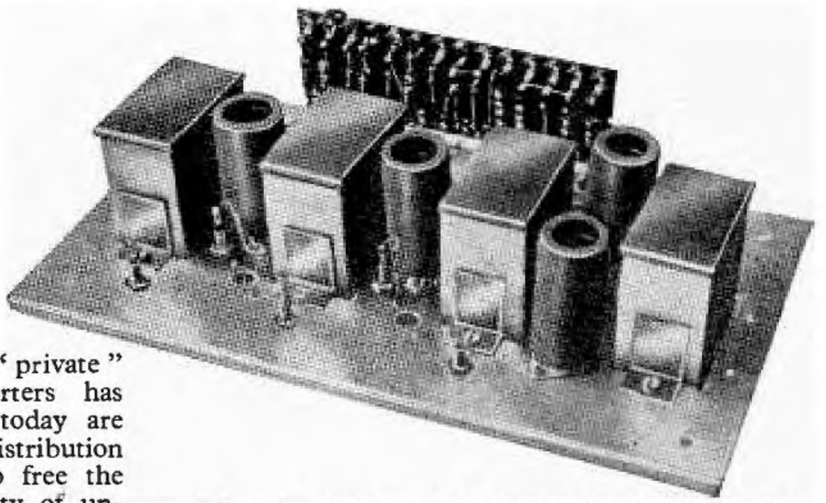
Wide-Band Linear R.F. Amplifier

For Use in Ships' Communal Aerial Systems

By B. F. DAVIES, A.M.I.E.E.*



SINCE the last war the need for the "private" broadcast receiver in the crew's quarters has been recognized and many new ships today are planned to be fitted with an aerial distribution system. In many cases this is done to free the ship's superstructure from the multiplicity of unsightly wires that would otherwise appear as more of the crew obtained receivers. In a small ship a multiplicity of "private" aerial wires may constitute a positive danger by altering the calibration of the d.f. loops and in oil tankers there is a fire hazard if sparking occurs in a badly installed aerial when the ship's MCW transmitter is working.

The obvious solution is to install one "official" aerial for broadcast reception and to serve all receiving points through a wide-band r.f. amplifier and coaxial cables.

Basic Requirements.—Considering the aerial first, it will be appreciated that the position, height and length of the aerial will be largely dictated by the layout of the ship and the position of the transmitting aerials. Difficulties are greater on the smaller ship and resort has often to be made to a smaller active length of aerial wire. It has not been uncommon with a transmitter of 500 watts and a main aerial length of 200ft to obtain with a receiving aerial of 100ft about 50 volts of r.f. across the 75-Ω coaxial cable termination. From statistics available this high voltage is to be found at transmitting frequencies somewhere in the region of 6 Mc/s. Length of aerial and spacing will modify this resonance considerably. At other frequencies usually less than 1 volt will be induced for the same transmitted power. No amplifier can handle 50 volts of input, and in practice a limit of about 25 volts would be set, adjusting aerial length or position to keep the "resonant" voltage down.

Some means of rejecting the transmitting frequencies has to be adopted and a standard form of filter was used. Assuming the worst case of 25 volts on the aerial coaxial lead-in cable and a minimum filter attenuation of 36 dB, the amplifier must handle 0.4 volt. From Table 1 it will be appreciated that

TABLE I.

MARINE TRANSMITTING BANDS	
415 —525 kc/s	M.F. Band
1.6 —3.8 Mc/s	I.F. Band
4.063—4.238 ,,	H.F. Band
6.20 —6.357 ,,	,,
8.195 —8.476 ,,	,,
12.33 —12.714 ,,	,,
16.46 —16.952 ,,	,,
22.00 —22.400 ,,	,,

* Formerly with the Marconi International Marine Communication Company Limited.

considerable band filtering would be required. It was decided that the standard amplifier should incorporate two band-stop filters covering the M.F. and I.F. bands only. In regions of dense shipping, transmissions in these two bands from ships close by can give rise to a total r.f. voltage on the aerial of as much as 50 mV. A filter unit for the H.F. bands used by the ship's transmitter would be separate and would be an optional extra. Where H.F. band transmissions are likely to be few in comparison with those in other bands, such a filter unit could be saved by using a remotely controlled relay to disconnect the aerial completely in order to keep high voltages out of the amplifier input during periods of transmissions in the H.F. bands.

To cover all the broadcast bands, the amplifier should possess a sensibly flat frequency response between 150 kc/s and 25 Mc/s. The input impedance to suit the aerial coaxial cable should be 75Ω and the outputs also suitable for feeding into 75Ω coaxial cables. The method of distribution is shown in Fig. 1, from which it is evident that although the loading, and consequent slight mis-match at each outlet point, is quite small a large number of receiver outlets could load the cable and introduce a cabling loss. The input impedance of each receiver differs considerably at any one frequency according to the tuning of the receiver and consequently the summation of all the loading effects is modified. It was decided to cater for up to 20 outlets and that four amplifier outlets would give a good economic compromise. In standardizing on equipment the circuit complexity and reliability have to be weighed against the average demands. In isolated cases, as indicated by present market demands, a ship requiring more than 80 outlets could be covered with two or more amplifiers working in parallel.

The amplification required was fixed at approximately 15 dB. As will be evident later, this again was a compromise between producing voltages at the receiver outlet points identical with that on the aerial terminated with 75Ω and keeping spurious responses within the amplifier to a minimum. In practice the overall gain of the system between aerial and the input terminals of the receiver also takes into account the fact that the aerial fitted for the distribution system will have a greater effective height than the average length of wire that can be slung haphazardly. Thus the equivalence of voltages at aerial and receiver can be disregarded. Emphasis was laid on making available in the cabin a "clean" signal free from machine noise interference, spurious responses, etc.

Effects of Amplifier Non-linearity.—The term a "linear amplifier" is, of course, only relative, but it does imply that in addition to producing an amplifier with a wide-band response attention has also been paid to obtaining the maximum degree of linearity. The wide-band amplifier suffers from the inherent disadvantage that in amplifying many signals simultaneously any non-linearity will produce other signals at the output. A power series can be used to represent the "curvature" and is in the familiar form $ax + bx^2 + cx^3$. This is equivalent to saying that the amplifier is capable of producing, as in audio techniques, the second and third harmonics of a fundamental sine-wave input. The two input signals are represented by $p \sin \alpha$ and $q \sin \beta$, and where one is considered to be modulated, p becomes $p\,(1 + m)$ where $m \not> \pm 1$. Analysis shows that the first term can be neglected, but the second term produces the expression bpq $[\cos (\alpha - \beta)t - \cos (\alpha + \beta)t]$. Thus the original carriers α and β have produced spurious carriers $(\alpha - \beta)$ and $(\alpha + \beta)$ both of which are dependent on the product $p\,q$. This means that modulation of either original carrier will produce modulation of the spurious carriers. This is called inter modulation (I.M.).

The third term produces the expressions $3c\,p^2 q$ $[\sin \beta\ t - \frac{1}{4}\sin\ (2d + \beta)t + \frac{1}{4}\sin\ (2\alpha - \beta)t]$ and $3c\,p\,q^2$ $[\sin \alpha\ t - \frac{1}{4}\sin (\alpha + 2\beta)t - \frac{1}{4}\sin (\alpha - 2\beta)t]$. The second and third terms in each of the brackets are similar to the I.M. terms discussed above except that the spurious responses are at different frequencies and that the spurious carrier modulation is proportional linearly to one original carrier and to the square of the other. This would mean that for low levels of original modulation the spurious carrier modulation would be double that for modulation of either p^2 or q^2 compared with modulation of p or q. Thus if p carrier is modulated the spurious carrier modulations at $(2\alpha + \beta)$ and $(2\alpha - \beta)$ will be double that at $(\alpha + 2\beta)$ and $(2\beta - \alpha)$. For modulation levels over 50%, severe distortion will be present in the spurious modulation of the first two frequencies. To discriminate between the two forms of inter modulation caused by the b and c coefficients in the power series expression, they will be called first-order intermodulation (1st I.M.) and second-order intermodulation (2nd I.M.) respectively.

The first parts of the third term expressions show that the original frequencies α and β are being produced, but whereas the modulation of carrier α was represented by an amplitude variation of p, this spurious response, at the same frequency, has

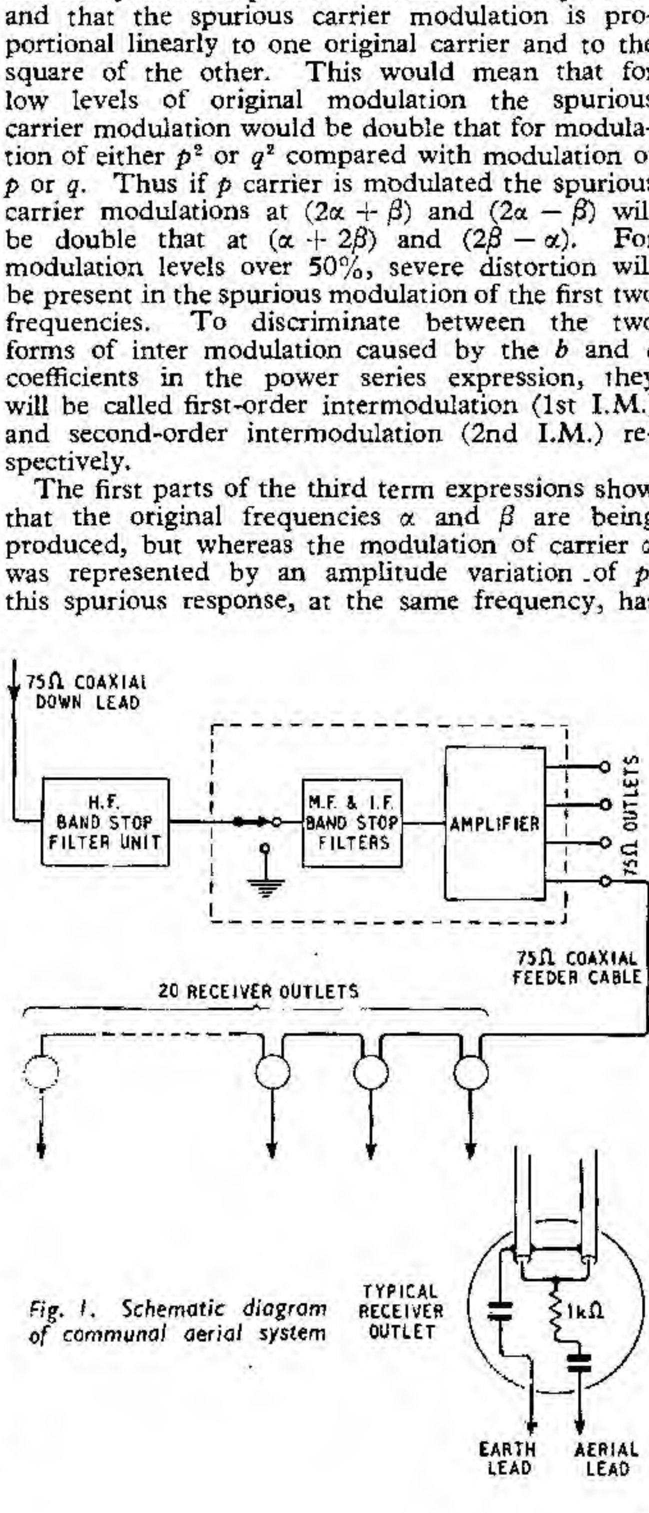


Fig. 1. Schematic diagram of communal aerial system

a modulation proportional to q^2. Thus the modulation of one carrier has become superimposed on the other. This effect is known as cross-modulation (X.M.). In this instance the unwanted effect will not be measured as the amplitude of an unwanted carrier, that would be coincident with a true carrier signal, but as the amount of background modulation existing on a true carrier having its own modulation.

In order to assess the linearity of an amplifier for this application, a limit on the unwanted responses enumerated above has to be set and a criterion based on the interpretation of test results has to be laid down. These matters will be referred to later.

Calculations have shown that in terms of the ultimate performance criteria laid down for I.M. products, the amplifier linearity is such that at the maximum voltage levels concerned, the 2nd harmonic production would be 0.0025% and for the 3rd harmonic 0.0004%. Although these figures are theoretical, they give some idea of the order of linearity required. Adoption of negative feedback methods would entail a considerable amount of feedback over the amplifier which, coupled with satisfying the Nyquist stability criterion over a very wide band, would be extremely difficult. The use of large valves working over a small portion of their characteristics would be most uneconomic if taken to the stage where linearity could be considered acceptable.

Current Amplification.—The amplifier circuit finally developed is actually one that amplifies the input current although to all intents and purposes it may look like a variation of a voltage amplifier.

The basic amplifier circuit is that of the grounded grid triode, as shown in Fig. 2. The input impedance at the cathode is $R_c' = \frac{r_a + R_L}{1+\mu}$. Now if $R_L \ll r_a$, R_c' becomes $\frac{r_a}{1+\mu}$ or approximately $1/g_m$ for a high-μ value. Now if the cathode is fed from a constant-current source ($Z = \infty$) any non-linearity of the $I_a - V_g$ characteristic resulting in a non-linear R_c', will have no effect and V_0 will be truly proportional to the input current. If now the impedance R_c' is presented to a valve, as in Fig. 3, and the impedance presented to the first triode is low compared with its r_a, then there is a good approximation to a constant-current feed. If V1 were a pentode, the approximation would be very good, but for reasons to be discussed later the pentode must be discounted. Further still, if the impedance in the cathode of V2 is high, the effective output impedance of V2 will become $r_a + (1+\mu)R_c$, where R_c is the impedance in the cathode. If, say, for V2 the anode lead to V1 is $r_a/10$, then the output impedance of V2 becomes $r_a + (1+\mu)\frac{10r_a}{(1+\mu)} = 11r_a$

The effect of R_c in V1 and the anode load in V2 is neglected for simplification. It should be noted that as the cathode and anode currents are one and the same thing in each valve, the amplification, as such, takes place in each transformer which serves to step up the current between valves. The amplification being substantially independent of the valve parameters, stability of h.t. and heater volts is unimportant; nor is the amount of h.t. ripple voltage, within reason, of any great importance. In the final circuit (Fig. 4) it was decided to use a push-pull arrangement because (a) greater linearity would be expected, (b) the circuit would be inherently more free from extraneous noise pick-up (i.e. mains borne interference), (c) it could provide means of adjusting the "balance" of the circuit and thus permit of some correction of dissimilar valves, and (d) the impedance presented to a previous stage would be held more constant.

As already mentioned, the pentode would appear to offer a very good approximation to a constant current source and might successfully be used as a grounded-grid amplifier in preference to the triode. Used in that manner the cathode current would be the sum of the screen and anode currents and any input current would be divided between screen and anode. In addition to reducing the current gain in the circuit (the screen circuit acting as a partial current bypass) the output current at the anode would no longer be truly proportional to the input current. Inspection of valve characteristic curves shows that the ratio of anode current to screen current is not constant for variations of grid-to-cathode potential and non-linearity would obviously be present.

Linearization Factor.—This term has been introduced as a way of expressing the capability of the circuit in reducing non-linearity of the valves. In Fig. 2, a generator of internal impedance Z is used to feed a current into the non-linear impedance R_c'. The greater Z compared with R_c', the smaller the effect of variation in R_c' will be on the relationship between I and E. The linearization factor is defined as $\frac{Z + R_c'}{R_c'}$ and if $Z = 9\,R_c'$ we can expect harmonic distortion products to be reduced by 10.

Circuit Considerations.—From the point of view of economy, the four 75-Ω outputs were arranged to be fed from one output stage. Only one output

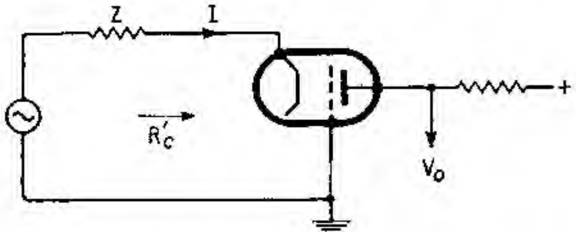


Fig. 2. Triode in the grounded-grid connection

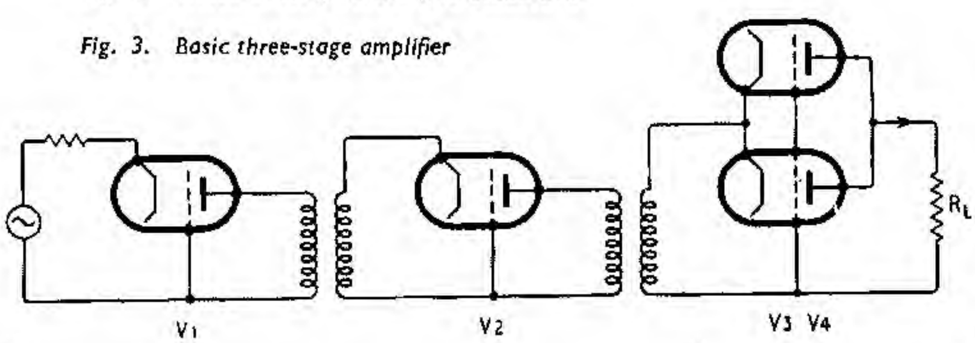


Fig. 3. Basic three-stage amplifier

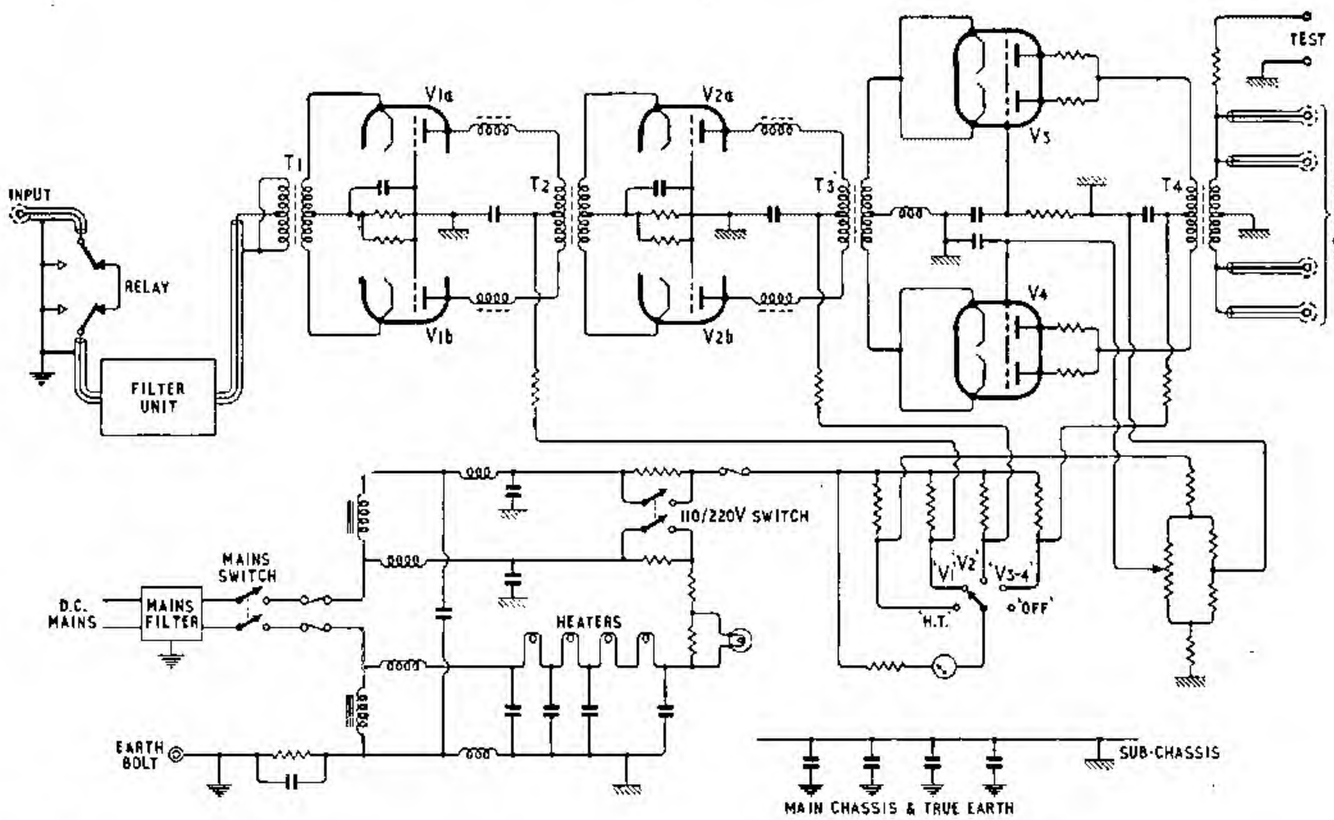


Fig. 4. Complete circuit diagram of final design

transformer was therefore required, but to obtain sufficient power gain in the last stage, a parallel push-pull arrangement of valves was adopted. Actually this meant a lower effective load impedance on the secondary of T3 and consequently a higher current amplification in the transformer could be obtained. This practice of paralleling valves to obtain greater current amplification by the preceding transformer cannot be carried out indefinitely owing to other limitations due to circuit capacities. The use of PCC84 valves enabled the whole amplifier to be constructed using four valves, which permitted the chassis to be packed in a very small space.

The success of the amplifier largely depended on the r.f. transformers which should have (a) good frequency response, (b) an effective high shunt loss impedance, and (c) freedom from non-linear hysteresis losses. An obvious choice of core material was Ferroxcube which proved to be very satisfactory. Shunt capacities across transformer primaries will serve as a current bypass at the anode of each valve with the result that at high frequencies the amplifier gain will be reduced. This effect sets an upper limit to the reflected primary impedances and, consequently, the turns ratio. From this basic consideration the gain per stage can be derived for the type of valve in question, and finally, the overall gain of the amplifier, which was calculated to be 15.2 dB.

It will be realized that shunt losses on the transformers will serve to bypass some of the current fed from the preceding valve with the result that the "constancy" of the current feed is impaired and the linearization factor will suffer. Alternatively, the impedance feeding the following valve is lowered. It is interesting to note that the linearization factor for the output stage, calculated for the valves alone, was 72.5. The transformers do not permit this figure to be fully realized in practice, but it indicates the measure of linearization that takes place.

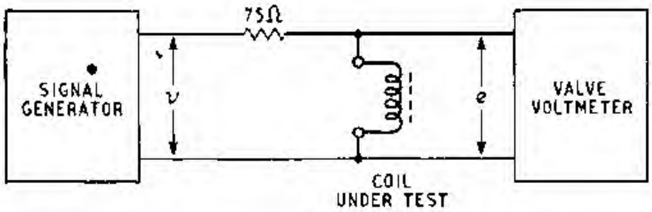


Fig. 5. Test circuit for effective shunt loss impedance

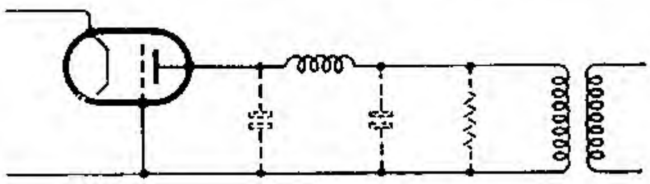


Fig. 6. Use of inductance to improve the gain at high frequencies

To assess the effective shunt loss impedance caused by core losses, a test circuit as shown in Fig. 5 was set up. Tests on various core and windings were carried out by measuring the change of voltage across the test terminals when the coil was connected in parallel. Shunt losses are of no particular consequence in T1 and T4 as they do not reduce the linearization factor by bypassing current fed into the cathode of a following amplifier. The shunt losses were found to be remarkably constant up to 30 Mc/s. In order to reduce the effects of winding capacities, which would be in parallel with the shunt losses, no secondary was coupled to the test winding.

When winding the transformers, care was taken to ensure a high coupling coefficient. Frequency response tests showed that up to 25 Mc/s, under conditions that were not as favourable as those in the valve circuit, the responses were within $\pm$ 1dB of the 1 Mc/s response.

As already mentioned, shunt capacities will cause loss of gain at high frequencies and to mitigate this compensating coils were introduced in the second stage. Fig. 6 shows the circuit capacities between which a coil of a few microhenries can give a useful "lift" up to 25 Mc/s.

The balance control in the grids of the output valves provides a differential bias voltage of $\pm$ 0.2 volt which was found adequate to deal with all normal dissimilarities between valves.

Amplifier Details

It was found on test that the signal-handling capacity of the amplifier was improved by introducing a high impedance into the common cathode connection. This results in the two cathode input impedances on each side of the circuit being in series instead of effectively in parallel. To explain the beneficial action of the choke, consider Fig. 7(a) in which a transformer is shown with series loss resistances across which must be developed a voltage proportional to the input current of the valve. Any non-linearity of R_c' will produce harmonic voltage components e_1 and e_2 across r_1 and r_2. From the point of view of even harmonics, if the valves were indentical e_1 would cancel e_2, but with dissimilar valves a difference voltage would exist. Now with the secondary centre tap earthed, a difference vcl.age $(e_1—e_2)$ acting in one cathode will effectively be applied between the cathodes (Fig. 7(b)), and will cause a current to circulate round the push-pull circuit. With a high impedance to earth from the centre tap, the resistances can be considered as being in series with the constant-current source on the primary side and will have no effect other than to cause half the difference voltage $(e_1—e_2)$ to appear across the choke. Although conclusive proof of this explanation, as to the advantageous effect of the choke, has yet to be established, the practice appears to be sound. The introduction of 10Ω into one cathode, for instance, was found to reduce considerably the signal handling capacity where 2nd harmonic distortion products were concerned.

The meter circuit enables the valve currents and the h.t. to be monitored, while various dropper resistances serve to reduce the voltages when working on 220 volt d.c. mains. The total consumption from the mains on 220 volts was 82 VA.

The amplifier was constructed as a small sub-chassis and mounted at the four corners by four v.h.f. type, bush-mounted, mica condensers. The amount of chassis work at d.c. mains potential was considerably reduced, while at the same time a satisfactory r.f. bypass to the main chassis at true earth potential was ensured. In addition to the r.f. chokes, a standard all-wave mains filter was incorporated to reduce the likelihood of mains-borne interference reaching the amplifier. The sub-chassis, measured overall, was 10½in. long, 6in wide and 4in deep. The greater part of the main chassis and space within the case was taken up with components associated with supply circuit and filtering arrangements.

The r.f. double band-stop filter circuit in Fig. 8 is conventional, each half consisting of a full constant-k section bounded by two half m-derived sections, having $m = 0.6$. Each band-stop filter has nine coils, a number which it was not possible to reduce, owing to the necessity of having to maintain a certain minimum attenuation between the resonant peaks. By judiciously laying out the filter to reduce magnetic coupling between the dust-iron cores to negligible proportions the filter was constructed within a box measuring 10in × 4in × 2in high.

(*To be concluded.*)

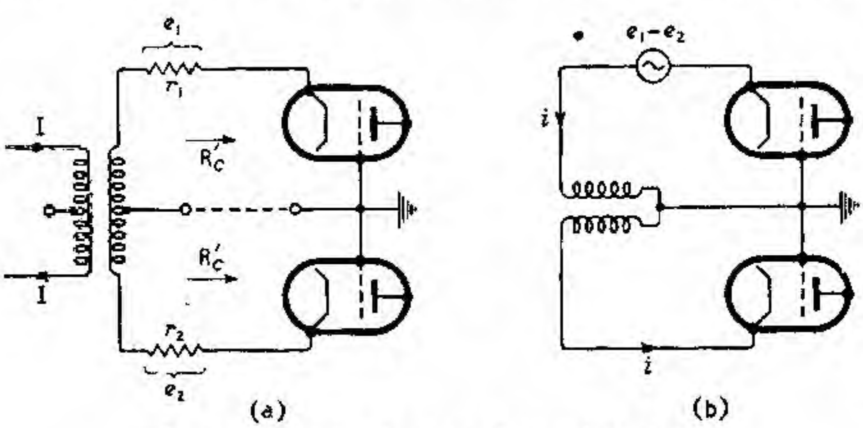


Fig. 7. Explaining the action of the cathode choke

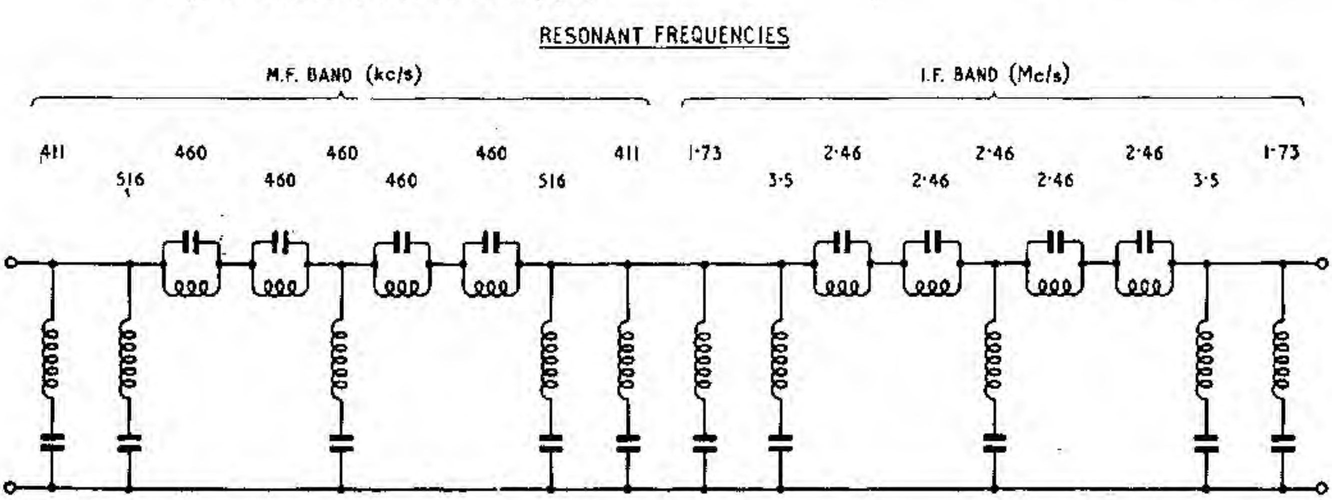


Fig. 8. Outline of double band-stop filter

Wide-Band Linear R.F. Amplifier

Results of Tests to Check

Performance of Amplifier and Filter

By B. F. DAVIES,* A.M.I.E.E.

(*Concluded from page 378 of the previous issue*)

SUMMARY OF AMPLIFIER CHARACTERISTICS		
Nominal Gain: 15 dB		
Frequency Coverage: 150 kc/s to to 25 Mc/s.		
Band-stop Filters: 400–535 kc/s: 1.6–3.8 Mc/s.		
Inlet: 75Ω coaxial.		
Outlets: 4 x 75Ω coaxial.		
Cross-modulation Limit:	0.35 V	Input
1st I.M. Criterion:	50 mV	Input
2nd I.M. Criterion:	25 mV	Input

Linearity of a high order is necessary in an amplifier covering all broadcast bands, if inter- and cross-modulation products are to be kept down. This article gives the results of measurements on an amplifier (described in the previous issue) which was designed as part of a ship's communal aerial system. The filter used to suppress currents induced by the ship's transmitter must also meet a stringent specification

MEASURED between 75-Ω impedances, the actual gain of the amplifier from input to any output was 14.6 dB, while the frequency response, which includes that of the filter, was as given in Fig. 9. Without the anode corrector coils in circuit, the 25-Mc/s response would have been some 6 dB down.

To make measurements utilizing two signals, the test circuit of Fig. 10 was set up. It was found necessary to have the isolating resistances in each signal-generator output in order to minimize reaction between the generators. Tests were commenced by first checking that, with the amplifier omitted and the generators delivering levels greater than that required for actual measurements, no unwanted inter- or cross-modulation effects could be detected on the receiver. The amplifier was then inserted, together with an attenuator to bring the voltage at the receiver back to the same level, and the measurement carried out.

The basis of the cross-modulation (X.M.) test was to find the level of a 30% modulated signal at the input that would produce modulation, 20 dB down (3% modulation level), on a small unmodulated carrier. A reference signal of 5μV at 1.4 Mc/s, 30% modulated, was applied to the input (the necessary allowance being made in all cases for the input test circuit loss), the receiver was tuned in and a reference output noted. The modulation was then switched off and the carrier increased to 50μV (+20 dB). The level of the other generator, 30% modulated, was then increased until the same reference output was read on the receiver output meter. The large interfering signal was, of course, kept off 1.4 Mc/s. From the results quoted, an input signal of 0.35 V could be handled. This means that if a filter is used which has an attenuation of 40 dB, the aerial input volts within the M.F. and I.F. bands could be as high as 35 volts.

At high frequencies the maximum input leve falls about 3 dB but at 20 to 25 Mc/s high voltages on the aerial are likely to be a rare occurrence and, from available statistics, do not exceed 5 V when the ship's transmitter is working in the 22-Mc/s band. Should an H.F. band filter be used in the installation, the attenuation requirements would therefore not be so stringent as for the filters for the M.F. and I.F. bands. It should be remembered, however, that in the short-wave broadcast bands H.F. transmissions are made by other ships which will induce voltages into the aerial which will pass unattenuated to the amplifier input. The H.F. band filtering unit could not be relied upon to rectify this, as in practice it would only be switched into the aerial circuit by the radio operator during his own H.F. transmissions. The main reason for keeping the H.F. band filter out of circuit as often as possible is to avoid attenuation of some broadcasting bands, which conflict with the communication bands and also to avoid introducing a general insertion loss into the aerial circuit, which would be the case if several conventional-type filters were permanently inserted. Nevertheless, induced voltages greater than 0.35 V, caused by such transmissions from neighbouring ships, are extremely unlikely and immunity from cross modulation on all carriers can largely be guaranteed.

Test Requirements

When we come to testing for both forms of intermodulation, we must first consider how the test results will be interpreted. The 1st I.M. has been shown to be proportional to the product of the two carrier strengths, pq. As the same result could be obtained with carriers of equal strength ($p = q$) or with one very strong carrier ($p \gg q$), it is evident that the total input voltage applied to the amplifier ($p + q$) will not be indicative of the signal-handling capabilities where intermodulation is concerned. Fig. 11 shows how for a constant pq (neither carrier being considered modulated) the total voltage, as

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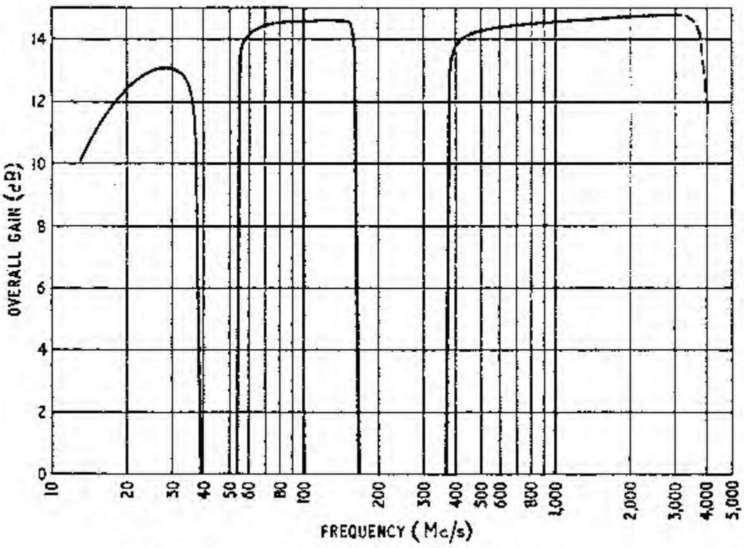

Fig. 9. Measured response of amplifier and filter.

would be measured on a peak reading r.f. voltmeter, varies with p. To establish a criterion, we can use $\sqrt{pq}$, which for two equal carriers will be equal to the level of each individual carrier. If, now, the tests are carried out by injecting the two signals simultaneously at the same level, we need only quote the signal level of either carrier entering the amplifier to ascertain the I.M. handling capabilities. With two equal carriers ($p = q$) for a given I.M. criterion voltage, the input level will be twice the criterion voltage. With only one carrier of high level (p), the total voltage level that can be accepted by the amplifier can, to a good approximation, exceed the criterion figure by the amount that the next highest carrier (q) is below the criterion figure.

The tests were based on ascertaining the I.M. criterion level at which the intermodulation product at ($\alpha + \beta$) or ($\alpha - \beta$) was equal to a carrier level of 5μV, a level which for entertainment purposes on a broadcast receiver would be considered an insignificant signal. It should be remembered that the noise field in the close vicinity of a ship, due to electrical machinery, is if possible kept down to 5μV/metre which demands signal strengths of 50μV/metre to obtain signal to noise ratios of 20 dB. On test a 5μV signal from one signal generator at 1.4 Mc/s, 30% modulated, was injected into the amplifier and the receiver set to a given reference output. The two generators, one of which was modulated 30%, were then set to 0.65 Mc/s and 0.75 Mc/s so that ($\alpha + \beta$) would correspond to 1.4 Mc/s and the levels increased simultaneously until the same reference output was obtained again. The push-pull balance control in this instance is used to obtain optimum results as it is capable of minimizing the 2nd harmonic distortion. Figures obtained for the I.M. criterion ranged from 50mV to 100mV according to the valves used. At 20 Mc/s and above, the results deteriorate due to the reduction in the linearization factor. For ($\alpha - \beta$) = 1.4 Mc/s where $\beta = 22$ Mc/s and $\alpha = 23.4$ Mc/s, the criterion figure was 25 mV. The question arises as to what happens when, say, α is at 22 Mc/s and β at 13 Mc/s, and the I.M. product is at 9 Mc/s. A series of tests have shown that, providing one of the carriers in question is at a sufficiently high frequency for the linearization factor to suffer, the I.M. criterion figure will be affected. Tables II and III show that 1st I.M. products, derived from frequencies in the common s.w. broadcast bands, can in some cases fall within the same s.w. bands. The greatest liability to I.M. from such causes

TABLE II

S.W. Broadcasting Bands					
49 m	5.95	—	6.20	Mc/s	a
41 m	7.15	—	7.30	,,	b
31 m	9.50	—	9.775	,,	c
25 m	11.70	—	11.975	,,	d
19 m	15.10	—	15.45	,,	e
16 m	17.70	—	17.90	,,	f
13 m	21.45	—	21.75	,,	g

TABLE III

(1)	(2)
a + d	f
a + e	g
c + d	g
b − a	M.W.
d − a	a
e − c	a
f − a	d
g − e	a
g − d	e

(1) Bands containing interfering signals.
(2) Band in which 1st I.M. product falls.

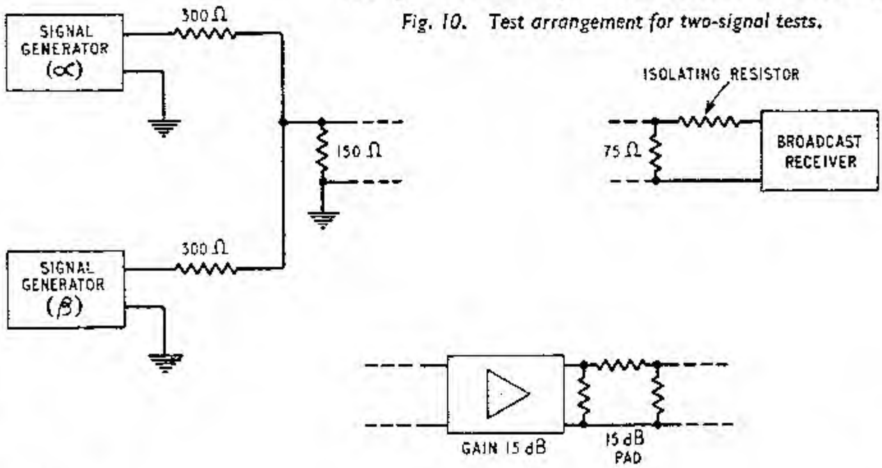

Fig. 10. Test arrangement for two-signal tests.

arises when there is a strong signal in the 13-metre band. Such interference from s.w. broadcasting bands is by no means the only interference due to 1st I.M. that could exist, there being many high-powered transmissions on communication frequencies, but such a consideration does serve to illustrate the possibilities of interference.

Second-order Intermodulation

When we come to the question of the 2nd order I.M., we must remember that the intermodulation term is proportional to p^2q or pq^2. Fig 12 shows how the total input level $(p + q)$ will vary as p is varied for a constant value of the product term. It will be seen that for the I.M. terms $(2\alpha + \beta)$ and $(2\alpha - \beta)$, where $k = p^2q$, the signal handling capacity $(p + q)$ rises more steeply below a certain minimum value: for $(2\beta + \alpha)$ and $(2\beta - \alpha)$ when $k = pq^2$, the rise below the minimum value is less. The converse case will apply for variations of q. Now in the 1st I.M. tests the criterion level corresponded to the minimum value of $(p + q)$, but in this case we can see from the graph that if we have $p = q$, it does not give a minimum for either sum of the 2nd I.M. terms. Actually, taking p^2q or $pq^2 = 1$, the two minima shown, for $p = 0.63$ or 1.26, give $(p + q) = 1.89$, while if $p = 1$, $(p + q) = 2$; a

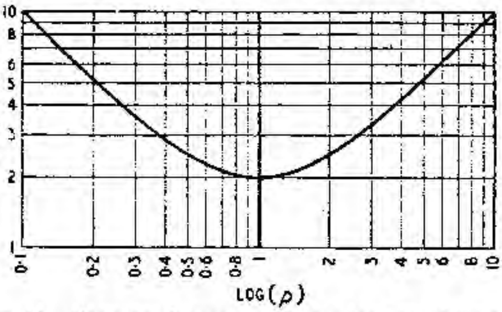

Fig. 11. Variation, for a given 1st I.M. product, of total peak input voltage (p + q) with p for the condition pq = 1.

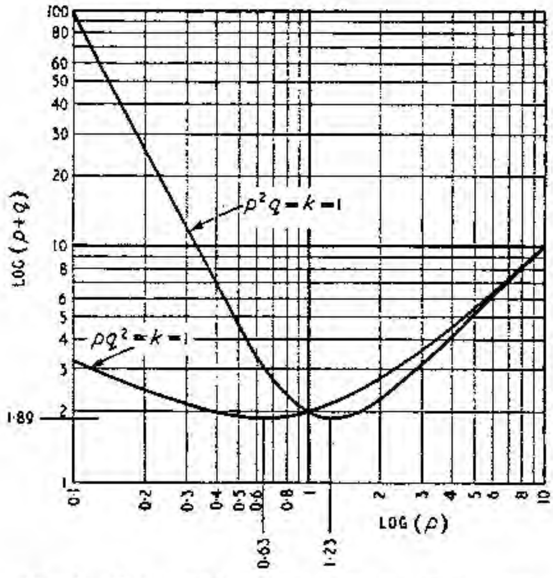

Fig. 12. Total input level for a given 2nd I.M. product.

difference of only 6%. We can, therefore, adopt the same method of expressing a 2nd I.M. criterion value for the input handling capacity as we used before with very little error, namely applying p and q of the same amplitude during test.

The other point that has to be considered is that if the 30% modulation is applied to the p carrier, for I.M. terms proportional to p^2q, we will have twice the modulation detected by the receiver on the I.M. carrier compared with having 30% modulation on the q carrier. By taking the worst possibility, we must ensure that the carrier representing p^2 or q^2 on test, is the one that is modulated.

The 2nd I.M. criterion level was found to be 50 mV, falling off only slightly above 20 Mc/s. Throughout the test, the reference frequency of 1.4 Mc/s was used for convenience, but there is no reason why other frequencies cannot be used, when similar results would again be obtained.

A point worthy of note is that 2nd I.M. products derived from $(2\alpha - \beta)$ can be important if α and β are close together, as would be the case of two signals in the same s.w. broadcast band. The $(2\alpha - \beta)$ product would then probably fall within the same band. This also applies to the M.W. band and in Table IV a few cases in the European area are shown where two signals could interfere with a third. On the practical side, the case of two 50 mV signals is hardly worth considering, since it is likely to arise only in some districts lying inland between transmitters.

Another interesting point concerning inter-modulation arises when one high-level input signal is present which is just at the cross-modulation level. From the final amplifier characteristics it is evident that the cross-modulation limit is 17 dB and 23 dB above the 1st and 2nd I.M. criterion limits respectively. Now, for the 1st I.M. case, if $\sqrt{pq} = 50$ mV and p is + 17 dB on 50 mV, then q is − 17 dB on 50 mV, or 7.1 mV, and with no other signals above this figure no 1st I.M. would be present. In practice, a few signals, especially when near port, may exceed this figure. For the 2nd I.M. case, we will have two figures, using the criterion level of 25 mV for the

TABLE IV
2nd I.M. PRODUCTS IN THE M.W. BROADCAST BAND

Interfering Stations		Station Experiencing Interference
(kc/s)	(kc/s)	(kc/s)
1376 Lille	1439 Luxembourg	1313 Stavanger
1151 B.B.C. Stagshaw	1088 B.B.C. Droitwich	1214 B.B.C. Brookman's Park
1007 Hilversum	926 Brussels	1088 B.B.C. Droitwich
908 B.B.C. Brookman's Park	1070 Paris	746 Hilversum
908 B.B.C. Brookman's Park	692 B.B.C. Moorside Edge	1124 Brussels
863 Paris	620 Brussels	1124 Brussels
746 Hilversum	620 Brussels	872 A.F.N. W. Germany

level of the smaller signal. When the input signal of 0.35 V corresponds to p^2, the small signal q must be $-$ 34dB on 25 mV, or 0.5 mV, but when the small signal corresponds to q^2 it need only be $-$ 8.5 dB on 25 mV or 9.4mV, to prevent 2nd I.M. occurring under these conditions.

Conclusion

The main requirement of an aerial amplifier and its associated distribution network is that it should operate with the minimum of attention and to all intents and purposes should be as reliable as the proverbial "piece of wire." Freedom from cross-modulation not only minimizes interference but ensures that secrecy is maintained in the transmission of messages from the ship. No aerial distribution system can be considered satisfactory if passengers' radio telephone calls are to be heard on every station on the cabin receivers!

The wide-band amplifier that has been developed has necessarily been a compromise, but experience has already indicated that it is meeting all requirements.