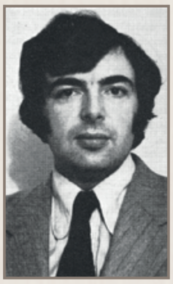
To celebrate the SCTE's 75th anniversary year, we will be running one archive article in each issue of *Broadband Journal* in 2020.

Proceedings at meetings of The Society of Cable Television Engineers on 17th October 1972, at the Institution of Electrical Engineers, Savoy Place, London W.C.2, with Mr. P. Scadeng occupying the Chair and on 31st October 1972, at the Post House Hotel, Bramhope, Nr. Leeds, with the President of the Society, Mr. R. J. Seacombe, in the Chair.

The Design Philosophy for a 5 - 300MHz Relay System - Part One

by Chris Swires (Fellow) Swires Research & Development, Hornchurch, Essex.



Chris Swires was born in Tadcaster near York, and educated at Tadcaster (Oglethorpe) Grammar School. He left Yorkshire to study at Enfield College of Technology and join Thorn Electrical Industries. He obtained his Higher National Certificate in Electrical Engineering and later, whilst still with Thorn, joined a group in its Transistor Applications Laboratory researching into the use of transistors in high frequency repeater amplifiers. As Senior Engineer, he developed a number of repeater amplifiers, for some of which he holds a patent. Before becoming Engineer in Charge of CATV Development, he developed Europe's first fully transistorised UHF to VHF converter. In early 1970, he joined Teleng as Manager of the Research and Development Laboratory, which has been responsible for a number of projects including development of the Teleng Derby Series, the Superverter (for which a patent is pending), and now the Oxford Series. He is married and lives in Essex.

The Paper outlines reasons for extending the bandwidth to 300MHz and the extra problems this involved; refers to the required distortion figures for the new equipment, with particular reference to second order intermodulation; and describes the methods used in obtaining the improved figures and the measurement methods. Further topics are the automatic level control requirement on large systems, the types of automatic level control available, and the system finally used on the range of equipment under discussion. The provision of reverse transmission in the Band 5-30MHz is outlined, and finally the mechanical design considerations and choice of housing is discussed.

Introduction

I would like to deal in some detail with the development of the Teleng Oxford range of Repeater Amplifiers. This Paper will be divided into three parts: the requirements of the system; the methods used to meet these requirements; and finally, the overall performance attained and the equipment which was produced as a result of this development project.

When the Oxford project commenced we knew we were undertaking a very expensive development programme. Great care therefore was taken in deciding the parameters of the equipment which we were to produce, to ensure that as wide a market as possible would be available to this equipment.

System Bandwidth

The first question which had to be answered was, could the present VHF Band of 40 - 230MHz carry the number of programmes likely to be required during the foreseeable future. This question was asked of Britain, the European continent and the American continent.

During our investigations it became apparent that the present 40 - 230MHz system could not cope with future requirements, particularly in continental Europe, America and Canada where the situation had already become difficult and where extra channel space would be required in the very near future.

In Britain the situation at that time was somewhat easier but the many developments which were being discussed, and indeed, actively pursued indicated to us that within the next few years demand for extra channel space was likely to occur.

Having decided that the 40 - 230MHz system was not adequate for all requirements we had to decide the Bandwidth to be used on the new equipment.

We examined several systems in use in America. It be- came apparent that a top frequency of at least 270MHz was required to enable 30 channel working to be achieved on the American system. The situation in Europe is made more complex by the number of different television standards in use, the result is that adjacent channel operation cannot be employed at present and greater spacing - for example about 14MHz - must be left between vision carriers on the system.

If we examine Fig.1, we can see that using a system with channels spaced at 14MHz even with 300MHz top frequency, we can only accommodate a maximum of 16 channels. This system is somewhat idealised as certain channels would have to be omitted due to local

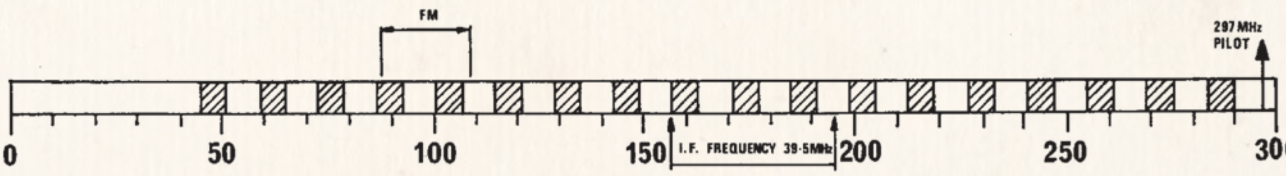


Fig.1. Channels spaced at 14MHz, 44 - 300MHz system

transmis- sions, and if an exact spacing of 14MHz could not be achieved, it may not be possible to utilise all the channels shown on this diagram. A spacing of 14MHz here enables local oscillator radiation from the television receiver to fall between channels, thereby obviating the need to eliminate certain channels because of local oscillator radiation.

As there were systems on the continent relaying as many as 10 television channels, and the likelihood was that this number would increase by at least 4, it became obvious that a bandwidth of about 300MHz would be required.

The upper frequency limit of 300MHz results in a cable loss 15% greater than that at 230MHz. Any further increase in top frequency will result in a greater loss being incurred, and for this reason it was felt that any increase in upper frequency should be kept as small as possible. This also avoids the need for complete re-engineering of systems already installed.

As a result of experimental work it became evident that existing transistors and techniques were quite capable of being extended to 300MHz. When upper frequency limits were taken very much higher, difficulties were encountered in the form of considerably reduced stage gain in the amplifiers and greater difficulty in obtaining good return loss performances. Satisfactory performance above about 300MHz would have resulted in a fairly sharp increase in cost.

After carefully considering all the points mentioned it was decided to develop the equipment with a top frequency limit of 300MHz.

Intermodulation Distortion

As Fig.1 indicates we considered transmission systems which employed the whole of the bandwidth available, that is from the lowest frequency of say 44MHz to 300MHz, encompassing not only band I and band III, but using all the inter-band space and the space above band III, commonly referred to as the Super band. The use of this continuous spectrum immediately poses a problem. This is that the carriers of any two channels can beat together to produce a frequency which falls in another used part of the band. The difficulty has been minimised up to the present time by avoiding areas where various second order sum and difference intermodulation products occur.

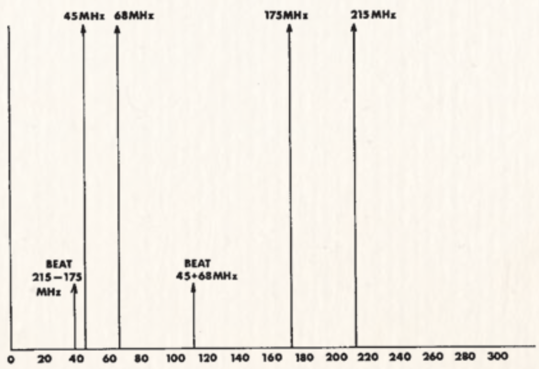


Fig.2.

Fig.2 shows that second order intermodulation products are not a particular problem using standard Band I and Band III frequencies. If we take two frequencies in Band I, say 45MHz and 68MHz, and feed these into a non linear device such as an amplifier, these beat together to produce a frequency of 113MHz, which is outside the normal tele- vision bands. Two channels in Band III, for example channel ‘D’ and channel ‘13’, frequencies 215MHz and 175MHz produce a beat frequency of 40MHz, which is just below Band I. If slight offsets of the carrier frequencies are employed on these two channels, it is possible that a beat frequency can be produced in channel 1. However, under normal conditions the standard Band I and Band III channels do not produce intermodulation products at frequencies which cause interference. Some troublesome products can be generated by the beat frequencies between Band I and FM, for example between 42MHz and 100MHz, resulting in a product at 58MHz, which can fall onto a television channel.

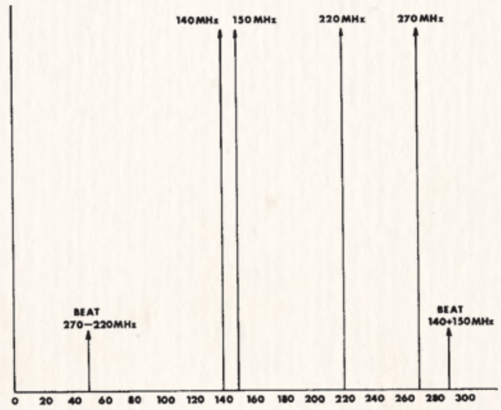


Fig.3.

It has been common practice to run FM channels well below the level of television channels and also to use slight frequency offsets, thus avoiding any major second order intermodulation problems when using standard channels.

It becomes apparent that if we use all the spectrum be- tween 40 and 300MHz, there are many possibilities of these beat frequencies falling within other channels.

Two examples are shown in Fig.3. Taking channels at 140 and 150MHz (the inter-band) a beat frequency of 290MHz is produced. Similarly, two channels in the super-band of say 220MHz and 270MHz, would produce a difference frequency of 50MHz. Therefore, in order to make full use of the extended band available and to use the inter-band and super-band channels we must limit the amplitude of the second order, or, intermodulation products. These must be held at a level which will prevent visible distortion of the carrier at the point where the inter-modulation product falls.

Most repeater amplifiers available prior to the develop- ment of the Oxford series had a somewhat poor performance in this respect. This in most cases resulted in visible inter- ference to carriers at or near the point where the beat frequency occurred. We could see that a very important requirement for the next generation of repeaters was that the second order intermodulation distortion should be such that the full bandwidth could be used, without the need for selecting and offsetting channels. The necessity for such selection and offsetting would limit the number of usable channels and the usefulness and versatility of the equipment.

Frequency Response

A further point which received our attention was the obvious need for close frequency response limits, if the maximum length of trunk network was to be achieved. This requirement was made more stringent by the extra bandwidth involved and it was decided that specification of plus or minus a quarter of a dB, per repeater, in response flatness was required. It was also noted that this response should be maintained during adjustment of both the gain and tilt of the amplifier. There is little point in designing an amplifier with a completely flat frequency response, when viewed in the maximum gain position, if the operation of either the attenuator/equaliser network, or the slope and gain controls, cause the response to be appreciably degraded.

Automatic Level Control Systems

As the equipment we were designing was for large networks, it was essential that a form of control be applied which would compensate for variations due to temperature effects, not only on the amplifiers but on the cable network. It is not sufficient to ensure that great care is taken with the stability of the amplifier with varying temperature, circuits must also be used which compensate for changes of loss in the cable used on the network. Several systems have been employed for compensating for the changing characteristics of the cable network on a VHF relay system. These can in the main be divided into two types, (1) passive networks, and (2) active networks. The passive type of network normally employed is the use of a temperature sensing device which senses the ambient temperature of the cable or amplifier, and adjusts the gain or slope of the amplifier to suit the prevailing temperature conditions. The principal difficulties encountered in this type of system, are in matching the compensation networks to the characteristics of the cable network, and in placing the equaliser at a point where it senses and compensates the cable temperature changes.

A cable may take many hours to respond to a change in ambient temperature, whereas a simple thermal sensor changes in minutes.

A further difficulty is that this type of network cannot sense and compensate for small changes in amplifier gain or even variations on the network accessories. It was considered therefore, that while a passive network would probably be adequate for small systems or as an additional compensation method, some form of active control would have to be employed on large systems.

The type of control which has become most popular in recent years is the use of a pilot carrier generator, and a pilot sensing device. As is well known this system transmits a pilot carrier which is detected at an amplifier, and used to control the gain of the amplifier. This is a form of servo control where the ALC unit corrects the system gain assuming that when the pilot is set to the correct output level all other signals will also be correct.

It is worth considering the workings of ALC systems in some detail since the use of a good ALC unit is the key to large system stability.

Fig.4 shows the change from nominal level of 5 cable lengths each with 20dB loss at 300MHz. At 44MHz these would have a loss of 7.7dB. If we study this plot we can see that the change of loss is greatest at 300MHz. This is because cable changes with temperature at the rate of .002dB per dB loss per °C. As the cable loss is greater at 300MHz the variation of attenuation due to temperature is therefore correspondingly greater. The reduction in loss for a 20°C drop in temperature is 1½dB at 44MHz and 4dB at 300MHz. Conversely when temperature rises by 20°C the loss increases by 1½dB at 44MHz and 4dB at 300MHz.

We can see in Fig.5 the effect of controlling the level by means of a 297MHz pilot. The lower line shows the change

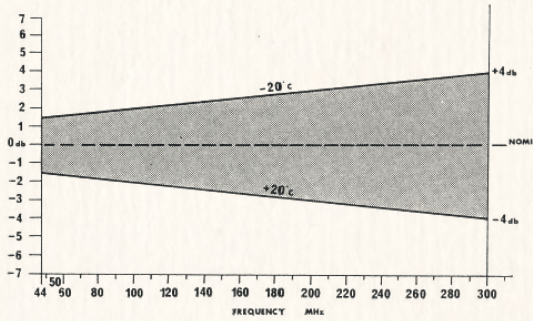


Fig.4. Five cable lengths, 20dB loss at 300MHz. Change from nominal due to +20°C.

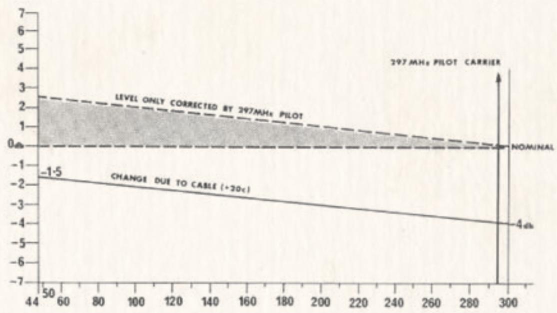


Fig.5. Effect of 297MHz pilot controlling a flat gain change.

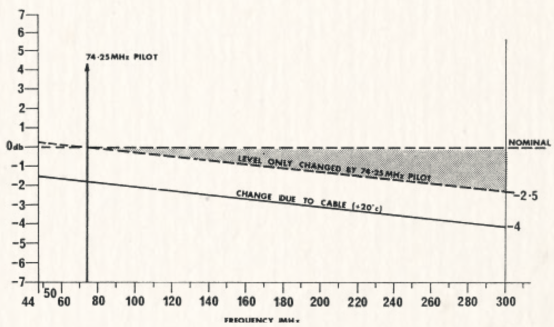


Fig.6. 74.25MHz pilot controlling a flat gain change.

due to a temperature rise of +20°C. In a single pilot carrier system with the pilot set at 297MHz, the ALC unit corrects the level at this frequency. However, it does not correct the slope which has been imposed by the cable's loss characteristic. The overall result therefore is that whilst the level is now corrected at 297MHz, the 44MHz end of the band has risen by 2½dB.

In Fig.6 the effect of controlling the level with a pilot at 74.25MHz, is evident. Whilst the pilot correctly adjusts the level at 74.25MHz, the change is not sufficient at 300MHz. Therefore this end of the band remains about 2½dB down on the nominal level and it is once again incorrect. From these graphs we can deduce that the best compromise which can be achieved, using a single pilot controlling a flat gain change, is that obtained by a pilot in the vicinity of the mid point of the band.

In Fig.7 after only five amplifiers the ALC has corrected the level at the mid point in the band, but raised the level of the 44MHz end of the band by 1.25dB and reduced the

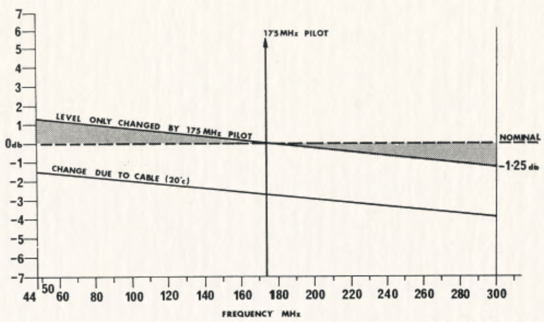


Fig.7. Best compromise for single pilot controlling a flat gain change.

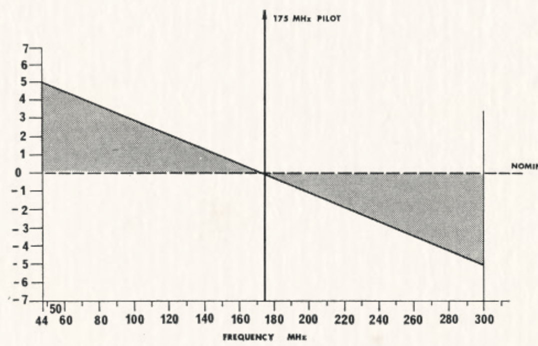


Fig.8. Single pilot flat gain control. Effect over 20 amplifiers for -20°C change in temperature.

300MHz end of the band by 1.25dBs. This effect is cumulative and the consequence can be seen in Fig.8. A 20 amplifier cascade with a +20°C temperature change gives the result shown. Single pilot carrier systems of this type increase the output at 44MHz by 5dB and cause a 5dB fall in output at 300MHz. This is a serious response change over 20 amplifiers.

One method which has been adopted with some success is that of using a single pilot carrier to operate a network which controls both the slope and the gain of the amplifier. In this way a change of the type shown earlier can be correctly compensated by using a combined gain and slope network. A network of this type is matched to compensate for cable loss and gives a slope of 2.5dB between 44 and 300MHz in addition to a change of gain of 4dB at 300MHz.

Use of this method presupposes that all the changes in a network are in the cable and follow the temperature law imposed by the cable. As we know the cable is not the only part of the system which undergoes a change of performance with temperature. A typical amplifier could change by perhaps .25dB in gain, over a 20°C temperature change.

The effect of amplifier gain change is shown on the lower line of Fig.9. If 20 amplifiers change by .25dB there will be an overall fall in gain of 5dB, which would be corrected by the ALC system giving the effect illustrated by Fig.10.

We can assume here that the cable change has been correctly compensated by the use of the slope/gain pilot system. However, there remains the -5dB change of gain caused by the amplifiers. The pilot which is at 297MHz increases the gain on the ALC unit and brings the 297MHz pilot to the nominal level. In doing this it adjusts the slope and gain networks as this is the only change which the unit is capable of making. Fig.10 shows the final result. Over this length of system the 44MHz signal has now fallen by 3.1dB. When it is considered that in an uncompensated system the gain change would have been -11dB at 44MHz and -21dBs at 300MHz, this can be regarded as a reasonably satisfactory result. However, even a 3dB change in level at 44MHz after 20 amplifiers can have significant effects on large system reach.

Therefore we concluded that an ALC system of this type would only be completely suitable for use on a maximum of 20 amplifiers in cascade.

As a result of our studies, it was clear that to achieve optimum network performance a system was required which would control not only the gain of the ALC unit but also the slope and would enable these two parameters to be independently adjusted. Gain change on the network and change due to slope - whether on cables or any other parts of the system - could then be independently corrected. This conclusion led us to investigate the use of a dual pilot carrier system.

In the dual pilot carrier system which I shall describe the 297MHz pilot controls the gain of the ALC unit, and a suitable range would be ±5dBs as shown in Fig.11.

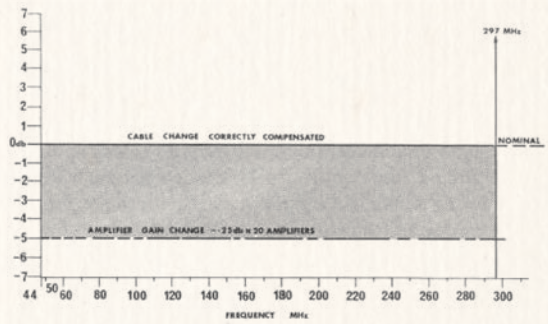


Fig.9. Single pilot controlling slope and gain network.

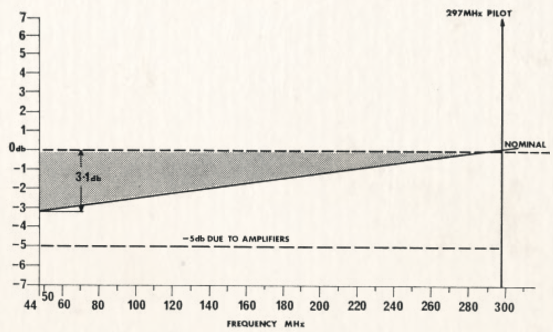


Fig.10. Effect of single pilot controlled tilt/gain network over 20 amplifiers and cables with .25dB of flat gain change due to each amplifier for + 20°C temperature change.

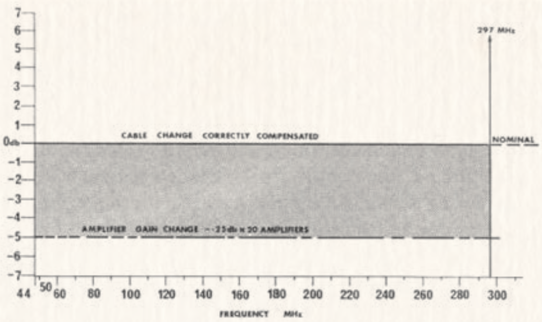


Fig. 11. 297MHz pilot controls gain by 15dB.

You will notice in Fig.12 the effect of the low frequency pilot which we placed at 74.25MHz. This controls only the slope of the amplifier, the gain at 300MHz remains fixed and the slope is varied to produce a ±3 dB change at 44MHz.

We now have two independent controls which can be operated by the pilot carriers; the 74.25MHz pilot controls the slope while the 297MHz pilot controls the gain.

Change due to temperature over 5 amplifiers and 5 cables is presented in Fig.13. The change due to the cable is shown on the upper line. The additional change due to the amplifiers is shown added to the cable change. This results in a complex change which could not be fully corrected by a single pilot system.

Fig.14 shows the correction due to the 297MHz pilot only. This increases the level until the original response is corrected at 297MHz but the 44MHz end of the band becomes 2.8dB high

In Fig.15 we see how the 74.25MHz slope pilot can then correct the residual slope. The 74.25MHz section of the ALC senses the increased level at this frequency and corrects the slope of the unit to bring the level to the correct nominal value throughout the band. In practice of course both these two operations would occur simultaneously.

The analysis to which you have just been subjected shows the real advantage to be gained from the dual pilot system and shows that it should certainly be employed on networks with more than 20 amplifiers in cascade.

It is important of course that the operation of the ALC system should be such that it causes as little degradation to the other system parameters as possible. That is, it should not seriously deteriorate intermodulation, cross modulation or noise. A further consideration when designing the ALC unit was that where practicable, it should be possible to replace a normal amplifier assembly with an ALC unit at a later date. Frequently large networks are modified to suit changes in population patterns, extra housing developments etc. It was therefore considered desirable that the ALC unit should control networks within the trunk amplifier rather than utilising a completely separate unit introducing only loss into the system.

Cross Modulation and Noise

Two other well known system parameters are cross modulation and noise. The cross modulation performance of modern amplifiers is largely limited by the performance of transistors currently available. Some improvement in cross modulation was possible during the development of this amplifier by very careful optimisation of all the relevant parameters.

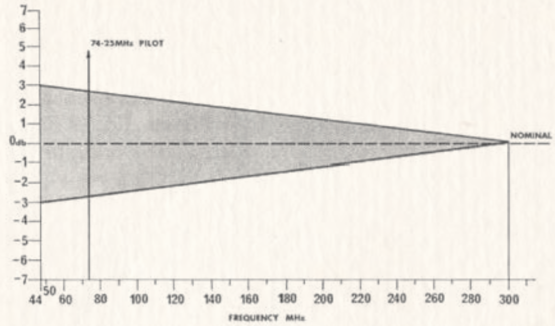


Fig.12. 74.25MHz pilot controls slope by ±3dB between 44MHz and 300MHz.

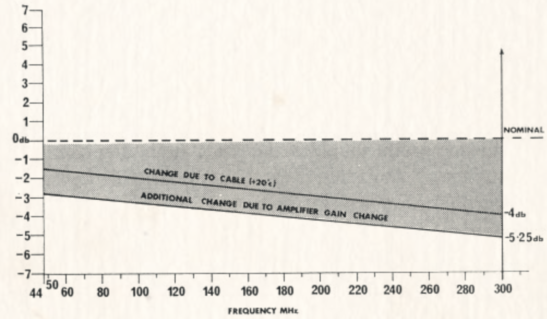


Fig.13. Changes due to temperature over five amplifiers and five cables.

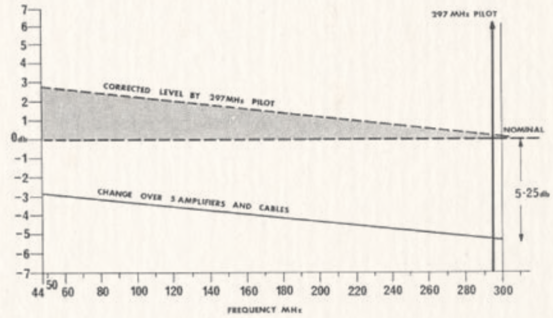


Fig.14. Correction due to 297MHz flat gain change pilot only.

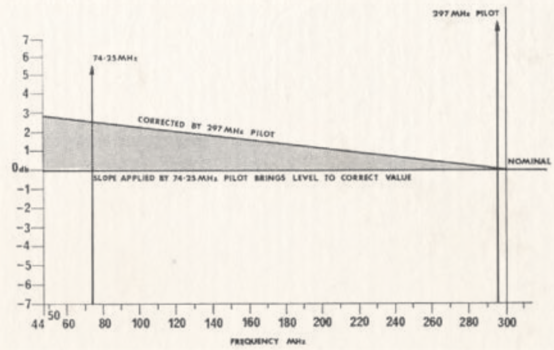


Fig.15. Correction by dual pilots. 74.25MHz slope pilot added to correction applied by 297MHz gain pilot.

The target was to obtain the best possible cross modulation figure from the devices. Noise figure is largely determined by transistor performance and care must be taken to use available semiconductors to their best advantage. One of the main decisions which can affect noise figure is the position of fixed or variable attenuators in the amplifier assembly. Any attenuation placed prior to the first transistor affects the noise figure.

However, as attenuation is normally applied only when the level of the signals is high, the overall signal to noise figure is not degraded. When the attenuation is placed after the first stage of the amplifier the network does not affect the noise figure of the first stage. It is not until the level of attenuation inserted is approaching the same value as the gain of the first stage that the overall amplifier noise figure falls.

When placing the attenuator after the first stage, great care must be taken to ensure that the cross modulation and intermodulation performance is adequate, otherwise overloading will occur due to the signals being at a higher level than that at which they would normally be applied to the first stage. If this distortion requirement can be met then it is advantageous to place the attenuator after the first stage.

Mechanical Design Considerations

We now move to the mechanical design considerations for a relay amplifier. From the outset it was considered that good mechanical housing and ease of maintenance was of the utmost importance. A considerable amount of thought was given to the mechanical construction of the units and many customer comments were taken into consideration during the mechanical design phase. The requirements which emerged from our studies were; that a plug-in modular system was desirable to facilitate fault finding and maintenance, and to make updating of components easier. Several forms of casing were considered, the main criteria being that the housing should be both robust and weatherproof.

Choice of Gain of Units

It is well known that the maximum reach of a relay system can be achieved using an amplifier of 8.7dB gain. However, the best practical value for long reach systems is between 15 and 25dB. As the Oxford Series is mainly intended for use on larger type networks, the two gains chosen for the assemblies were 21dB or 17dB. The 17dB system gives a particularly long reach. A very full description of the effects of gain on system reach was made in a paper given by Mr. Seacombe to this Society in 1971.

Reverse Transmission

One of the newest factors which has entered the sphere of the relay operator is the recently expressed interest in the use of reverse transmission on relay systems. At the present time reverse transmission generally means trans-mission from the opposite end of the system to the head end, whether this be on the trunk network of even from a subscriber's outlet back to the head end. This is normally considered in the band below Band I, that is approximately in the Band 5 - 30MHz.

There are several reasons for interest in this facility. The first is the possibility of feeding back television channels particularly those generated by the relay operator in the form of local origination.

This means that the system operator can use a camera and modulator to originate a programme at any point on his network. This programme may be fed back to the head end for reprocessing and then re-transmitted to the subscriber on another channel.

The second possible use for reverse transmission is for transmission of either data, or requests for information from each subscriber. Many possible projects have been outlined, some of them of an extremely advanced and expensive nature. Among these we can include the possibility of linking each home to a computer or information bank, which will enable the subscriber to call for data or computer time from his or her home. Other less ambitious schemes include the possibility of requesting specific information on local news, local events, etc. The possibility of using this reverse facility for measuring audience ratings and/or billing systems on pay television networks can also be envisaged.

The third use is the monitoring of the performance of the system itself. Data on the signal levels at various points on the system can be fed back to the head end. This information can be monitored and if the levels on any part of the network fall outside a pre-determined limit, suitable warning can be given.

It was therefore considered desirable that provision should be made in this new amplifier development for the inclusion of reverse transmission facilities. Owing to the somewhat futuristic nature of many of the reverse transmission projects, it was decided that it should be possible to make the amplifier assembly available, as an alternative, without reverse transmission, but with provision for adding this at a later date.

I have now outlined the system requirements in some considerable detail and it will probably be useful at this stage to underline the conclusions which we reached:

1. That the frequency range should be 5 to 300MHz.
2. That particularly in view of the extended bandwidth, great care would be required to reduce the inter- modulation distortion to a level where it would not affect system performance.
3. That the lowest possible noise and cross modulation should be achieved.
4. That great care should be taken that a flat response be obtained and that this be maintained at all settings of the gain and tilt control networks.
5. That a dual pilot ALC system should be developed for this equipment.
6. That reverse transmission facilities should be in- corporated.
7. That great care should be taken with the mechanical design of the unit to ensure ease of installation and maintenance.

These then are the requirements and I will now outline the methods used to meet these requirements.

The Trunk Amplifier Unit

Fig.16 is a block diagram of the amplifier assembly, and the final arrangement used in a trunk amplifier with a gain of 23dB can be seen. At the input to the unit is pro-vision for a plug-in attenuator and plug-in equaliser. These are selected according to the length of cable preceding the amplifier. The first stage of the amplifier follows these components and is designed to have as low a noise as possible whilst maintaining an adequate cross modulation performance. Good cross modulation performance is particularly necessary because the variable attenuator is placed between the first and second stage. The attenuator is electrically variable between approximately 0 and 10dB. This means that when the attenuator is turned to the -10dB setting, the first stage may have up to 10dBs more than its normal working level applied to it. Great care was taken in designing the first stage to minimise any distortion, due to the extra load placed on this stage when the attenuator is turned down by 10dB. The variable attenuator is the next component and this is controlled by a DC voltage. In the case of a manual amplifier the DC voltage is simply derived from the power rail and the level set by means of a potentiometer. This arrangement gives a fine gain control which can be used in addition to the plug-in attenuator to set the amplifier to the required gain. In the case of the automatic level and slope control assembly the variable attenuator is fed via a DC supply from the ALC.

The second stage incorporates the slope control network. Slope control is achieved by means of frequency selective feedback networks around this stage which apply the correct slope law to compensate for cable variations. Again the control is applied by means of a DC voltage. In the case of the manual amplifier which is an assembly without an automatic level control system, this voltage is set by means of a potentiometer forming the slope control. In an assembly containing automatic level and slope control the voltage is fed from the automatic slope unit to operate the slope network. Following the second stage are two further stages which are arranged in a push-pull configuration.

The reason for adopting a push-pull amplifier is in order to minimise second order intermodulation distortion, and it is worth studying in some detail the operation of such an amplifier at this stage. Firstly, we must consider the basic cause of second order intermodulation distortion in a transistor. This occurs due to the non-linear transfer characteristic of the transistor and it can be shown that second order intermodulation distortion is caused by the square-law part of the transistor characteristic. Like second harmonic distortion, intermodulation appears as a part of the squared term in the expansion of the transfer characteristic of an amplifier.

It is very difficult to demonstrate graphically the cancellation of inter-modulation products in an amplifier. This cancellation process is much more easily understood if we consider second harmonic distortion. This simplification is valid as both the second order inter-modulation products and 2nd Harmonic products are generated by the square law characteristic of the amplifier.

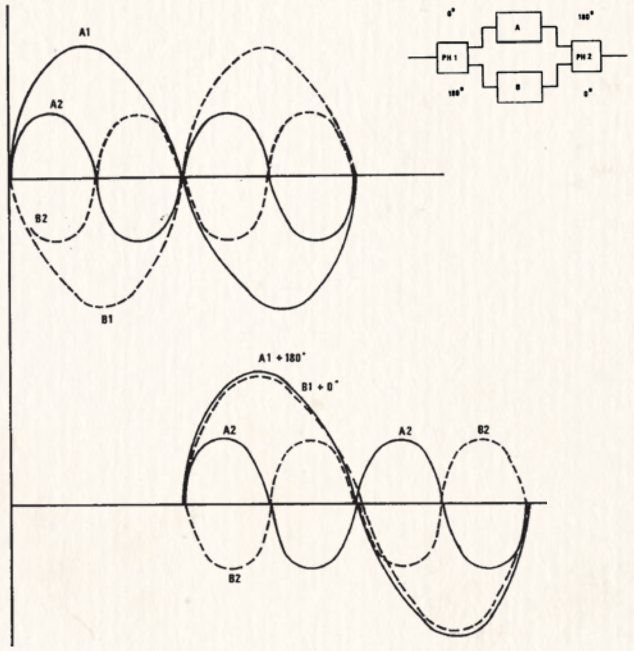


Fig.17. Cancellation of 2nd harmonic and intermodulation distortion.

Fig.17 show the output from a push-pull amplifier with a sinusoidal signal marked Al, and its second harmonic marked A2. The signal Al is fed into a phase splitting transformer which is shown on the diagram and marked PH1. Signal Al, after passing through the phase splitter, is then passed to the amplifier A without undergoing any phase shift. Out of the other arm of the phase splitter signal A is changed in phase by 180° and passed into amplifier B. Both these signals are fed into their respective amplifiers and at the output of these amplifiers, the relative phases will be as shown in the diagram, with A in phase with the original input. B1 is the A signal

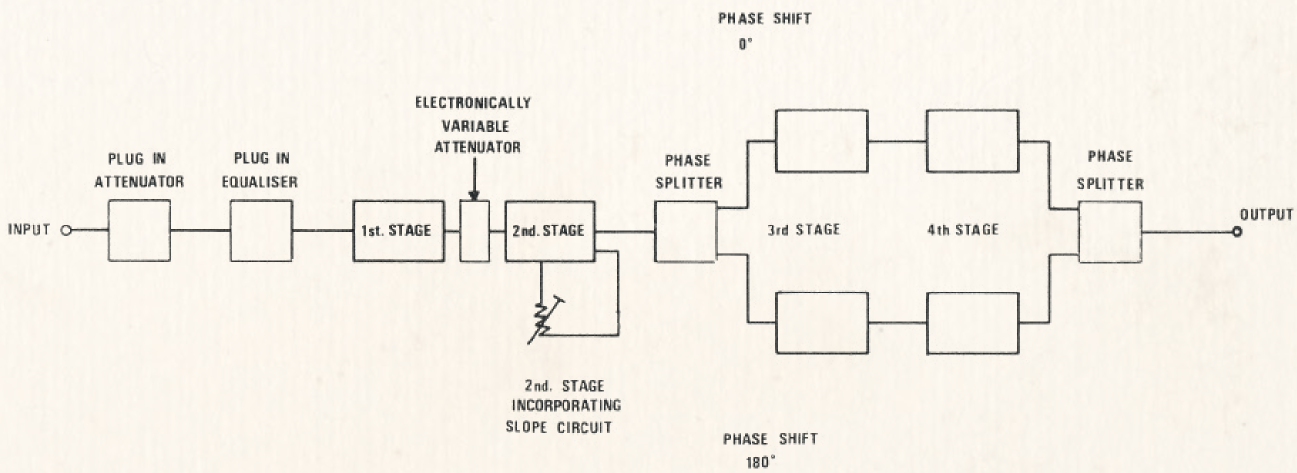


Fig.16. Oxford amplifier module.

which passed through the 180° phase shift and the B amplifier, B1 is shown by the broken line and is out of phase with the original input by 180°. The respective second harmonic products produced in the amplifiers are shown as: A2 which is the second harmonic of Al and B2 which is the second harmonic of B1. In the recombination process within the phase splitter PH2 the A1 signal is changed by 180° and the B1 signal remains in the same phase. This means that the two signals A1 and B1 are brought into phase and together form the output signal as shown on the lower diagram.

These would add to give an increased amplitude but for simplicity, this fact has been omitted from the diagram. The signals A1 and B1 have been suitably phase shifted and their second harmonics follow; A2 is not phase shifted in PH2 but B2 is phase shifted with B1. B1 is shifted by 180° in PH2 but as B2 is twice the frequency it is effectively shifted by 360°. This means that its relative position on the lower diagram referenced to that of the upper diagram is unchanged and it still appears in antiphase to signal A2.

Therefore, at the output we have a situation where A1 and B1 are adding but A2 and B2 are now in antiphase. This means that cancellation of second harmonic products A2 and B2 takes place. The cancellation of second order intermodulation distortion is of exactly the same nature. Unfortunately cross modulation, being a third order term, cannot be cancelled in this way. However, the use of a push-pull stage means that two transistors are being operated in the output stage and are sharing the output power. Consequently, up to 3dB improvement in output voltage can be obtained.

It may be asked if cancellation of second harmonic distortion or intermodulation occurs in a push-pull output stage, why use two stages in push-pull; why cannot the cancellation take place only in the output stage? The answer is that the push-pull stage can only cancel second harmonic distortion which is generated within itself. Any second order products which have been generated prior to this stage are passed through the amplifier in the same way as any other signal. Therefore, second order distortion generated in the single ended non push-pull stages is amplified along with the other signals in the output stage.

Why then not make the whole of the amplifier push-pull? If this were done cancellation could occur in every stage, and would result in no intermodulation distortion. Whilst this statement appears true in theory, in practice complete cancellation cannot be achieved. Furthermore, the first stages of the amplifier operate at relatively lower levels and no significant improvement in overall intermodulation is obtained in making them push-pull. To have stages in push- pull is obviously more expensive and consumes more current than to use non push-pull stages. On these grounds alone, therefore, it is desirable to keep the number of stages in push-pull to the minimum.

At the beginning of the design study it was decided that the required second order intermodulation level from the new amplifier would be a B2F of 113dBmV for the complete amplifier assembly. That is about 115dBmV from the output of the actual amplifier, allowing 2dB for losses in splitter networks to the bridger amplifier, feeds to ALC and test points.

For example, working at an output of +40dBmV, the second order intermodulation would be 73dB down, that is 40 + 73 = 113dBmV.

This B2F system of measurement has been used for some years at Teleng and considerably simplifies system and amplifier calculations. If we assume that the output stages contribute no second order intermodulation, any intermodulation distortion from the repeater amplifier is generated purely in the single ended stages. Now these stages are operated at a lower signal level than the output stages, and at this lower level contribute less intermodulation. For example, taking the gain of the output stages to be 12dB, and a required second order intermodulation distortion figure at the output of the amplifier of -73dBs at 40dBmV, then the allowable intermodulation at the output from the single ended stages at the same signal level is 12dB greater than this figure. This is because on reducing the output level by 12dB the intermodulation distortion also fall by 12dB. The resulting second order intermodulation requirement at the output of the driver stage becomes 101dBmV B2F or 61dB down at

+40dBmV out. Again we are assuming perfectly balanced push-pull stages are employed. With careful design it was found possible to produce a single ended driver stage which was capable of this order of second order intermodulation. We therefore designed the amplifier with two stages of push-pull. Had we made only a single stage push-pull, we would have achieved a 6dB worse second order intermodulation figure.

The foregoing remarks are based, for the purpose of explanation, on the assumption that a perfect distortion free push-pull stage can be achieved. The problems involved are quite significant and, of course, in practice a distortion free push-pull stage cannot be achieved.

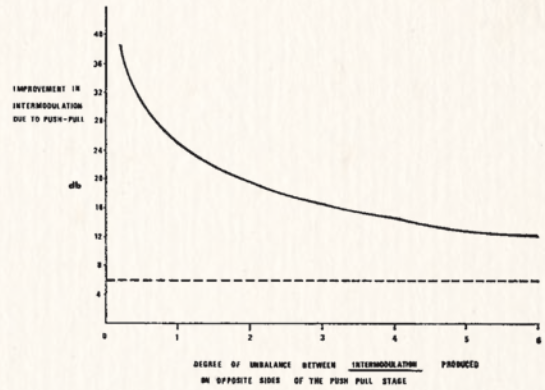


Fig.18.

Fig.18 shows the theoretical improvement possible using a push-pull stage with perfect phase balance. On the vertical axis the improvement in second order intermodulation in dB due to push-pull is plotted, while on the horizontal axis the difference in intermodulation between the two sides of the push-pull stage is shown.

If there are equal amounts of second order inter- modulation in each side of the push-pull amplifier, there will be complete cancellation and an infinite improvement in intermodulation can be achieved. With only 1dB difference in the intermodulation produced, this figure drops to 25dB and with 3dB difference to 16dB. This may not seem serious, but remember we are already talking of distortion of the order of 60dB down at 50dBmV output. This means that we are talking of differences in second order intermodulation between the two sides of the amplifier of 1 - 3dB in 60dBs, or as little as ten parts in 1000. Accurate matching must be achieved, not just at one frequency, but over the whole band from 44 - 300MHz. The one en- couraging factor in this result is shown by the dotted line, 6dBs above the zero line. With a push-pull circuit you win a 6dB improvement in second order intermodulation - free! - as in the soap powder advertisements. This is because, as the two stages are effectively in parallel they can be run at a 3dB lower level resulting in an overall improvement in second order intermodulation of 6dB.

By careful design, it was found possible to make a repeatable output stage which would allow us to achieve the design figure of 115dBmV B2F from the repeater amplifier. It is interesting to note that as further stages are added to the push-pull amplifier, that is for example a three section amplifier instead of a two section, good cancellation becomes more difficult to achieve. This was a further reason why every effort was made to optimise the first two stages to minimise their second order intermodulation distortion thus allowing us to use only two stages of push-pull amplification.

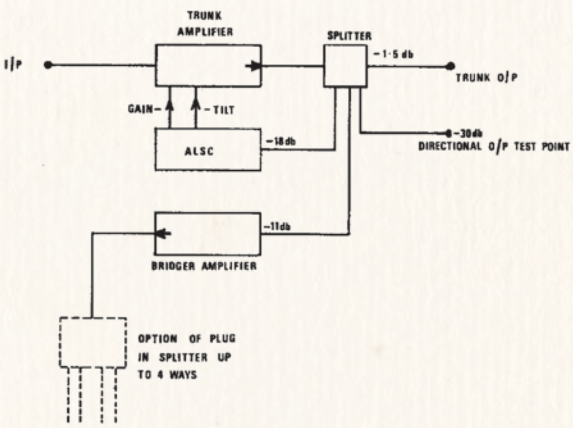


Fig.19.

A block diagram of a complete assembly is shown in Fig.19. Here we see that the output of the trunk amplifier module feeds into a 1½dB splitter which has three spurs. The first feeds signals to the automatic level and slope control unit from which this selects the two pilots necessary to control the gain and slope networks in the trunk repeater amplifier. The second output from this off-trunk splitter feeds to the bridger amplifier. From the bridger there is provision for plug-in splitters to allow up to four further distribution lines. The third output from this splitter is to a directionally coupled output monitor point. The use of a directional coupler ensures that the signal on this monitor corresponds to those on the output of the amplifier and that the reading is not seriously influenced by any cable or equipment following the amplifier. The trunk and bridger amplifier modules are in fact physically similar; are inter-changeable and are built in three gain versions. This means that any gain bridger unit may be used with any trunk unit. The gain range on both the trunk and bridger module lies between the 18½dB plug-in module and the 29dB plug-in module.

With the trunk working at the nominal correct operating level there is sufficient gain in the bridger amplifier to allow a launch from the bridger at its maxi- mum rated output. This high gain in the bridger means that in many cases the need for further line extension amplifiers is either eliminated or considerably reduced. In the case of less complete assemblies the modules which are not required are simply removed and where necessary 75 ohm load units fitted. The main chassis for the unit is standard. At any time a trunk amplifier can be raised to the status of an automatic level and slope control unit or have a bridger unit added. On a relay system where extensions and modifications to the network are quite often necessary, changes to the amplifier configurations are readily achieved.

This article will conclude in the next issue of *Broadband Journal*.