10

Power Amplification Using Wideband Techniques

10.1 THE CASE FOR WIDEBAND CIRCUITS

The loss of traffic time caused by the changes in frequency necessary to maintain communication by means of h.f. systems, is an ever-present incentive towards more rapid frequency changing. Limits to the minimum possible time are imposed by the number of circuits which have to be tuned, and by the physical size of the tuning components. The frequency-changing time with medium- or low-power transmitters can be less than that required for high-power transmitters, because of the number of stages involved and the smaller components used.

Even with medium-power transmitters there are applications, such as communicating with high-speed aircraft, where the minimum achievable time of 10 s or so is not fast enough, and there may be negligible warning of a need to change. To cover these applications with tuned equipments there are two solutions: (*a*) to incorporate a number of pre-tuned circuits and to change frequency by switching between pre-set components, and (*b*) to radiate the same intelligence on two frequencies by using two transmitters for a period covering the time necessary for a change to be made. Operators on transmitting stations call this system 'dualling'.

The disadvantages of both solutions stem from the use of tuned circuits, indicating that a satisfactory third solution would be the elimination of tuned circuits. This solution could be provided by wideband power amplifiers with linear characteristics, covering the whole h.f. spectrum without any need for tuning. Later in this chapter, it will be shown that this solution is quite practical with wideband amplifiers for output powers of about 1 kW, which is adequate for many applications, particularly if s.s.b. systems are used.

With wideband amplifiers, the time taken to change frequency is that required to change the frequency of the driving source only, and the choice of frequency is not limited by the number of pre-tuned circuits, as given by solution (*a*). 'Dualling' applications can be covered by radiating two carrier frequencies simultaneously from the same transmitter, and although this incurs a reduction in power on each, it is less than the 6 dB which would be expected.

Reliability is improved by the elimination of moving parts associated with tuning, for experience proves that a high percentage of faults on transmitting stations are of a mechanical nature. Where wideband amplification is in the form of distributed amplifiers, outage time is further reduced and the service can be continued at a lower performance, even with the loss of one or two valves.

Another advantage of wideband power amplifiers is their use as drives. They enable high-power transmitters to be constructed with only one tuned stage, thereby simplifying and reducing the time required for frequency changing; a particular advantage for self-tuned systems. When used as drives, there is also an advantage in the low and substantially constant impedance presented by wideband amplifiers to signals reflected from a driven stage. The reflected signals are absorbed and not re-reflected, so the drive performance is not degraded by distortion produced at the driven-stage input, such as by grid current. This problem is often very difficult to solve when dealing with tuned linear amplifiers in cascade.

10.2 GENERAL PRINCIPLES OF WIDEBAND AMPLIFIERS

Limiting factors

Spurious capacitances are associated with all amplifying devices, and with valves in particular. Such capacitances limit the power output, efficiency and gain of a wideband amplifier. The conditions existing at the three following points in an amplifier will be considered. (1) The output, as represented by Fig. 10.1(*a*). (2) The intervalve couplings, as represented by Fig. 10.1(*b*) and (*c*). (3) The input, as represented by Fig. 10.1(*d*).

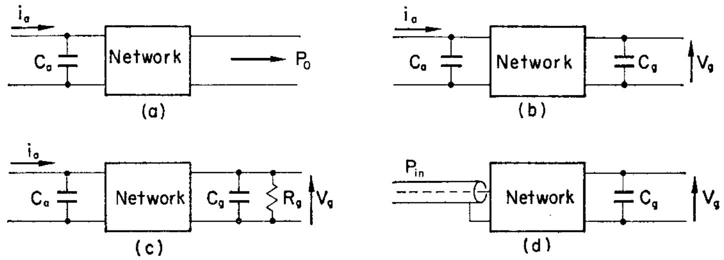

FIG. 10.1 Working conditions for a wideband amplifier. (*a*) At the output. (*b*), (*c*) At the interstage coupling. (*d*) At the input.

The networks of Fig. 10.1(*b*) and (*d*) must contain a resistive element.

In the first case, consider a current generator shunted by a capacitance C_a which cannot be removed. Here the aim is to produce the maximum output power for a given driving current. In any narrow-band amplifier, the load line is chosen to suit a particular valve. However, as the bandwidth is increased, the load resistance must be reduced, and the shunt capacitance sets a limit to the r.f. power that can be obtained from the generator.

The maximum power output can be calculated from the formula [1]

$$P_o \max = \frac{\pi}{2} \cdot i_a^2 \cdot XC_a \tag{1}$$

where P_o max = maximum power output (constant over the pass band);
i_a = r.m.s. value of fundamental component of anode current;
XC_a = reactance of anode capacitance at the edge of the pass band.

This expression assumes an infinite number of components in the network. When the number is restricted, the response is no longer flat and the power or the bandwidth is reduced. Owing to the low value of the load resistance, a suitable valve must be able to produce a large current at low anode voltage. Low anode capacitance is of course essential.

The problem at the interstage coupling is shown in Fig. 10.1(*b*). Here both the generator and the following valve are shunted by capacitances C_a and C_g. The aim is to produce a maximum driving voltage, V_g, from a given current, i_a, of the generator. Then

$$\left(\frac{V_g}{i_a}\right)\max = 2\sqrt{(XC_a \cdot XC_g)} \qquad (2)$$

where XC_g is the reactance at the edge of the pass band of capacitance C_g.

Equation (2) is a limiting equation, and assumes an infinite number of components in the network and a wideband transformer to allow for a different impedance level at anode and grid. A loading resistance must be present somewhere in the network. If both valves are of the same type, by multiplying equation (2) by the mutual conductance a well-known gain-bandwidth expression is obtained [2]

$$\text{maximum gain-bandwidth} = \frac{g_m}{\sqrt{(C_a \cdot C_g)}} \qquad (3)$$

Here again, with a restricted number of components the response will no longer be flat, and either the bandwidth or the gain will be reduced.

There are cases where the following valve absorbs power; i.e., where it presents a resistive termination, R_g as in Fig. 10.1(*c*). In this case the maximum possible power must be delivered to R_g, and R_g may then set the limit to the possible gain instead of C_g.

At the input to the amplifying stage [see Fig. 10.1(*d*)] the aim is to produce maximum driving voltage from the available power. Since a transmission line (a coaxial cable) will be used, the input network must be matched. Here again, the shunting capacitance and the maximum v.s.w.r. allowed will determine the limit of gain.

The above limiting equations cannot be exceeded in any simple amplifier, as the fundamental law of the charging rate of a capacitor is involved. It is easily seen that paralleling the valves, whether in phase or in push–pull, will not increase the power output per valve, nor will it increase the gain-bandwidth product. In all these cases the available charging current per capacitor is the same. It may be noted that class AB or class B operation will reduce the effective mutual conductance and hence will reduce the gain-bandwidth product.

Synthesis of wideband networks is fully covered in technical literature, and the normalized networks themselves are given in tabulated form in Refs [3] and [4].

Distributed amplifiers

In a distributed amplifier [5, 6, 7, 8] (see Fig. 10.2), the valve capacitances are not charged simultaneously. By placing valves along an artificial transmission line, the same instantaneous power will charge the valve capacitances in succession. To produce a given driving voltage across any of the valve-input capacitances, the same amount of power is required. In a simple amplifier, once this power

has served the purpose of producing a driving voltage it is dissipated in a termination. In a distributed amplifier, however, it is used to drive the following valve. This process can be repeated over and over again so that an unlimited number of valves (subject to losses) can be driven by the same power that is required to drive one valve.

The anodes feed a transmission line of the same delay characteristic as the grid line. Half the anode current of each valve will travel to the right and will add in phase in an output load. The other half will travel to the left, towards the terminating resistor. At low frequencies the phase delay contributed by the line is negligible, and these currents will add in phase so that half the power developed by the amplifier is dissipated in the terminating resistor. As the frequency is increased, however, standing waves will develop on the anode line, and although

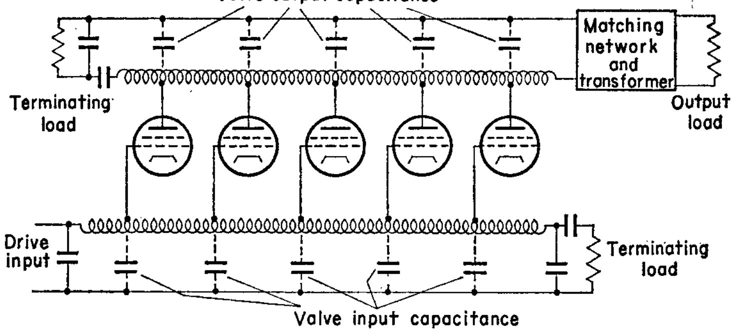

FIG. 10.2 Simplified diagram of a distributed amplifier.

the power dissipated in the resistor is reduced, the power output remains the same because extra power will be dissipated in some valves.

It is possible to direct the full anode current of each valve to the right, i.e., to the useful load, if the characteristic impedance of the anode delay line is tapered, i.e., lowered progressively at each valve connection [8, 9]. A simple calculation shows that under these conditions the r.f. voltage swing at each anode will be equal over the whole pass band—a very desirable property. However, this voltage swing will be determined by the delay-line impedance at the first valve of the chain, which is, in turn, determined by the anode capacitance of the valve. In fact, no more power per valve can be obtained than in a simple wideband amplifier.

The problem of power output per valve, and efficiency, is most important in transmitter applications. In an amplifier with a uniform anode line, the number of valves and the h.t. voltage are adjusted to permit optimum operation of individual valves, but half of the available power is lost. In the tapered-line amplifier, valves work inefficiently, but the addition of power is complete. In an amplifier with a uniform anode line for a bandwidth of about 30 MHz, only a few modern low-capacitance valves are required for optimum operation. By this is meant a condition under which the valves operate with a sufficiently high-impedance load line for the anode-voltage swing developed to be comparable with the h.t voltage applied. This optimum has, of course, no relation to the optimum gain per stage of a low-power distributed amplifier. Here, the first consideration is how to extract maximum power over a wide band from a given set of valves; the gain, provided that it is not unreasonably low, is of secondary importance.

Having reached the stage of sufficiently high anode-voltage swing, the anode line can now be tapered so as to keep a constant anode-voltage swing for all successive valves. To obtain a reasonable efficiency, most of the valves must be along the tapered line. The non-tapered section improves the load line for the tapered section, and must be sufficiently long. A compromise is necessary, and depends on the characteristics of a particular valve and bandwidth [10]. The power output is given by

$$P_o = mX'C_a\left(n-\frac{n_1}{2}\right)n_1 \cdot i_a^2 \tag{4}$$

as the frequency tends to zero,

where $X'C_a$ = reactance of the anode capacitance at the cut-off frequency;
i_a = r.m.s., value of fundamental component of anode current;
n = total number of valves;
n_1 = number of valves with uniform anode line, which need not be an integer.

The power output is almost constant up to 0·9 of the cut-off frequency if an m-derived anode line is used with $m = 1{\cdot}4$.

In designing an output stage of a wideband transmitter covering 2–30 MHz, one has a choice of a simple wideband stage or a distributed amplifier. An examination of available valves with suitable power ratings shows that both designs are possible, with some preference for a distributed amplifier. There is, however, a decisive factor in favour of a distributed amplifier if the transmitter is to be used with a varying load condition such as an antenna. If no loading or tuning adjustment is allowed, the transmitter must be able to operate with any impedance presented by an antenna feeder corresponding to a v.s.w.r. of 2 to 1, without an appreciable variation in power output and performance. This is only possible with a matched generator. A simple amplifier is not a matched generator, but a distributed amplifier behaves as one, provided that if the anode line is tapered it is sufficiently long. This in fact, together with the primary inductance of wideband transformers, sets the low-frequency limit of the pass band.

The amplifier will, for the lower frequencies of operation, amplify all harmonics which are generated at any point in the transmitter up to the fourteenth with no significant attenuation. Exceptional linearity is therefore required, as an harmonic content of –40 dB (barely enough in practice to meet the existing Atlantic City regulations) is still too large for any h.f. transmitter, although it corresponds to a distortion of only 1%, which is considered good even for an a.f. amplifier. A considerable reduction in harmonic distortion can be obtained by the suitable choice of an anode load line. This is often done in an a.f. amplifier, but in a distributed amplifier the slope of the load line is dictated by the r.f. bandwidth and by the valve capacitance, and is far too low from the point of view of distortion.

This initial difficulty is followed by the realization that it is impossible to extract a reasonable proportion of input power from a single-ended stage if this low second-harmonic content is required. Allowing for about 20 dB balance, a push–pull amplifier demands the second-harmonic component of an individual valve to be less than –20 dB with respect to the fundamental, as long as this second-harmonic frequency falls in the pass band of the amplifier. It is not

permissible to drive the valves below cut-off voltage, although the full current swing down to zero makes the operation of the amplifier, as far as changes in feed current, etc., are concerned, more like class AB than class A. There will, of course, be no cancellation of the third-harmonic components, so the linearity of the individual valve must be sufficiently good. A valve which will be found suitable as far as r.f. bandwidth and power are concerned has to be examined for linearity at different amplitude levels before it can be accepted. To satisfy the conditions for linearity as far as second and third harmonic is concerned will almost automatically satisfy the conditions required for the higher-order harmonics.

High v.s.w.r. on the output feeder introduces another limitation. Maximum voltage swing, which may be present on any valve of a distributed amplifier which is behaving as a matched generator, will be approximately equal to the product of $\sqrt{(\text{v.s.w.r.})}$ and the voltage swing under the matched condition. This must be allowed for in the design, and severely reduces the efficiency of the amplifier.

Push–pull operation creates still another problem—that of unbalanced power. This power will travel along the anode delay line, but it will be rejected by an output transformer. Again, in an a.f. amplifier or in a narrow-band r.f. amplifier, the anode circuits are so designed as to present a very low impedance at the anodes in the unbalanced mode of operation. This cannot be done simply here, because the length of the artificial delay lines will cause an effective parallel resonance of a high Q factor at some frequencies on some valves. There is no need to explain why such a resonance—even if it is not severe enough to cause a serious dip in the frequency response—is disastrous as far as harmonics are concerned. The only way to avoid this trouble is to use resistive loading of the unbalanced mode of operation. Fortunately, if this is done all along the anode lines only a negligible amount of useful power will be lost. This loading is also necessary to absorb second-harmonic current, which, if not absorbed, will have effects similar to those of the unbalanced power, in addition to the direct production of a high second-harmonic output.

Unless the grid-line impedance is very low indeed, the grid current will introduce high-order harmonics. It is not advisable to work with even a very limited grid-current, because even though the valves individually may be satisfactory, the harmonic powers so produced will travel along the grid line and the combined effect may be serious.

Because of the very large phase delay of the distributed amplifier, it is not possible to use wideband overall negative feedback. Individual feedback on each valve is very successful, especially cathode feedback. Here, the loss of gain is less than the amount of feedback applied, because the application of cathode feedback reduces the effective grid-input capacitance and the grid-line impedance can be increased.

The circuit arrangement for r.f. feedback on each individual valve is shown in Fig. 10.3. The ferrite-cored r.f. choke has a high r.f. reactance over the whole band, relative to the value of the feedback resistor, but a low d.c. resistance. The value of the resistor determines the level of feedback, but the low d.c. resistance of the choke ensures that the valve bias is not affected by its presence in the cathode circuit. Note that the design is simplified by the use of indirectly heated valves, because the heater current does not pass through the cathode choke.

The stability of the amplifier must also be considered. The fairly high gain of a power-distributed amplifier, combined with rather poor isolation between anode and grid of larger valves, makes the amplifier unstable unless the cut-off frequency of the anode and grid-delay lines are different. Grid lines with twice the cut-off frequency of the anode line (two sections for each section of the anode line) were found to give very stable amplifiers.

In the preceding discussion, the availability of a wideband transformer is assumed. Given ferrite-core material, very good transformers can be made if the leakage inductance and capacitances are arranged to form a part of a wideband low-pass network (see Section 10.7).

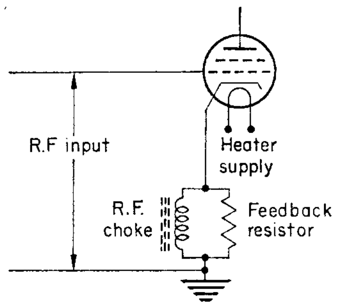

FIG. 10.3 Circuit for r.f. cathode feedback.

10.3 A 1 kW WIDEBAND TRANSMITTER, 2–28 MHz

The final amplifier

From the limitations described, it is apparent that the design of a high-power wideband amplifier depends on a careful compromise between contradictory requirements, the best amplifier being achieved with the best compromise. For this design, after a theoretical analysis a number of experimental tests were necessary to obtain the required performance, because of the absence of the correcting facilities provided by variable tuning and loading in conventional designs. Nevertheless, this work was justified by the performance being repeated on subsequent amplifiers made to the same design.

The final amplifier contains eighteen valves, type 4CX250B in two nine-valve distributed amplifiers in push–pull. The anode lines are uniform for two and a half valves per side, and then tapered to give a voltage swing of constant amplitude on the remaining valves. The balanced output of the two push–pull sections is converted to 50 Ω unbalanced by means of a wideband transformer on a ferrite core. The transformer is forced-air cooled and is rated at 4 kW c.w. to allow working into a feeder with a high v.s.w.r. The h.t. voltage of 1·1 kV allows for the anode swing produced by a v.s.w.r. of 2 to 1 on the feeder, but higher ratios may be accommodated in an emergency, because the valves are individually protected, even if the instantaneous anode voltage on some valves is driven below the screen voltage.

Two sections of the anode line are used for each valve to reduce the lowest frequency at which the amplifier will still behave as a matched generator. The 180° phase shift between successive valves, through the amplifier in the centre of the band, is avoided by putting only one section of the anode line in the middle

of each row. Two sections per valve through the remainder of the amplifier allow better suppression of the unbalanced mode.

The grid lines are also suppressed for any unbalanced mode and they are very accurately matched at the output end. In order to ensure the same drive voltage on all the valves, it is necessary to compensate for losses along the grid lines. This is arranged by driving most of the valves from tapping points on capacitance potentiometers of different ratios across the grid lines.

Design and construction of anode and grid lines

The most important feature of distributed amplifiers is the design of the anode and grid delay lines, together with the method of construction to ensure that they are made accurately in accordance with the theoretical design. In the case of push–pull distributed amplifiers, it is just as important to maintain the same overall phase delay on each side as it is to maintain the same delay along the grid and anode lines individually.

The criterion of line design is one which allows the maximum value of shunt capacitance for a given cut-off frequency, and has the most uniform voltage-frequency response across the capacitors along the line. Based on network theory, it can be shown that an m-derived filter with an m value of 1·4 has a fairly uniform voltage response and an almost linear phase delay. In addition, for a given capacitance the line impedance or cut-off frequency is increased by m times, i.e., 1·4 times.

If the line inductors are sections of a continuously wound coil, the mutual coupling between the sections is negative, so the mutual inductance is negative and this allows an m-derived filter to have an m value greater than unity. By suitably arranging the diameter of the continuously wound coil in relation to the turns per inch, the length/diameter ratio of the sections can be designed for an m value of 1·4. For the constant impedance of the uniform portion of the line, the shunt capacitors are equal and the series inductor sections are also equal.

For the tapered portion of each anode line, the same principle and method of construction are used, but the impedance reduction is made in a number of steps, with an increase in shunt capacitance and a decrease in series inductance. In considering the inductance sections, it is important that the leads connecting each valve to the inter-section tapping points should be as short as possible. All the sections must comprise a number of whole turns, so that all the tapping points can be in line longitudinally. This means that a careful choice of coil diameter, wire diameter and turns per inch must be made for each different impedance section, in order to maintain the correct inductance and m value with a whole number of turns. This is shown clearly in the photograph of the anode circuit of the 1 kW distributed amplifier shown in Fig. 10.4. This also shows the method of construction with the delay lines alongside the valves, which enables both short connections to be made and provides a ready means for valve replacement.

In order to achieve side-to-side symmetry, the direction of winding is different for each side, one being left-hand and the other right-hand. For the same reason, the grid lines are also wound in a different direction for each side, with the left-hand grid line driving the right-hand anode line.

While the shunt capacitors can be trimmed, there is no means of adjusting the inductor sections, so they must be correct within fine limits. The winding pitch can be accurately controlled, but the inductance is also critically dependent on the depth at which the winding sits in the threaded former. It has been found in practice that sufficient accuracy is given by measuring the overall diameter of the wound former to within specified limits.

It is apparent from Fig. 10.4 that the kilovolt-amp rating of the inductors and capacitors is not so high as usually found with 1 kW amplifiers. There are two reasons for this. First, these are line amplifiers having a Q factor of unity when matched. The only increase in component kilovolt-amp is due to the v.s.w.r. on the anode lines, produced by a mismatched termination. Second, the anode voltage and consequent anode swing are about half the amplitude of those in

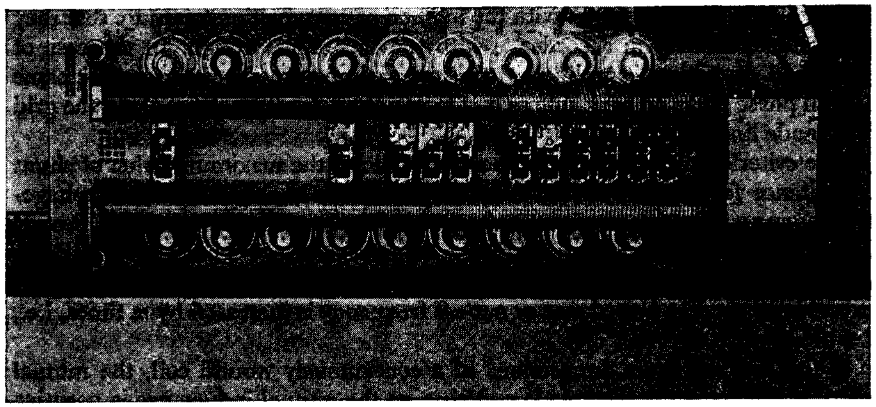

FIG. 10.4 Anode delay lines showing the tapering sections.

tuned amplifiers of the same order of power output. Assuming a maximum Q factor of 10 for a tuned amplifier, the component kilovolt-amp rating in the distributed amplifier need be only one-twentieth of the tuned case for the same circuit losses.

The penultimate amplifier

The penultimate stage in this case is also a distributed amplifier, containing two banks of miniature high-slope pentodes in a push–pull arrangement. Basically the principle is the same as for the final amplifier, but there are some differences which are worth noting.

As efficiency is not of primary importance at this power level, only the section of anode line associated with the last two valves on each side is tapered, the remainder of the line is uniform. This small amount of tapering is necessary to avoid excessive voltage swing on the anodes of these valves.

The load presented by the final-stage input is effectively constant over the frequency band, so the only mismatch seen by the penultimate amplifier is that produced by the interstage coupling transformer network. The v.s.w.r. resulting from this network is considerably less than 2 to 1, which means that the operating anode-swing can be nearer the maximum than it is on the final stage. In this respect the penultimate stage is the more efficient of the two amplifiers.

The construction of the most suitable pentodes available is such that the anode and control-grid connections are brought out of the base, so the grid and anode

lines are on the same side of the assembly deck. This necessitates very careful screening, and an important stabilizing feature is that the grid lines are wound in opposite directions to one another, and in the opposite direction to the anode lines on the same side of the circuit.

10.4 PERFORMANCE OF 1 kW WIDEBAND TRANSMITTER

The frequency range of this equipment is 2–30 MHz, and over the whole of this frequency range it will deliver 1 kW c.w., or p.e.p., although above 28 MHz this is subject to the v.s.w.r. at the output being rather less than 2 to 1. When used to drive a high-power tuned amplifier, the interstage matching network ensures a low v.s.w.r., and a drive power of 1·25 kW is available with low distortion.

In common with wideband networks in general, the frequency response of both penultimate and final amplifiers contains ripples within the passband, making it necessary for each amplifier to have enough gain to cater for the troughs. These troughs may be coincident at one or more frequencies, so the only significant response is that of the overall gain of the two wideband amplifiers. The response

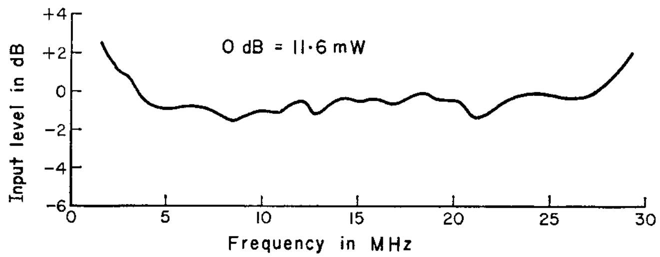

FIG. 10.5 Frequency response of two-stage distributed amplifier, in terms of drive level required for 1 kW output.

of all amplifiers made to any one design is not necessarily the same, but the variation is always within ±2·5 dB. In the typical example shown in Fig. 10.5 the drive required to give 1 kW output is +2·5 dB, −1·5 dB, about a mean level of 11·6 mW. This corresponds to an overall gain of between 48 dB and 52 dB, from 2 MHz to 28 MHz. The nominal gain of each amplifier is about 28 dB, which gives adequate coverage for contingencies.

For the full output to be obtained on all frequencies, the variation in gain over the frequency band makes it necessary for the drive level to be adjusted in accordance with the operational frequency. This does add a slight complication, but the adjustment can be made automatically by means of an automatic level control (a.l.c.) at the input, operated by a level indication picked up from the transmitter output.

It will be appreciated that a high order of linearity is required, because harmonics of the lower frequencies will be amplified and passed directly to the output in exactly the same way as the fundamental. It is for this reason that r.f. feedback, applied in the manner described in Section 10.2, is necessary. The result is that the third-order intermodulation products (as shown in Fig. 10.6) are quite low at output levels of 500 W and 1 kW p.e.p. Fifth-order products are rather lower than third-order, mainly because the peak voltage swing on the anodes does not approach saturation level, so peak flattening does not occur at the envelope peaks. Higher-order intermodulation products are sufficiently low to be ignored.

When operating on c.w. or f.s.k. at 1 kW output, at some frequencies the second and third harmonics are at a level higher than the 50 mW (– 43 dB) limit imposed by international regulations. To cover this type of operation, a small number of low-pass filters can be fitted in the output feeder. The appropriate filter is automatically switched in circuit by a patching system associated with the frequency-determining source. Alternatively, the switching can be associated with a particular antenna, for it is unlikely that a single antenna will be used for all frequencies between 2 and 28 MHz. These filters are not necessary for c.w. or f.s.k. working when the wideband transmitter is used to drive a high-power tuned amplifier, due to the harmonic attenuation provided by the final stage.

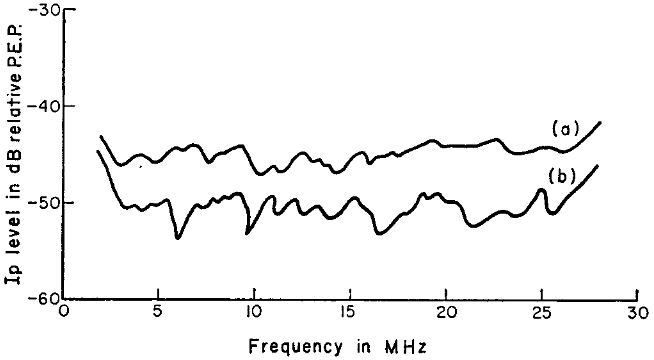

FIG. 10.6 Intermodulation products between two closely spaced tones with reference to p.e.p. (*a*) 1 kW p.e.p. (*b*) 500 W p.e.p.

The main disadvantage of the distributed amplifier system is the low conversion efficiency in terms of mains input to r.f. output, which is about 20% overall, i.e., at 1 kW output an input power of 5 kW is required. Low efficiency would be the case with any wideband amplifier covering nearly four octaves of the r.f. spectrum, and is not entirely due to the distributed amplifier as such.

10.5 MULTI-FREQUENCY OPERATION

In addition to being able to change frequency in the time taken to 'flick a switch', wideband transmitters enable several carrier frequencies to be radiated simultaneously. Further, the power per frequency need not be reduced in direct proportion to the number of frequencies in use, provided that the frequencies are not harmonically related. This is due to the infrequent occurrence of waveform peaks coinciding as the waves travel along the lines in the amplifier. The greater the number of frequencies, the less likelihood there will be of all the peaks coinciding. It is analogous to multi-channel speech on a single carrier.

It has been found in practice that the limiting factor is the *total mean power*, which should not exceed 500 W. Provided that this level is not exceeded, the linearity performance on each frequency is actually better than with a single frequency and inter-frequency cross-talk is at such a low level that it can be completely ignored. The real significance of multi-frequency operation is shown

TABLE 10.1

No. of frequencies radiated	1	2	3	4	5	6	7	8	9	10
f.s.k.										
Total mean power, W	1000	500	500	500	500	500	500	500	500	500
Mean and peak power at each frequency, W	1000	250	167	125	100	83	71	62·5	55·5	50
Total effective p.e.p., kW	1	1	1·5	2	2·5	3	3·5	4	4·5	5
s.s.b. two-tone										
Total mean power, W	500	500	500	500	500	500	500	500	500	500
Mean power at each frequency, W	500	250	167	125	100	83	71	62·5	55·5	50
Peak power at each frequency, W	1000	500	333	250	200	167	143	125	111	100
Total effective p.e.p., kW	1	2	3	4	5	6	7	8	9	10
a.m. 100% modulation										
Total mean power, W	375	500	500	500	500	500	500	500	500	500
Mean power at each frequency, W	375	250	167	125	100	83	71	62·5	55·5	50
Carrier power at each frequency, W	250	167	111	83	67	55·5	47·4	41·7	37	33·3
Peak power at each frequency, W	1000	667	444	333	267	222	190	167	148	133
Total effective p.e.p., kW	1	2·67	4	5·33	6·7	8	9·3	10·7	12	13·3
CONVENTIONAL AMPLIFIER										
f.d.m.										
Total mean power, W	1000	500	333	250	250	250	250	250	250	250
Mean and peak power at each frequency, W	1000	250	111	62·5	50	41·7	35·7	31·25	26·7	25
Total effective p.e.p., kW	1	1	1	1	1·25	1·5	1·75	2·0	2·25	2·5
s.s.b. two-tone										
Total mean power, W	500	250	166·6	125	125	125	125	125	125	125
Mean power at each frequency, W	500	125	55·5	31·25	25	20·8	17·9	15·6	13·4	12·5
Peak power at each frequency, W	1000	250	111	62·5	50	41·7	35·7	31·25	26·7	25
Total effective p.e.p.	1	1	1	1	1·25	1·5	1·75	2·0	2·25	2·5

in Table 10.1, where the total 'effective' power is given for three types of traffic. It can be seen that the total effective power can be many times the single frequency maximum of 1 kW p.e.p.

For multi-frequency operation, the mean and peak power capabilities of a conventional amplifier are governed by the coincidence of signals on a time basis, whereas the multiplicity of valves in the tapered-line amplifier, which are electrically separated on a frequency basis, avoids coincidence of signals in any one valve under multi-frequency conditions. The mean and peak power levels of a conventional amplifier operating with multi-channel f.d.m. telegraphy and two-tone s.s.b. are also shown in Table 10.1, for comparison.

The facility of being able to radiate simultaneously on more than one frequency is particularly advantageous for ground-to-air communications with high-speed aircraft. It is the same as 'dualling', but with a single transmitter. This enables the operator in the aircraft to maintain communications simply by changing the frequency of his own equipment at the most appropriate time.

There are many other uses which are possible by exploitation of this multi-frequency operation capability. One that springs to mind is as a means of anti-jamming, or secrecy. By a pre-arranged and readily changeable coding, the transmission can be radiated by a number of sequential frequencies, changed in a fairly rapid and apparently random manner.

10.6 FREQUENCY EXTENSION TO COVER THE M.F. BAND

The distributed amplifiers themselves are capable of operating quite satisfactorily at frequencies well below 2 MHz. This also applies to the small wideband transformers used at the input to the penultimate stage and for interstage coupling. The limit is imposed by the output transformer network, the physical size of which presents design problems if required to give a satisfactory performance over a frequency range much in excess of 15 to 1.

To cover the whole frequency spectrum from 300 kHz to 30 MHz it is only necessary to provide two output transformer networks and a double-pole changeover switch, as shown in Fig. 10.7. As the h.f. transformer T_1 is satisfactory down to 2 MHz and m.f. transformer T_2 covers 300 kHz to 3 MHz, the changeover by

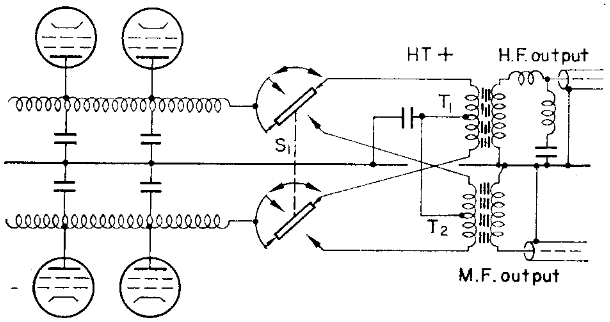

Fig. 10.7 **Output circuit arrangement for h.f. and m.f. version of distributed amplifier.**

means of switch S_1 can be anywhere between 2 MHz and 3 MHz. With this arrangement it is not possible to radiate frequencies simultaneously in both the m.f. and h.f. bands, but it enables a single wideband transmitter to take the place of two conventionally tuned transmitters; an obvious economic advantage to user organizations who employ both bands for their communications.

A good example of the need for operating in both bands is provided by shipborne applications, where space limitation is a further incentive to use one transmitter instead of two. Wideband equipment has further advantages for these applications. Access to the transmitter is not necessary during operation and the only maintenance required is an occasional valve change. Consequently, it can be housed in any convenient space on the ship, with only the frequency and traffic controls in the operators' cabin. The complete absence of variable tuning elements and complicated mechanisms combined with the general structural rigidity, mean that operational reliability is of a very high order—even under the vibration encountered on warships.

10.7 WIDEBAND TRANSFORMERS

Wideband transformers are essential for the design of wideband systems for power amplification. It is not proposed to include all the necessary detailed information to design these transformers, but to give salient features, together with an example of their practical application in the design of a transformer for 40 kW in the h.f. band.

Existing literature covers both the basic principles [11] and the use of ferrite magnetic cores [12, 13].

At the low-frequency end of the band, the transformer bandwidth is limited by the low value of shunt inductance, while the upper-frequency limit is determined by a low-pass Π network, consisting of leakage inductance and spurious shunt capacitances. This low-pass network may be so designed as to be part of a more complicated network used for broadbanding the input, interstage, or output circuit of a tuned amplifier. Alternatively, it may be part of a grid or anode delay line in a distributed amplifier.

With ferrite transformers at r.f., the core and $I^2 R$ losses cannot be reduced by careful design, as in the case of lower-frequency transformers. In order to keep the leakage inductance small, the winding must have the minimum possible number of turns, and in consequence the core material is very heavily loaded. This means that the transformer rating depends on the effectiveness of the core cooling, the temperature of which often becomes quite high. The thermal conductivity of ferrite material is very low.

Magnetic material retains its magnetic properties up to the Curie temperature, but the working temperature of a ferrite core is limited to a much lower value. As the temperature increases above ambient, the amount of heat removed by cooling increases, because, whilst the thermal conductivity increases with temperature the losses also increase with temperature. There will be a temperature above which any increase in cooling is more than offset by the increase in losses, with a resultant rapid and continued rise in temperature. This is known as the run-away temperature, and it is obvious that the working temperature must be below this value. The run-away temperature depends on the particular grade of ferrite in use, but with the grade used for h.f. power transformers, the temperature rise at the hottest part of the core should not exceed 60°C. It has been found in practice that the most satisfactory method of cooling is by conduction, which can be provided by the core being clamped between cast-aluminium cooling plates for powers up to 5 kW.

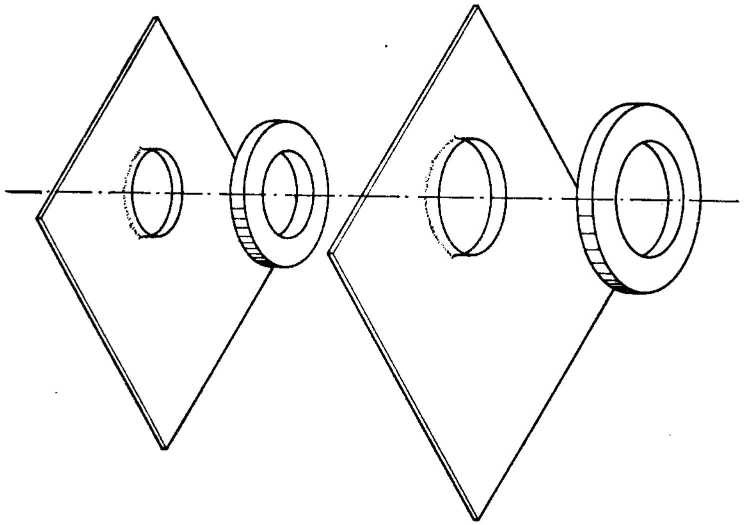

FIG. 10.8 Method of ferrite core assembly for 40 kW h.f. transformer.

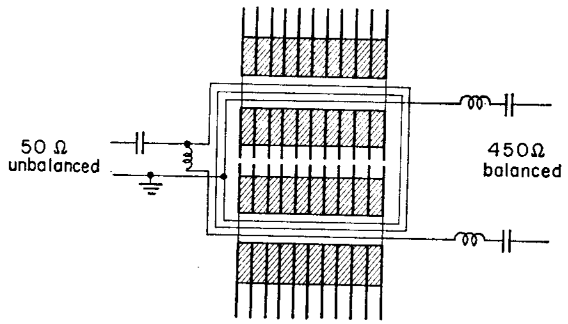

FIG. 10.9 Circuit arrangement of a 40 kW wideband transformer.

An example of an improved method of heat extraction by conduction is given in the design of a 40 kW h.f. transformer. In this design the core is composed of two stacks of twenty-five thin ferrite 'washers' (4 in. o.d., 2 in. i.d., 0·25 in. thick) sandwiched between aluminium cooling plates (Fig. 10.8), with the winding passing through the centre of the ferrite cores, as shown in Fig. 10.9. The cooling plates serve the additional purpose of reducing the leakage inductance and of screening the ferrite from any electric field, thus avoiding high dielectric loss. In spite of the large size of this transformer, the match provided between 450 Ω balanced and 50 Ω unbalanced is very good, being practically within a v.s.w.r.

of 1·2 to 1 over the frequency range of 3–28 MHz. A plot of the matching performance on a Smith chart is shown in Fig. 10.10. One of the virtues of this method of construction which makes this matching performance possible is due to the large spurious capacitances and leakage inductances being distributed along the winding; they are not in the form of single lumped elements.

The power loss in transformers of this type is between 1% and 3%, but this represents about 1 kW at full rating, so they should always be mounted in a position where the air can circulate freely.

There are two important points to consider in connection with the use of wideband transformers. The first concerns power rating and v.s.w.r. With a mismatched output, the maximum throughput power allowed is the maximum rated power divided by the v.s.w.r. produced by the mismatch. This means that the maximum mean throughput power of the 40 kW transformer is limited to 20 kW if the v.s.w.r. of the mismatch is 2 to 1.

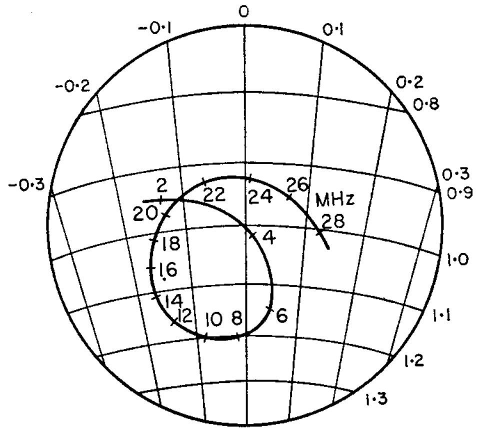

FIG. 10.10 Smith's chart presentation of input matching of a 40 kW wideband transformer.

In the case of the 1 kW wideband transmitter, the output transformer is designed for a maximum rating of 4 kW, to allow working into a feeder with a high v.s.w.r. In addition, the mismatch of the transformer must be taken into account when considering the total mismatch which a final stage can tolerate. This is because the feeder and transformer mismatches will add and subtract in a random manner over the frequency band, but there will always be some frequencies at which they will add.

The second point is concerned with the direction of power flow through the transformer. The difficulty arises when the power flow is from the balanced side to the unbalanced side. A balanced generator usually produces a certain amount of unbalanced power which will not be loaded by the transformer. Balanced-feeder resonance in an unbalanced mode is inevitable at some frequency or frequencies, so the unwanted unbalanced power can give rise to very high voltages and currents. Unless the unbalanced mode is either eliminated or damped, there is a distinct probability that the transformer will be seriously damaged if used on a frequency at which this type of resonance is present.

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11

Intermediate-Stage Amplifiers

11.1 GENERAL CONSIDERATIONS

The purpose of intermediate stages in any transmitter is to provide sufficient gain to amplify the low-power input to the level required to drive the final amplifier. In addition, the quality of the signal must be better than required at the transmitter output in order to allow for some degradation in the final stage. This is particularly important where linear amplification is concerned. Because the power consumed by the intermediate stages is only a small portion of the total, conversion efficiency is not a prime objective. On the other hand, stage gain is important because the most economical design is likely to be that with the smallest number of stages.

In the case of high- and medium-power transmitters it is usual, and more convenient, for the drives to be housed separately, especially on multi-transmitter stations. As there is always a possibility of stray r.f. from the final stage being picked up on the drive cable, any detrimental effect on this pick-up will be minimized by raising the power level in the drive cable. Typically, this is of the order of 2–5 W. Therefore the gain required from the intermediate stage or stages must be adequate to give the final-stage drive power from an input of about 2 W.

With the exception of high-power high-quality linear amplifiers, where grounded-grid triodes are used in the final stage and wideband drive systems are preferred, tetrodes are used in the final stage for practically all other high- and medium-power applications. By using tetrodes in the final stage, the drive power required is quite low and mainly due to resistance loading of the driving stage, in order to maintain the desired operational load line. Therefore, in the majority of applications, adequate gain can be provided by a single tetrode or pentode intermediate stage between the incoming drive and the final stage.

Two intermediate stages may be required for very high-power applications, but in these cases the tendency is to use the more economical approach of increasing the drive level, and again use only one intermediate stage.

The main difference between intermediate and final amplifiers is that the former operate into a substantially constant load, so tuning and loading adjustments are not necessary during service.

11.2 THE INPUT CIRCUIT

For a number of reasons it is important to terminate the incoming drive cable, which in all probability will have an impedance of 75 Ω. If the cable is terminated directly with a 75 Ω resistor, the peak r.f. voltage from a 2 W source will be only