are given in the graph of Fig. 6.

In conclusion, a new transconductance triangle to sinewave shaper has been thoroughly investigated with a prototype, realizing t.h.d. figures that are an order of magnitude improvement on readily available equipment. The circuit itself is simple in conception, and readily lends itself to integration. Special care must be exercised in realizing a good practical lay-out and low-drift, low-noise power supplies are necessary to achieve the best results. Although lying outside the scope of this paper, the purity of the triangle wave plays a vital role in the final outcome of the sinewave. In the experience of the authors, this factor is often overlooked in commercial designs.

The results of this work are significant inasmuch as the potential for generation of a truly low distortion sinewave (better than 0·01%) is possible with further refinements to this technique. This raises the possibility of swept frequency measurements to new accuracies, since an oscillator based on this method is readily programmable (the test generator had a 10000:1 sweep). A limit of 10^5 hertz has been quoted for this work, with the possibility of extension in upper frequency response.

The sensitivity of the circuit temperature is conceded and an estimate of this effect is shown in the graph of Fig. 7 where the circuit itself has been previously set up for minimum distortion at the miniature oven temperature of 63°C and then subjected to temperature variations in a second temperature enclosure with the miniature oven removed. Hot substrate techniques or an input potential divider having the requisite temperature coefficient to cancel the u_T term in the transfer function are possible solutions to this problem. A third solution is to vary the peak to peak values of the triangle wave with temperature. This calls for temperature sensitive trip levels in the triangle generator itself and is currently the subject of further investigation.

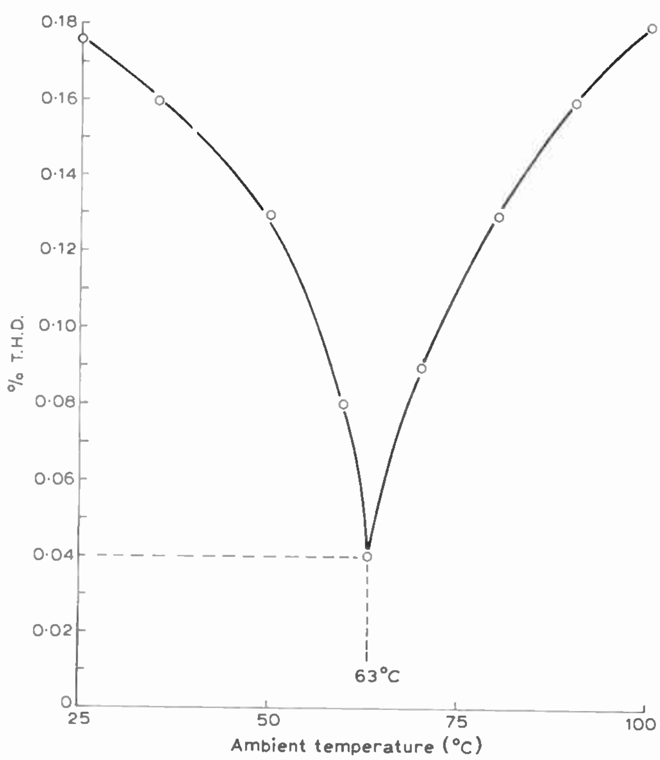


Fig. 7. Percentage total harmonic distortion v. ambient temperature for shaper pair at 10 kHz.

6 Acknowledgments

The authors wish to thank Senior Research Assistant, Mr. E. M. Davies, for many useful suggestions and contributions during the building and testing of the prototype.

7 References

1 Barnes, L., 'Linear segment approximations to a sinewave', *Electronic Engineering*, **40**, p. 502, 1968.

2 Schiffer, V. and Evans, W. A., 'The trapezoidal synthesis of a tri-wave to sine converter'. (To be published.)

3 Ritchie, C. G. and Young, R. W., 'The design of biased diode function generators', *Electronic Engineering*, **31**, p. 347, 1959.

4 Klein, G. and Hagenbeuk, H., 'Accurate triangle-sine converter', *Electronic Engineering*, **39**, p. 700, 1967.

5 Paull, C. J. and Evans, W. A., 'Waveform shaping for the design of signal sources', *The Radio and Electronic Engineer*, **44**, p. 523, 1974.

6 Middlebrook, R. D. and Richer, I., 'Non-reactive filter converts triangle waves to sines', *Electronics*, **38**, No. 5, p. 96, 1965.

7 Grebene, A. B., 'Monolithic waveform generation', *IEEE Spectrum*, **9**, p. 34, 1972.

8 Meyer, R. G., Sansen, W. M. C., Lui, S. and Peeters, S., 'The differential pair as a triangle-sine wave converter', *IEEE J. Solid-State Circuits*, SC-11, No. 3, p. 418, 1976 (Corresp.).

9 Gilbert, B., 'Log circuit applications', in 'Non-Linear Circuits Handbook', Chap. 4, Sheingold, D. H. (Ed.) (Analog Devices Inc., Norwood, Mass., 1900).

8 Appendix 1: Total harmonic distortion of low distortion cosine waves

The calculation of t.h.d. requires the solution of four integrals:

$$[\text{t.h.d.}]^2 = \frac{\frac{1}{T}\int_0^T f^2(t)\,dt - \frac{a_0^2}{4} - \frac{1}{2}(a_1^2 + b_1^2)}{\frac{1}{T}\int_0^T f^2(t)\,dt - \frac{a_0^2}{4}} \qquad (18)$$

where the distorted cosine wave $f(t)$ is given by:

$$f(t) = \frac{a_0}{2} + \sum_{n=1}^{\infty} a_n \cos\left(2\pi n \frac{t}{T}\right) + \sum_{n=1}^{\infty} b_n \sin\left(2\pi n \frac{t}{T}\right)$$

n = integer and

$$a_0 = \frac{2}{T}\int_0^T f(t)\,dt, \quad a_1 = \frac{2}{T}\int_0^T f(t)\cos\left(2\pi\frac{t}{T}\right)dt,$$

$$b_1 = \frac{2}{T}\int_0^T f(t)\sin\left(2\pi\frac{t}{T}\right)dt$$

Since the d.c. terms do not influence distortion figures, a_0 can be offset by addition of a constant and may thus be neglected. Equation (18) is simplified to:

$$[\text{t.h.d.}]^2 = \frac{\frac{1}{T}\int_0^T f^2(t)\,dt - \frac{1}{2}(a_1{}^2 + b_1{}^2)}{\frac{1}{T}\int_0^T f^2(t)\,dt} \qquad (19)$$

If f(t) is decomposed into a pure cosine wave and the residual distortion D(t) then:

$$f(t) = A\left[\cos\left(2\pi\frac{t}{T}\right) + D(t)\right] \qquad (20)$$

$$a_1 = \frac{2}{T}\int_0^T A\left[\cos\left(2\pi\frac{t}{T}\right) + D(t)\right]\cos\left(2\pi\frac{t}{T}\right)dt$$

$$= A\left[1 + \frac{2}{T}\int_0^T D(t)\cos\left(2\pi\frac{t}{T}\right)dt\right] \qquad (21)$$

$$b_1 = \frac{2A}{T}\int_0^T \left[\sin\left(2\pi\frac{t}{T}\right)\cos\left(2\pi\frac{t}{T}\right) + {} \right.$$

$$\left. {} + D(t)\sin\left(2\pi\frac{t}{T}\right)\right]dt$$

$$= \frac{2A}{T}\int_0^T D(t)\sin\left(2\pi\frac{t}{T}\right)dt \qquad (22)$$

and

$$\frac{1}{T}\int_0^T f^2(t)\,dt = \frac{A^2}{T}\int_0^T \left[\cos^2\left(2\pi\frac{t}{T}\right) + {}\right.$$

$$\left. {} + 2\cos\left(2\pi\frac{t}{T}\right)D(t) + D^2(t)\right]dt$$

$$= \frac{A^2}{2} + A(a_1 - A) + \frac{A^2}{T}\int_0^T D^2(t)\,dt \qquad (23)$$

Substituting into (19).

$$[\text{t.h.d.}]^2 = \frac{-\frac{1}{2}(A - a_1)^2 - \frac{b_1{}^2}{2} + \frac{A^2}{T}\int_0^T D^2(t)\,dt}{\frac{A^2}{2} + A(a_1 - A) + \frac{A^2}{T}\int_0^T D^2(t)\,dt}$$

$$= \frac{-\left[\frac{2}{T}\int_0^T D(t)\cos\left(2\pi\frac{t}{T}\right)dt\right]^2 - \left[\frac{2}{T}\int_0^T D(t)\sin\left(2\pi\frac{t}{T}\right)dt\right]^2 + \frac{2}{T}\int_0^T D^2(t)\,dt}{1 + \frac{4}{T}\int_0^T D(t)\cos\left(2\pi\frac{t}{T}\right)dt + \frac{2}{T}\int_0^T D^2(t)\,dt} \qquad (24)$$

For low values of distortion $D(t) \ll 1$ so that

$$[\text{t.h.d.})^2 \simeq \frac{2}{T}\int_0^T D^2(t)\,dt - $$

$$-\left[\frac{2}{T}\int_0^T D(t)\cos\left(2\pi\frac{t}{T}\right)dt\right]^2 - $$

$$-\left[\frac{2}{T}\int_0^T D(t)\sin\left(2\pi\frac{t}{T}\right)dt\right]^2 \qquad (25)$$

Equation (25) is more useful than (18) for computation purposes since in the former case significant digits are lost in the subtraction operation. With equation (25) no significant errors occur during the integration since the value of D(t) always has less significant digits than the number of digits used by the calculator.

Manuscript first received by the Institution on 2nd August 1976 and in final form on 23rd December 1976 (Paper No. 1766/CC271).

UDC 621.396.673

Broadband RC-loaded microwave cylindrical antenna with approximately real input admittance

Dj. S. PAUNOVIĆ, M.Sc.*

and

Professor B. D. POPOVIĆ, D.Sc.*

SUMMARY

A method is presented for theoretical synthesis of loaded broadband cylindrical antenna with minimized input susceptance. Such an RC-loaded antenna was synthesized and realized, having the v.s.w.r. with respect to a real impedance of about 1·1 in a frequency range of almost 3:1, thus being probably the best cylindrical broadband antenna available. The antenna radiation pattern was found to be reasonably stable in the whole frequency range.

* Department of Electrical Engineering, University of Belgrade, P.O. Box 816, 11001 Belgrade, Yugoslavia.

1 Introduction and Remarks on Broadband Antenna Matching to Feeder

Cylindrical broadband antennas are of considerable practical interest in various applications where an omnidirectional broadband antenna is required. To improve their broadband properties, cylindrical antennas have been loaded along their length with either capacitive[1,2], resistive[3] or combined[4] loading. Antennas having remarkable broadband characteristics in both admittance and radiation pattern were thus obtained.

Unfortunately, all these antennas have rather large susceptance in the whole frequency range of interest. While in the case of a narrow frequency band this difficulty can be easily overcome by connecting a convenient reactive matching network between the antenna terminals, this is not in general possible for broadband antennas. The reason is that, by Foster's reactance theorem, reactance, X, and susceptance, $B = -1/X$, of a linear reactive network always increase with frequency, (i.e., if negative, with increasing frequency they become smaller in magnitude), while the antenna susceptance does not generally decrease with frequency, and cannot in all cases be compensated by a reactive compensating network connected between the antenna terminals. Thus a broadband matching of an antenna is in general a complicated task, which might even require synthesis and realization of an active network. This fact makes the above mentioned broadband cylindrical antennas much less attractive for practical applications than their, otherwise excellent, broadband properties seem to indicate.

The present paper is aimed at describing a simple method for obtaining a loaded broadband cylindrical antenna with virtually zero susceptance in a relatively wide frequency range. To understand better the novelty contained in the method, note first that it is possible to analyse either experimentally or numerically the frequency behaviour of a number of antennas with the desired kind of loadings. All those antennas having susceptance which decreases in a certain frequency range then can, in principle, be matched in that frequency range to a feeder by means of a reactive compensating network (which has to be synthesized) connected between the antenna terminals, and a Chebyshev transformer. Unfortunately, this is usually not a simple task, particularly at microwave frequencies, because network synthesis and its realization at these frequencies is most often very intricate.

However, by varying the antenna loading, frequency dependence of its susceptance can be modified substantially. Therefore, basically opposite procedure is also possible, i.e. first to adopt a convenient simple reactive element with known frequency behaviour, and then to synthesize the loading along the antenna in order to obtain the antenna susceptance which compensates the reactive element susceptance. A certain insight into possibilities offered by variation of the loadings along a cylindrical antenna in this case is of course necessary prior to adopting the reactive element, but for microwave antennas considered in the present paper this method is nevertheless more appropriate, because synthesis and realization of a microwave broadband compensating reactive network is thereby avoided. It will be shown that in this manner loaded cylindrical antennas can be obtained having excellent broadband properties and quite small input susceptance in a relatively wide frequency range.

2 Brief Theory and Experimental Model

Theoretical determination of current distribution $I(z)$ along a monopole antenna of radius a and height h, its axis coinciding with the z-axis of a co-ordinate system and driven by a coaxial line, for different magnitude and distribution of the loading along the antenna can be achieved starting from the integral equation for current of Hallén's type. For an antenna with n concentrated loadings and arbitrary distribution of continuous loadings this integral equation is of the form

$$\int_{-h}^{h} I(s)F(z,s)\,\mathrm{d}s + \int_{0}^{z} I(s)R(z,s)\,\mathrm{d}s + \sum_{i=1}^{n+1} I(z_i)H(z,z_i) = G(z). \quad (1)$$

In this equation,

$$F(z,s) = \frac{1}{r}\exp(-\mathrm{j}\beta r) - \frac{\cos\beta z}{r_0}\exp(-\mathrm{j}\beta r_0) \quad (2)$$

$$r = \sqrt{[a^2+(z-s)^2]}, \quad r_0 = r(z=0) \quad (3)$$

$$R(z,s) = -\mathrm{j}2\frac{2\pi}{\zeta_0}Z_p(s)\sin\beta(z-s) \quad (4)$$

$$H(z,z_i) = \begin{cases} \mathrm{j}Z_{ci}\dfrac{2\pi}{\zeta_0}2\sin\beta(z_i-z), & z_i<z \\ 0, & z_i \geqslant z \end{cases} \quad (5)$$

$$G(z) = -\mathrm{j}\frac{2\pi}{\zeta_0}VD(z). \quad (6)$$

$D(z)$ for the so-called belt-generator approximation to coaxial-line excitation is given in Reference 7. In equations (2)–(6) ζ_0 is the intrinsic impedance of a vacuum and β the free-space propagation coefficient. Further, z_i designates the positions of the concentrated loadings, Z_{ci} are the impedance values of the concentrated loadings and $Z_p(z)$ stands for the function describing the distribution of the continuous loading. Equation (1) can be solved numerically by approximating, for example, current $I(z)$ by a set of polynomials with unknown complex coefficients $I_{i,k}$,

$$I(z) \approx \sum_{k=1}^{n_i+1} I_{i,k}\left(\frac{z}{z_{i+1}}\right)^{k-1},$$
$$z\in[z_i, z_{i+1}], \quad i=1,2,\ldots(n+1) \quad (7)$$

and then applying the point-matching method. This can be done by combining the methods described in References 5, 6 and 7, and will not be further elaborated here. Once $I(z)$ is determined, we obtain the antenna admittance as $Y_{ant} = G_{ant} + \mathrm{j}B_{ant} = I(0)/V$.

It is found theoretically and confirmed experimentally that the RC-loaded (as well as only C or R loaded) cylindrical antennas usually have driving-point susceptance positive in a wide frequency range. In accordance with the method outlined in the preceding Section, an inductive coil was adopted as the compensating network, being the simplest for realization. Thus, theoretically,

$$B_{\text{comp.net.}} = -1/\omega L_c. \quad (8)$$

Following the proposed method, we next adopt the antenna dimensions, and determine the distributed and concentrated loadings along the antenna which minimize the minimization functional

$$P = \int_{f_1}^{f_2} |B_{ant}(f) - 1/(2\pi f L_c)|\,\mathrm{d}f. \quad (9)$$

In this equation L_c is also considered to be an unknown parameter to be determined from the optimization process, and f_1 and f_2 are the lower and upper frequencies of the frequency range considered. In this manner both the loading along the antenna and the magnitude of the compensating coil inductance, which result in approximately real input antenna admittance, are obtained. Of several optimization methods that can be used, a slight modification of the simplex method[8] was adopted and found satisfactory.

Of particular interest in practice is that the antenna also has as constant as possible input conductance, in order that the Chebyshev transformer matching can be applied. Therefore the optimization functional given in equation (9) can be modified to require also that the antenna conductance be as close as possible to the average conductance in the frequency range $[f_1, f_2]$.

Although the method described above can be applied for any type of the loading along the antenna, in practice it is restricted to cases which can more or less easily be realized. At microwave frequencies it is very difficult to obtain a pure concentrated or distributed inductive loading, as well as pure concentrated resistive loading or distributed capacitive loading. On the other hand, concentrated capacitive loading can easily be obtained,[2] and not difficult to realize is also a distributed resistive loading.[4] It has been already shown that cylindrical antennas with concentrated capacitive and distributed resistive loading can offer very interesting possibilities,[4] and are relatively easy to make. An experimental model was therefore adopted similar to that described in Reference 4. Briefly, it consisted of a row of cylindrical resistors with metallic endings between which teflon disks were inserted to obtain lumped capacitors. The first segment was made of brass, i.e. with practically zero internal resistance. Contrary to the model described in

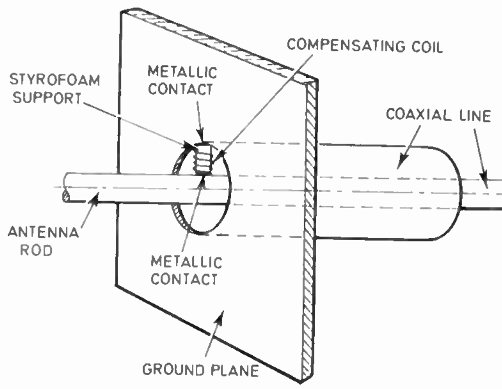


Fig. 1. Monopole antenna excitation zone with compensating element.

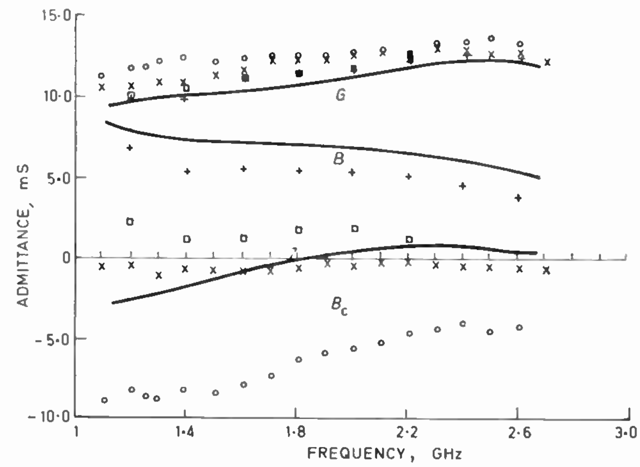


Fig. 2. Real (G), imaginary (B) and imaginary compensated (B_c) parts of the RC-loaded cylindrical monopole antenna admittance versus frequency, $a = 3{\cdot}5$ mm, $h = 17{\cdot}75$ cm.

——— theoretical; + + + + experimental, without compensation; × × × × experimental, optimally compensated; □ □ □ □ experimental, under-compensated; ○ ○ ○ ○ experimental, over-compensated.

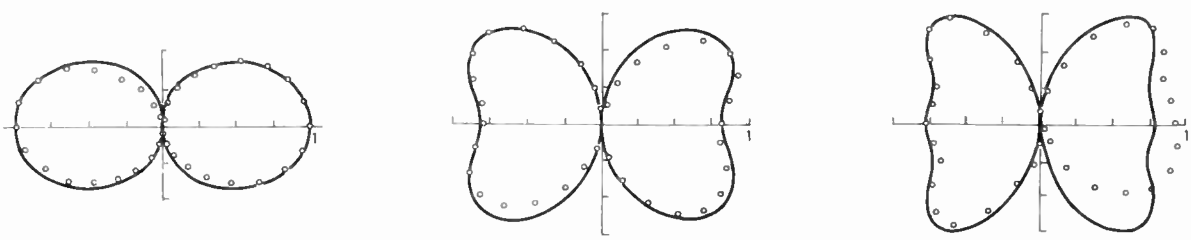
Reference 4, the present antenna was made as a self-supporting structure. To achieve this, a small hole was drilled along the axis of all the antenna elements, and the antenna parts were held together by means of a stretched nylon filament passing through the hole. By a simple device the nylon filament could be relaxed, the antenna dismounted and then assembled again from new desired elements.

3 Numerical and Experimental Results

As an example of the method outlined above, consider a cylindrical RC-loaded monopole antenna of diameter $2a = 0{\cdot}7$ cm, made of four segments (i.e. three lumped capacitors), driven by a coaxial line of the inner diameter of the outer conductor $2b = 1{\cdot}38$ cm. The values of the three lumped capacitors were considered as variables, as well as the length of the first brass segment and the values of the continuous resistive loading of the other three segments, each of the length 4·8 cm. Available resistance values of the resistive segments were 50, 100, 200 and 400 Ω, i.e. about 1050, 2100, 4200 and 8400 Ω/m, respectively, which were incorporated into the optimization process as the only possible resistance values.

Numerical optimization resulted in the length of the first brass segment equal to 3·11 cm, and in resistances of the other three segments of 100, 200, and 400 Ω, respectively. The optimal capacitances of the concentrated capacitors were found to be approximately 1·43 pF, 0·65 pF and 0·4 pF, counted from the excitation zone towards the antenna end. Finally, the optimal value of the compensating coil inductance was found to be approximately $L_c = 15$ nH. The compensating coil was made in the form of few turns of very thin wire (of radius 0·09 mm), wound on a short styrofoam cylinder of radius 4 mm. At both ends of the coil very small metallic contacts were made and the coil was fixed at the coaxial line opening, between the antenna rod and the outer coaxial line conductor (Fig. 1). By a rough calculation it was found that the coil should have approximately 2 turns of wire, but the precision of this result was quite doubtful. Therefore several coils were made and the optimal one, having approximately 1·75 turns, was determined experimentally.

Shown in Fig. 2 are the computed and measured antenna driving-point conductance G and susceptance B versus frequency. For comparison, curves are also shown for under-compensation (too high value of L_c, approximately 3 turns) and over-compensation (too small value of L_c, approximately 1 turn), as well as the measured results without compensation. Excellent broadband properties of the optimally compensated antenna can be observed, comparable to those of a much more complicated structure like a log-periodic antenna with over ten elements.[9] Figure 3 shows radiation patterns of the



1·1 GHz, 1·8 GHz, 2·7 GHz.

Fig. 3. Dipole antenna radiation patterns. ○ ○ ○ ○ experimental; ——— theoretical.

Table 1. Theoretical and experimental average antenna parameters

	Without compensation		Optimal compensation	
	theory	experiment	theory	experiment
Frequency range GHz	1·1–2·7	1·2–2·6	1·1–2·7	1·1–2·7
Average admittance mS	11·04+j6·84	11·29+j5·25	11·04−j0·52	12·07−j0·51
Average reflection coefficient %	30·0	23·1	6·74	3·94
Average v.s.w.r.	1·86	1·60	1·14	1·08

antenna at three frequencies. The patterns have the expected shapes, typical for travelling-wave linear antennas, and are quite stable in a wide frequency range.

By comparing theoretical and experimental susceptance curves it is obvious that the compensating coil susceptance does not vary with frequency exactly as $(-1/\omega L_c)$. Also, in all the cases with compensating element the measured G-curves are affected as well. Fortunately actual frequency behaviour of the coil appears to be more favourable for our purpose than the theoretical one.

For convenience, Table 1 summarizes some of the results shown in Fig. 2 in a different form. The v.s.w.r.s and the reflection coefficients were computed corresponding to the average values of conductance G in the considered frequency range. It should also be noted that theoretically obtained efficiency of the antenna was over 80% in the whole frequency range. If a higher antenna efficiency is required, the optimization functional can of course be modified to include efficiency as a parameter to be maximized. However, to achieve performances comparable to those of the antenna described above, a smaller frequency range than in the present example should be adopted.

4 Conclusion

Matching of broadband antennas to a feeder by a reactive matching network is possible only if the antenna susceptance decreases with frequency in the frequency range of interest. By appropriately loading a cylindrical antenna this condition can in some cases be achieved and the corresponding broadband reactive matching network then synthesized and constructed. A more suitable method at microwave frequencies is to adopt a compensating element of known susceptance frequency behaviour, and then to synthesize the loading along the antenna resulting in antenna susceptance which compensates the adopted element susceptance, thus avoiding complicated synthesis and realization of a microwave reactive broadband compensating network. As an example it was shown theoretically and experimentally that excellent RC-loaded broadband cylindrical antennas can be obtained in this manner, having average v.s.w.r. with respect to a real admittance of the order of 1·1 in a frequency range of about 3:1. The authors are unaware of any other cylindrical antenna having comparable broadband properties.

5 References

1 Hallén, E., 'Electromagnetic Theory', pp. 501–4 (Chapman and Hall, London, 1961).

2 Popović, B. D. and Dragović, M. B., 'Capacitively loaded thin cylindrical antenna', *Proc. Instn Elect. Engrs*, **118**, p. 101, February 1974.

3 Wu, T. T. and King, R. W. P., 'The cylindrical antenna with non-reflecting resistive loading', *IEEE Trans. on Antennas and Propagation*, **AP-13**, p. 369, September 1965.

4 Popović, B. D., Dragović, M. B. and Paunović, Dj. S., 'Broadband cylindrical antenna with continuous resistive and concentrated capacitive loadings', *Electronics Letters*, **11**, Nos. 25/26, p. 611, 11th December 1975.

5 Popović, B. D., 'Theory of cylindrical antennas with lumped impedance loadings', *The Radio and Electronic Engineer*, **43**, No. 4, p. 243, April 1973.

6 Popović, B. D., 'Theory of cylindrical antennas with arbitrary impedance loading', *Proc. Instn Elect. Engrs*, **118**, No. 10, p. 1327, October 1971.

7 Popović, B. D., 'Thin monopole antenna: finite-size belt-generator representation of coaxial-line excitation', *ibid.*, **120**, No. 5, p. 544, May 1973.

8 Nelder, J. A. and Mead, R., 'A simplex method for function minimization', *Computer J.*, **7**, p. 308, 1965.

9 Mittra, R. (Editor), 'Numerical and Asymptotic Techniques in Electromagnetics', pp. 25–27 (Springer-Verlag, Berlin, 1975).

Manuscript received by the Institution on 26th October 1976. (Short Contribution No. 185/Com. 147).

UDC 621.673: 621.677.8

Synthesis of broadband cylindrical monopole antenna with parasitic elements

Professor B. D. POPOVIĆ, D.Sc.*

M. B. DRAGOVIĆ, B.Sc.*

and

A. R. DJORDJEVIĆ, M.Sc.*

SUMMARY

A method is presented for synthesis of the optimal broadband cylindrical monopole antenna with two closely spaced identical parasitic elements. The proximity and end effects are taken approximately into account, which results in a better agreement between theoretical and experimental results. A very flexible experimental model is described, which even allows an experimental synthesis of the antenna.

* Department of Electrical Engineering, University of Belgrade, PO Box 816, 11001 Belgrade, Yugoslavia.

1 Introduction

Already for some time it has been known that by adding two parasitic elements at a small distance from and parallel to a dipole antenna near resonance a relatively good broadband antenna could be obtained.[1] In a more recent paper[2] it has been shown that the polynomial approximation of current[3] together with the so-called delta-function representation of the excitation zone can be used for analysis of such structures. The synthesis problem of determining the optimal dimensions of such an antenna by an optimization procedure has not, however, been considered.

The present paper is aimed at describing a simple method for synthesis of optimal monopole antenna with two symmetrical, closely-spaced parasitic elements with respect to the monopole input admittance, taking into account approximately the proximity and end effects. In order that experimental results for monopole antennas driven by a coaxial line could better be compared with theoretically obtained values, the so-called belt-generator approximation to coaxial-line excitation[4] has been used in conjunction with piecewise-polynomial approximation for current[5] along the driven element and polynomial approximation for current along the parasitic elements. In addition, a very flexible experimental antenna model had been constructed, which enabled easy changes of lengths and distances of parasitic elements and length of the driven element. It was found that better agreement between theoretical and experimental results was obtained if the end and proximity effects were taken approximately into account as proposed in the paper.

2 Outline of the Theory

Consider the monopole antenna driven by a coaxial line and with two identical, symmetrically positioned parasitic elements, shown in Fig. 1(a). As demonstrated in Ref. 4, the coaxial-line feed can be approximated by a belt-generator, with the coaxial-line opening closed, and then the image theory applied to obtain the equivalent dipole antenna with two parasitic elements shown in Fig. 1(b).

Approximate simultaneous integral equations for currents $I_1(z)$ and $I_2(z)$ along the driven and the parasitic dipole elements have the following form:

$$\begin{aligned} \int_{-h_1}^{h_1} I_1(s)K_{11}(z,s)\,\mathrm{d}s + \int_{-h_2}^{h_2} I_2(s)K_{12}(z,s)\,\mathrm{d}s &= D(z) \\ \int_{-h_1}^{h_1} I_1(s)K_{21}(z,s)\,\mathrm{d}s + \int_{-h_2}^{h_2} I_2(s)K_{22}(z,s)\,\mathrm{d}s &= 0. \end{aligned} \tag{1}$$

The kernels $K_{mn}(z, s)$ are known functions, and $D(z)$ is a function describing the belt-generator excitation. This system of integral equations can be approximately solved, for example, by assuming current distribution in the form of polynomials with unknown complex coefficients and applying the point-matching method, as in Refs. 2, 4 and 5. On the driven element we approximate current

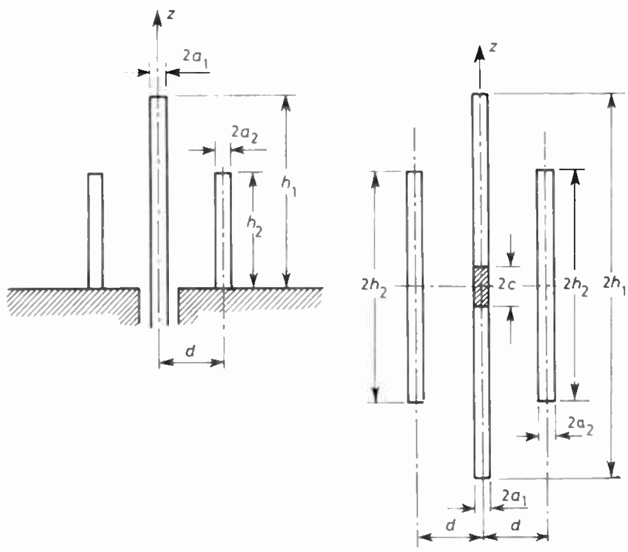


Fig. 1. Sketch of the antenna with two identical parasitic elements.
(a) Coaxial-line feed;
(b) Belt-generator feed of equivalent dipole.

distribution by two polynomials (one along the belt-generator, and the other along the rest of the antenna),

$$I_1(z) = \begin{cases} \sum\limits_{\substack{i=0 \\ i \neq 1}}^{n_1} I_{1i}(z/c)^i, & 0 \leqslant z \leqslant c \\ \sum\limits_{i=0}^{n_2} I_{2i}(z/h_1)^i, & c \leqslant z \leqslant h_1 \end{cases} \tag{2}$$

with constraints that values of the polynomials and their first derivatives at $z = c$ be equal, and that $I_1(h_1) = 0$. Along the parasitic elements we adopt simply

$$I_2(z) = I_p(1 - z^2/h_2^2), \tag{3}$$

because the parasites are electrically short. A higher-order approximation for current along the parasitic elements was found to be unnecessary.

Substituting the expressions (2) and (3) into equation (1) and applying the point-matching method, a system of linear equations in the unknown coefficients I_{1i}, I_{2i} and I_p is obtained. Solutions of these equations can be considered as functions of independent variables h_1, h_2, a_1, a_2, d and frequency f. A convenient optimization function can be constructed and minimized (or maximized) by varying all or some of the independent variables.

Of particular interest in the present case was to synthesize an antenna with approximately real and constant input admittance in a given frequency range. To that aim, first for n frequencies in the desired frequency range the antenna conductances were computed, and their geometric mean value G_0 determined. The moduli $|\rho_i|$, $i = 1, 2, \ldots n$, of the reflection coefficients were then found at the n frequencies with respect to G_0, and the corresponding v.s.w.r.s, σ_i, calculated. The mean value of the v.s.w.r. was then defined as

$$\sigma_{\text{mean}} = \left(\frac{1}{n}\sum_{i=1}^{n} \sigma_i^{\,m}\right)^{1/m}, \tag{4}$$

which served as the minimum function, with $m = 8$.

Determination of $(\sigma_{\text{mean}})_{\text{min}}$ was performed on a digital computer by the pattern search[6] in the plane of the variables d and h_2, with $a_1 = a_2 = a$ and h_1 kept constant. The search was programmed to terminate when simultaneously the step size in d was less than 1·6 mm and in h_2 less than 2·5 mm.

The elements of the antenna considered can be relatively thick with respect to their lengths and distances between them. For example, in the experimental model described below $(d/a) \approx 6$, $(h_1/a) \approx 25$ and $(h_2/a) \approx 15$. Therefore certain corrections of the end and proximity effects were considered to be necessary when comparing the theoretical and experimental results.

Concerning the lengths of the elements, in the experimental model the ends of all the elements are flat and charged, whereas in the theoretical model electric charge exists only on the cylindrical antenna parts. To take approximately into account the flat charged surface of the experimental model, the theoretical antenna length was adopted to be for $a/2$ longer, which results in additional surface of area $a^2\pi$, equal to the area of the flat end surface in the experimental model.

It was more difficult to decide on the kind of the proximity effect correction. Preliminary theoretical results indicated the current in the driven and the parasitic elements to be approximately opposite in phase. Therefore the quasi-static approximation for the equivalent distance between the conductors of a two-wire line was adopted, which amounted to taking a somewhat smaller d in the theoretical model.

3 Experimental Model of the Antenna

A very flexible experimental model was searched for, which could enable a correlation to be made between corrected and uncorrected theoretical and experimental results. In the final version of the antenna, both the driven and the parasitic elements were made of several cylindrical pieces of radius $a = 3$ mm screwed one into the other. In this manner it was possible to change the lengths of the elements in steps of $\Delta h = 0{\cdot}5$ mm. The parasitic elements were mounted onto thin strips which could slide along a radial slot made in the ground plane. The driven element represented a simple protrusion of the inner conductor of the coaxial line through the ground plane. The inner radius of the outer coaxial line conductor was $b = 6{\cdot}9$ mm.

4 Experimental and Numerical Results

Theoretical synthesis of broadband cylindrical antenna with two parasitic elements was performed for a system with $h_1 = 7{\cdot}5$ cm and $a_1 = a_2 = 0{\cdot}3$ cm, requesting that the

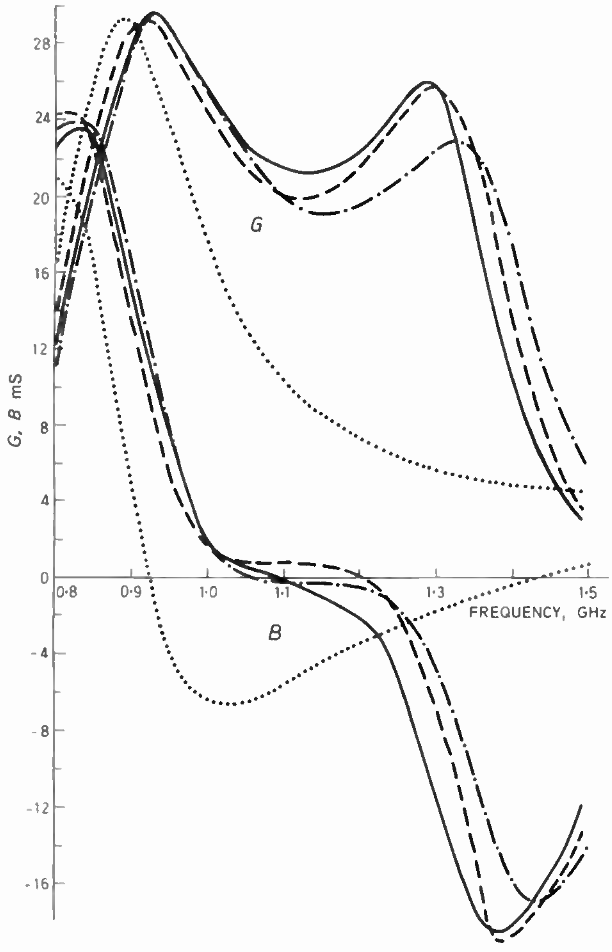


Fig. 2. Conductance (G) and susceptance (B) of the monopole antenna shown in Fig. 1(a), with $a_1 = a_2 = 0{\cdot}3$ cm.

———— experimental, $h_1 = 7{\cdot}35$ cm, $h_2 = 4{\cdot}55$ cm, $d = 2$ cm
– – – – – computed, $h_1 = 7{\cdot}5$ cm, $h_2 = 4{\cdot}7$ cm, $d = 1{\cdot}9$ cm
– · – · – computed, $h_1 = 7{\cdot}35$ cm, $h_2 = 4{\cdot}55$ cm, $d = 2$ cm
· · · · · computed, $h_1 = 7{\cdot}5$ cm, $h_2 = 0$ (antenna without parasitic elements)

optimization function given in eqn. (4) be minimal for $n = 3$, with $f_i = [1 + (i-1) \times 0{\cdot}1]$ GHz, $i = 1, 2, 3$. Optimization resulted in $h_2 = 4{\cdot}7$ cm and $d = 1{\cdot}9$ cm, with $|\sigma_{mean}|_{min} = 1{\cdot}09$, with respect to $G_0 = 21{\cdot}9$ mS.

This antenna was then realized and checked experimentally. The results are shown in Fig. 2. Very good agreement between theoretical and experimental results can be observed. For comparison, theoretical results are also plotted for the antenna without correction of the end and proximity effects, showing worse agreement with experimental data than those with corrected effects, as well as for the antenna without parasitic elements.

The radiation pattern of the antenna was found to be practically identical with that of a half-wave dipole in the whole frequency range, as expected. Thus, a broadband antenna in both the input admittance and the radiation pattern was obtained.

Several other cases were synthesized theoretically and checked experimentally, all of them showing the same degree of agreement between theory and experiment.

5 Conclusions

A simple method for theoretical synthesis of a broadband cylindrical monopole antenna driven by a coaxial line and with two closely spaced identical parasitic elements is proposed, in which the end and proximity effects are approximately taken into account. The theoretically optimized antennas were made and checked experimentally. Agreement between theoretical results and experimental data was found to be good, and the antenna behaved as a good broadband antenna in both the input admittance (optimized to be approximately constant and real in the specified frequency range), and in the radiation pattern.

For a frequency range of $1{\cdot}2{:}1$ the v.s.w.r. with respect to the average antenna conductance was found to be only 1·09. If a wider frequency range is required, however, v.s.w.r. cannot be kept so low.

6 References

1 Weeks, W. L., 'Antenna Engineering', p. 265 (McGraw-Hill, New York, 1968).

2 Surutka, J. V. and Veličković, D. M., 'Theoretical and experimental analysis of a broadband antenna', Proc. of the XVIII Yugoslav ETAN Conf., p. 359, June 1974.

3 Popović, B. D., 'Polynomial approximation of current along thin symmetrical cylindrical dipoles', *Proc. Instn Elect. Engrs*, **117**, No. 5, p. 873, May 1970.

4 Popović, B. D., 'Thin monopole antenna: finite-size belt-generator representation of coaxial-line excitation', *ibid.*, **120**, No. 5, p. 544, May 1973.

5 Popović, B. D., 'Theory of cylindrical antennas with lumped impedance loadings', *The Radio and Electronic Engineer*, **43**, p. 243, April 1973.

6 Adby, P. R. and Dempster, M. A. H., 'Introduction to Optimization Methods', p. 48 (Chapman and Hall, London, 1974).

Manuscript received by the Institution on 22nd September 1976. (Short Contribution No. 186/Com. 148.)

Contributors to this issue*

Geoffrey C. Bagley (Member 1961) studied electrical engineering at Wolverhampton and Staffs. College of Technology and Norwich City College, obtaining the Higher National Certificate in electrical engineering and City and Guilds Full Technological Certificate in telecommunications engineering. He joined the Civil Service in 1951, and has held various teaching and engineering posts. In 1965 he moved to the Royal Radar Establishment to work on radar techniques; in 1968 he received the Institution's Lord Brabazon Premium for a paper on radar-pulse compression by random phase coding. During 1972–73 Mr. Bagley was Exchange Scientist at the US Naval Research Laboratory, Washington DC, and he is now back at the Royal Signals and Radar Establishment, Malvern, where he holds the grade of Principal Scientific Officer.

Robert Corner served a five-year craft apprenticeship with Ferranti Ltd. (Edinburgh) from 1962 to 1967 as an electronics fitter, receiving the City and Guilds final certificate in electronics servicing. He subsequently worked in the production test facilities, testing avionic electronics assemblies. In 1972 he was transferred to the laser laboratory to assist in the development of high power switching power supplies. In 1974 he joined the University of Edinburgh as an electronics technician assisting in the design and construction of v.h.f./u.h.f. s.a.w.o. modules incorporating phase locked loops; and other modules based on charge coupled device components.

Kassim Hussain graduated with a B.Sc. degree in electrical engineering from the University of Baghdad in 1966, and for the next nine years worked as an electronics engineer in the Iraqi Navy. In October 1975 he commenced Ph.D. studies at the University of Edinburgh, researching into new methods of realizing digital frequency synthesizers.

* See also page 206

Peter Grant received the B.Sc. degree in electronic engineering in 1966 from the Heriot-Watt University and the Ph.D. degree in 1975 from Edinburgh University. From 1966 to 1970 he was employed as a development engineer with the Plessey Radio Systems Development Unit, Havant, designing frequency synthesizers and standards for mobile military communications systems. Following a year as senior m.o.s. applications engineer with Emihus Microcomponents, Glenrothes, he joined the University of Edinburgh in 1971 on an S.R.C. research fellowship studying the design and application of surface acoustic wave programmable matched filters. He is currently responsible for a research team studying the application of s.a.w. and charge coupled device components in communication systems. Following appointment to a lectureship in 1976, he now also teaches electronic circuits and systems.

Professor Jeff Collins (Fellow 1972) received the B.Sc. in physics and the M.Sc. in mathematics from the University of London in 1951 and 1954, and his experience in microwave tubes and ferrite parametric amplifiers was obtained during employment at the GEC Hirst Research Centre, Wembley, and with Ferranti in Edinburgh between 1951 and 1956. For the next ten years he was with the Electrical Engineering Department of the University of Glasgow, where he taught network theory and materials science, and researched microwave ferrites and microwave acoustics. During the scholastic years 1966–1968, he was a research engineer in the W. W. Hansen Laboratories of Physics, Stanford University, California, engaged in acoustic interactions, and he was then appointed Director of Physical Sciences at the Electronics Division of Rockwell International, California. In 1970 he returned to Scotland, where he has a chair in Industrial Electronics at the University of Edinburgh; he is currently on sabbatical leave at the University of Texas at Arlington. Professor Collins has contributed previous papers to this Journal and in 1974 a paper on programmable s.a.w. devices gained the Bulgin Premium for himself, Dr. Grant and two other members of the Department.

Alun Evans is a Senior Lecturer in the Department of Electronic and Electrical Engineering at the University College of Swansea; he graduated in physics from the College in 1957 and subsequently obtained an M.Sc. degree in information engineering at the University of Birmingham. Following industrial experience in Canada he was for two years a Lecturer at the University College of North Wales before returning to Swansea in 1968. He has contributed previous papers on instrumentation to this Journal and shared the J. Langham Thompson Premium in 1973 and the Sir Charles Wheatstone Premium in 1975.

Viktor Schiffer received the Diplom-Ingenieur degree in electrical engineering from the Rheinisch-Westfälische Technische Hochschule, Aachen, Germany, in 1974 and the M.Sc. degree in electronic instrumentation from the University of Wales in 1976, following work with the Department of Electrical Engineering at the University College of Swansea. He is now with Wandel & Goltermann, Reutlingen, West Germany, working on the design of level measuring instruments.

Professor Branko D. Popović received the B.Sc., M.Sc. and D.Sc. degrees of the University of Belgrade in 1958, 1963 and 1967 respectively. From 1959 to 1966 he was employed as an Assistant, and in 1966 he was appointed a 'Docent' with the Department of Electrical Engineering of the University of Belgrade. He is at present an Associate Professor in the same Department, and from 1973 to 1975 was Associate Dean. Professor Popović has made numerous contributions to this Journal in the past few years, and he was awarded the Heinrich Hertz Premium for 1973.

Djordje Paunović received the Dipl.Eng. degree from the University of Belgrade in 1962. In 1972/3 he held a scholarship at the Polytechnic Institute of Brooklyn, New York, and was awarded his M.S.E.E. degree. At present he is completing his doctorate in the Department of Electrical Engineering, University of Belgrade. Since 1963 he has been a Teaching Assistant in the same Department and he is mainly concerned with research and computer-aided design in radio communications.

Antonije Djordjević received the B.Sc. degree in electrical engineering from the University of Belgrade in 1975. During his studies he won several prizes in national and international competitions in physics, mathematics and electrical engineering and was awarded the October Prize for his Diploma work. Mr. Djordjević is at present a teaching assistant with the Department of Electrical Engineering, University of Belgrade, and is working towards his M.Sc. degree in antenna theory and techniques.

Momčilo Dragović received the B.Sc. degree in electrical engineering from the University of Belgrade in 1957, and from 1957 to 1968 he was employed as a teaching assistant with the Department of Electrical Engineering at the University of Belgrade. At present he is a Lecturer in the same Department and Head of the newly formed Laboratory for Antennas. In 1960 he held a one-year scholarship awarded by the Swiss School Council at the ETH in Zürich. His main fields of interest are antennas and electromagnetic theory.

Standard Frequency Transmissions—March 1977

(Communication from the National Physical Laboratory)

March 1977	Relative Phase Readings in Microseconds NPL—Station (Readings at 1500 UT)		
	MSF 60 kHz	GBR 16 kHz	Droitwich 200 kHz
1	7·0	10·2	1·8
2	7·2	9·8	1·6
3	7·1	9·8	1·5
4	7·0	9·4	1·4
5	7·0	9·7	1·2
6	6·8	10·0	—
7	6·9	9·4	1·1
8	7·0	9·8	1·1
9	7·0	9·6	0·9
10	7·2	9·8	0·7
11	7·0	9·6	0·2
12	7·0	9·9	−0·5
13	7·1	9·9	−1·3
14	7·1	9·6	−1·7
15	6·9	9·6	−2·9
16	7·0	9·7	−4·0
17	7·0	9·6	−5·2
18	6·9	9·7	−6·3
19	6·9	—	−7·3
20	6·7	—	−8·3
21	6·7	9·6	−9·0
22	6·7	—	−9·9
23	6·7	—	−10·8
24	6·5	—	−11·6
25	6·5	9·5	−12·7
26	6·5	9·3	−13·0
27	6·5	9·4	−13·4
28	6·3	9·2	−13·9
29	6·3	9·3	−14·3
30	6·3	9·3	−14·5
31	6·1	9·3	−14·5

Notes: (a) Relative to UTC scale (UTC_{NPL}–Station) = +10 at 1500 UT, 1st January 1977.

(b) The convention followed is that a decrease in phase reading represents an increase in frequency.

(c) Phase differences may be converted to frequency differences by using fact that 1 μs represents a frequency change of 1 part in 10^{11} per day.

Experimental direct broadcast reception of 12 GHz television signals from the Canadian Communications Technology Satellite

K. G. FREEMAN, B.Sc., M.Inst.P., C.Eng., M.I.E.R.E.

In order to gain practical experience of reception of television from satellites, a long-term experiment has been set up in Canada. The ground receiving station equipment and the monitoring arrangements are described.

Introduction

The feasibility of using high-power geostationary satellites for the broadcasting of television and radio has been extensively studied in recent years and the 1971 World Administrative Radio Conference of the I.T.U. authorized such broadcasting in a number of frequency bands.[1] For countries, such as those in Europe, which already employ the u.h.f. band for terrestrial broadcasting, the allocation of up to 800 MHz at 12 GHz is of particular interest and a Planning Conference met in January 1977 to make detailed proposals for the use of this band. However, the use of frequency modulation has already for some time been regarded as the obvious choice on the grounds of satellite power, co-channel interference and receiver complexity.

Although it is probable that a 12 GHz satellite television broadcast service would largely be received on a community basis where equipment costs could be shared, it is accepted that the service should also be capable of satisfactory and economical direct reception in the individual home where desired or where community reception is impracticable. Current planning studies[2] are therefore based on the assumption that the individual receiving system would employ a 75 cm diameter parabolic receiving aerial (or its equivalent) and a 12 GHz receiver with a noise figure of ~ 7 dB. This is considered a reasonable compromise on the grounds of cost, sensitivity, aerial pointing accuracy and protection from interference. For good quality of the received picture at the edge of the service area with an adequate margin (~ 6–7 dB) to allow for aerial pointing errors, rain attenuation etc., this would require a satellite e.i.r.p. of ~ 65 dBW, which in turn for a 1° transmitting beam (adequate to cover the UK) would require about 300 watts of r.f. signal power.

For some years investigations have been in progress at the Mullard Research Laboratories (MRL) aimed at the development of microwave sub-assemblies and complete receiving systems suitable for individual reception at reasonable cost. Much of this work has been described previously[3–6] and advanced product development is now in progress. It is generally envisaged that an individual receiving system would consist of a wideband fixed-tuned converter, located at or near the aerial, which converts all signals from the 12 GHz band to an i.f. band in the region of 1·2 GHz. A second unit located indoors would provide selection of the wanted channel in a second, tuned, converter, with an i.f. of ~ 120 MHz, followed by demodulation of the f.m. video signal (and accompanying sound sub-carrier signals). This could be used to amplitude modulate a carrier for feeding to the aerial input of a standard receiver or could be fed directly at video if a suitable interface were available.

Currently the 12 GHz front-end converter is based upon the use of Schottky barrier-diode balanced mixers executed in microstrip with a Gunn diode local oscillator source. However, other approaches can be envisaged for the future, such as the use of f.e.t. preamplifier/mixer configurations.[7]

Until recently, in the absence of any suitable high-power 12 GHz satellite signal, all experimental work has been limited to the use of a laboratory-based low-power transmitter. However, it is evident that some aspects of receiver design and performance can only be properly investigated under long term conditions of reception from an actual satellite of adequate power. The opportunity to do this was provided by the experimental Canadian Communications Technology Satellite (CTS) which was launched in January 1976. The object of this short paper is to report briefly upon the MRL participation in this experiment and the experience to date.

The Canadian Communications Technology Satellite

CTS is a joint venture of the Canadian Department of Communications (DOC) and the US National Aeronautics and Space Administration (NASA) which is being used to explore the applications to communications of advanced satellite technology.[8] The satellite weighs 700 kg and has 3-axis stabilization and is located in a geostationary orbit 36 000 km above the equator at longitude 116°W. It has two steerable 2·5° beamwidth aerials, a 1·2 kW solar array and a 200 W travelling-wave tube giving an e.i.r.p. of 58·5 dBW at 12·08 GHz. It is currently being used for a variety of experiments and the Canadian programme includes experiments in 2-way telephony, teleconferencing, interactive educational television and television broadcasting to remote communities.

For most of these experiments the DOC has provided suitable terminals of a professional design[9] but there was also interest, particularly on the part of the Canadian Broadcasting Corporation (CBC), in experimental television reception using relatively cheap and simple terminals with small aerials. Following discussions with the DOC and CBC it was agreed that MRL and the Video Group of Philips Eindhoven under Ir J. W. Edens should construct a number of receiver converters, suitable for use with standard NTSC colour receivers. These would be loaned to the DOC for a period of about 1 year and by the use of suitable recording equipment were

Ken Freeman (Member 1961, Graduate 1958) obtained his B.Sc. with honours in mathematical physics from Birmingham University in 1953. After National Service he joined the Mullard Research Laboratories renamed Philips Research Laboratories on 1st June 1977 where he has been concerned with research into many aspects of television, particularly colour display devices. He is now a Principal Scientist in the Circuit Physics and Applications Division and is currently involved in the study of direct broadcast reception of television from satellites and of television data systems such as Teletext.
Mr Freeman has served on the Papers Committee since January 1973. The author of a number of papers published in the Journal, he received the Rediffusion Television Premium in 1959 for a paper entitled 'A Gating Circuit for Single Gun Colour Tubes'.

expected to yield valuable information regarding the performance of domestic-type receiving equipment under practical conditions.

At the time that construction was undertaken, the possibility of achieving a converter noise figure of 7 dB with a mixer input and a 1·2 GHz first i.f. seemed remote, but it was expected that this figure could be reached by recourse to a first i.f. in the region of 400 MHz. To compensate for the fact that the transmitter e.i.r.p. would be about 6·5 dB less than the value referred to earlier as being necessary for reception with a 75 cm aerial, it was therefore decided to employ a 1·6 m diameter aerial. As shown from the power budget calculations summarized in Table 1, this was expected to yield good picture quality at beam centre under ideal clear weather conditions, with a margin of ~8 dB above the f.m. threshold to allow for pointing errors, reception off beam centre, rain attenuation, etc.

MRL Receiving Equipment

A block diagram of the MRL receiver is shown in Fig. 1. (The Video Group's receivers are similar in conception but differ in some design details.) The outdoor unit is enclosed, but not hermetically sealed, in a cylindrical housing with an integral waveguide horn and is located at the focus of a 1·6 m diameter metal-coated parabolic aerial made of glass-fibre reinforced polyester (Fig. 2). Since the satellite position is not fully stabilized a remote-control single-axis tracking facility is provided.

The outdoor unit (Fig. 3) consists of a fixed-tuned Schottky-barrier-diode balanced mixer executed in microstrip followed by some 46 dB amplification provided by a thin-film amplifier and a hybrid amplifier module in cascade. As already mentioned, to achieve the target noise figure of 7 dB the first i.f. was chosen to be ~410 MHz rather than the higher value likely to be preferred in the future. The local oscillator is a Gunn device located in an aluminium waveguide cavity integral with the rest of the converter. By insertion of a post of temperature-dependent dielectric material in the cavity the oscillator frequency of 11·67 GHz is maintained stable to within ~5 MHz over the temperature range −40°C to +40°C. This is well within the range of the automatic frequency control (a.f.c.) applied to the second mixer located in the indoor unit. D.c. power to the outdoor unit is supplied via the coaxial i.f. signal downlead.

The indoor unit provides further conversion to the second i.f. of ~120 MHz, at which frequency limiting and frequency discrimination are carried out to yield the NTSC System M

Fig. 2. 1·6 m aerial for C.T.S. television receivers.

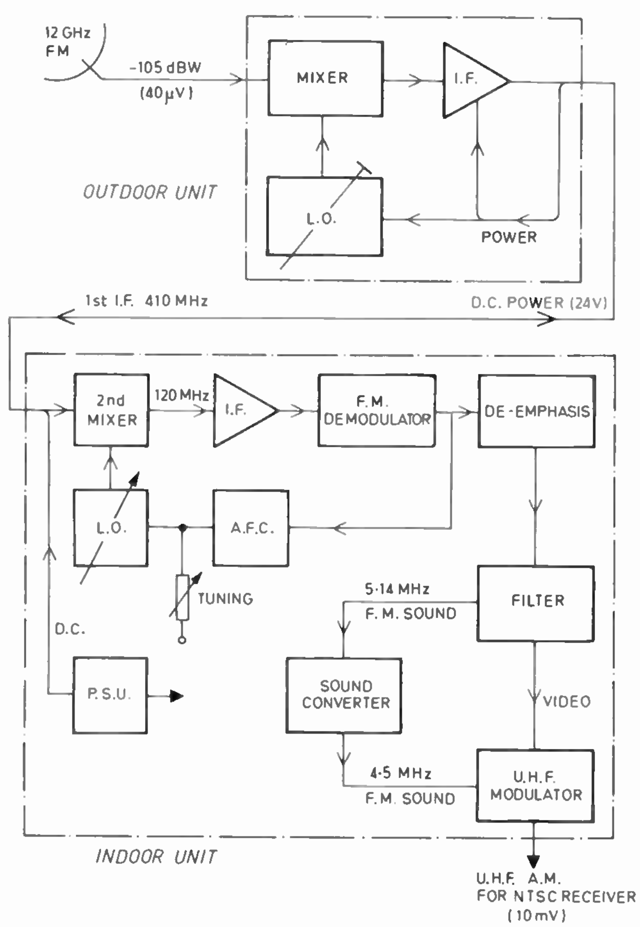


Fig. 1. Simplified block diagram of MRL C.T.S. television receiver.

Table 1. Summary of power budget calculations under ideal conditions

Satellite t.w.t. power	23 dBW (200 W)
Output filter loss	0·8 dB
Satellite aerial gain	36·3 dB
E.i.r.p.	58·5 dBW
Free space loss	206·3 dB
Clear weather atmospheric loss	0·7 dB
Receiver aerial gain (1·6 m)	43 dB
Received signal (clear weather at beam centre with no pointing losses)	−105·5 dBW
Receiver noise figure	7 dB
Receiver bandwidth	20·4 MHz
Total receiver noise (including aerial)	−124·3 dBW
Downlink carrier noise	18·8 dB
Received carrier noise (including uplink)	~18·0 dB
Approximate video signal/noise (unweighted)	~36·0 dB

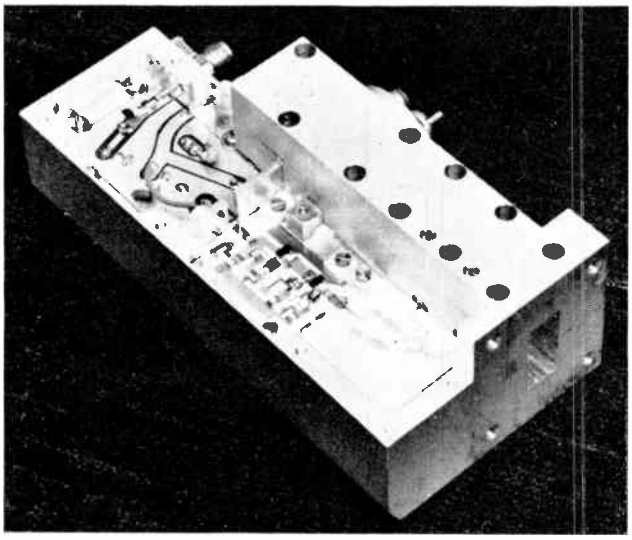


Fig. 3. 12 GHz front end for receiver.

composite video colour signal and the satellite intercarrier f.m. sound signal at 5·14 MHz. To provide a signal suitable for a standard System M colour receiver the sound signal is converted to the normal intercarrier frequency of 4·5 MHz and this together with the composite video signal is fed to a simple diode modulator which yields a double sideband u.h.f. amplitude-modulated signal at ~ 500 MHz and a level of ~ 10 mV r.m.s.

A feature of the equipment is the provision of facilities for monitoring performance during setting up and operation. This includes the recording at five minute intervals day and night of the received signal strength (maximum, minimum and instantaneous), the indoor unit a.f.c. tuning voltage (as a measure of Gunn oscillator frequency) and the Gunn diode current, together with such environmental conditions as temperature (outside ambient and inside the indoor and outdoor units), humidity (inside the outdoor unit), rainfall and windspeed

During the expected 1 year duration of the experiment each parameter will be recorded about 100 000 times. To facilitate subsequent computer analysis (e.g. to search for correlations in the event of any untoward behaviour) use has been made of a multichannel data logger employing digital magnetic tape cassettes which are filled and returned to MRL at 2 week intervals.

Installation and Operational Experience

All receivers were installed and commissioned in June and July 1976. The MRL receiver and one Video receiver were installed at the DOC Communications Research Centre (CRC) Ottawa. Other Video receivers are located at the CBC studios and the Philips Electronic Industries buildings in Toronto and at the CBC Engineering Headquarters and Maison de Radio in Montreal. The receiving aerials were mounted on the roofs of the buildings concerned in a rudimentary manner and final adjustment of the aerial pointing and axial orientation and position of the outdoor unit was made using the satellite signal itself in conjunction with a signal strength meter. At Ottawa the satellite elevation and bearing are respectively 24·4° and 230° and the signal polarization (linear) is $\sim 30°$ to the vertical.

With the satellite beam directed at Ottawa the received signal strength under clear weather conditions was measured to be the expected value of ~ -105 dBW and in these conditions good picture and sound quality was achieved—picture noise being only just perceptible. During the commissioning period it was also possible to assess the performance margin available, as a very heavy rainstorm occurred at a time when the satellite beam was directed to the southern end of Hudson Bay, i.e. when Ottawa was at the beam edge. Under these circumstances the received signal was reduced to the neighbourhood of the f.m. threshold (~ -113 dBW) and the picture, although still recognizable, was extremely noisy.

Data cassettes received subsequent to commissioning confirm that apart from minor problems the equipment has continued to operate satisfactorily. It is hoped to give a more detailed report of the operational experience when the experiment is concluded later this year.

This successful outcome of an activity which began some 10 years ago as a paper study at MRL is very gratifying and promises well for the future of 12 GHz satellite broadcast reception, although clearly many detailed problems remain to be solved before individual reception in the home is a *fait accompli.*

Acknowledgments

Acknowledgment is gratefully made to many colleagues at MRL for their important contributions and especially to Ir Edens of Philips, Eindhoven, who organized our joint participation in these experiments and has separately reported his own experiences.[10] Thanks are also due to Mr R. Baylis of Philips Toronto and to the various staff of DOC, CRC and CBC for their help and co-operation.

References

1 'I.T.U. Radio Regulations 1976.' International Telecommunications Union, Geneva, 1976.

2 Mertens, H., 'Satellite broadcasting. Design and planning of 12 GHz systems', EBU Technical Document 3220-E, March 1976.

3 Freeman, K. G., 'Reception of 12 GHz f.m. television from broadcast satellites', *EBU Rev. Tech.*, No. 138, p. 70, April 1973.

4 Freeman, K. G., 'A 12 GHz f.m. converter for individual reception of television from satellites', Proceedings of ESRO Symposium on Satellite Broadcasting, Frascati, Italy, 12th–14th February 1975, p. 53.

5 Spencer, D. B., 'A comparison of f.m. and digital modulation for direct television broadcasting from geostationary satellites', *EBU Rev. Tech.*, No. 137, p. 23, February 1973.

6 Linnecar, T. P. R. L., 'Maximum tolerable local oscillator noise in a 12 GHz f.m. satellite television receiver', *The Radio and Electronic Engineer*, **44**, No. 2, p. 77, February 1974.

7 Magarshack, J., 'Advanced Techniques for Future Satellite TV Receivers at 12 GHz', ESRO Symposium on Satellite Broadcasting, Stockholm, 22nd–24th November 1976.

8 Davies, N. G. and Blevis, G. C., 'Communications experiments using the Communications Technology Satellite', Computers, Electronics and Control Symposium, Calgary, 23rd–25th May 1974.

9 Huck, R. W., 'Communications Experimenters Guide—Communications Technology Satellite', Dept. of Communications, Canada. CRC Technical Note No. 671, February 1975.

10 Edens, J. W., 'Satellite-to-home television', *Canadian Electronics Engng*, **20**, No. 12, p. 14, December 1976.

Manuscript first received by the Institution on 13th January 1977 and in final form on 22nd February 1977.
(Short Contribution No. 187/Com.149.)

IERE News and Commentary

That Academic Test—A Further Explanation

It appears, from enquiries received by the Institution following the publication of the notice to Graduate members in the March issue of the Journal about the CEI Part 2 Academic Test, that there is still widespread misunderstanding of the present situation. A more detailed explanation may therefore be helpful.

Prior to the end of 1973, those who had passed the Institution's former Graduateship examination, or obtained recognized exempting qualifications before the end of 1970, could be admitted to Graduate or Corporate membership without further examination. Those admitted to Graduate membership who had not met the 'training and experience' requirements for transfer to Corporate membership by the end of 1973, have been, and still are, required to pass the CEI Part 2 Academic Test before an application for transfer can be considered. The Test consists of two papers from the CEI Part 2 examination, which must be passed at the same sitting. Two attempts are allowed: these must be made within two years of each other. There are now only two opportunities left to make the first of these—namely in May 1978 and May 1979. The aim of the notice in the Journal was to alert the Graduate members concerned to the situation before it is too late.

Graduate members who have already passed the Academic Test, or have since their election obtained qualifications *which the Institution has confirmed* will exempt them from the CEI examinations need read no further, even if these qualifications were obtained before the end of 1973. The notice in the Journal does *not* mean that a new requirement has been placed on everyone admitted to Graduate membership before the end of 1973: it applies *only* to those whose academic qualifications do not meet the requirements for exemption from the CEI examinations.

A number of anxious enquirers have asked whether their Graduate membership will be terminated if they do not attempt, or fail to pass, the Academic Test. The answer is 'No'. But they will not be eligible for transfer to Corporate membership until they have obtained whatever further qualifications are needed to meet the academic requirements for C.Eng. registration. They can remain as Graduate members indefinitely (there is no upper age limit) or apply for transfer to the classes of Associate or Associate Member: admission to the latter class carries with it entitlement to registration as a Technician Engineer [T.Eng(CEI)].

Some enquirers have stated that they thought that they could not attempt the Academic Test until they were ready to apply for transfer to Corporate membership. The Institution is at a loss to understand how this unfortunate mistaken impression has arisen: anyone who is required to take the Academic Test can enter for the first time in either of the two years remaining. Formal course attendance is not compulsory: preparation by private study or by correspondence tuition is permitted. The Institution's Education Department will be pleased to give any help or advice it can on this or any other matter connected with academic qualifications. Therefore, in the immortal words often inscribed on engineering drawings, 'If in doubt, ask!'

New Director of The Appleton Laboratory

Dr. Frederick Horner, D.Sc., C.Eng., F.I.E.E., Deputy Director of the Appleton Laboratory, Slough, Berkshire, is to succeed Dr. J. A. Saxton, C.B.E., D.Sc., Ph.D., C.Eng., F.I.E.E., F.Inst.P., as Director with effect from 1st July 1977, following Dr. Saxton's retirement from the post he has held for eleven years.

Dr. Horner joined the Radio Division of the National Physical Laboratory in 1941. It became a separate organization after the war and has evolved into the present Appleton Laboratory, of which he became Deputy Director in 1969. Dr. Horner's research has been concerned mainly with radio direction finding and with the study of radio emission from thunderstorms. He has been active in international affairs in radio science since 1950, holding various offices in international organizations.

IEE Conference Publications

Members of the IERE may purchase single copies at reduced rates of the following volumes of papers read at conferences for which the Institution was a co-sponsor:

'Satellite Communication Systems Technology', London, April 1975 (IEE Conference Publication 126), £7.40 (normal price £11.30).

'Trends in On-line Computer Control Systems', Sheffield, April 1975 (IEE Conference Publication 127), £6.70 (normal price £10.20).

'On-line Operation and Optimisation of Transmission and Distribution Systems', London, June 1976 (IEE Conference Publication 140), £5.30 (normal price £7.95).

'Automobile Electronics', London, July 1976 (IEE Conference Publication 141), £5.30 (normal price £7.95).

'Advances in Magnetic Materials and their Applications', London, September 1976 (IEE Conference Publication 142), £5.50 (normal price £8.40).

'Gas Discharges', Swansea, September 1976 (IEE Conference Publication 143), £10.80 (normal price £16.20).

'International Broadcasting Convention', London, September 1976 (IEE Conference Publication 145), £6.60 (normal price £9.90).

'Millimetric Waveguide Systems', London, November 1976 (IEE Conference Publication 146), £6.95 (normal price £10.35).

'The Future of Aircraft All-weather Operations', London, November 1976 (IEE Conference Publication 147), £6.40 (normal price £9.20).

Orders from members wishing to take advantage of these special rates should be placed through the IERE Publications Sales Department, 8-9 Bedford Square, London WC1B 3RG.

Mr. F. W. Appleby

On April 10th last, Mr. Frank W. Appleby, the Institution's Publications Manager, retired after 18 years' service during which he has been responsible for the production side of all IERE publications. A printer by profession, Mr. Appleby was for many years Works Manager of the Press which printed the Journal during the 'fifties.

The President and other officers of the Institution as well as senior staff members, wished Frank Appleby a long and happy retirement at a dinner party in London on March 24th.

Colloquium Report

The Future of Higher Diplomas

An IERE Education and Training Group Colloquium held at Southampton College of Technology on 2nd March 1977

The Colloquium was the second in which the primary subject of discussion has been the education and training of Technician Engineers and Technicians.* Its title reflected the widespread concern felt in both educational and industrial circles regarding the impending demise of the Higher National Diploma in Electrical/Electronic Engineering—currently threatened with slow death as a consequence of the progressive implementation by the Technician Education Council of the recommendations of the Haslegrave report.

At the morning session, three speakers from the field of education explained in turn the history and structure of H.N.D. courses, the anticipated character of the Higher Diploma courses being developed under the auspices of TEC, and the changes in shape and concept of courses for the Dip.H.E.—originally conceived as a non-vocational award and having its origins in the special needs of the teaching profession. The afternoon session was devoted to exploration of the reaction of representatives of the electronics industry to the question 'Which of the 3 Diplomas described this morning seems to you most likely to produce the kind of Technician Engineer you are seeking?' By the end of the Colloquium, the morning's plea that the IERE should do everything in its power to save the H.N.D. had been slightly modified—perhaps what was needed for the H.N.D. was not a reprieve but a stay of execution.

It would be impossible in a brief report to recount all the arguments which were presented, challenged and modified during a very lively and frank presentation and discussion: since it is often claimed by educationalists that industry does not contribute much to the discussion of matters of vital importance to it, it is proposed to devote more space to summary of the comments of the representatives of industry than to the educational arguments, of which the first—those for the reprieve of the H.N.D., were presented by Mr F. Dellow (Portsmouth Polytechnic).

Mr Dellow may well have surprised many of those present by his reminder that the H.N.D. began in 1924, that the number of H.N.D.s awarded reached a peak in 1970, and that the decline in the number of awards which took place in the following years until 1975 was replaced in 1976 by a small rise. It is presumably not necessary to reproduce his account of the typical content of H.N.D. courses in Electrical/Electronic Engineering, or of the manner of their organization and administration: it should be equally unnecessary to remind readers that the H.N.D. is still highly regarded by many companies in the electronics industry and that it has only quite recently ceased to be accepted as meeting the academic requirements for Corporate Membership of many Chartered Institutions. It was clear, however, that Mr Dellow, for all the eloquence of his pleading, knew that sooner or later the H.N.D. would die: what he sought was that it should be allowed to die a natural death and that, far from doing anything to speed its demise, we should seek by all possible means to ensure that it survived until such time as we had found a successor which met the same criteria—that it should have the confidence of industry, meet the needs of the nation, and inspire its students.

The Chairman of TEC Programme Committee A2, Mr A. J. Kenward, in arguing that TECs projected Higher Diploma could well be made a worthy successor to the H.N.D., pointed out that it was no longer true that the possession of an H.N.D. secured direct admission to Corporate Membership of Chartered Institutions: there was a big obstacle—Part 2 of the CEI examination—which the H.N.D. holder must surmount to qualify as a Chartered Engineer. It was therefore illogical to fault TEC's projected award because it did not meet the academic requirements of the Chartered Institutions. Moreover, unlike H.N.D. courses, TEC Higher Diploma courses had a built-in flexibility: if it was clear that industry wished to continue to have available something with a strong family resemblance to the H.N.D., the TEC Diploma structure could provide it. In reply to a question regarding admission of holders of 'A' levels to TEC Higher Diploma courses, Mr Kenward explained that, although it was expected that the main feed to these courses would come from holders of the TEC Certificate or Diploma, those awards themselves had variations in structure to meet the needs of particular regions and industries. The first year studies for the Higher Diploma would therefore of necessity be designed to accommodate the consequent variations—for example, the level to which the student had studied Mathematics in his previous Certificate or Diploma course. They could therefore undoubtedly be organized to meet the somewhat different needs of holders of 'A' levels instead of basic engineering qualifications.

He also reminded the audience that it was the declared objective of TEC to produce courses and awards which met the requirements of the student and of industry. This being so, the issue of whether or not the awards were satisfactory to CEI, the ERB, the Institutions and the Universities was, though an important one, a secondary consideration. Consultation with all these bodies was undertaken regularly, and he was optimistic that the 'bridges' which existed in the present system of awards, to enable the gifted individual to migrate from one type of course to another, could be preserved. It was not expected that every TEC award would be acceptable to other examining authorities: it would be wrong to distort the basic concept in an attempt to achieve what could well be a totally unrealistic objective. What TEC hoped to do, and believed that, with sufficient goodwill on the part of other parties could be achieved, was to ensure that every student could, from the range of options available at all levels, choose at least one which would be acceptable to any other examining authority or qualifying body whose recognition he might wish to obtain.

'The case for the Dip.H.E.' was presented by Dr J. Salmon, Deputy Chief Officer and Registrar of CNAA, who first explained how the original concept had been widened. It was no longer envisaged as a non-vocational award: vocational and non-vocational, specialist and generalist variants were possible: in the last respect it had an advantage over both the H.N.D. and the TEC Higher Diploma. In particular, it might well be to the advantage of the engineering industry if such people as its company accountants had studied engineering to Dip.H.E. level.

He proceeded to make comparisons between the three Higher Diplomas being paraded for inspection by the audience, in the course of which he drew attention to the many similar or common features and to the distinguishing characteristics

* The first, held in November 1977, was 'Technicians and technician engineers in the electronics industry'. See *The Radio and Electronic Engineer*, 47, No. 1/2, p. 81, January/February 1977.

which gave each contender for their support its special appeal. Of particular interest was his comment that the TEC Higher Diploma might bear closer resemblance to an H.N.C. with endorsements than to an H.N.D., which was in many ways a unique qualification. Also noteworthy was his suggestion (no doubt anathema to the tidy-minded) that the first-year courses for all the Diplomas in question had so much in common that they could probably be largely shared. He also drew attention to the fact that the major difficulties encountered by 'A' level students entering all courses of the kind being discussed arose in the first year. They were taken from an environment in which their companions and rivals were persons of similar background and ability, into one where they met and had to communicate with people of widely different abilities and backgrounds, and had to learn how to involve themselves in coherent groups of what might well be a completely unfamiliar kind. There was clear evidence that ability to adjust to the new environment was not in any way linked to previous academic achievement: the holders of good 'A' levels were often so much slower in adapting to the change in climate that others apparently less gifted rapidly overtook and passed them. This was, perhaps, one more facet of the universal communication problem—ot the solution of one well-known aspect of which he felt the Dip.H.E. might make a useful contribution. Industry frequently complained that there was a shortage of people qualified at just below first degree level who could act as 'interpreters' between designer and executor and so help to ensure that mistakes were not made for the too-common reason that the objectives and principles of the design were not fully understood by those trying to implement it.

Dr Salmon caused some amusement by a reference to attempts to obtain from industry any clear indication of what it hoped the Technical Colleges and Polytechnics would teach its future employees, and commiserated with the Head of any College attempting to develop courses to meet the needs of the kind of hotch-potch distribution of industry which has developed alongside the Great West Road in the Chiswick–Brentford area. His own experience of attempts to obtain the views of industry on proposed course structures was that these fell into three categories: 'Don't know, but it looks O.K.'; 'It's all right if it teaches the same things as I was taught'; or 'It's quite useless unless it contains at least 90% of ... (whatever the individual's particular fad may be).'

The discussion following the morning session produced little by way of further food for thought, being centred mainly on comment on the claims and counter-claims of the principal speakers. One contributor, however, raised an important issue. O.N.C./O.N.D. in Engineering courses currently have an annual output of about 2,000: they are expected to continue in existence for a further three years and so will probably produce about 6,000 possible entrants for H.N.C./H.N.D. courses. The O.N.D. in Technology is to continue indefinitely: although primarily envisaged as an alternative to 'A' levels as a University entrance qualification, it has a significant output of students who appear unlikely to survive beyond H.N.D. level. Surely, therefore, these 'feeder' courses should be discontinued before the H.N.D. is given the final *coup de grâce*.

The first address of the afternoon session was given by Mr R. W. S. Hewitt (Training Manager, EMI) who described the company's training methods in considerable detail, and, like the following speaker, drew attention to respects in which existing courses failed to meet all the desirable objectives, and left the Training Departments of the companies, great and small, in the industry, with the task of remedying deficiencies which they believed should not exist. He reminded the audience that his company had established the EMI Institutes, which once offered full-time, part-time and correspondence courses leading to a variety of awards, including external degrees, and might well have eventually been transformed into the EMI Polytechnic. His problem was that of continuing to provide a company which, like many others at the present time, had a low order book and was forced by shortage of money to try to make cuts, with the well-educated and trained people whom it needed: this was a process which could involve the company in providing partial or total financial support for trainees for anything from 2 to 5 years. In the process he had to try to persuade busy engineers to make an active contribution to the training, to the point where they could 'stand on their own feet', of young people who had often been accustomed to spoon-feeding and were lacking in competitive spirit.

This meant that training objectives must be clearly defined: trainees had to learn not just 'how to do it', but how to do it within given constraints in respect of financial or other resources; they next must understand that having decided the method it was necessary to define what was required for its execution; they must appreciate the need to record and report, and to consult those responsible for their training about problems. The practical difficulties which had been encountered in trying to fulfil these objectives showed that there was an almost universal shortfall in the training in communication given in existing courses, and that the need for and the merits of joint effort were often not realized. In the area of 'recording and reporting' the maintenance of the traditional log book seemed to be regarded as a 'chore' both by the trainee and those responsible for scrutinizing it. Some surprising facts came to light—one, seldom realized by the trainees themselves, was that in the course of their training they had learned more about people than about things.

On the question of the usefulness of 'General Studies' as a means of remedying weaknesses in communication skills, he felt that all too often the topics selected were unsuitable. No useful contribution to the understanding by the student of the vital importance of communication as the means by which one could ask, listen, observe and report, was likely to be made by any 'general studies' which produced the reaction 'What has all this got to do with engineering?'.

With these considerations in mind, Mr Hewitt felt that the view of his company regarding the relative merits of the three Higher Diplomas would probably be conditioned by the fact that those trainees thought suitable for sponsorship for further study were opting for degree courses rather than H.N.D. courses, because the latter no longer met the requirements for Corporate Membership of the Chartered Institutions. The Dip.H.E. seemed not yet to have 'found its bearings' and might well have no future. Since the company was particularly interested in courses so structured that progress could be made from stage to stage on a part-time basis, it might well be that, once the dust had settled and the letters TEC could no longer be claimed to stand for 'Technical Education Chaos', the new Higher Diploma would be regarded as a useful award.

The next speaker, Mr J. Fredericks (Independent Broadcasting Authority), having startled the educationalists by announcing that he did not care much for any of their schemes, softened the blow a little by adding that his company's requirement was so small and of such a specialized nature that he did not really expect to find a suitable 'ready made' supply. He believed, nevertheless, that the company was being obliged to spend more time and money modifying 'products' carefully selected from the available range than should be necessary. At the present time, it employs about 250 Technician Engineers on the maintenance of 353 transmitters. By 1980 the number of transmitters would have risen to 650: it was nevertheless hoped that the maintenance force could be reduced.

Currently, his company recruited its maintenance staff from holders of specialist awards in Telecommunications which covered the whole range from Full Technological Certificates to degrees. Having selected suitable candidates (typically 15 out of 400 applicants), it subjected them to an 18-month training course which included attendance at a specialized Diploma course at Plymouth Polytechnic and cost approximately £7,000 per student. It was therefore highly important that the selection procedure should keep wrong choices to a minimum. To this end it incorporated visits by 'short listed' applicants to a transmitter site, where they could obtain an impression of what they would be expected to do under what conditions. This was followed by a 35-minute written test, after which candidates were interviewed by a three-man panel which was provided with their answers to the test questions and could use those answers as focal points for further questioning.

None of the questions in the test required elaborate answers: a diagram or a few words only would be sufficient. The questions covered fundamental concepts: the constant voltage/constant current generator approach; properties of resonant circuits; 'magnification' in such circuits; differentiating and integrating circuits; the relationship between time-constant and attenuation; properties of transmission lines; on the 'power' side, transformer losses and motor characteristics; and that old favourite, the condition for maximum power transfer and the efficiency at maximum power. The great weakness found (in reply to a question, Mr Fredericks stated that this was general and not linked to the candidates' academic background) was a deficiency in grasp of fundamental principles and ability to relate a problem to them. Although candidates' answers to the test were not marked, he thought the average mark would be about 30%—a situation which was sometimes excused by the candidates on the basis that 'they had not expected anything quite as simple: they did all that stuff in the first year, and had forgotten most of it.'

The company's problems were not confined to those of finding people with good diagnostic abilities: personality was of considerable importance. Its maintenance engineers were sent to 'trouble spots' in pairs: at those sites they were often required to work in cramped and uncongenial conditions. Consequently the company was obliged to reject anyone who, however great his diagnostic ability, gave the slightest hint of being any kind of 'odd-ball'.

Turning from the special problems of his company to general educational considerations, Mr Fredericks commented that the effect of present overall policy was to create too many low-grade technologists and too few high-grade Technicians and Technician Engineers. There appeared to be a need right across the board for a more practical approach and for greater attention to fundamentals. While, in later discussion, there was general support for this view, it was stressed that there were dangers in an exclusively utilitarian approach to education. Although the structure of TEC courses generally might facilitate the inclusion of specialist modules suited to the needs of a particular industry, such modules could not, and should not, completely relieve firms of the need to supplement their employees' education by appropriate training.

In summing up the discussions which had taken place at both sessions, Dr Salmon showed proper impartiality and did not attempt to 'sell' the Dip.H.E. as an answer to the problems. He stressed once again, however, the reorientation problem experienced by first-year students moving from a 'peer' environment into a widely diverse one. He also pointed to the possible value of the Dip.H.E. as a 'consolation prize' for the undergraduate who found that the final year of a degree course took him beyond his academic 'ceiling'—and who might otherwise have nothing left but his 'A' levels with which to enter the employment market. As to the issue of 'which Higher Diploma?' he felt that what had emerged was that the levels of academic competence typified by the O.N.D., the H.N.D. and the degree had proved to be a reasonable match to the needs of the three main classes of employee in the engineering industry, and that any future system of awards should seek to preserve similar standards and equally effective migratory paths both for the 'high-flyer' and the individual who 'couldn't quite make it'.

It was perhaps not to be expected that the Colloquium would pass a resolution urging the IERE to campaign for preferential recognition of one or other of the Higher Diplomas which had been presented for discussion. It could almost have been said of the 45 or so delegates present that *quot homines, tot sententiae*—a situation which must have encouraged anyone fearful of being crushed under some particular educational 'bandwaggon'.

There are still problems to be solved: it is already clear that the Institution's efforts to set people thinking about them are much appreciated by those most closely involved, whether as educators or employers. The communication problem clearly remains a major one: it is one already proposed for a future Colloquium, and the relevance to it of 'The Engineer in Society' paper in Part 2 of the CEI examination has already been the subject of a Paper submitted by the Institution for consideration by CEI. The Institution has a Working Group busily investigating the structure and content of degree courses in electronic engineering: is it mere coincidence that this Group has indicated that it expects to identify three types of structure, each with its own special merits, or is there a common factor with the three kinds of Higher Diplomas which the Colloquium considered? The inadequate understanding of fundamentals complained of: why is it so common? Are the courses which produce designers unsuitable for diagnosticians? The Institution was recently told of a Brazilian professor who said that 'The difference between a Professional Engineer and a Technician Engineer is that the Professional Engineer knows more Mathematics.' Dare one ask whether perhaps, particularly in the diagnostic field, the Technician Engineer may even so be expected to know too much Mathematics? There are many intensely practical people who find Higher Mathematics in particular, very hard work. Have they, perhaps forgotten their fundamentals because, in their struggles to understand the higher flights of circuit and waveform analysis, they have seen them as exercises in Mathematics instead of as applications of those fundamental principles which they once understood (or thought they did)? Perhaps, under circumstances in which technological developments are of a kind which is likely to lead to a demand for fewer designers and more diagnosticians, this will soon be the question of the hour.

This being so, it was fitting that the Colloquium should have been closed by a quotation from an ancient Chinese philosopher. 'If you wish to provide for next year, sow corn. If you wish to provide for the next decade, plant trees. If you wish to provide for posterity, train men.'

K. J. COPPIN

Conference on 'Digital Processing of Signals in Communications'

Organized by THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS with the association of the Institution of Electrical Engineers and the Institute of Electrical and Electronics Engineers.

University of Technology, Loughborough—6th to 9th September 1977

PROVISIONAL PROGRAMME

(Note: Changes may be made within sessions)

Tuesday 6th September 1977

Keynote speaker: Dr. K. BENJAMIN (*G.C. HQ*)

DIGITAL FILTERS IN COMMUNICATIONS

'Lowpass digital filtering with host windowing design techniques'
By A. H. ABED and Dr. G. D. CAIN (*Polytechnic of Central London*)

'Digital filters with differing ripple characteristics'
By Dr. A. M. TELLESI (*University of Tripoli*) and Dr. L. G. CUTHBERT (*Queen Mary College, University of London*)

'A survey of FIR digital Hilbert filters'
By A. A. ABU-EL-ATA and Dr. G. D. CAIN (*Polytechnic of Central London*)

'Computation rate reduction in digital narrow filter banks'
By G. BONNORET and M. COUDREUSE (*T.R.T., France*)

'A digital correlator-simulator using c.c.d. memory for sonar applications'
By Dr. J. MAVOR *et al.* (*University of Edinburgh*)

'Digital filter bank for real-time speech analysis and synthesis using logarithmically quantized signals'
By Dr. N. G. KINGSBURY (*Marconi Space and Defence Systems*) and L. C. KELLY (*Government Communications HQ*)

'Digital single-voice frequency receiver for telephone signalling'
By P. F. ADAMS (*Post Office Research Centre, Ipswich*)

ADAPTIVE SYSTEMS AND CHANNEL EQUALIZATION

Review Speaker: R. K. P. GALPIN (*Plessey*)

'Channel equalization using a stack algorithm'
By J. GORDON and N. MONTAGUE (*Hatfield Polytechnic*)

'Improved detection processes for distorted digital signals'
By Dr. A. P. CLARK, J. D. HARVEY and J. P. DRISCOLL (*University of Technology, Loughborough*)

'Signal distortion in a sampled digital signal'
By Dr. A. P. CLARK (*University of Technology, Loughborough*)

'An iterative detection process for a wide class of linearly distorted data signals'
By Dr. J. D. DALEY (*University of Technology, Loughborough*)

'Some aspects of the use of orthogonal groups of signal elements for detection of distorted digital signals'
By A. CLEMENTS (*University of Technology, Loughborough*)

'The implementation of Viterbi decoding on satellite communication circuits'
By Dr. G. FOLEY and B. D. DAVIES (*RSRE, Christchurch*)

Wednesday 7th September 1977

DIGITAL TELEVISION AND IMAGE PROCESSING

Review Speaker: Dr. J. E. THOMPSON (*P.O. Research Centre*)

'Non-linear quantization of colour difference signals using data on colour differential thresholds'
By F. KRETZ and J-P. DE LA TRIBONNIERE (*CCETT, France*)

'Optimization of d.p.c.m. video coding scheme using data on colour differential thresholds'
By F. KRETZ and J. L. BOUDEVILLE (*CCETT, France*)

'A subjective investigation into the use of d.p.c.m. for transmission of visual telephone signals'
By J. A. KITCHEN (*Post Office Research Centre, Ipswich*)

'2Mbit/s coder for transmitting visual telephone signals'
By R. C. NICOL (*Post Office Research Centre, Ipswich*)

'Interframe predictive coding of television signals with computer control'
By M. A. BROWN, T. J. DENNIS and Dr. D. E. PEARSON (*University of Essex*)

'Multidimensional delta modulation combines intra- and interframe coding'
By R. FORCHHEIMER (*Linkoping University, Sweden*)

'Differential encoding for monochrome television'
By J. C. CANDY (*Bell Laboratories, USA*)

'Adaptive bit rate reduction by unitary transform'
By R. MOTSCH (*CCETT, France*)

'Error detection and correction of d.p.c.m. video signals using Walsh-Hadamard transform technique'
By D. FENWICK and Dr. R. STEELE (*University of Technology, Loughborough*)

'Application of error correcting codes to digital television'
By S. HARARI and A. RANQUET (*CCETT, France*)

'High bit rate, low error rate, magnetic tape recorder for digital signals'
By D. NASSE (*CCETT, France*)

'Digital image reconstruction by a circulant matrix algorithm'
By Dr. C. GOUTIS and Dr. T. S. DURRANI (*University of Strathclyde*)

'Applications of a new pattern distance function to automatic real-time reading of handwritten symbols'
By J. A. GORDON and R. J. DRISCOLL (*Hatfield Polytechnic*)

Thursday 8th September 1977

Review Speaker: Dr. J. D. GOODMAN (*Bell Laboratories*)

COMMUNICATION TECHNIQUES AND SYSTEMS

'High-reliability data transmission to mobile vehicles'
By J. R. EDWARDS (*Royal Military College of Science*)

'Optimal design of a data transmission channel containing jitter noise'
By Dr. L. F. LIND and S. E. NADER (*University of Essex*)

'Bit timing for telecontrol communication systems'
By J. D. MARTIN and C. SINGH (*University of Bath*)

'Applications of soft decision techniques to block codes'
By C. N. HARRISON (*Royal Military College of Science*)

'Microprocessor-controlled soft-decision decoding of error-correcting block codes'
By Dr. R. M. F. GOODMAN and A. D. GREEN (*University of Hull*)

'Sampling of narrow band sequences with application to digital detection of stochastic signals'
By J. F. Bohme (*University of Bonn, Germany*)

'A constant level modulation system'
By Dr. V. J. Phillips and D. J. D. Doherty (*University College of Swansea*)

'A useful class of pre-sampling filter functions for digital processing—the Butterworth-Thompson method'
By R. J. Rechter (*Hughes Aircraft Co., California*)

'A real-time simulator for digital signal processing'
By Dr. G. B. Lockhart, B. M. G. Cheetham and B. M. Manson (*University of Leeds*)

'A simulation of a hybrid multiplex system'
By M. J. Carey (*Post Office Research Centre, Ipswich*) and A. G. Constantinides (*Imperial College*)

POSTER BOOTH PRESENTATIONS

'Decision errors in adaptive equalizers in transmission'
By Dr. C. J. McLeod and E. Ciapala (*University of Strathclyde*)

'Telecontrol code formats/economical digital simulation'
By J. D. Martin (*University of Bath*)

'Quantized pulse duration modulation—new multi-level digital transmission technique'
By D. G. W. Ingram and K. S. Chung (*University of Cambridge*)

'The digital generation of 'chirp' signals'
By Dr. R. L. Brewster (*University of Aston, Birmingham*)

'Digital automatic gain control for p.c.m. signals'
By Prof. J. E. Flood and G. Coates (*University of Aston, Birmingham*)

'Analogue and digital implementation of matched filters for Barker coded signals'
By T. Bucciarelli and C. Picardi (*Istituto di Fisica, Universita di Perugia*)

'Implementation of an asynchronous communications time division multiplex using microprocessors'
By D. C. Laval (*Computer and Systems Engineering*)

'Soft decision threshold decoding of convolutional codes'
By W. H. Ng and Dr. R. M. F. Goodman (*University of Hull*)

'Deltamodulator coder using a parallel realization'
By Dr. M. J. Hawksford (*University of Essex*)

'Digital tomographic restoration for radiographs'
By A. N. Venetsanopoulos (*University of Toronto*)

'A microprocessor controlled speech communications link'
By S. K. Das, H. F. Silverman and C. C. Tappert (*IBM, New York*)

'Timing extraction from noisy biphase signals'
By Dr. P. Vanyai (*Research Institute for Telecommunications, Budapest*)

Friday 9th September 1977

COMMUNICATION TECHNIQUES AND SYSTEMS

'Medium-speed digital data transmission over h.f. channels'
By Dr. M. Darnell (*SHAPE Technical Centre, Netherlands*)

'Simulation of fast fading in quasi-synchronous area coverage schemes for mobile radio'
By Dr. J. P. McGeehan and K. Gladstone (*University of Bath*)

'The generation of codes for d.p.s.k. channels'
By C. R. Telfer (*Marconi Electronics*)

'The hardware realization of the d.d.c. system for p.c.m. signals'
By W. C. Wong, Dr. R. Steele and Dr. D. J. Goodman (*University of Technology, Loughborough*)

SPEECH ENCODING

'An application of error protection to a vocoded speech channel'
By N. Montague (*Hatfield Polytechnic*)

'Speech digitization by delayed encoding with adaptive and fixed predictive codes'
By J. Uddenfeldt (*Royal Institute of Technology, Sweden*)

'Applications of digital speech processing in broadcasting'
By R. Rawlings (*Independent Broadcasting Authority*)

'Digital generation of speech'
By J. C. L. Brazier, P. H. C. Madams and I. H. Witten (*University of Essex*)

Further information and registration forms may be obtained from the IERE Conference Department.

CEI to Investigate Innovation in Industry

The Council of Engineering Institution' Committee on Creativity and Innovation is setting up a small full-time team to investigate the process of innovation in medium-sized companies in order to make an effective contribution towards improving the performance of British Industry. Sponsorship for the project for three years in the first place has been forthcoming from the Department of Industry, NRDC, the Foundation for Management Education, and industry, as well as from CEI's resources, while the CBI and TUC have expressed interest in the project and offered assistance.

Mr. J. G. Dawson, Chairman of the Committee on Creativity and Innovation, has pointed out that there is still a substantial part of British industry which is not making sufficient use of the technology available to it to introduce new products and processes and to market these effectively throughout the world. Bars to the acceptance of new ideas and innovation in general exist unrecognized in management structures and attitudes and he believes that if these bars and their antidotes can be identified then more managements will be assisted in improving their company's performance.

The approach being followed is to do critical case studies in companies which recognize the need for innovation and are mindful that this may call for a change in staff attitudes. From these it is expected that a comprehensive study can be published on the best methods of establishing a suitable climate and organization within a company to facilitate innovative development.

The head of the project is Dr. R. C. Parker, who in his previous position as Technical Director of the Ferodo Group made many contributions and became an acknowledged expert in this field. The investigating team will be based on Ashridge Management College, Hertfordshire.

Members' Appointments

CORPORATE MEMBERS

Mr. L. W. Barclay, B.Sc. (Fellow 1967, Member 1960, Graduate 1955) has joined the Home Office as Head of the Propagation Section in the Directorate of Radio Technology in London. He has been with the Research Division of the Marconi Company at Baddow since 1960, for the past eleven years as chief of the Ionospheric Propagation Section. Mr. Barclay has served on the Papers Committee and the Communications Group Committee for several years and has recently been appointed Deputy Chairman of the Papers Committee. A fuller note on his career was published in the Ionospheric Radio Wave Propagation issue of the Journal (January/February 1975) for which he was Guest Editor.

Mr. R. M. Denny (Fellow 1974, Member 1955) has been appointed Chairman of the Council of the Cable Television Association of Great Britain. Mr. Denny is on the Board of the Rediffusion Group and chairman or director of a number of companies in the Group, which he joined in 1970. He had previously held appointments with ATV Network Ltd. and before that with BBC Radio and Television.

Mr. D. W. Heightman (Fellow 1948, Member 1942) retired on 31st March last from the position of Technical Director of Thorn Television Rentals Ltd., which he has held for the past twelve years; he is continuing as a part-time non-executive director/consultant to the company for at least a year. Mr. Heightman is now serving a second three-year term of office as a Vice-President, and he is currently a member of the Executive Committee and was appointed Chairman of the Finance Committee in April. He was recently appointed Deputy Chairman of the Cable Television Association.

Mr. T. W. Welch (Fellow 1966, Member 1952) has been appointed Vice-Chairman of the Council of the Association of Further and Higher Education. A member of the Board of Governors of the Guildford County College of Technology for the past 15 years and now its Chairman, Mr Welch is a consultant in radar and navigational aids. He has contributed papers to the Institution's Journal and meetings programmes on subjects in his field, and he is at present Chairman of the Aerospace, Maritime and Military Systems Group Committee.

Mr. F. Grimm (Fellow 1973, Member 1952, Graduate1949) has been appointed Technical Co-ordinator of Pye's Mobile Radio Management Group. He joined Pye in 1950 and from 1970 until this year he was Technical Director of Pye Telecommunications, responsible for the company's extensive research and development facilities.

Mr. I. Andrews (Member 1974, Graduate 1971) who has been with Brandenburg Limited, Thornton Heath, Surrey, since 1973, has been appointed Development Manager with full responsibility for all aspects of innovation and development. He was previously Manager of the company's Nuclear Engineering Division.

Captain A. D. Barlow, M.A., RN (Member 1953) has been appointed Chairman of the Naval Nuclear Technical Safety Panel of the Ministry of Defence in London. For the past two years he has been Superintendent Base Support in HMS *Neptune* at the Clyde Submarine Base, Faslane, Dumbartonshire.

Mr. R. I. Carrington-Smith (Member 1973, Graduate 1966), previously Supervisor, Customer Engineering and product Development, with Corning Glassworks, Rayleigh, North Caroline, has become President and Chairman of the Board of Intervideo Inc., at Rayleigh. He left the United Kingdom in 1966 to take up employment with the Avco Corporation, working on p.c.m. telemetry design for the Lunar Excursion Module, and after two years with the Raytheon Company, joined the Memory Products Department of Corning Glassworks in 1970.

Mr. M. J. Clements (Member 1969, Graduate 1966) has joined Grundy and Partners Ltd., Stonehouse, Glos., as Manager, Data Communications Products. He was previously Group Engineering Manager with RS Components Ltd.

Mr. P. J. Cracknell (Member 1973, Graduate 1971) who has been working as a research and Development Engineer with the South African Atomic Energy Board in Pretoria for the past eighteen months, has returned to the UK to take up the post of Project Engineer with Radiation Dynamics Ltd.

Mr. P. W. Crane, M.Sc. (Member 1966, Graduate 1963) has taken up the appointment of Principal Lecturer and Deputy Head of the Department of Engineering of the South Devon Technical College at Torquay. He was previously Senior Lecturer in Electronics and Communications at the Riversdale College of Technology, Liverpool.

Squadron Leader J. Cumming, RAF (Member 1973, Graduate 1970) has been appointed Electrical Engineer Communications 2B at Headquarters Strike Command, RAF High Wycombe. He was previously Officer Commanding Ground Radio Flight at RAF Marham.

Mr. M. W. J. Hailstone (Member 1974, Graduate 1971) is now a Nuclear Inspector (P & TO I) with the Nuclear Installations Inspectorate of the Health and Safety Executive in London. He had previously been a Professional and Technology Officer 2 with the Royal Aircraft Establishment at Farnborough, Hants, from 1974 to 1977, and he was with the British Aircraft Corporation at Weybridge from 1964 to 1974.

Mr. C. Hooper (Member 1972, Graduate 1967) has been appointed Senior Systems Engineer with Ferranti Ltd. at Bracknell, Berks. He was formerly Senior Engineer with Rank Radio International, Chiswick.

Flight-Lieutenant D. A. Longden, RAF (Member 1974, Graduate 1969) has taken up the post of Officer Commanding Engineering Squadron at RAF Fylingdales. He had previously been serving in West Germany as Elec.Eng. 2C1 with RAF Headquarters at Rheindahlen.

Mr. K. Muirhead (Member 1972, Graduate 1970) has been appointed Senior Lecturer in Electrical Engineering at Birkenhead College of Technology, Merseyside. He joined the College in 1967 as Lecturer Grade I.

Mr. S. G. Piper (Member 1973, Graduate 1971) who was formerly a Development Engineer with Plessey Radar, Cowes, has taken up the post of Project Engineer with Marine Exploration Ltd., at Cowes.

Mr. M. L. Raab, B.Tech. (Member 1974, Graduate 1970) has been appointed Chief Engineer at Rediffusion (East Midlands), with effect from August 1977. Mr. Raab, who joined Rediffusion London Ltd. in 1969 as a regional test engineer, was appointed engineer-in-charge of the Berkshire/Oxfordshire branch of Rediffusion London Ltd, in 1972 and two years later became Assistant Chief Engineer.

L. W. Barclay

D. W. Heightman

T. W. Welch

Mr. P. Sample, B.Sc. (Member 1970) has been appointed Manager of the new Radar Systems Division of Microwave and Electronic Systems Ltd., at Linlithgow. He was previously Group Leader with Racal Instruments Ltd., Bracknell, which he joined in 1967.

Mr. W. B. Smith (Member 1938) retired at the end of last March after more than 19 years with Rediffusion Engineering Ltd., as Head of the Test Section. He had been concerned with radio relay since 1930. During the war Mr. Smith was with the Aeronautical Inspection Department as examiner, mainly engaged in work on radar. He then worked with EMI for a time, until in 1957 he joined Rediffusion.

Mr. R. C. Theobald (Member 1972, Graduate 1968) who has been a Development Engineer with Motorola Funkgeräte GmbH in Mainz-Kastell, West Germany, since 1974, has returned to the UK to take up the post of Senior Development Engineer in the Obstetrics Division of Sonicaid Ltd. Chichester.

Dr. L. A. Trinogga, Ing.(grad.), M.Sc., Ph.D. (Member 1969) is now Senior Lecturer in Electronics in the Department of Physics and Physical Electronics, Newcastle upon Tyne Polytechnic, which he joined in 1968 as a Lecturer. He has served on the Committee of the North Eastern Section for several years and is at present Chairman.

Mr. J. Wade (Member 1969, Graduate 1964) has been appointed Manager, Computer Operations and Service, with GEC Electrical Projects at Rugby, which he joined in 1972 as Senior Section Engineer, Systems Test.

NON-CORPORATE MEMBERS

Flight-Lieutenant C. Cho-Young M.Sc., RAF (Graduate 1972) who since 1972 has been an Education Officer with RAF Locking, has been posted to RAF Cosford, where he is now Radar Course Officer, Advanced Electronics Squadron, with No. 2 School of Technical Training.

Mr. D. T. Goodwin, B.Sc. (Graduate 1971) has been promoted to Senior Field Engineer with Digital Equipment Ltd., London.

Mr. S. Keophysavong, B.Sc. (Graduate 1975) is now working as a Maintenance Technician with the Canadian Broadcasting Corporation in Halifax, Nova Scotia. He left Laos in 1975 and spent two years working in France and the UK before emigrating to Canada.

Mr. P. W. Walters, B.Sc. (Graduate 1974) previously a Development Engineer with EMI Medical Ltd., has taken up the appointment of Senior Systems Engineer with GEC Computers Ltd., Boreham Wood, Herts.

Mr. M. I. M. Cham (Associate Member 1976) is now a fully qualified Telecommunications Officer with the Civil Aviation Department of the Ministry of Works and Communications, Banjul, The Gambia.

Lieutenant Commander J. W. Smith, RN (Associate Member 1974) has been posted to HMS *Charybdis* as Weapons Electrical Officer. He was previously Officer in Charge at the Royal Naval Wireless Station, Crimond, Aberdeenshire.

Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at its meeting on 19th April 1977 recommended to the Council the election and transfer of the following candidates. In accordance with Bye-law 23, the Council has directed that the names of the following candidates shall be published under the grade of membership to which election or transfer is proposed by the Council. Any communication from Corporate Members concerning the proposed elections must be addressed by letter to the Secretary within twenty-eight days after publication of these details.

Meeting: 19th April 1977 (Membership Approval List No. 233)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Graduate to Member

BARNES, Melvyn. *Leigh, Lancashire.*
FAIRS, William George. *Rainham, Essex.*
JACKSON, Derek. *Stockport, Cheshire.*
RENWICK, Neil Herbert. *Mundon, Essex.*

Direct Election to Member

*GILHAM, Frank John Alfred. *Cardiff.*
HOPKINS, David Wyn Gruffydd. *Swansea, West Glamorgan.*
MITCHELL, Barry John. *Romsey, Hampshire.*
PACEY, Adrian John. *Bracknell, Berkshire.*
SHEPPARD, Martin Henry. *Blandford Forum, Dorset.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

EVANS, Paul Edwin. *Harlow, Essex.*
WHITEHEAD, Richard Donaldson. *Guildford, Surrey.*

Transfer from Student to Associate Member

ADASA, Nathaniel Emamomo. *London.*

*Subject to Mature Candidate Procedure.

Direct Election to Associate Member

ELPHINSTONE, Neil. *Swansea, West Glamorgan.*
LOVICK, Stephen. *Edgware, Middlesex.*
O'SULLIVAN, David Peter. *Dublin, Republic of Ireland.*

STUDENTS REGISTERED

HODGKINS, Alan Mark. *Coventry.*
NAISBITT, Donald Anthony. *Farningham, Kent.*
VONTAS, Dimitrios. *Cardiff.*
WOODFIN, Harold. *Manchester.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Member to Fellow

LEUNG, Wai Sun. *Hong Kong.*

Transfer from Graduate to Member

CHAN, Chack-Kuen. *Hong Kong.*
KHOO, Poon Tong. *Singapore.*

Direct Election to Member

LEUNG, Woon Yin. *Hong Kong.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

KAZIPE, Leonard Rufeo. *Zomba, Malawi.*

Direct Election to Graduate

WONG, Ping-Kwong. *Hong Kong.*

Transfer from Student to Associate Member

LEE, Tai Khin. *Singapore.*

Direct Election to Associate Member

ABEYESUNDERE, Sunil. *Rajagiriya, Sri Lanka.*
TEE, Lit Teong. *Kuala Lumpur, Malaysia.*
TEO, Ah Moy. *Singapore.*
YAP, Eng Soon. *Singapore.*

Transfer from Student to Associate

THIRUGNANASAMPANTHAR, Ampalavanar. *Jaffna, Sri Lanka.*

STUDENTS REGISTERED

CAULACHAND, Sunjeev. *Mauritius.*
KOO, Kock Beng. *Singapore.*
KWAN, Oi Lin. *Singapore.*
LAM, Ping Wah. *Hong Kong.*
LEE, Kin Lam. *Singapore.*
NG, Heung Sun. *Hong Kong.*
NG, Kwing Leung. *Hong Kong.*
ONG, Ka Sin. *Singapore.*
UDOKPOH, Okon Peter. *Opobo, Nigeria.*
WONG, Teck Boon. *Singapore.*
YUEN, Mei Leng. *Singapore.*

P

Peter Grugeon

HER MAJESTY QUEEN ELIZABETH THE SECOND

Patron of the Institution

and

His Royal Highness Prince Philip, Duke of Edinburgh, K.G., K.T., F.R.S.

Honorary Fellow of the IERE

1952 SILVER JUBILEE 1977