The Marconi Series

Covering Advances in Radio and Radar

Series Editor:

P. S. Brandon

Manager, Research Division,
The Marconi Company Limited

G. E. Beck—*Navigation Systems: A Survey of Modern Electronic Aids*

J. A. Scarlett—*Printed Circuit Boards for Microelectronics*

J. K. Skwirzynski—*Design Theory and Data for Electrical Filters*

V. O. Stokes—*Radio Transmitters: R.F. Power Amplification*

R. A. Waldron—*Ferrites: An Introduction for Microwave Engineers*

R. A. Waldron—*Theory of Guided Electromagnetic Waves*

D. W. Watson and H. E. Wright—*Radio Direction-Finding*

Radio Transmitters

R.F. Power Amplification

V.O. STOKES

The Marconi Company Limited

VAN NOSTRAND REINHOLD COMPANY
LONDON

NEW YORK CINCINNATI TORONTO MELBOURNE

Contents

PART 2

MEDIUM AND LOW POWER

Illustrations

Figure

Figure

Introduction

The essential characteristic of any transmitter design is that the end-product shall fulfil all the technical requirements of its specification regarding performance, reliability, ease of operation, etc. The introduction of new techniques and components is desirable not only when they offer operational advantages to the user, but also because the most modern equipment has considerable customer appeal. However, since the commercial success of a design depends mainly on its initial cost and running expenses, innovations should be treated with caution, partly because new components generally have unknown operational reliability, and partly because there is a tendency towards increased complication and cost in striving for the perfect technical solution. In fact, economic aspects are of major importance at all stages of design, particularly regarding ease of manufacture, with consequent low production cost and operational simplicity.

Power amplifiers constitute that section of a radio transmitter which provides amplification at the final radiated frequency, with or without modulation, but do not include the modulator and frequency generating equipment. However, the modulation system with which the amplifier will be used must be considered in relation to such requirements as linearity and bandwidth. Amplifiers can be classified into a number of categories, but are usually identified by the type of service, frequency and power.

The intended application is the most important consideration initially, because it determines the frequency range to be covered and the power to be radiated from the transmitting antenna. For any particular application, the output power required from the final amplifier is determined by the antenna efficiency, the propagation path loss and the signal-to-noise ratio which will be necessary at the receiving station to give an adequate service. The computation of these factors from published graphs or by calculation, is not part of this project, but the importance of accuracy cannot be overstressed in relation to specifying the power required out of the final amplifier. If the estimate is too low, the service will be inadequate, and to increase the power output after completion of an installation can be a very costly business. On the other hand, a serious over-estimate will mean that both initial and running costs are higher than those of competing manufacturers, resulting in fewer sales.

Apart from the economic aspect, excessive radiated power is more likely to interfere with other radio transmissions, yet the power output of transmitters differs considerably for different applications, even for the same transmission path length in the same frequency band. So, it is important that the reasons for this apparent discrepancy are fully appreciated prior to the commencement of any transmitter design, in order that the proposed equipment shall meet the optimum objective of providing an adequate service with a minimum of radiated power. Consequently the first chapter of this book is devoted to explaining the reasons why the power level indicated by the term 'high power' often varies very considerably in relation to different applications.

It may appear unorthodox that the last stage of amplification is considered first, but the major contributory factor to the success of a transmitter, both technically and economically, is provided at the high-power stage. Further, as the driving power required by the final stage cannot be known until all the operating conditions of this stage have been determined, the design of an amplifier chain must begin at the final stage.

The second chapter is mainly a recapitulation of known-art regarding the class of valve operation, but emphasis is placed on the power consumption because the major portion of the total input power to a transmitter is fed into the final stage. Eventual running costs are mainly dependent on the power consumption of this stage. The stage gain is also important, in that it determines the level of driving power required.

Throughout all stages of transmitter design, consideration must be given to the potential reliability of the equipment when put into operational service. Breaks in transmission can be annoying during entertainment broadcasting, more serious in civil communication links, but possibly disastrous in military communications. To cover such contingencies it is not unusual to provide duplicate low-power equipment, with automatic changeover facilities, for restoring service when a fault develops. With high-power equipment, frequently due to space limitation but also on economic grounds, duplication is the exception rather than the rule.

It follows that a high order of operational reliability should be one of the most essential characteristics of high-power design. This requirement is often in conflict with cost considerations, but cheap equipment is doomed to failure if operational breakdowns occur too frequently. The real skill of a good design engineer is proven by his choice of tolerances to give the best compromise between reliability and cost *for a given application.* Where there is a choice between a component which will initially just comply with the required specification, and one which is apparently somewhat under-rated, a good guiding principle is 'fit and forget', rather than 'fit and fret'.

In the present stage of development, the active elements for high-power amplifiers are valves, so solid-state devices are only considered for lower-power applications. In connection with communications, it should not be forgotten that the transmitter is only one item of the total equipment required to form a complete communication system. As such, the design must be engineered to be fully in accordance with the system requirements. With the possible exception of the antennas and feeders, the transmitter is usually the most expensive item. Consequently, the technical performance and cost of the transmitter influence potential customers towards purchasing the whole-system equipment from the manufacturer who can supply the best transmitters.

Part 1
High Power

1

Power Relationships

1.1 THE MEANING OF HIGH POWER

The power output of a radio transmitter is normally specified as the radio-frequency power delivered by the final or high-power amplifier.

The actual power at the output terminals varies considerably for different applications, both in level and by definition. This means that the power input and size of high-power amplifiers also cover a wide range, but the following comparisons show that, in many cases, these differences are only apparent when considered in terms of the effective level of intelligence power radiated. In other cases the differences are substantial, particularly when comparing transmitters for operation in the v.l.f. and the h.f. bands, where attenuation of the propagation paths also differs considerably.

It is normal practice in telecommunications to express power, voltage and current ratios in decibels. For receiver and low-power applications, a requirement for increasing the gain by, say, 6 dB, can be met with equipment which is relatively easy to construct and inexpensive. When comparing levels in the high-power region, ratios expressed only in decibels are liable to give a false impression of the real significance in terms of equipment size, power consumption and cost. For example, to increase the power output of a 100 kW amplifier by 6 dB, means an extra 300 kW. Hence, it is not surprising to find that overall system gain achieved by increasing transmitter power is sometimes described as 'gold plated decibels'.

Double-sideband and single-sideband transmissions

Consider a comparison between transmissions of audio-frequency programmes for broadcasting and for point-to-point communications in the h.f. band between 3 MHz and 30 MHz. Both types of transmission are used in this band for long-haul circuits, with similar frequencies for any particular distance range so that a direct comparison can be made on a system basis.

In the generally accepted meaning of the term, broadcast transmissions are aimed at providing good-quality reception by anyone within the service area. So to reach the largest possible audience, the radio receivers must be simple to operate and relatively inexpensive. To meet this condition, the most appropriate transmission system is amplitude modulation, with a high-level carrier and a pair of lower-level sidebands which convey the intelligence. Amplitude modulation is classified as double sideband (d.s.b.) and is used for most broadcast applications. The power rating of d.s.b. transmitters is specified in terms of the level of carrier power at the output terminals.

On the other hand, communication transmissions are for conveying information between two specific points. They are mainly used by civil and military organizations, where the prime objective is to maintain continuous communication and quality need not be so high as in broadcast systems. Experienced operators are employed, so the receivers can be more sophisticated, enabling the use of more appropriate transmission systems, such as the single sideband (s.s.b.). The cost of s.s.b. receivers is much higher than that of d.s.b. receivers, but when considered in relation to the overall cost of a communication network, the extra cost is relatively insignificant.

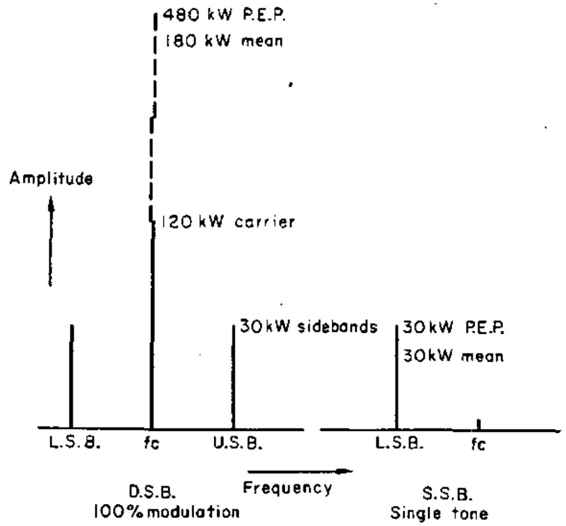


FIG. 1.1 Comparison between radio-frequency spectrum with d.s.b. and s.s.b.

In conventional d.s.b. working, the intelligence is contained in full in each of the two sidebands. Advantage can be taken of this redundancy by transmitting on one sideband only. After amplitude modulation at a low level, the carrier and one sideband are filtered out, leaving the other sideband to be amplified and radiated. For s.s.b. transmitters the power output is specified as peak envelope power (p.e.p.) or peak sideband power (p.s.p.).

Peak envelope power is the power which would be developed by a continuous carrier, the amplitude of which is equal to the peak amplitude of a signal consisting of a pilot carrier and one or more sideband components.

Peak sideband power is the power which would be developed by a continuous carrier, the amplitude of which is equal to the peak amplitude of one or more sideband components, with no pilot carrier.

The two terms are practically synonymous because if a pilot carrier is radiated, it is at a low level—normally −16 dB or −26 dB with reference to peak level.

The output power required for the final amplifiers of d.s.b. and s.s.b. systems, to give the same radiated power in one sideband, is shown in Fig. 1.1. It will be seen that the total power output is 180 kW (120 + 30 + 30) in the d.s.b. case at 100% modulation, compared with 30 kW for s.s.b. at full level. Taking into account the power required by the modulator, a more comprehensive comparison between the two systems is shown in Table 1.1.

TABLE 1.1

	d.s.b.	s.s.b.
r.f. output, no signal	120 kW	0
Final amplifier efficiency	80%	66%
Final amplifier d.c. input	150 kW	10 kW (approx.)
r.f. output, full signal	180 kW	30 kW
Power output of modulator	75 kW (50% of 150 kW)	negligible
Modulator efficiency	50%	unimportant
d.c. power input to modulators full signal	150 kW	negligible
Total d.c. input, full signal	300 kW	45·5 kW
Difference between d.c. input and one r.f. sideband	270 kW (300 − 30)	15·5 kW (45·5 − 30)
Peak voltage on anode of final amplifier valve	H.T.V. × 4 (approx.)	H.T.V. × 2 (approx.)

When audio programmes are being transmitted in operational service, the average levels of input and radiated power will be less than those given in the Table 1.1, by an amount depending on the type of programme, but the maximum conditions must be considered at the design stage.

For long-range transmission systems of any type in the h.f. range, it is usual to employ directional antennas, which concentrate the power into a beam in the desired direction. Typically, antenna gain is of the order of 10 dB, so the effective radiated power (e.r.p.) of one sideband in the foregoing comparison is about 300 kW for both d.s.b. and s.s.b. transmissions.

In the case of communication circuits, it is usual to employ directional antennas for reception as well as transmission. Elementary antennas are used with broadcast receivers, whereas complex antennas are used by operating organizations to enhance reception, enabling the use of much lower transmitting powers of the order of 1 kW or so in some instances. One of the main advantages of directional antennas for reception is the reduced level of unwanted signals and noise from other directions which improves the signal-to-noise ratio.

Broadcast transmissions in the m.f. band

The majority of transmissions in this band are used for local services, and have, in fact, formed the backbone of most broadcasting systems for many years. The transmitter power required is governed by the area to be served and limited to a level which will cause as little interference as possible with m.f. broadcasts in other countries. In this context, high-power transmitters fall within the range of 50–100 kW carrier power. Double-sideband systems are used, so the level of each sideband is one-quarter of the carrier power at 100% modulation, i.e. between 12·5 kW and 25 kW. Relative to the power ratios being considered, the gain/loss characteristics of suitable antennas can be neglected, so the effective radiated power is also between 12·5 kW and 25 kW—considerably less than the 300 kW e.r.p. of a 30 kW s.s.b. transmitter.

This, again, emphasizes the price that has to be paid in transmitter power in order that good-quality reception can be obtained with simple and inexpensive receivers.

Long-range broadcasting in the m.f. band

Long-range broadcasting is also carried out in the m.f. band, using very high-power transmitters. These programmes are mainly for propaganda purposes, so d.s.b. transmission is employed in order that reception can be made with very simple and inexpensive receivers. In terms of received signal strength, it is not possible to give a general comparison with h.f. broadcasting, because of the different propagation characteristics in the two bands, but a comparison can be made in terms of e.r.p.

In the m.f. band it is a very expensive proposition to erect directional antennas and the gain achievable is small, so it is usual to employ omni-directional antennas, for which the efficiency is about 90%. With a d.s.b. system and an omni-directional antenna, a radiated power of 300 kW in one sideband, at 100% modulation necessitates a carrier power of 1330 kW and a total output of 2000 kW at the transmitter terminals. This is a typical power-level used for this type of service. It is not surprising that the final stage of such a transmitter is called a super-power amplifier, yet the effective power of intelligence radiated is only the same as that given by the final amplifier in a 30 kW s.s.b. system.

From these comparisons it is obvious that s.s.b. transmissions are far more economical at the transmitting station, both initially and during operation. But this advantage is probably less important than the greater use that is made of the frequency spectrum and the less likelihood of causing interference with other transmissions, because the total power radiated is so much lower.

Transmission in the v.l.f. band (below 30 kHz)

Transmitters operating in the v.l.f. band are in a different category for a number of reasons, so that a direct comparison with transmitters in other bands can only be made in respect of amplifier output and radiated power. The available bandwidth is too low for speech, limiting the traffic to telegraphy. Transmissions at v.l.f. are subject to very much less ground-wave attenuation than those at higher frequencies, and relatively low power need be radiated for world-wide communication. This is just as well, because it is not practical to erect antennas of high efficiency at these frequencies. A good example is at the G.P.O. station at Rugby, where the figure-of-eight antenna, mounted on a number of 850 ft masts, has an efficiency of less than 20% at 16 kHz. This means that the radiated power is less than 20 kW for every 100 kW appearing at the transmitter output.

The bandwidth of v.l.f. transmission is further limited by the Q factor of these large antennas, which is often between 200 and 300. Where bandwidth is of prime importance in order to operate with telegraph rates of up to fifty or seventy bauds, it may be necessary to add damping to the antenna circuit, thereby reducing the radiation efficiency and radiated power for a given input.

Tropical broadcasting

The h.f. band is also used for broadcasting in tropical regions, where atmospheric noise is less than in the m.f. band. The radiated power required is relatively low, because communities tend to be grouped in small but widely separated areas. Another power limit is imposed by the need to keep skywave radiation to a minimum, and considerable attention is given to antennas designed for this purpose.

'Typically high power' refers to carrier levels between 10 kW and 20 kW, with one-quarter of the power in each sideband. As omni-directional antennas are generally employed, the effective radiated power is approximately the same as at the amplifier output.

Compatible single sideband in the h.f. band

Compatible single sideband is the term applied to a system in which a full-level carrier and only one sideband are radiated. It provides a ready means of adapting s.s.b. transmitters to enable them to be used for communication with receiver stations fitted with d.s.b. equipment only, hence the term 'compatible'. It is

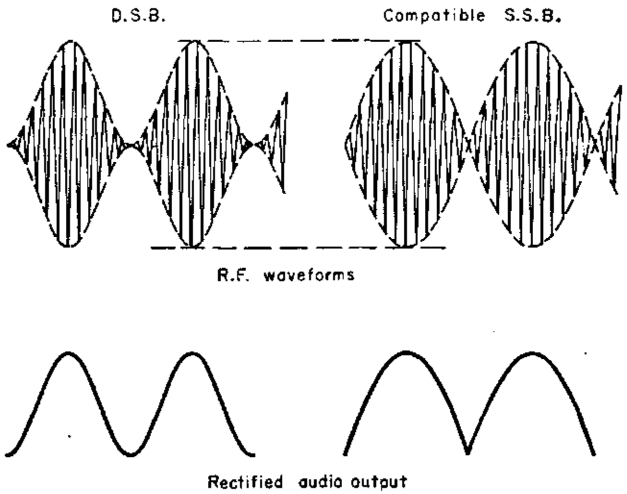


Fig. 1.2 Waveforms produced by d.s.b. and compatible s.s.b. systems.

limited to single-channel operation, but it is in quite general use for low-grade commercial-quality speech circuits.

At full modulation, the sideband amplitude is equal to the carrier amplitude. The radiated power in the one sideband is half the total mean power and one-quarter of the peak power. In terms of the proportion of intelligence power radiated, this is obviously an improvement on d.s.b., where the power in one sideband is one-sixth of the total mean power and one-sixteenth of the peak power.

The disadvantage is the high level of distortion, which increases with modulation depth to about 30% when the sideband is the same amplitude as the carrier. Consequently, its use is normally limited to low-power transmitter applications.

In its simplest form it is not suitable for high-quality programmes, so is not applicable to high-power transmitters for broadcasting.

In efforts to improve the quality and take advantage of the power and spectrum saving possibilities, particularly for high power, several modifications to the basic system have been proposed, notably by Kahn [1]. These have all been tried experimentally by a number of operating organizations, but the performance has not been acceptable and the system is not in general use for broadcasting.

The reason for a high level of distortion with this system can be seen in Fig. 1.2, where the waveforms of d.s.b. and compatible s.s.b. are compared. In the compatible case this type of waveform is produced when the carrier and one sideband are of equal amplitude. When the d.s.b. signal is rectified, the resultant audio-frequency envelope will be sinusoidal. On the other hand, when the compatible s.s.b. signal is rectified, the resultant audio-frequency envelope will be a series of half waves, containing fundamental and a high harmonic content.

1.2 MULTI-CHANNEL OPERATION

It is well known that for many years amateurs have achieved world-wide communication with only a few watts of r.f. power. There seems little doubt that the initial professional interest in the h.f. band was stimulated by the results obtained by amateurs. So it may seem curious that e.r.p. levels of the order of 300 kW are considered necessary for long-range communication links by civil and military organizations, particularly bearing in mind the more efficient receiving equipment which they have available.

There are two reasons for this apparent anomaly. First, amateurs can pick the best time to take advantage of the existing propagation conditions, whereas professional organizations require their communications circuits to be available at any time of day, at any season of the year and for any year of the sunspot cycle. Here higher power is obviously an advantage in maintaining circuits open during marginal conditions. Second, by using s.s.b. techniques, several speech and/or telegraph channels can be radiated from a single transmitter, the total power being divided between the channels so that the power on each channel is only a fraction of the total. In considering power relationships, the importance of radiating a number of channels within the bandwidth allocated to a single assigned frequency should not be overlooked. This facility is an invaluable asset to all organizations operating communications services in the h.f. spectrum, which is already overcrowded.

In the case of multi-channel operation, the signals on the two sidebands are quite independent of one another except in relation to power level, and the system is known as independent sideband (i.s.b.). The power per channel must be reduced to a level which ensures that the peak power capability of the transmitter is not exceeded to such an extent as to cause excessive cross-talk or out-of-band radiation. The amplitude of the composite signal is the sum of the amplitudes of the individual signals and the amplitude ratios are the square root of the power ratios; so the total peak power is the square of the sum of the roots of the individual powers. As an example, consider four frequency-division multiplex channels of equal power, W, operating on a transmitter of 30 kW peak power, then $(4\sqrt{W})^2 = 30$ kW; from which $W = (30/16)$ kW, which is 1·9 kW per channel, or 12 dB below p.e.p.

In the case of multi-channel speech, the channel level and transmitter loading cannot be clearly defined in terms of p.e.p., since the duration of signals exceeding nominal rating is very short and not of great consequence. G.P.O. telephony circuits (other than Lincompex) are operated at a speech-signal level of −8 to −10 dB RTP at a 0 dBm reference point [2], where a 1 mW tone-signal produces −6 dB relative to p.e.p. output at the transmitter. Speech-signal level is measured on a volume indicator in accordance to C.C.I.F. specification (1936). Other operating concerns often use the American VU meter or a speech voltmeter [3].

Speech-channel level and transmitter loading continues to be a matter for study and data directly relating to tests on actual transmitters is scanty. Some tests have been made using speech or white-noise signals, but the difficulty remains of correlating tests with actual traffic conditions. Such tests have clearly indicated the advantage of Lincompex circuits [4].

The present G.P.O. practice for transmitters carrying Lincompex circuits is (taking V = p.e.p. voltage):

Single-channel Lincompex

Pilot carrier:	−16 dB relative to p.e.p. =	0·158 V
Speech level:	−6 dB relative to p.e.p. =	0·500 V
Control-signal level:	−11 dB relative to p.e.p. =	0·282 V
		0·940 V

Four-channel Lincompex

Pilot carrier:	−16 dB relative to p.e.p. =	0·158 V
Speech level:	4 at −16 dB relative to p.e.p. =	0·632 V
Control-signal level:	4 at −21 dB relative to p.e.p. =	0·356 V
		1·146 V

i.e. +1·2 dB relative to p.e.p.

Lincompex is the term applied to a system developed by the British Post Office for improving h.f. radio telephone circuits. The system employs compression of the speech level for transmission, with expansion on reception to restore the original level variations by means of a separate control signal. The name Lincompex is derived from the expression *lin*ked *comp*ressor and *ex*pander. The compressor is in the audio circuit of the transmit side, the expander is in the audio circuit of the receive side and linking is carried out by a narrow band f.m. channel situated above the speech band, but within the normal 3 kHz channel bandwidth.

With this system, the audio input level to the transmitter is substantially constant and the transmitter loading can be specified in terms of p.e.p.

For multi-channel telegraphy the power per channel depends on the type of system employed. The two main systems in general use are frequency-division multiplex (f.d.m.) and time-division multiplex (t.d.m.). Both systems use two sideband frequencies to represent the mark and space conditions, but the manner in which they are transmitted is different. In f.d.m., one of the sideband frequencies is always present on each channel, so four frequencies will be present simultaneously in a four-channel system, and the power per channel must not exceed

one-sixteenth of the total rated output power. In the case of t.d.m., the channel frequencies are separated on a pre-arranged time basis, only one frequency being present irrespective of the number of channels. This means that each channel can be operated at the full output power of the transmitter. Taking four-channel operation as an example, the permissible output power per channel on t.d.m. is sixteen times that on f.d.m. for the same transmitter.

When more than four f.d.m. channels are being radiated, the C.C.I.R. Recommendation [5] is that the power per channel can be increased to:

$$\frac{\text{Total admissible peak power}}{4n}$$

where n is the number of simultaneous channels. In this way, whenever there are more than four channels, rather more than the theoretical channel power can be radiated without exceeding the total peak power for more than 1–2% of the time.

In the case of one or two speech channels (except Lincompex), and a number of f.d.m. telegraphy channels being radiated simultaneously, the level of each speech channel is set as for telephony only. The level of each telegraph channel is reduced by a further 3 dB for one channel of speech and 6 dB for two, relative to the channel level without speech.

Table 1.2 gives the recommended level at which each telegraph channel frequency should be set, in relation to the number of telegraph channels and the absence or presence of speech in one or two channels.

TABLE 1.2 *Frequency level relative to peak power*

Number of f.d.m. channels	No speech channel, dB	One speech channel, dB	Two speech channels, dB
1	0	−3	−6
2	−6	−9	−12
3	−9·5	−12·5	−15·5
4	−12	−15	−18
6	−13·8	−16·8	−19·8
8	−15	−18	−21
12	−16·8	−19·8	−22·8

It is interesting to note the level of each of the twelve telegraph channels, in the presence of two speech channels, i.e., −22·8 dB. Relative to 300 kW e.r.p., −22·8 dB is approximately 1·5 kW. Allowing for an antenna gain of 10 dB, the transmitter output per channel is 150 W, which is the same order of power as that used by amateurs.

An important feature of multi-channel telegraphy operation is the ability to increase the power per channel by reducing the number of channels. This often prevents a complete breakdown of the circuit under abnormal conditions by using only one or two channels.

1.3 VALVE CONVERSION EFFICIENCY

When contemplating the purchase of high-power transmitters, one of the main features considered by operating organizations is total running cost, of which an appreciable portion is attributable to power consumption. As most of the power is fed into the final stages, the d.c. to r.f. conversion efficiency of power-amplifier valves must be a major consideration in all high-power designs.

The result of obtaining the optimum conversion efficiency is a threefold advantage. First, the total power consumption will be at a minimum. (This is of great importance where the power has to be generated on site.) In the case of high-power amplitude modulation, the modulator must supply sufficient audio-frequency power to modulate the d.c. input to the r.f. amplifier. The power input to both modulator and r.f. stages is reduced by improving the conversion efficiency of the r.f. valve.

Secondly, the mains power not converted into r.f. is converted into heat; so the arrangements for disposing of the unwanted heat will be simplified by improving conversion efficiency.

The third advantage is the result of the first and second. The overall size and cost of the equipment, including mains transformers, rectifiers and cooling arrangements, will be lower and more attractive to prospective purchasers.

The conversion efficiency of a valve depends on the dynamic operating conditions and these are grouped into three main categories, classes A, B and C. There is also a recently introduced form of class C operation, which is sometimes called class D. Amplifiers of this type use rectangular, or curved–rectangular waveforms and have higher conversion efficiencies than class C types.

Table 1.3 has been compiled to show the relative conversion efficiencies of triode and tetrode valves in classes A, B, C and D under typical operating conditions, in order to emphasize the effect of conversion efficiency in terms of power.

TABLE 1.3 *100 kW r.f. output*

Class	Triodes				Tetrodes			
	A	B	C	D	A	B	C	D
Typical conversion efficiency, %	38	66	83	89	38	64	82	88
Power input, kW	263	152	121	113	263	156	122	114
Anode loss, kW	163	52	21	13	163	56	22	14

From this table it is apparent that if conversion efficiency were to be the overriding factor in all transmitter designs, then class D operation with triode valves would be the universal choice. But this is not so. There are many applications where other factors are equally important, or even paramount, such as linearity, rapid change of frequency, available drive power and suitable valves. The reasons for the choice of valve type and class of operation are discussed in Chapter 2 in relation to various high-power applications.

There are two other comments on conversion efficiency which are appropriate. First, the calculation of power output and efficiency for any particular valve will be based on the *average* characteristics for that type, as supplied by the valve manufacturer. Although all valves of the same type have very similar characteristics (especially high-power types), they will not be identical. Consequently, some tolerance must be allowed to accommodate probable differences. This point is particularly important when testing a new design for performance, which cannot be fully proven until checks have been made with a number of valves.

The second point is that the calculated output power is that obtained at the valve anode. The transmitter output will be lower than that at the valve anode, due to the inherent loss in the circuits associated with tuning and coupling. The circuit loss may not appear to be appreciable when expressed in decibels, but it can be a serious embarrassment if a design will not quite meet a specification in terms of actual power in kilowatts. For example, consider an amplifier in which the combined anode and output circuits have loaded and unloaded Q factors of 20 and 500, respectively. The power lost in the circuits is one-twenty-fifth of the valve output, corresponding to −0·17 dB. This ratio represents a loss of 4 kW in 100 kW, so the amplifier output would be 96 kW for 100 kW at the valve anode.

REFERENCES

[1] KAHN, L. R. 'Improved compatible single-sideband equipment for standard broadcast service.' *I.R.E. natn. Conv. Rec.*, 7, 55–60 (1958).

[2] BRAY, W. J. and MORRIS, D. W. 'Single-sideband multi-channel operation of short-wave point-point radio links'. *P.O. elect. Engrs' J.*, **45**, Part 3 (October 1952).

[3] SHEARME, J. M. and RICHARDS, D. L. 'The measurement of speech level'. *P.O. elect. Engrs' J.*, **45**, Part 3 (October 1954).

[4] WATT-CARTER, D. E. and WHEELER, L. K. 'The Lincompex system for the improvement of h.f. radiotelephone circuits'. *P.O. elect. Engrs' J.*, **59**, Part 3 (October 1966).

[5] C.C.I.R. Recommendation 326, 'Power of radio-transmitter relationship between the peak envelope power, the mean power, and the carrier power of a radio transmitter'. *Documents of the Xth Plenary Assembly*, Vol. 1 (Geneva 1963).

2

Amplifier Classification and Valve Characteristics

2.1 GENERAL CLASSIFICATION

Valve amplifiers can be grouped into a number of categories, depending on their power rating, application, frequency range, valve types and their dynamic operating conditions, circuit configuration, etc. Most of these classifications are interdependent, that used at any one time being determined by the context of the subject under discussion.

Within the context of high-power amplification, the intended application is the important aspect initially, for it directly implies the frequency range and order of power level required. The frequency range is then a *fait accompli*, but the method of obtaining the power level is determined by the dynamic operating conditions of the power valves appropriate to the application. Valve type (triode, tetrode or pentode), is determined partly by the application and partly by the type available at the power level required.

The circuit configuration associated with the valve (grounded cathode, grounded grid, neutralized, etc.) is also determined by both the application and the valve type, but the configuration of the anode-output circuits is often dependent on the availability of suitable components in respect of value, voltage and current rating.

From the foregoing observations, it is clear that the basis of a successful power-amplifier design depends on a full appreciation of the dynamic operating conditions of valves in relation to various applications. It follows that it is necessary to apply the correct interpretation to valve characteristics in terms of performance capability.

A number of textbooks [1–3] give a fairly exhaustive description and analysis of class A, B and C amplifiers, together with sub-divisions, indicating the absence or presence of grid current by subscripts 1 and 2, respectively. A degree of recapitulation is given, but only in sufficient detail to point out the salient features affecting high-power applications. Class D operation is also included.

2.2 DEFINITION OF CLASS A, B, C AND D OPERATION

Class A

A valve amplifier is operating in class A when the grid bias and alternating voltages are such that the anode current is flowing continuously throughout every electrical cycle.

Class B

By definition, the bias of a class B amplifier is adjusted so that the anode current flows for 180° of the alternating voltage, i.e., for half the electrical cycle. In high-power applications, the bias voltage is rather less, allowing some anode current to flow in the no-signal condition (static feed). When an alternating grid voltage is applied, the anode current flows for rather more than 180°. In accordance with conventional definition, this is class AB operation. However, compared with the peak anode current at the crest of a driving waveform, the static anode current is relatively small and power amplifiers in this state are still said to be in class B.

Valves operating in class B are more efficient than those in class A. The peak and mean-power output is also higher, but the peak driving voltage for full output is approximately twice as high.

Class C

For class C operation, the bias is set at a level well above that required to cut off the valve in the static condition, and anode current flows during less than half of the alternating cycle, normally for about 120°.

Class C is more efficient than class B in terms of d.c. to r.f. conversion. Also, the required peak driving voltage is higher for full output and the drive power is higher, due to the presence of grid current.

Class D

Class D is a modified form of class C, first used by Tyler [4] for high-power applications. The biasing condition is the same, but the applied waveform is rectangular instead of sinusoidal. The efficiency is higher than class C, an important feature for high-power transmitters. The peak driving power is the same as for class C, but the mean driving power is much higher. In fact, available grid dissipation can be the factor which limits the output in class D operation.

At radio frequencies, it is not practical to derive a truly rectangular waveform at the power level required to drive a final amplifier. It has been found in practice that a 'squared' waveform, consisting of fundamental plus a second or third harmonic in the correct proportions, is sufficiently rectangular to give efficiencies approaching 90%.

2.3 VALVE CHARACTERISTICS

Preliminary considerations

For many years the electrical characteristics of power valves were plotted in the form of anode current against anode voltage (I_A/V_A) for constant grid voltages, V_g. It is now more usual to use 'constant current' characteristics, in which V_A is plotted against V_g for constant values of I_A. While the former method enabled a fair estimate of linear performance to be obtained from a cursory examination, the latter is accepted as being more suitable for the computation of performance under all operating conditions.

Additional information is supplied regarding the limiting values of voltage, current and power dissipation which can be permitted for each electrode, when using the cooling arrangements recommended by the valve manufacturers.

Typical operating conditions are also given for the appropriate class and application, for which the particular valve is suitable. These typical conditions are very useful at the pre-design stage when scanning data for the most appropriate valve for the application under consideration. They not only enable a 'short-list' of valves to be selected quickly, but they also provide guide lines for verifying subsequent calculations based on the plotted characteristics.

2.4 CHARACTERISTICS OF POWER TRIODES

The constant-current characteristics of a power triode are shown in Fig. 2.1, together with the abridged data for maximum ratings permitted. Based on this information, the r.f. performance can be derived from an analysis of the anode current waveform, when an alternating voltage is applied between grid and cathode. Normally the waveform of the grid voltage is assumed to be sinusoidal.

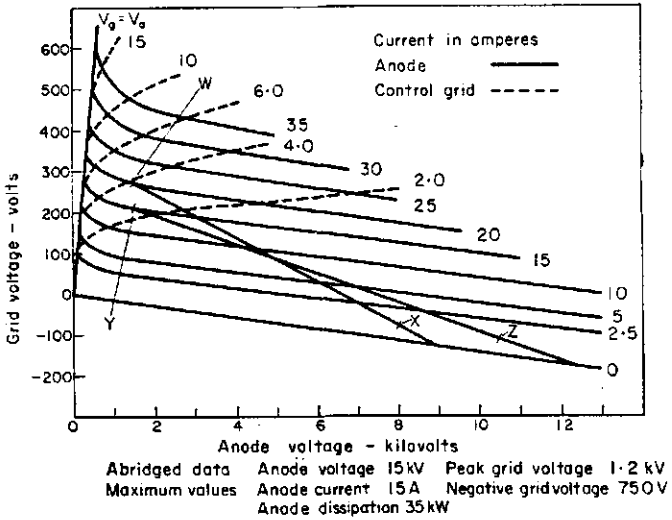


FIG. 2.1 Load lines on constant-current characteristics of triode valves type BR 1161 (English Electric Valve Co.).

When the amplifier load appears as a pure resistance at the valve anode, i.e., when the anode circuit is in tune, the excursion of the anode current follows a straight line, called the operating line or load line. This line joins the point given by the bias and anode voltages at zero signal, to the peak value of anode current reached at peak grid voltage. These points must be settled first.

The reasons for the selection of these limiting points initially and a method of deriving the performance from the resulting load line, can be clarified by means of a simple example. Consider a requirement for a linear class B amplifier, with a power output of 30 kW, from a valve having the characteristics shown in Fig. 2.1 and an available h.t. supply of 8 kV. The obvious non-linearity near cut off and in the region of peak anode current at low anode voltage, means that these

areas must be avoided. The latter area limits the minimum anode voltage at peak current to about 1·5 kV, which will give a peak anode voltage excursion of 6·5 kV (8 – 1·5).

If the linearity is perfect in the operating region, the power output will be half the product of the r.m.s. values of r.f. voltage and current (half the product because the valve is conducting for only half of each cycle), i.e.,

$$\frac{PV_A}{\sqrt{2}} \times \frac{PI_A}{\sqrt{2}} \times \tfrac{1}{2} = 30 \text{ kW}$$

or

$$PV_A \times PI_A = 4 \times 30\,000 = 120\,000$$

As the PV_A has been selected at 6500,

$$PI_A = \frac{120\,000}{6500} = 19 \text{ A (approx.)}$$

Assuming a static feed of 1·0 A, the peak anode current required is 20 A and point W is given by this value at an anode voltage of 1·5 kV. The lower end of the load line is 1·0 A at 8 kV. The approximate power output, when operating on this load line is

$$\frac{6{\cdot}5 \text{ kV} \times 19 \text{ A}}{\sqrt{2} \times \sqrt{2} \times 2} = 31 \text{ kW}$$

The conditions given by operating on load line WX, is sufficiently near the required output power for the performance to be examined in greater detail. Before the load line can be drawn, the location of point X must be found, because the 1·0 A curve is not given in this example.

To determine point X, plot a curve of grid voltage and anode current on either side of 1·0 A at 8 kV, as shown in Fig. 2.2. Interpolation shows that a

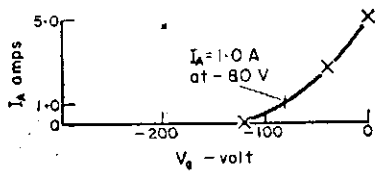


FIG. 2.2 Interpolation of I_A from V_g/I_A graph.

bias of –80 V is required to allow 1·0 A of static feed, so point X is at the intersection of –80 V and 8 kV. This method will also be found advantageous in determining other V_g/I_A points with greater accuracy for subsequent analysis.

The next step is to determine the performance obtained by operating on this load line in terms of power output at fundamental and low-order harmonics, d.c. anode current, power input, anode dissipation, conversion efficiency, d.c. grid current and required driving power. A practical method consists of tabulating a list of anode current levels, corresponding to grid voltages at regular intervals of the drive waveform, throughout the conducting period of the anode cycle. From this list the required information can be obtained by any well-known method of waveform analysis, such as that given in Appendix I. Similarly, an analysis of the grid current waveform can be obtained. Intervals of 15° are normally adequate, except where there is a considerable departure from sinusoidal waveforms, as in class D operation.

A method of performance computation

A simplified method of analysis can be performed by the use of the 'tube performance computor' first produced by Eitel McCullough Inc. [5]. A copy of this computer is shown in Fig. 2.3, although it should be realized that it must be

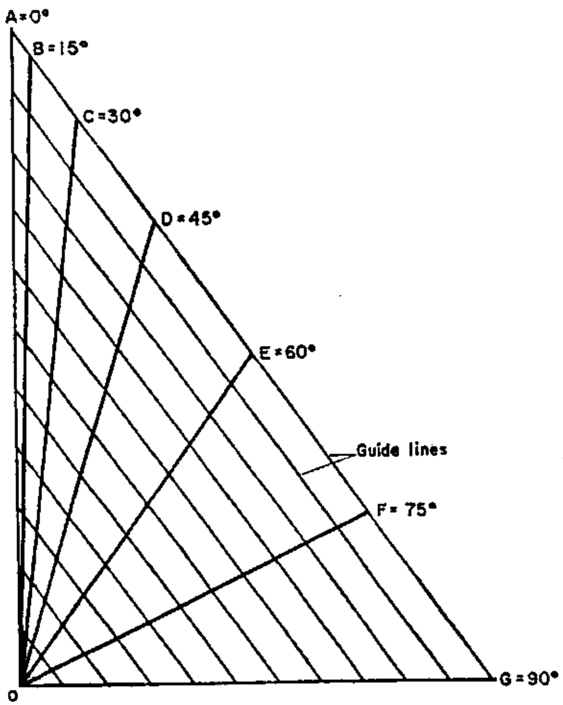


FIG. 2.3 Tube performance computor (Eimac).

provided on a transparent sheet, so that the valve characteristics can be seen through it.

The computor is placed on top of the characteristics, with the guide lines parallel to the load line and with lines OA and OG passing through the peak and quiescent points, respectively.

Lines OA, OB, OC, OD, OE, OF, OG, correspond to 15° intervals of the drive waveform, so that the points where they intersect the load line are the

current amplitudes at these intervals. These amplitudes are recorded as A, B, C, D, E, F, G, and the performance is calculated from the following formulae.

Anode current, d.c., meter reading =

$$\frac{(0{\cdot}5A + B + C + D + E + F)}{12}$$

Grid current, d.c., as for anode current, from grid characteristics.
Peak fundamental (r.f.) =

$$\frac{(A + 1{\cdot}93B + 1{\cdot}73C + 1{\cdot}41D + E + 0{\cdot}52F)}{12}$$

Output power =

$$\frac{(\text{peak r.f. anode current} \times \text{peak r.f. anode voltage})}{2}$$

Driving power = grid current, d.c. × peak r.f. grid voltage
Peak second harmonic r.f. (approx.) =

$$\frac{(A + 1{\cdot}73B + C - E - 1{\cdot}73F)}{12}$$

Peak third harmonic r.f. (approx.) =

$$\frac{(A + 1{\cdot}41B - 1{\cdot}41D - 2E - 1{\cdot}41F)}{12}$$

The harmonic formulae are suitable for tetrode and pentodes only, so another method such as that described in Appendix I must be used for triodes.

Examples using the performance computer

Referring to Fig. 2.1, the performance behaviour of the valve has been worked out for two load lines, WX and YZ, to show both the use of the computer and the effects of a change in h.t. voltage for approximately the same output power. These results are shown in Table 2.1.

From Table 2.1, it will be seen that by increasing the h.t. voltage from 8 kV to 10·5 kV, the fundamental output is increased slightly, but the required driving power is reduced by more than two-and-a-half times. As the r.f. output is higher, the anode conversion efficiency is also higher, because the input power is practically unaffected.

In both conditions, this valve is operating well below its maximum permissible levels, so the factors determining which load line to use depend on the relative merits of low h.t. voltage with high drive, or high h.t. voltage with low drive. This surely must be a decision based on economic grounds.

It was mentioned earlier that the characteristics are only typical for that type and that the calculated level of output power does not take circuit losses into account. It is at this early design stage that some allowance must be made for valve tolerance and circuit losses, in order to avoid an embarrassing shortage of power at a later stage.

TABLE 2.1

Triode	Load line *WX*	Load line *YZ*
h.t. voltage, d.c.	8 kV	10·5 kV
Anode voltage trough	1·5 kV	1·5 kV
Peak r.f. voltage	6·5 kV	9 kV
Peak of anode current	20 A	15 A
Static anode current	1·0 A	1·0 A
Anode current excursion	19 A	14 A
Peak fundamental r.f. current	9·6 A	7·27 A
Anode current, d.c.	6·0 A	4·57 A
Power input, d.c.	48 kW	47·95 kW
Fundamental r.f. output power	31·2 kW	32·75 kW
(Fundamental r.f. output power of initial assessment, for comparison)	(30·88 kW)	(31·5 kW)
Anode dissipation—neglecting harmonic power	16·8 kW	15·2 kW
Power conversion efficiency, d.c. to r.f.	65%	68·2%
Grid current, d.c.	0·7 A	0·29 A
Peak r.f. grid voltage	350 V	320 V
Drive power required, valve only	245 W	93 W

Before leaving considerations of the load line, there are two other matters which are not always apparent. First, as a plot of anode current against anode voltage, the slope of the load line represents a resistance value, known as the resistance of the load line. This is *not* the resistance value used when calculating the anode/output circuits for matching the load to the valve. For this purpose the effective value of matching resistance required is determined by the peak fundamental r.f. current and the output power, i.e.

$$R = \frac{\text{power output}}{I^2}$$

Referring to the specimen load line *WX*, its V_A/I_A resistance is 342 Ω, whilst the effective resistance value for matching purposes is 677 Ω. If this difference is not appreciated, the anode/output components selected will cause the valve to operate on quite a different load line from that required and the performance will suffer in a number of respects, even though the circuits are in tune.

The second matter is concerned with the effect on performance of the anode circuit being off-tune. In this case the load line is elliptical instead of in a straight line, the width of the ellipse representing the departure from tune. Nobody would be likely to set up an amplifier in this condition, but the effect of weather conditions on antennas and feeders can cause a considerable departure from tune. Many modern high-power transmitters are fitted with equipment which automatically corrects tuning and loading under these conditions, in order to maintain the optimum performance with mismatched feeders. However, a high proportion of h.f. transmitters are not fitted with automatic tuning facilities, so the effect on valve loading is worth noting. Allowance must also be made to cater for the effects of mismatch in new designs.

Elliptical load lines

In general, a change in feeder or antenna impedance will alter the mean slope of the load line, as well as producing a reactive component across the valve. Whether the slope of the load line increases or decreases, depends on the distance of the mismatch from the valve anode, in terms of wavelength at the operating frequency. It follows that a change of frequency will also alter the valve conditions, due to the 'wavelength' distance changing.

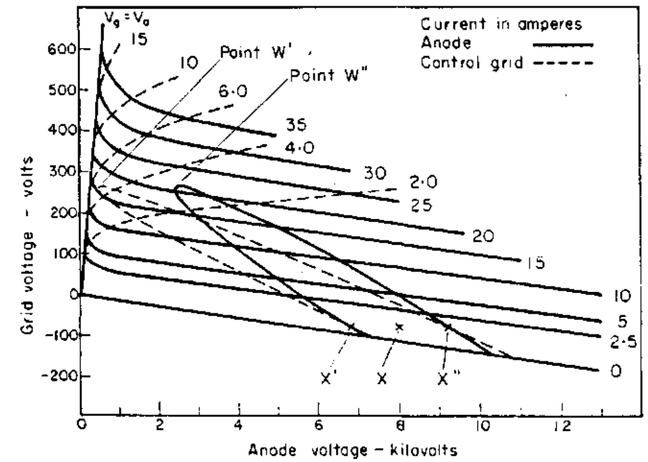


FIG. 2.4 Elliptical load lines on characteristics of triode type BR 1161.

In order to show the effects of mismatch, two elliptical load lines have been drawn on triode characteristics in Fig. 2.4. The characteristics are the same as those shown in Fig. 2.1. Both elliptical plots represent changes from load line *WX* of this figure, produced by a voltage standing wave ratio (v.s.w.r.) of 1·5 on the feeder. Load line *X′ W′ X″* shows a slope decrease and *X′ W″ X″* shows a slope increase.

Obviously the performance computor cannot be used directly, due to the load line curvature, but the grid voltages recorded at 15° intervals for load line *WX* can be used to determine the anode current on both sides of each ellipse. The formulae to calculate the performance from the derived list of anode currents must also be modified. If the current levels from *X′* to *W′* are indicated by *B′*, *C′*, *D′*, etc., and those from *W′* to *X″* are indicated by *B″*, *C″*, *D″*, etc., the modified formulae are given below:

Anode current, d.c. meter reading =

$$\frac{(A+B'+B''+C'+C''+D'+D''+E'+E''+F'+F'')}{24}$$

Grid current, d.c. meter reading; as for anode current but using the grid current characteristics.

Peak fundamental r.f. =

$$\frac{\{2A+1{\cdot}93(B'+B'')+1{\cdot}73(C'+C'')+1{\cdot}41(D'+D'')+(E'+E'')+0{\cdot}52(F'+F'')\}}{24}$$

Output power and driving power are determined as with the original formulae.

Using these formulae, the result of operating on load lines $X'W'X''$ and $X'W''X''$ are tabulated in Table 2.2, together with those for the original load line WX, to show the significant differences.

TABLE 2.2

Triode	Load line WX	Load line $X'W'X''$	Load line $X'W''X''$
h.t. voltage, d.c.	8 kV	8 kV	8 kV
Anode voltage trough	1·5 kV	0·5 kV	2·5 kV
Peak r.f. voltage	6·5 kV	7·5 kV	5·5 kV
Peak of anode current	20 A	17·2 A	21 A
Static anode current	1·0 A	1·0 A	1·0 A
Anode current excursion	19 A	16·2 A	20 A
Peak fundamental r.f. current	9·6 A	8·54 A	10·1 A
Anode current, d.c.	6·0 A	5·38 A	6·35 A
Power input, d.c.	48 kW	43 kW	50·8 kW
Power output, r.f.	31·2 kW	32 kW	27·8 kW
Anode dissipation—neglecting harmonic power	16·8 kW	11 kW	23 kW
Power conversion efficiency, d.c. to r.f.	65%	74%	54·7%
Grid current, d.c.	0·7 A	0·94 A	0·67 A
Peak r.f. grid voltage	350 V	350 V	350 V
Drive power required, valve only	245 W	330 W	235 W

The main differences from a power viewpoint are the changes in anode dissipation and efficiency, the change in power output being of a lower order. The significant factor is the increase in anode loss in the heavier loaded conditions. This accounts for the fact that the valves used in r.f. high-power class B amplifiers, which must cater for mismatched conditions, so often appear to be well underrated in normal operation. In the lightly loaded condition, the required increase in drive power could cause significant deterioration in performance if the drive stage was capable of supplying just sufficient power under normal conditions.

The anode current waveforms shown in Fig. 2.5 correspond to the conditions given in Table 2.2. From these waveforms the relative distortion is immediately apparent, even without an analysis for harmonic content. In the lightly loaded condition the distortion is partly due to the ellipse and partly to running into the non-linear region of the characteristics. In the heavily loaded condition, the

distorting effect of the ellipse is reduced by the improvement obtained by operating on a load line having an increased slope.

These waveforms and calculations are based on the assumption that the grid voltage remains constant in the three conditions. In practice, unless the drive power available is excessive, there will be some rounding at the peak of the grid voltage waveform, due to the rapidly increasing grid current with voltage amplitude. The greater the grid current, the greater will be the reduction in

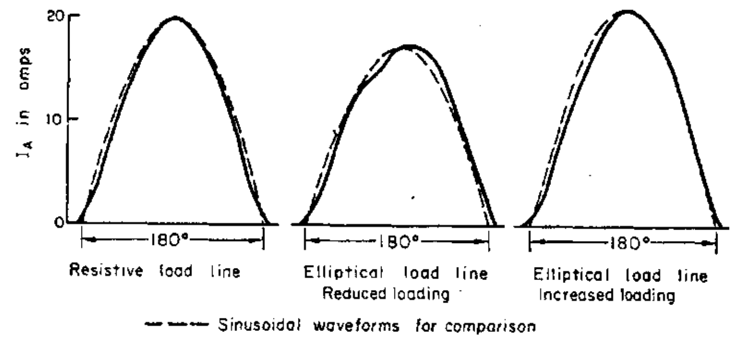


FIG. 2.5 Effect of feeder mismatch on anode current waveform of triode.

peak voltage. The effect of a slightly rounded waveform at the grid will be a slight rounding at the peak of the anode current waveform. Referring to Fig. 2.5, it will be seen that a degree of rounding will make the anode current waveform more nearly sinusoidal, in both the normal and more heavily loaded conditions, reducing distortion. In the lightly loaded condition, the greater demand on drive power will cause the anode current waveform to depart further from a sinewave and so increased distortion will result.

Operation at reduced anode excursion

Linear amplifiers are normally used for multi-channel operation, during which the mean power output is considerably less than the peak power. It follows that the mean input power under multi-signal conditions is less than that required for a single r.f. signal giving full anode excursion. Advantage is often taken of this fact to economize in the design of the d.c. power supply. In these cases, the power will not be adequate for a single-channel constant-level signal, such as frequency shift keying (f.s.k.), so the anode swing is reduced by reducing the drive level.

The power output and anode dissipation obtained at various percentages of the total anode excursion is shown in Fig. 2.6 plotted from load line WX on Fig. 2.1. It will be noted that the anode dissipation increases with reductions from full anode swing, rising to maximum dissipation in the region of 80% of the anode swing for full p.e.p. output. Typically, an economically designed d.c. power supply for 30 kW peak power, would be adequate to supply an output of 20 kW on f.s.k. It can be seen from Fig. 2.6 that the anode dissipation is a maximum at 20 kW, about 8% higher than it would be at 30 kW continuous output.

Adding this fact to the increased anode dissipation likely to be encountered due to feeder mismatch, further emphasizes the necessity of using a valve with an anode dissipation appreciably greater than that calculated for a continuous signal at full excursion on a straight load line. In the case of the specimen load line, *WX* on Fig. 2.1, the anode dissipation of 16·8 kW calculated for a steady signal output of 31·2 kW, would increase to about 25 kW, if operated at 20 kW with a mismatched feeder.

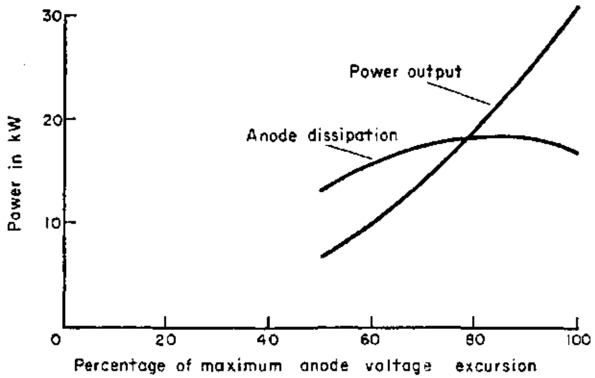


FIG. 2.6 Anode dissipation and power output in relation to anode voltage excursion.

2.5 CHARACTERISTICS OF POWER TETRODES

The effect of a screen between the anode and grid of a valve, changes the characteristics in several ways. The two main results are an increase in power gain and a considerable reduction in the interelectrode capacitance between anode and grid. This means that tetrodes can be operated in grounded cathode circuits without the need for neutralizing, and can give considerable output power for very small drive levels. In fact, for linear amplification, the tetrode valve can be fully exploited without running into grid current, so the power required from the drive is only that necessary to overcome the losses of the input circuit. Additionally, grid current is a major factor in causing non-linearity, so its avoidance is particularly of value in linear amplification.

Another feature of tetrodes is that the characteristics can be optimized for any particular application by suitably adjusting the screen supply voltage. Also, for amplitude modulation, the screen supply can be modulated with considerably less power than that required to modulate the anode supply.

Over the normal operating portions of the characteristics, tetrodes are not quite so linear as triodes, but non-linearity increases rapidly as the anode voltage approaches that of the screen supply, limiting the trough of anode voltage at peak anode current. The conversion efficiency of tetrodes tends to be less than that of triodes.

As the screen is effectively at cathode r.f. potential in grounded cathode circuits, the input and output capacitance of tetrodes is higher than that of

similarly rated triodes. With power tetrodes the higher capacitance can be an embarrassment when designing circuit components for the high-frequency end of the operating range.

It is important that the screen supply is not applied to the valve in the absence of the anode supply, and interlocks must be provided to prevent this happening, either when running up or during anode trips. In making triode/tetrode decisions on economic grounds, the total cost of the screen supply and associated interlocking arrangements must be considered in relation to the cost of the drive and neutralizing requirements.

Typical linear operating conditions

In order to make a direct comparison with triode operating conditions, explanatory examples are given for a tetrode having the same anode dissipation as the triode exemplified in Section 2.4. Typical characteristics for screen supplies of 1·0 kV and 1·5 kV are shown in Figs 2.7(*a*) and 2.7(*b*), respectively.

Considering linear operation and referring to Fig. 2.7(*a*) load line *KL* has the same anode current and voltage excursions as those on load line *YZ*, Fig. 2.1. Because of the obvious non-linearity at peak current in the region of 15 A, the anode voltage trough is limited to 2 kV, so the d.c. anode supply is increased from 10·5 kV to 11 kV, in order to obtain the same voltage excursion.

The change in characteristics and greater capability of this valve, produced by increasing the screen voltage from 1·0 kV to 1·5 kV, is shown by comparing Fig. 2.7(*b*) with Fig. 2.7(*a*). By operating the screen at 1·5 kV, with a d.c. anode voltage of 20 kV, it is possible to obtain an output of more than 100 kW in a linear condition, at a conversion efficiency of 70% and without grid current. In this condition the anode dissipation is at the upper limit of 35 kW, with no allowance for differences in the characteristics of individual valves, or for operating tolerances required to cover cases of feeder mismatch.

Apart from the anode dissipation limit of such operation, the high level of d.c. voltage has disadvantages, both technical and economic. From a technical aspect, higher d.c. voltages mean greater voltage clearances and higher losses in the anode circuit. As the r.f. anode voltage is approximately proportional to the d.c. voltage, at a given frequency the r.f. current in the anode circuit is also proportional to the d.c. voltage; but the losses in the anode circuit are proportional to the square of the r.f. current, so the circuit losses increase approximately as the square of the d.c. voltage increase.

The economic disadvantage is more obvious, particularly in the case of solid-state rectifiers, where the number of rectifier units depends on the voltage, so the cost is roughly proportional to the total d.c. voltage. Current rating is not likely to have an appreciable effect on cost.

With these considerations in mind, typical operating load lines have been marked on Fig. 2.7(*b*), for a screen voltage of 1·5 kV. The result of using load line *MN* is designed to show the greater output available, when compared with operating at a screen voltage of 1·0 kV on load line *KL* Fig. 2.7(*a*). Load line *MP* exemplifies the linearity improvement obtained by increasing the static anode feed from 1·0 A to 2·0 A.

Using the tube performance computer described in Section 2.4, the results obtainable by operating this tetrode on load lines *KL*, *MN* and *MP* are tabulated in Table 2.3, in order to make a direct comparison between the three conditions.

Comparing the results obtained by operating the tetrode on load line *KL*, Table 2.3, with those given for a triode on load line *YZ*, Table 2.1, shows remarkably little difference. The slightly better conversion efficiency of the triode is offset by a lower grid excursion and zero driving power for the tetrode itself.

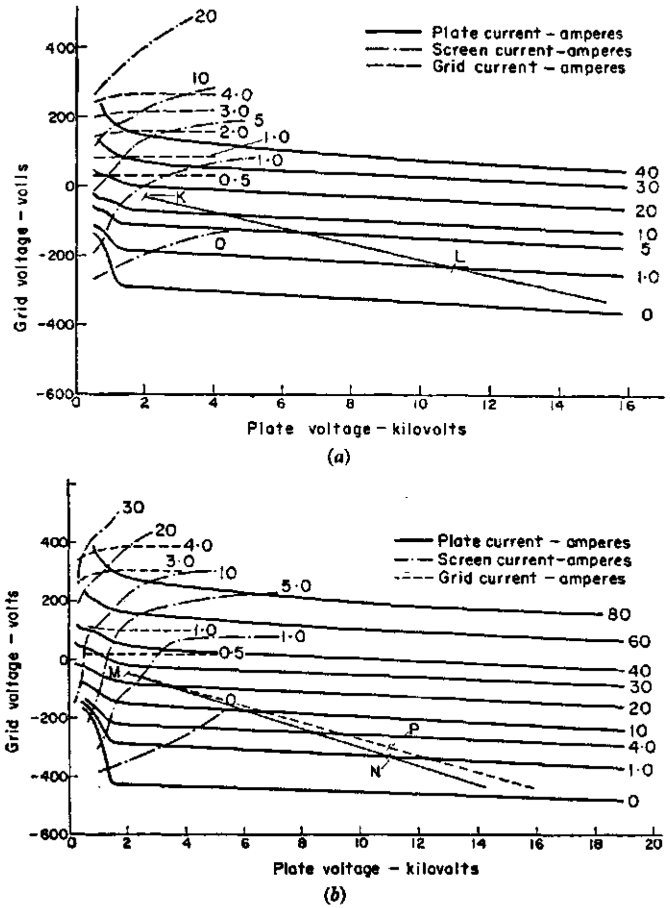


FIG. 2.7 Load lines on constant-current characteristics of tetrode valve type 4C X 35 000 C (Eimac). (*a*) Screen voltage = 1000 V. (*b*) Screen voltage = 1500 V.

Zero drive power is not realizable in practice, due to the losses in the input circuit, which are higher in the case of tetrodes of high input capacitance. The average input capacitance of this tetrode is 465 pF, which has a reactance of 14 ohms at 25 MHz. With a peak voltage drive of 200 V peak, the r.m.s. current is 10 A, giving a V_A of 1400. Assuming a Q factor of 350 for the input tuned circuit, the losses will be 4 W at 25 MHz. As the power loss varies with frequency,

TABLE 2.3

Tetrode	Load line *KL*	Load line *MN*	Load line *MP*
h.t. voltage	11 kV	11 kV	11 kV
Screen voltage	1·0 kV	1·5 kV	1·5 kV
Anode voltage trough	2·0 kV	2·0 kV	2·0 kV
Peak r.f. voltage	9·0 kV	9·0 kV	9·0 kV
Peak of anode current	15 A	25 A	25 A
Static anode current	1·0 A	1·0 A	2·0 A
Anode current excursion	14 A	24 A	23 A
Peak fundamental r.f. current	6·88 A	11·45 A	11·83 A
Anode current, d.c.	4·2 A	7·0 A	7·34 A
Anode power input, d.c.	46·2 kW	77 kW	80·74 kW
Power output, r.f.	30·95 kW	51·5 kW	53·24 kW
(Fundamental r.f. output power of initial assessment, for comparison)	(31·5 kW)	(54 kW)	(51·75 kW)
Anode dissipation—neglecting harmonic power	15·25 kW	25·5 kW	27·5 kW
Power conversion efficiency, d.c. to r.f.	67%	67%	66%
Mean screen current	180 mA	210 mA	225 mA
Screen dissipation	180 W	315 W	338 W
Peak r.f. driving voltage	200 V	280 V	250 V
Drive power required, valve only	nil	nil	nil

the loading on the driving amplifier will also vary with frequency. This is not a desirable feature, so it is usual to swamp the loss variation by means of a fixed resistor across the input circuit. In the case reviewed, a resistor of 1000 Ω would be a typical value, thereby limiting the total drive power variation of 21–24 W

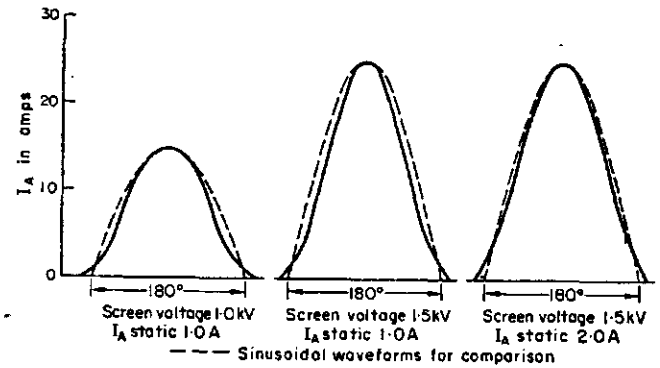


FIG. 2.8 Effect of changing screen voltage and static anode feed on anode current waveform of tetrode.

over the h.f. range. This is still considerably lower than the drive power of 93 W required for a triode under similar conditions.

The results obtained from load line *MN*, Fig. 2.7(*b*) and *KL*, Fig. 2.7(*a*), show the greater output obtainable by increasing the screen voltage from 1·0 kV to 1·5 kV.

Load lines *MN* and *MP* represent typical operating conditions, bearing in mind the probable increase in anode dissipation which may be encountered in service. Increasing the static anode feed from 1·0 A to 2·0 A, load lines *MN* and *MP*, respectively, gives an increase of 3·3% in output power, for a 1% decrease in conversion efficiency and an increase of 8% in anode dissipation. However, as linear operation is being considered, the most useful effect of increasing the static feed is the improvement in linearity. This is shown diagrammatically in Fig. 2.8, in which the waveform of the anode current is plotted for two static feed conditions, compared with half sinewaves. The anode waveform at 1·0 kV screen supply also indicates that linearity would be improved by increasing the static feed above 1·0 A.

REFERENCES

[1] TERMAN, F. E. *Radio Engineering*. McGraw-Hill, New York (1937).

[2] SANDEMAN, E. K. *Radio Engineering*, Vol. 1. Chapman and Hall, London (1947).

[3] GRAY, L. F. and GRAHAM, R. *Radio Transmitters*. McGraw-Hill, New York (1961).

[4] TYLER, V. J.: 'A new high-efficiency high-power amplifier'. *Marconi Rev.* **21**, 3rd quarter (1958).

[5] 'Tube performance computor'. Eitel McCullough Inc. application bulletin No. 5.

3

Application and Type of Operation

3.1 GENERAL CONSIDERATIONS

There are a number of factors influencing the selection of the most appropriate circuit configuration for a particular application, such as frequency band to be covered, valve type, balanced or unbalanced feeders, etc. If the frequency is in the v.l.f., l.f. or m.f. bands, groundwave propagation will be used and the transmitter will operate on a single assigned frequency when in service. On the other hand, transmissions in the h.f. band employ skywave propagation, with the accompanying need for a number of frequencies to maintain the service. An exception to this general rule is the use of h.f. and groundwave propagation for local broadcast transmissions in the tropics.

For fixed-frequency transmitters, the majority of components can be of fixed value with a small number of variable components of only sufficient value to be used as trimmers. The result is that the choice of circuit configuration is often governed by the most simple and inexpensive method of construction, rather than by technical elegance. At an early stage in design, development effort can be saved by considering adequate component values to cater for any frequency likely to be used in one band, but frequency changing is not an operational requirement. The operating frequency is nominally set up by the manufacturers, prior to installation. This means that the fixed-value components in any one transmitter can be limited to those required for one frequency and the value of variable components can be reduced by providing pre-set tapping points on inductors, with an arrangement of links for fixed capacitors. The adoption of these methods of construction enables the cost of a fixed-frequency transmitter to be kept at a minimum.

For transmitters operating in the h.f. band, the type of service is an important factor in determining the arrangements needed for changing frequency, and hence the circuit configuration. For broadcasting applications, transmissions are permitted only in limited sub-bands of the h.f. spectrum, so only that equipment required to cover these sub-bands need be supplied; complete frequency coverage is not necessary. To be effective, the times of transmission on specific frequencies must be pre-programmed, to enable listeners to have the information beforehand. Also, most broadcasting organizations allow a certain time for changing frequency between programmes. While changes of frequency should be of relatively short duration, they need not necessarily be rapid. Where rapid changes are deemed to be necessary, by adding a degree of mechanical complexity, they can be made by switching between pre-set component values.

In the h.f. band, communication transmissions are also *normally* limited to certain sub-bands of the spectrum, but in two major respects the transmitter requirements differ from those of broadcasting. ('Normally' is used advisedly, for military organizations are liable to use any convenient frequency in wartime, regardless of peacetime regulations, so military transmitters should cover any frequency within the h.f. spectrum.) In other respects the requirements of communication transmitters for civil use are the same as those for military purposes. It would be uneconomical to design different transmitters for the two types of user, and so it is usual for all communication transmitters to be able to radiate on any frequency in the h.f. band.

While a degree of frequency/time planning is employed in r.f. communications, particularly on the main traffic circuits, it is quite practical, and not uncommon, to change frequency at short notice. Also, a change of route often accompanies a frequency change, so the new frequency required can be in any of the sub-bands. For civil organizations, loss of traffic time means loss of revenue, especially when multi-channel operation is in progress. The time taken to change frequency should be as short as possible, and rapid frequency changing is an economic necessity.

Loss of traffic time is probably most important to military organizations where frequency changes are more spasmodic, and delayed messages could be calamitous in dealing with high-speed aircraft and military vessels. So again, rapid frequency changing is a necessity to reduce message delays to a minimum. This also applies to civil aircraft.

From the foregoing considerations it is clear that the circuit configuration depends both on the operational frequency band and the type of service within that band.

Valve type and class of operation are so closely allied to circuit configuration that it is convenient to classify application in terms of class A, B, C and D operation.

3.2 CLASS A APPLICATIONS

Class A amplifiers are not normally considered for high-power applications, owing to their low conversion efficiency, but there is a v.l.f. application where the final amplifier conversion efficiency is not the most important consideration.

It is pointed out in Chapter 1, Section 1.1, that some form of antenna damping may be necessary at v.l.f. to increase bandwidth, by reducing the antenna Q factor, which can be of the order of 300. It can be shown that the damping provided by the low anode impedance of a final amplifier operating in class A, can give an effective improvement in bandwidth. By using this method, the damping losses are dissipated at the valve anode and no additional cooling equipment is required. This is a more simple and less costly arrangement than first generating the r.f. power more efficiently in a class B stage, subsequently dissipating it in an external resistor requiring additional cooling arrangements. Class A stages also require less driving power than class B stages.

Although this is a desirable application, it is also very limited. The amount of damping provided is not controllable and the antenna Q factor reduction may be required only at the low-frequency end of the band. Consequently it is more general to use class B or class AB, and to associate the required damping with that part of the inductor used for the lower frequencies only. The large surface area of this portion of the inductor can ensure that the conductor temperature rise is not great and the cooling requirement for considerable power dissipation need be little more than that provided by convection. By these means the damping is more controllable, the equipment required for the extra cooling is negligible and advantage can be taken of the higher conversion efficiency at the upper end of the band.

3.3 CLASS B APPLICATIONS

The main purpose of using valves in class B for power amplifiers is to provide the linear amplification necessary for most communication circuits. Linear amplification is not necessary for plain c.w., single-channel f.s.k. or multi-channel telegraphy using t.d.m., but for most other types of multi-channel traffic it is essential.

For two major reasons, the importance of linearity cannot be overstressed in connection with communication transmitters. The first concerns multi-channel traffic on the system being operated. The distorting effects of non-linearity cause undesirable cross-talk between channels, which can result in a complete loss of intelligence under poor conditions. The second is that the resulting intermodulation products increase the bandwidth, thereby producing out-of-band radiation and interference with other communication circuits.

Operationally it is usual for transmitters to be set up in a linear condition all the time, leaving the type of signal, number of channels in use and the level per channel in the hands of the traffic controller. With the advent of automatic tuning, combined with remote control of transmitters and antenna selection, it is becoming more general for the entire operation of transmitters to be also in the hands of the traffic controller. This is much more efficient operationally and enables transmitting stations to be manned by maintenance staff only. Not only is it an economic advantage in reducing the number of staff required, but it overcomes the difficulty that many user organizations have in being able to obtain an adequate number of trained staff at transmitting stations, due to the sites being in relatively remote areas.

All automatic systems are inevitably accompanied by complexity. The greater the complexity the greater the liability of something going wrong, and greater skill is required for rapid diagnosis and fault correction. It follows that for successful automatic control, both the transmitters and the control system itself must be as simple and reliable as possible. In the present context of high-power amplifiers for communication purposes, the circuit configuration must be as simple as possible, consistent with the capability of being controlled and tuned automatically over the complete h.f. band, without manual intervention.

With simplicity and reliability as the keynotes, the number of circuits to be tuned, or adjustments made for different frequencies, must be limited to those absolutely essential to obtain the required performance. In this respect it is preferable to have the same setting-up procedure as regards operational levels, allowing for sufficient power loss at the highest frequencies and obtaining rather more output at lower frequencies. As the circuits depend on the type of valve used, the relative merits of tetrodes and triodes for linear amplification are considered first.

Choice of valve for linear amplifiers

The relative merits of tetrodes and triodes are considered on the assumption that both types are available at the power level required, but for power levels up to about 10 kW it is probable that triodes are not available, so the use of a tetrode is inevitable. There are also a few power pentodes, but they tend to be available for lower-power applications only, and in any case the circuits are similar to those required for tetrodes.

The method of setting up both tetrodes and triodes for linear operation in class B, based on typical characteristics, has been given in Chapter 2. It was not pointed out that in r.f. applications the energy-storage capacity of the tuned anode circuit (flywheel effect) is sufficient to permit linear operation with single valves. This is distinct from the push–pull arrangement necessary at audio frequencies.

Linear amplification with tetrodes

For single valves in grounded-cathode circuits, stable operation of tetrodes is quite practicable without neutralization, due to the isolating effect of the screen between anode and grid. Even so, with the very high power gain of tetrodes, some positive feedback between anode and grid is usually present, due either to the residual interelectrode capacitance or to incomplete isolation between the external circuits. Unless this positive feedback is cancelled by neutralization, it will not be possible to obtain the optimum linear performance. The level of the neutralizing signal required is quite small and not at all critical. In consequence the associated circuit can be sufficiently wideband to cover the whole h.f. spectrum, without the need for adjustment during frequency changes. It is quite practical for tetrodes to be used in class B for linear amplifiers in grounded-cathode circuits, without the complication of controls for neutralization. In this way tetrodes can be used for high-power output levels with very small drive levels, without grid current.

Linear amplification with triodes

On the other hand, triodes must be neutralized when used in grounded-cathode circuits. With push–pull configurations neutralizing is relatively simple, although it is unlikely that a single setting will be adequate over the h.f. band, and adjustments will be required for changes of frequency. With single valve stages, additional circuits are necessary, adding to the complexity. Even the simplest form of neutralizing by means of a split grid circuit, requires an increase in drive level. Bearing in mind the need for rapid frequency changing, neutralizing by itself means that grounded-cathode circuits are not recommended for triodes in the h.f. band, but there is an additional disadvantage.

To obtain high-power outputs from triodes, they must be operated in the positive grid region, with accompanying grid current, and grid current is a major cause of distortion in linear amplifiers. The effect of grid current on linearity is twofold. First, as the grid voltage excursion increases positively, the rise in grid current is very rapid, particularly when the trough of anode voltage approaches the level of the grid voltage, i.e., when operating as efficiently as possible within linear limits. Unless the drive power available from a low-power source is considerably in excess of that required to drive the valve, the grid voltage peaks will be depressed, with resultant amplitude distortion. The second

effect is not so obvious. The rapid change of grid current with positive grid voltage means that the effective input resistance of the valve changes rapidly over the positive region of the driving cycle; but the input capacitative reactance remains constant at any one frequency. A change of phase occurs during each half cycle, causing a phase change in the modulation envelope, and producing a phase distortion which is just as detrimental to linearity as amplitude distortion.

The alternative arrangement of grounded-grid connection is more suitable for triodes as linear amplifiers in the h.f. range. Neutralizing is not normally required, because the grounded grid behaves as an effective screen between anode and cathode. As in the case of tetrodes, there will be occasions when a small amount of neutralizing will be needed to counteract unwanted positive feedback, but again the circuit can be made sufficiently wideband to cover the whole h.f. band at one setting.

On the score of linearity, the performance of triodes in grounded-grid configuration is very good. The feedback effect of the circuit improves the inherently good linearity of the valves themselves. On the input side, the anode–cathode load presents a substantially constant resistance over the driving half cycle, of a much lower effective value than that caused by grid current.

The variation of grid current is swamped by the steady load and the effects of amplitude and phase distortion are reduced to a low level. The actual distortion level depends on the ratio of peak anode current to peak grid current, which depends both on the type of valve and the relation between anode voltage trough and peak grid voltage. This means that if there is more than one triode of suitable rating for the application being considered, the type having the lower grid current requirement should be selected. Also, as the slope of the load line affects both the peak grid current and the anode voltage trough, for better linearity the slope of the operating load line should be steeper than that which would give optimum efficiency.

It is obvious that the r.f. power required to drive grounded-grid triodes is much greater than that needed by the same valves in grounded-cathode circuits, but the additional power is not lost. The drive power into the anode–cathode load is in series with the anode output power and so appears as part of the amplifier output. This accounts for the anode–cathode drive power being known as 'throughput', and it is interesting to note that this arrangement was known originally as a series amplifier.

The high level of power required to drive grounded-grid triodes compares very unfavourably with tetrode drive power, but in general, triodes enable a better linear performance to be obtained more readily, due to the inherent feedback with grounded-grid circuits. Apart from the probability of a slightly inferior linear performance, the main disadvantage of tetrodes is that an additional d.c. power supply must be provided for the screen, with associated interlocking circuits. In deciding which valve type to use, all these factors must be considered very carefully from both technical and economic aspects.

Methods of improving tetrode linearity

Probably the most effective method of obtaining better linearity from tetrodes is to use a grounded-grid configuration. As well as giving improved performance, neutralizing is quite unnecessary. However, both a high drive level and a d.c. screen supply has to be provided, so it is a costly arrangement. It is not commonly used and is not readily applicable as a means of changing from an existing grounded-cathode circuit.

There is another method which is simple to apply to existing amplifiers and gives quite an improvement in linearity. In principle, correction is applied to the audio-frequency envelope by means of a resistance in the cathode circuit, which is effectively in circuit at audio frequencies only. It is sometimes called an 'i.p. improver'.

The circuit diagram is shown in Fig. 3.1. Capacitor C_1, typically 0·01 μf, provides a low-impedance by-pass to earth for r.f. The choke L_1 has an inductance of about 1·0 Hy and a d.c. resistance of 1·0 Ω offering a high impedance to audio frequencies and a low impedance to d.c. (a choke of high d.c. resistance would

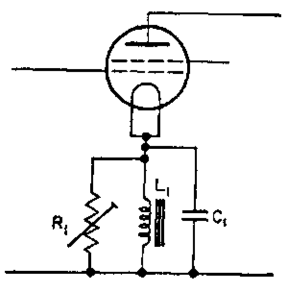


FIG. 3.1 Circuit for improving tetrode linearity.

ruin linearity by causing changes in grid bias, in sympathy with the audio envelope). The resistance value of R_1 determines the level of applied correction, usually of between 20 Ω and 100 Ω, and is adjusted for optimum performance. It may be necessary to change the value of the resistor over the h.f. band, but the value is not critical, and as only audio frequencies are involved simple switched resistors are quite satisfactory.

This circuit should not be considered as a cure for all forms of distortion, but should only be applied after obtaining the best performance by conventional methods. Then the improvement possible can make all the difference in i.p. level being a few decibels worse and a few decibels better than a specification.

Linear power amplifiers for broadcasting

Double sideband transmitters with a carrier power of about 10 kW are used for local broadcasting in both the m.f. and h.f. bands. The h.f. band is used in tropical regions where the atmospheric noise is less than at m.f. As groundwave propagation is used in both frequency bands, rapid frequency changing is not a requirement.

Modern transmitters for these services employ grid modulated high-power tetrodes for the final amplifier, which must operate in a linear condition. Adequate peak power output can be obtained without grid current, so both the r.f. drive and modulator are low-power units. In fact, these power levels are determined by the value of the resistor, which is normally fitted across the input to the final stage in order to provide a constant load for both drive and modulator. Typically,

the drive power required is about 10 W, so both drive and modulator can be solid-state devices. This means that the only valve in the transmitter is the output amplifier, and as it is operated under conditions well below maximum ratings a long life can be expected. Although the cost of the single valve is high, this is a very economical type of transmitter as regards initial cost and running expenses. It is also a transmitter of high performance.

The characteristics of a tetrode suitable for being grid modulated and giving a carrier power of 10 kW, are shown in Fig. 3.2. In order to obtain the right setting-up conditions, a load line is first selected to give a peak power of 40 kW as a linear

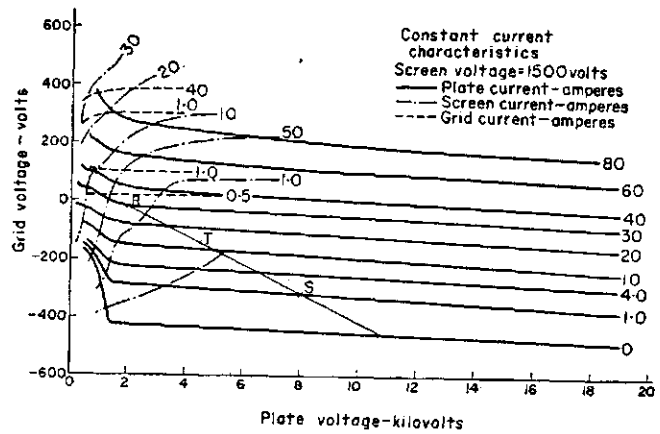


FIG. 3.2 Load line on constant-current characteristics of tetrode valve for low-level modulation.

amplifier. From this load line, the operating conditions are determined by using the tube performance computer described in Chapter 2, Section 2.4. The operating conditions obtained by operating on load line RS, Fig. 3.2, are given in Table 3.1.

There are several points to be noted from the results given in Table 3.1.

(*a*) The anode current does not change when modulation is applied, so the anode dissipation decreases and the efficiency increases. Hence grid modulation is a type of efficiency modulation.

(*b*) The carrier power is determined by selecting point T on the load line, at such a position that about 10 kW output is obtained on section TS of the load line.

(*c*) The difference in grid voltage between points T and R on the load line gives the positive grid excursion required to give a peak of 100% modulation, hence the drive power required.

(*d*) With the grid voltage determined from (*c*) −120 V in this case, it will be seen that the negative peak of grid voltage will not cut the valve off. The trough of modulation is not quite 100%. This is, of course, due to the curvature of the valve characteristics in the low anode current region.

TABLE 3.1

(Anode voltage: 8 kV; screen voltage: 1·5 kV; grid bias: –320 V; static anode current: 1·0 A)

	Carrier	100% Modulation
Peak anode current	13 A	30 A
Anode voltage trough	4·5 kV	2 kV
Anode voltage excursion	3·5 kV	6 kV
Peak r.f. power output		41·2 kW
Mean r.f. power output	10·4 kW	16·2 kW
Anode current, d.c.	3·72 A	3·72 A
Power input, d.c.	29·8 kW	29·8 kW
Anode dissipation	19·4 kW	13·6 kW
Conversion efficiency	35%	54·3%
Grid voltage peak	–140 V	–20 V
Grid voltage excursion		120 V relative to –140 V

(*e*) Because of the non-linearity there will inevitably be some distortion at 100% modulation, so it is normal practice to specify the distortion at either 90% or 95% modulation. This is a common feature of specifications for all methods of producing d.s.b. signals.

(*f*) As these calculations are based on typical characteristics, it will be appreciated that minor adjustments will be required to obtain the optimum performance from individual valves, mainly in relation to the levels of the r.f. drive modulating voltage.

Before concluding this section on low-level modulation for d.s.b., it should be pointed out that modulating the grid voltage is only one method. It is equally effective, and in some instances more convenient, to apply modulation to the r.f. drive. But in the context of the final stage valve the conditions are no different.

3.4 CLASS C APPLICATIONS

Transmitters for the m.f. and h.f. bands

The most usual application for class C amplifiers is d.s.b. transmission for broadcasting in the m.f. and h.f. bands. In both bands a typical carrier power level is 100 kW, although there is an increasing tendency to increase the power in the m.f. band to 500 kW, or even 1000 kW. Thus there is an obvious advantage in using class C to obtain a high d.c. to r.f. conversion efficiency, especially in cases where very high-power m.f. transmitters are installed in remote areas, or on islands, and the power supply has to be generated locally.

The main difference between transmitters for the two frequency bands, apart from component values, is that m.f. transmitters normally operate on one frequency only, whilst operational frequency changing is an essential feature of h.f. transmitters. Although it is usual to design the transmitters for both purposes with sufficient components to cover their respective frequency bands, the mechanical arrangement must be governed by the ease with which the frequency can be changed in the h.f. case.

The choice of valve type for class C amplifiers

A resumé of the valves used for class C high-power amplifiers since h.f. broadcasting began is appropriate as a means of showing how valves have developed. Early h.f. transmitters had a power output of about 15 kW, from a pair of triodes in a grounded-cathode push-pull circuit with cross neutralization. The whole circuits were well designed, but the valves were the same types as used for m.f. applications, with a single lead connection for the grid and having long internal connections. The single grid connection was inadequate to carry the higher r.f. current at h.f., so local overheating occurred, resulting in the glass envelop cracking and a high mortality rate among valves. This was overcome by making valves with a ring seal for the grid connection.

The long internal connections necessitated long external connections, the combined effect being that the setting of the neutralizing capacitors had to be changed considerably over the frequency range. At frequencies well above the operating band, the combined reactance of the neutralizing arms became inductive instead of capacitative, providing a ready-made circuit for spurious oscillation. To counteract the tendency to oscillate, it was not unusual to find high-power amplifiers bristling with anti-squegger devices.

A big advance towards amplifier stability was the advent of the grounded-grid connection in the 1930s. This arrangement required much more drive power when compared with the grounded-cathode arrangement, but not as much as might be expected. In order to achieve 100% modulation, the throughput power had to be reduced at the envelope trough by modulating the driving stage. Although some complication was caused, it had the compensating effect of increasing the drive level at modulation peaks.

The next step forward was the production of triodes of almost squat construction, in which the inductance of the internal connections was kept quite low by making them short and wide. This type was developed mainly for use at v.h.f. and was quite effective in reducing the tendency of neutralized triodes to oscillate in a grounded-cathode circuit.

In recent years, the development of valve techniques has made it possible to produce tetrodes for very high power. The use of high-power tetrodes, operating class C in a grounded-cathode circuit, combines the advantages of low drive power, high stage gain and stable operation, without the need for neutralizing, thereby simplifying frequency changing. In order to achieve anode modulation at 100% it does mean that the screen must also be modulated, but the screen-modulating power required is relatively low. As a result of this development, it is almost universal to use tetrodes for the final modulated amplifiers of high-power broadcast transmitters.

Setting up tetrodes in class C for anode-modulated r.f. amplifiers

There are several factors to be taken into account before selecting the most suitable load line on which to base the performance calculations for an anode-modulated r.f. amplifier in class C.

(*a*) Although the d.c. anode current remains constant when the anode is modulated at 100% modulation, the total input power to the valve (d.c. + audio) is increased by 50%, so the anode dissipation increases by 50%. In the carrier condition, the anode dissipation must not exceed 66% of the maximum permissible. It might be argued that a steady modulation at 100% is not a normal operating condition, but it is quite usual for a customer to check the performance at 100% modulation, by specifying a load run at this level for 10–15 min.

(*b*) In order to obtain high efficiency, the anode voltage excursion should be high, but grid current increases rapidly as the anode voltage trough approaches the grid voltage level. Hence the grid dissipation imposes a limit on the efficiency attainable.

(*c*) When triodes are anode modulated, the grid must be overdriven in the carrier condition in order that the drive level will be adequate to sustain the peak anode current at 100% modulation. Alternatively, the drive must be modulated. With tetrodes the same effect is achieved by modulating the screen, enabling the anode current peaks to be attained with the same drive level as that required for the 'carrier only' condition. To some extent this eases the grid dissipation limit mentioned earlier.

(*d*) It might appear that the limits imposed by grid dissipation on the one hand and adequate drive level on the other mean that the drive level must be adjusted within fine limits; a very undesirable condition. Fortunately, this can be overcome by the self-compensating action of automatic grid bias, obtained from the voltage set up across a resistor by the grid current. It follows that the compensation will be proportional to the level of bias produced automatically, but there must be some steady d.c. bias to prevent excessive anode dissipation in the event of drive failure.

(*e*) When calculating the drive power required by means of the tube performance calculator, the value obtained is that required for the valve alone. To obtain the total drive power required, the power dissipated in the grid resistor and bias circuit must be added, together with the grid circuit losses, which can be appreciable with class C because of the high level of r.f. grid voltage. Allowance must also be made for valves with characteristics which differ from the typical characteristics published. Bearing in mind these additional requirements, it is not unreasonable to allow a drive power of three times that required by the valve itself in order to avoid the embarrassment of being short of drive later.

(*f*) Limits are specified for maximum screen voltage and dissipation, so allowance must be made for the audio modulating voltage and screen dissipation at 100% modulation. The peak of audio voltage on the screen will be about 80% of the d.c. voltage, which means an increased screen dissipation of about 30% above that calculated for the carrier condition $(0{\cdot}8 \times 0{\cdot}707)^2 = 32\%$.

(*g*) The total r.f. plus d.c. voltage on the valve anode is another limiting factor, which means that the d.c. voltage must not be too high. This is an advantage in reducing anode circuit losses, for the r.f. losses increase as the square of the r.f. voltage increase.

(*h*) In selecting the bias voltage for the carrier condition, i.e., the lower-end of the load line, consideration must be given to the angular portion of the positive grid excursion during which anode current will flow. For a given peak anode current the valve conversion efficiency increases as the angle of current flow is reduced, but the power output falls. Therefore, within the capabilities of any

valve type, there must be a compromise between efficiency and output. As a general guide, the angle of current flow is usually chosen between 120° and 140°.

Typical operation for 100 kW carrier output

With due consideration to paragraphs (*a*) to (*h*), operating load line *MO* has been drawn on a set of tetrode characteristics in Fig. 3.3 to give a carrier output of 100 kW, suitable for 100% anode modulation.

A significant feature of class C operation is that the power output is more critically dependent on the slope of the load line than it is for class B operation. The output coupling arrangement must be capable of being adjusted within fine limits. However, as linear operation is not required, changes in loading caused

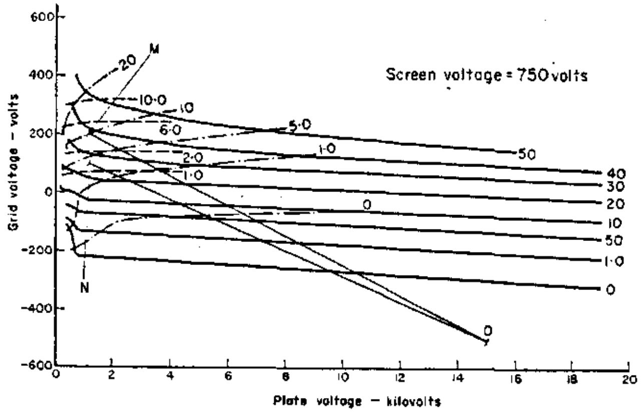


FIG. 3.3 Load lines on constant-current characteristics of tetrode valve for anode modulation.

by feeder impedance changes during programmes do not affect the performance as much as they do in class B amplifiers, even if the load line is somewhat elliptical.

For load line *MO*, the d.c. anode voltage is higher than might be expected from condition (*h*), but it is within the valve limits and has been chosen deliberately. The characteristics are for the same valve as used for the class B tetrode calculations, with the screen voltage reduced to 750 V, not because it is the only valve available, but to demonstrate the versatility of modern valves. This feature is particularly useful during experimental transmitter work, for it permits the same valves to be used for several applications, thereby reducing the expenditure on expensive valves.

Operating on load line *MO*, calculation by means of the tube performance computor gives a carrier power output of 103·5 kW at 83·7% conversion efficiency. In order to avoid duplication, the full performance figures are given in Section 3.6, where they are compared with those obtained for class D operation, for the same valve, on load line *NO*.

3.5 CLASS D APPLICATIONS

The principle of class D operation

Class D is the name given to class C operation with waveforms that are more rectangular than sinusoidal, as a means of obtaining improved conversion efficiency. The highest efficiency is obtained by the use of truly rectangular waveforms, but the r.f. voltages required by high-power amplifiers are not readily available in rectangular form. Worthwhile improvements can be obtained with

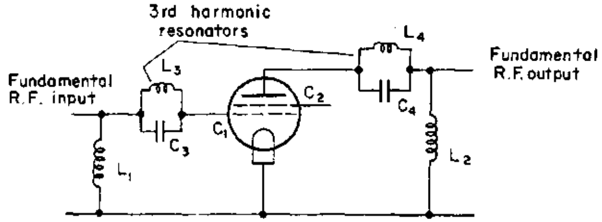


FIG. 3.4 Basic circuit arrangement for class D operation.

waveforms consisting of fundamental and third harmonic components in the right proportion and in the correct phase relationship.

The method of deriving the input waveform from the fundamental and of deriving the fundamental output from the resultant complex waveform at the anode, can best be described by reference to the simplified circuit diagram shown in Fig. 3.4.

A third harmonic resonator L_3C_2 is inserted as part of the input tuned circuit L_1C_1, resulting in a waveform at the tetrode grid of fundamental plus third

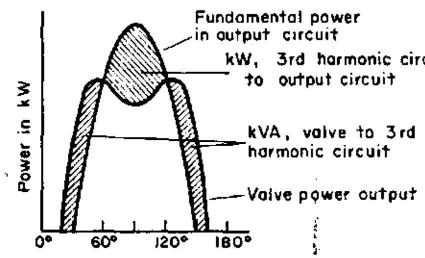


FIG. 3.5 Energy transfer in class D operation.

harmonic, in the correct relative phase. Both resonant circuits are similarly, but lightly, loaded by the tetrode input impedance, so the amplitude relationship is substantially as required. It follows that the anode current and anode voltage have waveforms similar to those at the valve input.

The anode tuned circuit L_2C_2 also includes a third harmonic resonator, L_4C_4, the purpose of which is to restore the sinusoidal waveform of the fundamental at the output of the anode circuit. This occurs in the following manner. The initial portion of anode power output feeds power into the fundamental circuit, and wattless energy into the third harmonic circuit. By the resonant

action of the harmonic circuit, from a fundamental angular position of 60°–120° the phase of the harmonic energy is reversed and transferred into the fundamental circuit. At 120° the phase is again reversed and for the remainder of the conducting cycle the harmonic circuit again receives energy from the valve. This action is shown clearly in Fig. 3.5, the two types of cross-hatching representing the change of phase between the fundamental and harmonic energies. This diagram also shows a side-effect of operating the valve with a complex waveform, which is to increase the angle of current flow, relative to using a sinewave under similar conditions. In the diagram the increase is from 120° to 140°, which is typical.

Operational frequency band

The addition of the third harmonic resonators does add a degree of complication to the tuning procedure, which although relatively simple makes frequency changing more difficult. As operational changes of frequency are necessary in the h.f. band, class D is not recommended for h.f. There is also the difficulty of designing the third harmonic circuits for high power at frequencies approaching 90 MHz, for use at the upper end of the h.f. band (with a total anode tuning capacitance of 100 pF, the r.m.s. circulating current is likely to be in excess of 100 A at the upper end of the h.f. band).

It would be exceptional to require anode modulation in the v.l.f. and l.f. bands, so the use of class D is limited to the m.f. band. In this band it is particularly useful as a means of saving power consumption, because of the very high power often used.

Method of calculating class D performance

Owing to the complex waveform, the 15° intervals used for calculating the operating conditions by means of the tube performance calculator are inadequate to obtain sufficient accuracy. It is necessary to obtain levels of anode current and anode voltage from the load line at intervals of 10° of the drive waveform.

A waveform with a substantially flat top is obtained by adding a third harmonic component at 20% of the fundamental level, in phase opposition at the fundamental peak. The voltage at each 10° of the positive grid excursion is calculated on the basis that the two peaks of the combined waveform are of the correct value to drive the anode along the full extent of the load line. These voltages are used to determine the instantaneous levels of anode current and anode voltage at 10° intervals along the load line. By subtracting the instantaneous voltages from the d.c. voltage the anode voltage excursions are obtained, corresponding to the instantaneous values of anode current. Assuming that the instantaneous values of anode current I and anode voltage excursion V at 10° intervals from 90° to 0° are indicated by the suffixes A to J, respectively, the performance can be calculated from the following formulae:

$$\text{Anode current d.c.} = \frac{0{\cdot}5 I_A + I_B + I_C + \cdots + I_I}{18}$$

The d.c. values of grid current and screen current are obtained similarly from their instantaneous values at 10° intervals.

$$\text{Output power, r.m.s.} = \frac{0{\cdot}5 I_A \times V_A + I_B V_B + I_C V_C + \cdots + I_I V_I}{18}$$

The r.m.s. power of drive required by the valve only must be calculated in the same manner as that used for output power, but using the instantaneous values of grid current and voltage.

3.6 COMPARISON BETWEEN CLASS D AND CLASS C—100 kW OUTPUT

A power output level of 100 kW carrier has been chosen as an example of class D operation, in order to give a direct comparison with class C on the same valve and under similar conditions, as in the example given in Section 3.4.

Referring again to Fig. 3.3, load line *NO* represents the operating line for class D, to give approximately the output power as class C on load line *MO*. Details of the carrier performance of the two classes of operation on their respective load lines are given in Table 3.2. Significant differences are also given for 100% modulation.

TABLE 3.2

	Class C		Class D	
Valve type 4C X 3500	Carrier	100% Modulation	Carrier	100% Modulation
Anode voltage, d.c.	15 kV		15 kV	
Screen voltage, d.c.	750 V		750 V	
Grid bias voltage total	−500 V		−500 V	
Anode voltage trough	1·2 kV		1·2 kV	
Anode voltage excursion	13·8 kV		13·8 kV	
Anode current peak	38 A		22 A	
Anode current, d.c.	8·4 A	8·4 A	8·1 A	8·1 A
Power input, d.c.	126 kW		121·5 kW	
Power output, r.f.	103·5 kW		108 kW	
Conversion efficiency	82·2%		88·8%	
Anode dissipation	22·5 kW	33·75 kW	13·5 kW	20·25 kW
Grid voltage excursion	700 V		600 V	
Peak of fundamental component of grid voltage	700 V		690 V	
Grid current, d.c.	0·5 A		0·27 A	
Grid drive, valve only	350 W		190 W	
Screen current, d.c.	1·4 A		0·65 A	
Screen dissipation	1·05 kW	1·4 kW (approx.)	0·49 kW	0·65 kW

The main features are that class D is 6·6% more efficient, giving 4·5 kW more output for 4·5 kW less input and a reduction of 40% in anode dissipation. As the maximum permissible anode dissipation for this valve is 35 kW, class C is practically on the upper limit at 100% modulation. For class D an appreciable

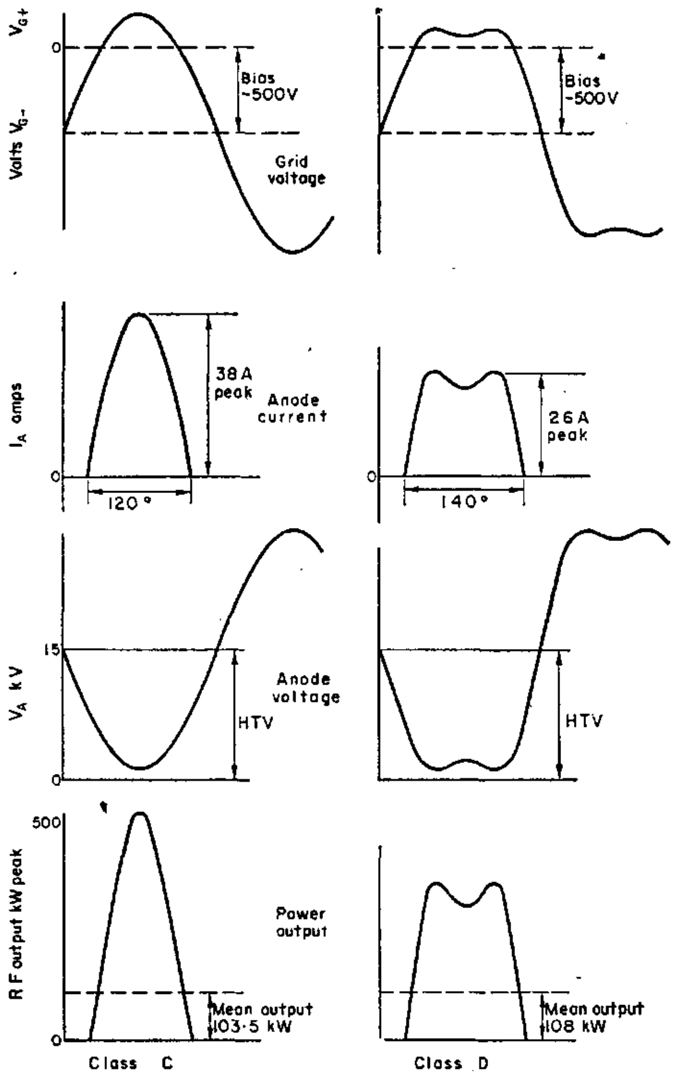


FIG. 3.6 Comparison between class C and class D waveforms.

increase in output could be obtained before approaching this limit. The driving power and screen dissipation are also less in the case of class D, but if the output is increased to load the anode more fully, the grid or screen dissipation will most probably determine the upper limit.

In order to give a better mental picture of the two classes of operation, their waveforms are compared in Fig. 3.6. The reason for the high efficiency of class

D is clearly shown by the fact that anode voltage remains low during the major portion of the anode current pulse.

3.7 CLASS D FOR VERY HIGH POWER

Tetrode valves are the obvious choice for very high-power applications in the m.f. band, but there is a very limited number available from which to make a selection. In fact, for carrier powers much in excess of 250 kW they do not exist and their production would be uneconomical because of the small number of transmitters likely to be used at these power levels. It is also more economical to use the same type of valve in the modulator as in the r.f. amplifier, thereby limiting the spares which have to be stocked by the operating organization. Valves are available for 250 kW carrier output and capable of being anode modulated at 100%. On this basis the most economical transmitters can be produced for carrier powers of 250 kW, 500 kW, 750 kW, 1000 kW and 1500 kW, using one, two, three, four or six valves, respectively for the r.f. amplifiers. Two valves are necessary for high-power modulators, and using the same type of valve, would be able to supply sufficient audio power to modulate r.f. amplifiers up to 750 kW of carrier power. The capability of being able to fully modulate a 750 kW carrier stage is due to the high conversion efficiency of r.f. amplifiers in class D. For carrier powers of 1000 kW and 1500 kW, four modulator valves would be required.

By taking advantage of the concepts outlined, it is quite practical to develop transmitters with carrier powers of 250 kW, 500 kW, 750 kW, 1000 kW and 1500 kW from two basic designs, all using the same type of valve. This not only reduces development and maintenance costs, but simplifies operational procedures, because in all cases the r.f. valves operate in the same condition.

One design is for 250 kW only, containing a single tetrode in the r.f. amplifier and a pair of the same type in the modulator. Because it is being operated well below maximum rating, the modulator valve life will be extended, and the essence of the design will be the same as for higher-power applications.

The other basic design is for 500 kW or 750 kW. By making the initial provision for two or three valves to be accommodated in the r.f. amplifier, the same design will be suitable for both power ratings. The difference in initial cost of the d.c power supplies will be negligible, provided that the higher current requirement is considered at the design stage.

Although it is not normally economical to double the power output by means of combining two transmitters, in this case it is so. This is because the demand for transmitters of 1000 kW and 1500 kW is very limited indeed. For the somewhat rare requirement of transmitters for these power ratings, it is considered economical to supply two transmitters, with a circuit for combining the two r.f. outputs into one antenna. Alternatively, each transmitter can feed a separate antenna, by which means a degree of directivity can be obtained if the two antennas are suitably positioned.

Thus it is seen that the most suitable valve for giving a carrier power output of 250 kW, when operated in class D, is the essence of all very high-power designs for m.f. broadcasting. The constant current characteristics of such a valve are shown in Fig. 3.7, with the preferred load line for this purpose, *PQ*. The calculated performance obtainable when operating on load line *PQ* is given in Table 3.3, compared with the maximum permissible ratings of the valve.

The actual power output at the valve is 260 kW, so allowance has been made for circuit losses and valve differences to enable 250 kW output at the transmitter terminals. Although the anode dissipation is only one-third of the maximum

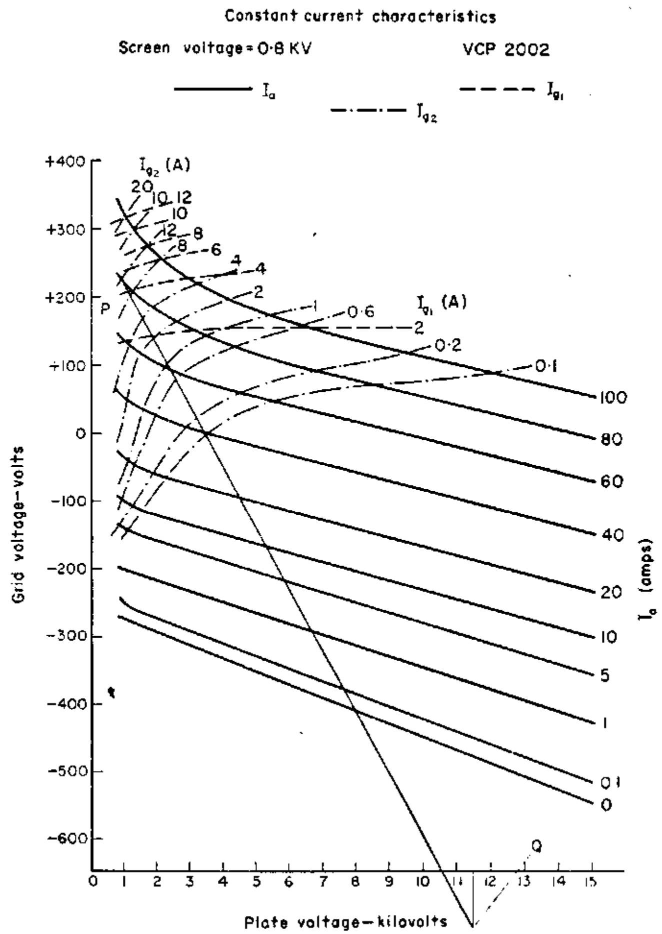


FIG. 3.7 Load line on constant-current characteristics of tetrode for very high power at high efficiency in class D. Valve type VCP 2002 (English Electric Valve Co.).

allowed, both the screen and control grids are near their limits. This is typical of class D operation.

For 260 kW output from the same valve in class C would require an additional d.c. input of 27 kW on carrier. To this must be added the extra power into the modulator and the a.c. mains to d.c. conversion efficiency. This means that on

TABLE 3.3

Tetrode type VCP 2002	Maximum ratings	Class D Carrier	Class D 100% Modulation
Anode voltage, d.c.	11·5 kV	11·5 kV	
Screen grid voltage, d.c.	1·2 kV	0·8 kV	
Control grid voltage d.c.	–800 V	–725 V	
Grid voltage excursion		900 V	
Mean cathode current, d.c.	60 A	29·6 A	
Anode dissipation	180 kW	42 kW	63 kW
Screen grid dissipation	2·7 kW	1·82 kW	2·37 kW (approx.)
Control grid dissipation	1·2 kW		
Drive power, valve only		1·08 kW	
Anode current, d.c.		26·3 A	
Anode power input, d.c.		303 kW	
Power output, r.f.		261 kW	
Conversion efficiency		86%	
Angular width of current pulse		145°	

broadcast transmission, with an average modulation depth of about 30%, the saving in a.c. mains power obtained by operating the final r.f. amplifier in class D instead of class C, is about 40 kW/h for a 250 kW carrier. As the high-power output levels proposed are based on 250 kW units, the saving in mains power by using high-efficiency class D amplifiers in the final stages is proportional to the number of 250 kW units. The approximate saving in mains power by the use of class D for these high-power transmitters will be of the order of 40 kW/h at 250 kW, 80 kW at 500 kW, 120 kW at 750 kW, 160 kW at 1000 kW and 240 kW at 1500 kW.

In terms of daily running costs, it can be seen that the real advantage of class D operation is more apparent at very high power.

4

Circuit Configuration and Component Selection

4.1 PUSH–PULL AND SINGLE-SIDED CIRCUITS

Before the advent of high-power tetrodes, the push–pull or balanced circuit was almost universally employed for triodes in final amplifier stages. The main reason for this was that the balanced circuit formed a simple and convenient means of neutralizing the effect of the internal anode-grid capacitance, by cross-connecting neutralizing capacitors from the anode of each side to the grid of the other side. There were a number of other reasons.

(*a*) The anode-earth capacitance of the two valves was in series, so that it was possible to tune to higher frequencies, with sufficiently large inductors for magnetic coupling between the anode and output circuits.

(*b*) The use of magnetic coupling made it possible to use the same anode and output circuit for either balanced or unbalanced feeders. For balanced feeders the output capacitor and inductor were connected in parallel, while for unbalanced or coaxial feeders they were connected in series.

(*c*) Even with the biggest triodes available there were instances where power output was required in excess of that obtainable from a single valve. Two valves in push–pull supplied double the power and it was not unknown for the power to be further increased by using valves in parallel on each side of the push–pull circuit.

(*d*) With a balanced circuit, the physical centre-point of the anode inductor is at a very low r.f. potential to earth. In consequence, the d.c. supply could be fed in at this centre-point, with very little fundamental potential across the h.t. choke.

There were, of course, disadvantages, such as the tendency to self-oscillation at frequencies above the operating band, when the neutralizing arms become inductive. This type of oscillation could be particularly prevalent at the second harmonic frequency, due to the fact that the pulses of anode current through the d.c. feed choke were at twice the fundamental frequency.

With modern high-power tetrodes, internal anode-grid capacitance is so low that even if neutralizing is required, it can be accomplished across individual valves, by means of a very low external capacitance, combined with a wideband transformer for changing phase. The elimination of the need for high-level neutralizing with tetrodes is undoubtedly a major reason for the tendency of designers to depart from push–pull in favour of single-sided circuits.

Triodes can also be operated without neutralizing, provided that they are grounded-grid circuits, so again push–pull has given way to the greater simplicity of unbalanced circuits. However, the high drive level required by triodes in grounded-grid circuits must be taken into account in deciding which type of circuit to use.

The type of input required by the antenna must also be considered. In the v.l.f. and l.f. bands, the antennas are monopoles requiring unbalanced or concentric feeders. Consequently, single-sided circuits are preferred for applications in these bands.

In the m.f. band, monopole antennas are also more general, though there are applications where balanced feeders are required, for example in split-mast radiators. In the h.f. band, dipole antennas with balanced inputs are most usual.

Where single-sided output circuits are used for coupling into balanced feeders it is quite a problem to obtain a well-balanced output over a complete frequency band. If the output is not balanced, the twin-wire feeder behaves as a single radiator for the unbalanced component. The effect is not only a loss of power in the antenna, but also undesirable radiation which can be detrimental to the radiation pattern, particularly in the vicinity of the antenna.

For communication transmitters with power outputs of 30 kW or less, the problem has been solved by the use of wideband unbalanced-to-balanced feeder transformers. These transformers have ferrite cores, with a substantially flat response over the h.f. band, and are capable of carrying 40 kW r.f. power. Wideband baluns with similar characteristics, but capable of carrying rather more power are also available. Either of these devices enables coaxial feeder switching, with obvious advantages over balanced feeder switching, to be used for the operational antenna changes required in the h.f. band.

For h.f. applications up to 30 kW output, single-sided circuits are preferred on all counts, with either grounded-cathode tetrodes or grounded-grid triodes.

For broadcasting applications in the h.f. and m.f. bands, with antennas requiring a balanced input and power levels of the order of 100–250 kW, wideband transformers and baluns are not available for unbalance-to-balance conversion. For this purpose push–pull anode-output circuits are recommended. It means that balanced feeder switching is necessary for antenna changing. But the problem is less difficult than in the communications case, since the number of antenna changes required is usually less, and more time can be allowed for frequency and antenna changing. For these applications, tetrode valves in grounded-cathode circuits are preferred to triodes in grounded-grid because of the lower driving power necessary.

4.2 SINGLE-SIDED ANODE-OUTPUT CIRCUITS IN THE H.F. BAND

The two most usual forms of single-sided anode-output circuits are shown in Figs. 4.1 and 4.2. Both contain triodes in a grounded-grid arrangement, but the points raised in this section are just as applicable to grounded-cathode tetrodes.

The combination of a shunt-tuned anode circuit $L_1 C_1$ and a series-tuned output circuit $L_2 C_2$ (Fig. 4.1), has the advantage that the h.t., d.c. supply is fed into a point effectively at zero r.f. potential, with the output circuit and feeders isolated from this supply. Loading control can be either by adjusting the mutual coupling between L_1 and L_2, or by adjusting the feeder inductor L_3, leaving the mutual coupling between L_1 and L_2 in a fixed position.

For an amplifier in the h.f. band, the main disadvantage is that three controls must be operated for each frequency change. Also, the indication of output circuit resonance is not sufficiently simple for operational frequency changes. Output-circuit tuning can only be set to predetermined positions, making the whole circuit unsuitable for completely automatic tuning systems.

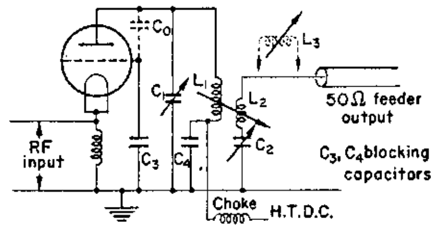


FIG. 4.1 Tuned-anode tuned-output circuits with magnetic coupling for grounded-grid triode.

The circuit shown in Fig. 4.2 is a combination of Π and L circuits, with a common shunt element C_2; normally called a ΠL circuit. With only two controls for tuning and loading, it is very simple to operate and particularly suitable for automatic tuning. The whole h.f. band can be covered in a number of ranges, by means of tapping points on the tuning and loading inductors L_1 and L_2. In addition, the required inductance value of L_1 is greater in the case of Fig. 4.2

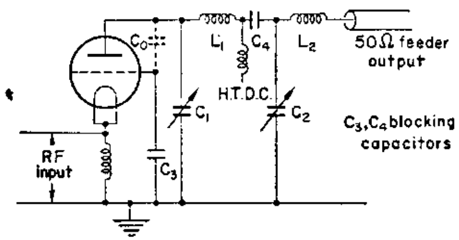


FIG. 4.2 Basic ΠL circuit for grounded-grid triode.

for the same frequency, making it more practical to construct the low value of inductance required for the upper frequencies.

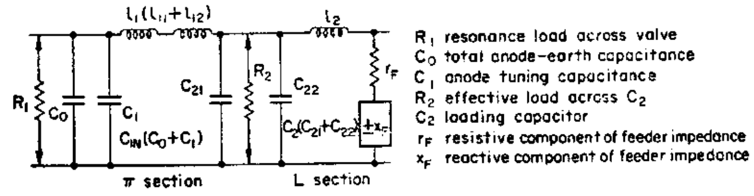
An alternative arrangement of a ΠL circuit is obtained by eliminating the variable capacitor C_1, and making inductor L_1 variable for anode tuning. While this has some technical advantages, variable inductors for high power are not only difficult to construct but are more liable to give trouble in service and are quite expensive.

It is for this reason that the circuit arrangement shown in Fig. 4.2 is most generally adopted for h.f. power amplifiers and why it is used to demonstrate a method of calculating the component values for single-sided circuits.

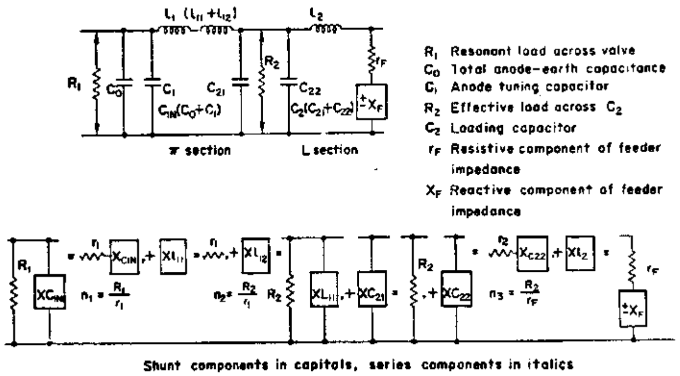
4.3 A METHOD OF CALCULATING COMPONENT VALUES FOR *ΠL* CIRCUITS

Information required for calculations

As explained in Section 4.2, *ΠL* circuits consist of *Π* and *L* circuits in series (Fig. 4.3(*a*)). For calculation purposes this is further divided into a series of three *L* circuits (Fig. 4.3(*b*)). By this means it can be seen that the calculations consist of a number of conversions from resistance and reactance components in shunt to equivalent series components, or vice versa.



(*a*)



(*b*)

FIG. 4.3 (*a*) Circuit arrangement for *ΠL* calculations. (*b*) Breakdown of *ΠL* circuit into three *L* circuits.

The first essential is a knowledge or estimate of certain basic values on which to base the calculations.

(1) R_1. The effective resonant load across the valve at the output power required.

(2) C_{IN}. The total internal and external capacitance of the valve anode to earth, including allowances for valve mounting, strays and the minimum tunable capacitance of C_1.

(3) R_2. The effective resonant resistance across the coupling capacitor, C_2.

(4) r_F. The nominal resistance of the feeder.

(5) v.s.w.r. The likely standing wave ratio to be encountered in service, or the limits given in a specification, in order to determine the reactive component X_F of the corresponding feeder impedance.

(6) The frequency range to be covered.

(1) Determination of R_1

$$R_1 = \frac{PVA \times PVA}{\sqrt{2} \times \sqrt{2}\ \text{W output}}$$

Thus, for a 31 kW output, if the d.c. voltage is 8000, the anode swing *PVA* will be about 6000 V for a linear amplifier. Then

$$R_1 = \frac{6000 \times 6000}{2 \times 31\,000} = 580\ \Omega$$

(2) Estimate of C_{IN}

In a grounded-grid circuit, the internal anode-grid capacitance of a triode, capable of giving 30 kW output, is likely to be of the order of 60 pF. To this must be added the external capacitance to earth of the valve and its mounting, say 20 pF, giving $C_0 = 80$ pF.

Assuming that the minimum tunable capacitance of C_1 is 20 pF, then $C_{IN} = 100$ pF minimum.

(3) The value of R_2

An interesting point comes to light in estimating R_2. It would be possible to couple the anode resistance R_1 directly into the feeder impedance with a *Π* circuit only. The addition of the *L* circuit has two advantages over the *Π* only arrangement. First, with the *Π* circuit alone, the maximum value of C_2 at the low-frequency end of the band would be excessively large, particularly to cover a v.s.w.r. of 2 to 1. Second, the *L* section reduces the harmonic content in the feeder by providing attenuation at frequencies above the fundamental.

As a compromise between capacitance value of C_2 and the peak r.f. voltage across it, R_2 is often selected to be the geometric mean between R_1 and r_F i.e., $R_2 = \sqrt{(R_1 \times r_F)}$. For a 50 Ω feeder and R_1 of 580 Ω, $R_2 = 170$ Ω approx.

(4) and (5) r_F and v.s.w.r.

The most usual feeder impedance is either 50 Ω or 60 Ω, with provision for working into a mismatched feeder with a v.s.w.r. of 2 to 1. For a 50 Ω feeder, a v.s.w.r. of 2 to 1 at 3 MHz means that about ±150 pF must be allowed on the value of C_2 calculated for a purely resistive 50 Ω feeder.

(6) The frequency range

For the explanatory example given later, the h.f. band is considered from 3 MHz to 30 MHz.

Formulae for deriving component values from the known or estimated information

Although monographs are normally available for series-parallel impedance conversion, their accuracy decreases as the ratio x/r increases. For upper frequencies in the h.f. band, the x/r ratio is likely to be of the order of 10 to 20, making monograph conversions too inaccurate. Consequently, conversion by means of basic formulae is recommended. Desk calculators are invaluable for this purpose.

Resistance and reactance components in shunt are expressed in capitals, with series components in lower case. The operator j has been omitted for clarity.

$$n = \frac{R}{r}, \qquad X_{(\mathrm{L\,or\,C})} = \frac{R}{\sqrt{(n-1)}}, \qquad x_{(\mathrm{L\,or\,C})} = r\sqrt{(n-1)}$$

From which:

$$r = \frac{R}{n} = \frac{x}{\sqrt{(n-1)}} \qquad \text{and} \qquad R = rn = X\sqrt{(n-1)}.$$

Incidentally, $\sqrt{(n-1)}$ is the Q factor of the circuit, and in subsequent calculations n_1, n_2, n_3 refer to different parts of the circuit.

If monographs are not readily available for reactance inductance, or reactance-capacitance conversion, the following formulae might prove useful, particularly as they contain units appropriate to the application

$$X_{\mathrm{L}} \text{ (or } x_{\mathrm{L}}) = 6{\cdot}283 \cdot L\,(\mu\mathrm{H}) \cdot f\,(\mathrm{MHz})$$

$$X_{\mathrm{C}} \text{ (or } x_{\mathrm{C}}) = \frac{15{\cdot}93 \cdot 10^4}{C\,(\mathrm{pF}) \cdot f\,(\mathrm{MHz})}$$

Example of the calculations required to obtain component values for a 30 kW amplifier supplying power into a 50 Ω feeder, via a ΠL circuit from 3 MHz to 30 MHz

It is assumed that a d.c., h.t. supply of 8 kV is contemplated, and as linear operation is typical for this application, the peak anode swing is taken as 6 kV. It is also assumed that the correct load line has been selected to give 31 kW output at the valve anode, as described in Chapter 2, Section 2.4. Component values for 30 MHz and 3·0 MHz are given in Table 4.1, for the breakdown circuit shown in Fig. 4.3(*b*).

An examination of the results in Table 4.1 indicates the large range of capacitance and inductance required to cover the 10 to 1 frequency range of 3–30 MHz. Consequently, it is not unusual for high-power communication transmitters to have a limited coverage of 4–27·5 MHz. Apart from the advantage of lower range of component values, there are operational reasons for this limitation.

Transmissions between 3 MHz and 4 MHz are for short-range links, with low propagation attenuation, so it is usual for operating organizations to use lower power on these frequencies.

Between 27·5 MHz and 30 MHz there are only two very narrow bands for fixed communication services and they are of little value for long-range links except at very limited periods of the sunspot cycle.

Frequency ranges

While it is obviously an operational advantage for frequency changing to have the smallest number of ranges possible, as well as the limited coverage on each range imposed by the capacitance range, there is a limit due to r.f. circulating current.

TABLE 4.1

Frequency	30 MHz	3 MHz
C_{IN}	100 pF	300 pF
XC_{IN}	−53·1 Ω	−177 Ω
R_1	580 Ω	580 Ω
$\sqrt{(n_1-1)} = \frac{R_1}{XC_{\mathrm{IN}}}$	10·92	3·28
n_1	120	11·76
$r_1 = \frac{R_1}{n_1}$	4·83 Ω	49·32 Ω
$xC_{\mathrm{IN}} = r_1\sqrt{(n_1-1)}$	−52·74 Ω	−161·77 Ω
$xl_{11} = -xC_{\mathrm{IN}}$	52·74 Ω	161·77 Ω
R_2 (selected)	170 Ω	170 Ω
$n_2 = \frac{R_2}{r_1}$	35·2	3·45
$\sqrt{(n_2-1)}$	5·85	1·56
$xl_{12} = r_1\sqrt{(n_2-1)}$	28·26 Ω	76·94 Ω
xl_1 (total) $= xl_{11} + xl_{12}$	81·00 Ω	238·71 Ω
$XC_{21} = \frac{R_2}{\sqrt{(n_2-1)}}$	−29·06 Ω	−108 Ω
r_{F}	50 Ω	50 Ω
$n_3 = \frac{R_2}{r_{\mathrm{F}}}$	3·4	3·4
$\sqrt{(n_3-1)}$	1·53	1·53
$XC_{22} = \frac{R_2}{\sqrt{(n_3-1)}}$	−111·11 Ω	−111·11 Ω
XC_2 (total) $= \frac{XC_{21} - XC_{22}}{XC_{21} + XC_{22}}$	−23·06 Ω	−54·8 Ω
$xl_2 = r_{\mathrm{F}}\sqrt{(n_3-1)}$	76·5 Ω	76·5 Ω
Allowing 2 to 1 v.s.w.r. on C_2, l_2		
XC_2	−20 to −26 Ω	−43 to −70 Ω
xl_2	36–120 Ω	36–120 Ω
C_1	100 pF	300 pF
l_1	4·3 μH	127 μH
C_2	265–205 pF	1240–760 pF
l_2	1·9–6·4 μH	19–64 μH

The circulating current is determined by the anode impedance of the resistive and reactive series components,,which are 4·83 Ω and 52·74 Ω, respectively, at 30 MHz (Table 4.1). With a peak r.f. voltage of 6 kV, the r.m.s. circulating current at this frequency is:

$$\frac{6000}{\sqrt{2}\cdot\sqrt{(4{\cdot}83^2 + 52{\cdot}74^2)}} = 80 \text{ A r.m.s.}$$

The use of the full three to one capacitance range of C_{IN} (100–300 pF) would make it possible to tune from 30 MHz to $(30/\sqrt{3})$ MHz, i.e., 17·3 MHz. As the reactance is the dominating factor in the impedance, the impedance would fall by $\sqrt{3}$ at 17·3 MHz, and the circulating current would increase by $\sqrt{3}$ to 138 A r.m.s. Also, the loss increase being proportional to the current increase, the losses would increase by three times.

This means that the size of inductor material would have to be increased to carry a circulating current which will not be present elsewhere in the band, or the upper frequency ranges must have a limited frequency coverage. In practice the number of ranges is a compromise between operational simplicity, current-carrying capacity and manufacturing costs.

4.4 FIXED CAPACITORS IN PARALLEL

Referring to blocking capacitors C_3 and C_4, Fig. 4.2, there are reasons why it might appear to be advantageous to use two or more in parallel instead of a single unit, but this is not a good principle for transmitters covering a 10 to 1 frequency range.

Consider capacitor C_3, the dual purpose of which is to isolate the d.c. grid supply from ground and at the same time to provide a very low impedance path to r.f. To reduce further the r.f. impedance and to spread the anode circulating current round the valves, the grid connections should be via a surrounding ring. The most suitable capacitor for this purpose is formed by isolating the ring from the ground plane with a thin film of insulating material. Even with a film having a high dielectric constant and high breakdown voltage, the capacitance achievable is often too small to provide a sufficiently low r.f. impedance path at the low-frequency end of the band.

One solution is to replace the thin film with a number of fixed capacitors distributed round the ring, but the value of the individual capacitors must be selected with care. When capacitors are connected in parallel there is an inevitable inductance both in the inter-connecting leads and in the capacitors themselves. At some frequency this inductance will resonate with the capacitance of the individual units in series.

Experience indicates that typical loop inductances are between 0·04 μH and 0·08 μH, and these values have been used in Fig. 4.4 to show likely resonant frequencies. As an example, suppose that two capacitors of 0·01 μF are used to obtain a sufficiently low reactance at 3 MHz. From Fig. 4.4 it can be seen that resonance is probable in the region of 8–12 MHz; when the grid-to-ground r.f. impedances will be far too high for satisfactory operation.

There are two practical solutions to the problem. First, a multiplicity of capacitors, each of less than 0·001 μF (see Fig. 4.4), can be distributed round the

valve. With this arrangement, an early check should be made that there are no similar resonances within the band.

The other solution is to connect the grid directly to ground, providing the lowest possible impedance path at all frequencies, without the possibility of this type of resonance occurring. This solution does add a degree of complexity to the d.c. power supplies, but it is known science and by far the easier problem to solve.

Returning again to Fig. 4.2, consider the reasons for requiring capacitors in parallel for C_4 and the likely resonant effect of so doing. At low frequencies the reactance should be low, otherwise the inductance of L_1 must be increased and

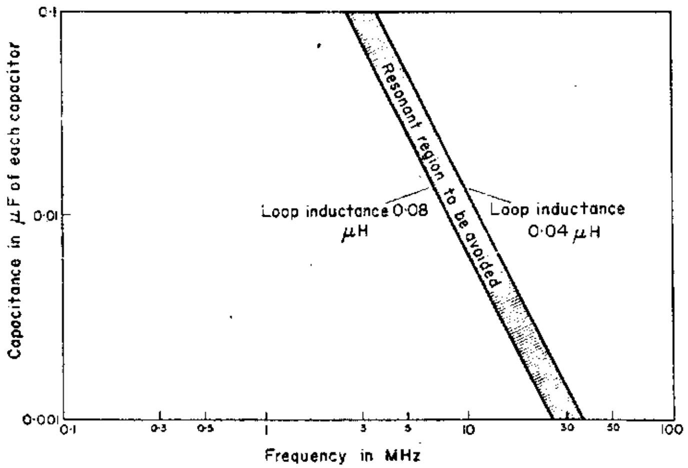


FIG. 4.4 Probable resonance of capacitors in parallel.

the r.f. voltage across the capacitor must be added to the d.c. voltage for rating purposes. At high frequencies capacitor C_4 has to carry the total r.f. current circulating in the Π section. Replacing C_4 by two capacitors in parallel is likely to cause an in-band resonance, resulting in a circulating current within the loop, limited only by the loop losses. This current can be considerably greater than that in a single capacitor of the same value carrying the maximum circulating current of the Π section. Not infrequently such loop resonances will cause the capacitors to explode, combined with a short circuit. The d.c. voltage will be applied to C_2, and a flashover in the capacitor is not improbable.

Two solutions are possible. First, every effort should be made to obtain a single capacitor of adequate current rating, even if the capacitance is such that a slight increase in L_1 is required.

The other solution is to connect the d.c. blocking capacitor in series with the feeder inductor L_2 and to provide a loading capacitor C_2 of adequate voltage rating to withstand the d.c. plus r.f. voltages.

These two examples should serve to show that paralleled capacitors should not be used in high-power applications where a range of frequencies is to be covered.

4.5 THE CURRENT-CARRYING CAPACITY OF COPPER CONDUCTORS AT R.F.

For frequencies below 30 MHz, unplated copper is the normal material used for inductors and connector leads. The improvement obtainable with silver plating is negligibly small and other plating metals only increase the circuit losses.

The r.f. current-carrying capability of conductors of various sizes depends entirely on the permissible temperature rise and the amplifier cooling system, so

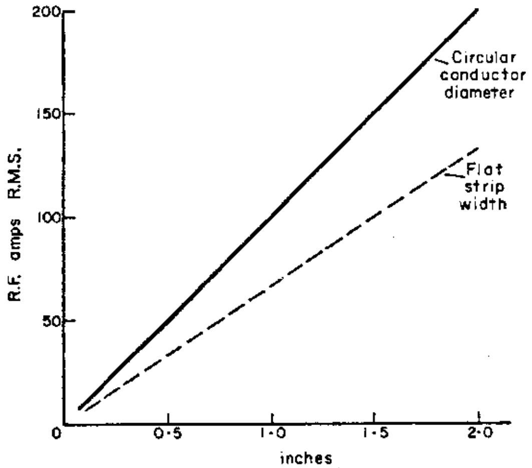


Fig. 4.5 R.F. current-carrying capacity of conductors for frequencies below 30 MHz.

conductor size is tailored to the cooling arrangements. However, experience has shown that with the average cooling systems of transmitters operating below 30 MHz, circular conductors of the size indicated in Fig. 4.5 can be used without excessive temperature rise.

Flat conductors are not so easy to define and their position in the circuit affects the cooling available to them. Furthermore, their width is more often determined by the inductance permitted; Fig. 4.5 can therefore be used as a guide only.

4.6 INDUCTOR CONFIGURATION

On the basis of r.f. current determining the minimum size of material to use for any particular inductor, other relevant factors must be taken into account in the inductor design.

(1) The minimum spacing between any surrounding metal or screens and the ends or sides of the inductor, should be at least half the inductor diameter, in order to avoid excessive eddy current losses in the metal sheets. It is advantageous

further to increase this spacing at the high-potential end in order to reduce minimum circuit capacitance.

(2) The low-potential end of the inductor should not be too close to the valve anode, because such an arrangement enables harmonics to be capacitance coupled directly into the output circuit.

(3) On the upper-frequency ranges, when the anode inductor is likely to consist of one turn or so, the full anode r.f. voltage will appear across that turn, so the spacing between the turns must be adequate to prevent flashover. Remember also that any insulating material in the vicinity of the gap between turns will reduce the voltage at which breakdown will occur.

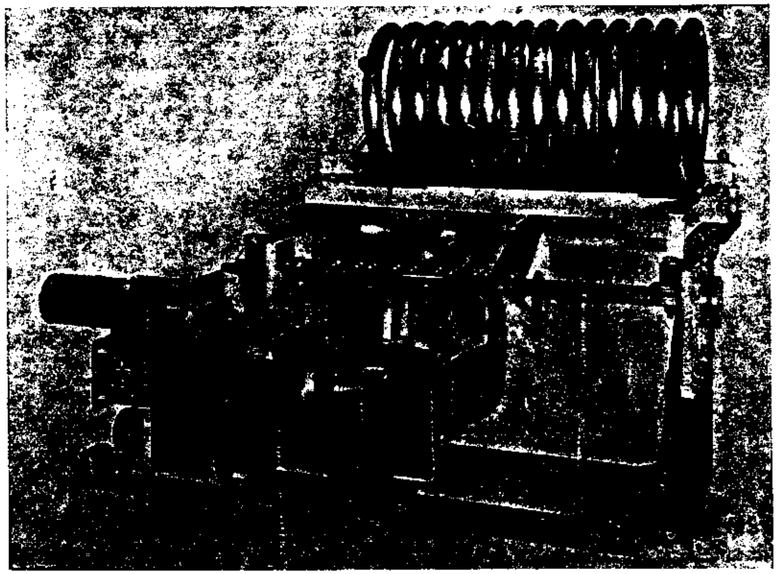


Fig. 4.6 Example of tapped-inductor variable-capacitor anode circuit.

(4) For equipment covering a wide frequency range by means of tapping points on the inductor, it is a definite advantage in the mechanical design for all tapping points to be in line longitudinally. This can often be achieved by making the inductor diameter such that suitable ranges can be obtained by tapping points being at intervals of one or more whole turns.

(5) For the higher-frequency ranges the tapping points on the unused section should remain in contact with the inductor, in order to avoid 'back-end' resonance at either fundamental or harmonics.

(6) When the tapping point for the next lower range is separated from the active contact by only one turn, there will be an appreciable rise in current through the active contact, due to 'shorted-turn' effect. This current rise can be reduced to manageable proportions by opening that contact for the range in use.

(7) The essence of a low-resistance r.f. contact is high pressure on a single

point, 6 lb being a reasonable minimum value. Even with the best possible arrangement, heat will be generated at the contact point, so the body of the contact should be bulky and preferably fitted with cooling fins to assist heat radiation.

(8) Two point contacts in series should definitely be avoided by making a clamped connection for the remote end of the contact arm. This means that the contact arm must be of a spring material of low r.f. resistance, such as beryllium copper, and of sufficient width that the temperature rise is maintained below the level required to affect the temper of the spring.

A photograph of a tuning assembly with the inductor designed on the basis of the points outlined is shown in Fig. 4.6. It should be noted that provision has been made for air cooling to be applied to the two turns at the high-potential end of the inductor. This enabled a material of smaller diameter to be used, so reducing the overall size of the inductor. This is an important feature of this particular design, because of the limited space available for the tuning assembly.

Another method of making more effective use of the current-carrying capacity of smaller conductors over the whole frequency band is to use two inductors in parallel for the higher-frequency ranges. This is of particular advantage where available space limits the inductor length but not the diameter. This is both technically and economically preferable to using a large-diameter inductor with increased material size, which would make the mechanical arrangement of preferred tapping points more difficult and expensive.

4.7 VARIABLE CAPACITORS

There is such a wide range of variable-vacuum capacitors available, that there are few applications in high-power amplifiers where they are not suitable as variable elements for tuning and loading. This is due to their low minimum capacitance, high capacitance range and small bulk, combined with high voltage and current rating. The only precautions necessary are the provision of adequate safety factors under normal conditions, so that the maximum specified ratings are not exceeded under fault conditions. A typical example of a vacuum capacitor for anode-circuit tuning is shown in Fig. 4.6.

One of the applications for which the available range of vacuum capacitors is not suitable, concerns very high-power amplifiers in the m.f. range. In this case the very high voltage and current ratings required are of prime importance, whereas the minimum capacitance and capacitance range are a secondary consideration, because frequency changing is not an operational requirement. For this application the most usual type of variable capacitor is mounted in a metal tank filled with nitrogen under high pressure. As the voltage at which breakdown occurs is determined by the nitrogen pressure, an external gauge monitors the pressure continuously and provision is made to re-pressurize the tank if a specified lower limit is reached.

4.8 VOLTAGE FLASHOVER AND THE USE OF CORONA RINGS

The importance of taking every precaution to prevent voltage flashover cannot be overstressed in high-power applications. In most cases the highest voltages are between the high-potential parts of the anode circuit and earth, so that an

r.f. flashover often takes the d.c. with it. Even with the fastest possible power-tripping facilities, the energy stored in the d.c. smoothing capacitors is sufficient to do quite a lot of damage under fault conditions. Therefore, an adequate safety factor must be allowed between the working voltage across any two electrodes and the theoretical breakdown voltage.

A reasonable safety factor to allow is 3 to 1, relative to the likely breakdown voltage given in Appendix IV. Also, it should be noted that the curve is for ideal electrode shapes, and due allowance must be made for practical departures from these shapes, together with a further allowance for any insulating material in the vicinity.

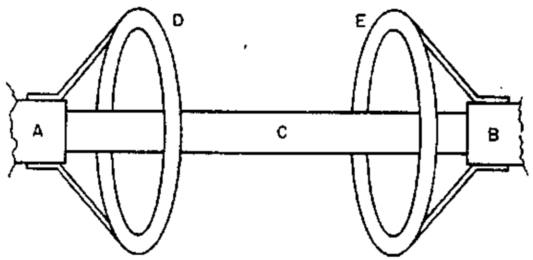


FIG. 4.7 The use of corona rings to increase breakdown voltage.

Each case must be treated on its own merits, but as a general guide for early estimates only, reasonable safety factors will be given by allowing not more than 8 kV/in. in air, or 4 kV/in. across insulating material.

In connection with the very high r.f. voltages associated with high-power amplifiers amplitude modulated, it frequently happens that the voltage across the insulator is too high to allow adequate safety factors. In many cases also, either or both of the electrodes have sharp edges which are not easily removable, such as those produced by fixing bolts, and ionization would start irrespective of the length of insulator.

This defect can be overcome by fitting corona rings as shown in Fig. 4.7. Although the faces of the corona rings *DE* are nearer together than the electrodes *AB*, the steep potential gradient near the ends of the insulator *C* is removed and replaced by a more evenly distributed potential gradient in air. The result is that the effective voltage for breakdown to occur between the electrodes is considerably increased.

This is only one example of the use of corona rings as a means of increasing the voltage safety factor, but the principle finds wide applications in high-power r.f. designs.

4.9 HARMONIC ACCENTUATION IN ANODE CIRCUITS

In the construction of a capacitor-tuned anode circuit for a high-power amplifier, the valve and tuning capacitor, because of their size, and allowances which must be made for voltage clearance, cannot be as close as theoretically desirable. The result is an unavoidable inductance in series with the valve, across the terminals of the tuning capacitor. This inductance is distributed between the external and internal connections of the valve, but for simplification it is shown diagrammatically as a lumped inductor, L_3 in Fig. 4.8.

The effect of this inductance is twofold. First, the anode-to-earth capacitance appears higher at higher frequencies, limiting the top frequency to which the circuit will tune. The second effect causes the level of certain harmonics to be accentuated during operational service and in consequence deserves serious consideration. It applies particularly at the highest operational frequencies when accentuation is given to harmonics of a relatively low order. This occurs when the effective reactance of the L_3C_0 path (Fig. 4.8) is inductive and equal and opposite to the reactance of the capacitor C_1, at a particular harmonic of the fundamental frequency. It is obvious that the value of the inductance L_3 depends on the

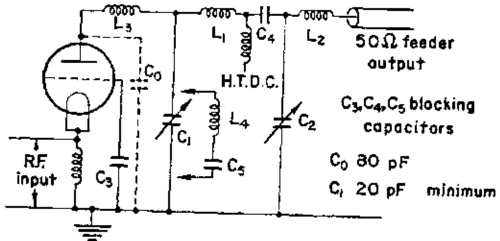


FIG. 4.8 Modified ΠL circuit arrangement for high power.

construction of any particular amplifier, but it has been found in practice that in a well-designed amplifier for 20–30 kW output it is likely to be between 0·03 μH and 0·1 μH.

Using values within this range, examples have been worked out to show the widespread nature of this effect and means of overcoming it. The results are plotted in Fig. 4.9(*a*), (*b*), (*c*) and (*d*), for the 22–30 MHz range of a power amplifier, with an effective anode-earth capacitance C_0 of 80 pF and tuned with a variable capacitor C_1 set at 20 pF for 30 MHz. Table 4.2 gives an analysis of the results. It should be noted that accentuation will occur at frequencies in the near vicinity of the crossover point, as well as at actual resonance.

Reference to Fig. 4.9 (*b*) shows two methods of avoiding this type of resonance if the fifth harmonic is troublesome. Where harmonic resonance occurs at the higher frequencies of the range, it indicates an inductance in excess of 0·05 μH (in the example). This can be avoided by connecting an inductor L_4 in shunt with the tuning capacitor C_1, so that the actual capacitance of C_1 is increased and XC_1 reduced. This new value of XC_1 is shown as a broken line. However, if the inductance is of the order of 0·04 μH, the fifth harmonic will be accentuated over the whole of the frequency range, so the parallel inductor makes matters worse.

For inductance values of 0·05 μH or below, it is preferable to change range at, say, 28 MHz by increasing the main inductor L_1. The new value XC_1 is also shown on the graph, but it also indicates that fifth harmonic resonance might appear near the low-frequency end of the range.

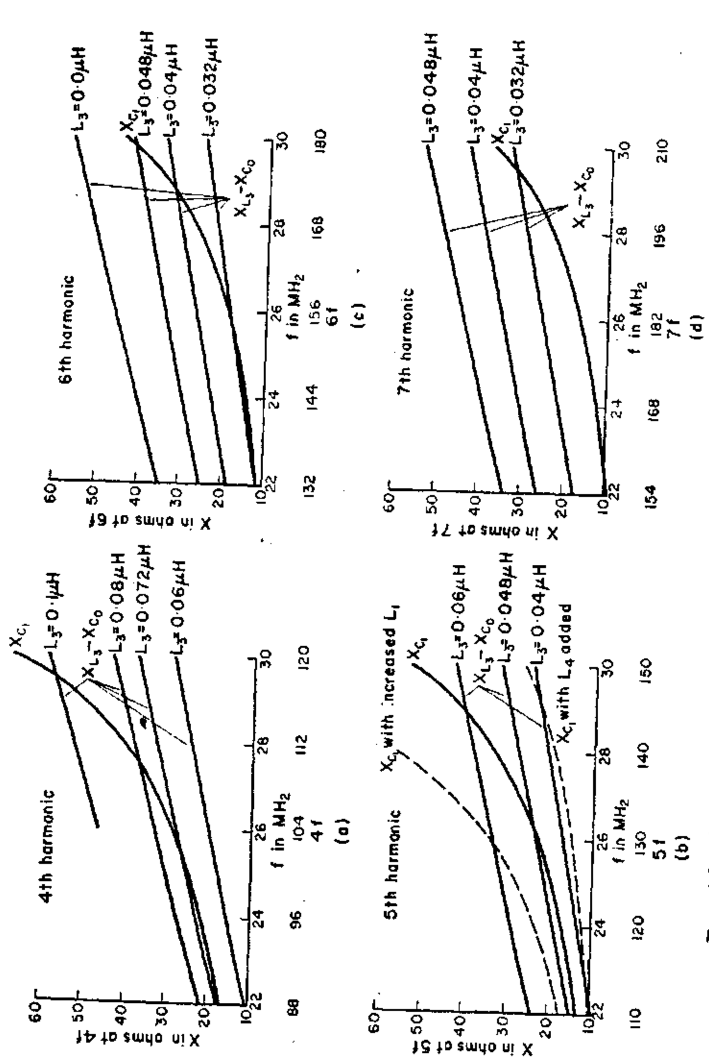


FIG. 4.9 Graphs showing liability for harmonic accentuation in anode circuits.

This emphasizes the point raised at the beginning of this section, that the cathode connections must be effective at r.f. by means of by-pass capacitors with short and wide connectors.

Another effect is the high circulating current in the input circuit, again due to the low input reactance of the valves. With a peak grid voltage of 500 V, the circulating current in the input circuit will be between 40 A and 50 A r.m.s. at 26 MHz for valves with the capacitance values quoted. This means that the input-circuit components must be designed to carry this circulating current without overheating. Even so, the losses are likely to be appreciable and more drive power will be required at the higher frequencies of the h.f. band.

The proportional increase depends on the relative values of the loss-loading and the loading applied to reduce the effects of grid current. In many cases the drive stage will have sufficient power in hand to supply the extra power. However, with the object of providing a substantially constant load, the variation can be reduced to some extent by connecting a small inductor in series with the loading resistor. The effect of the inductor will be to reduce the resistor loading at the higher frequencies and so compensate for the loss-loading, which increases with frequency.

Grounded-grid circuits with triode valves

The most usual application for triodes in a grounded-grid circuit is for linear amplification in the h.f. band. One of the main design problems is to maintain a cathode-ground reactance of sufficiently high value over the whole h.f. spectrum. Although the reactance has to be high only in relation to the low input resistance, it must be remembered that the cathode-heater current will be of the order of 150–300 A for high-power valves.

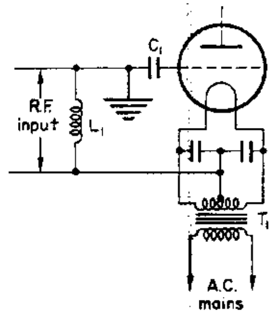


FIG. 4.10 Grounded-grid circuit with low-capacitance heater transformer secondary at r.f. potential.

In one arrangement of the grounded-grid circuit, the cathode-heater supply is fed from a low-capacitance mains transformer (T_1, Fig. 4.10) the secondary of which is at the r.f. potential of the cathode. In this arrangement the cathode choke L_1 has to carry the d.c. plus r.f. cathode current, but not the heater power supply. This circuit is quite satisfactory for a limited range of frequencies at the lower end of the h.f. band, but not the whole h.f. spectrum. However well designed, the transformer will have a complex impedance which will cause a disturbing

resonance either in itself or with other grid-cathode circuit components, particularly towards the higher frequencies.

The choke L_1 also presents a problem. If the inductance value is high enough to have adequate reactance at low frequencies, the effective length of the inductor winding with its self-capacitance will be a half wavelength at some higher frequency within the band. This means that at the half-wave resonant frequency, the choke will present a short circuit to the r.f. drive, thereby making the amplifier unsuitable for a band of frequencies around this resonance.

Therefore, on two counts it is considered that the grounded-grid arrangement shown in Fig. 4.10 is suitable for only a limited range of the h.f. spectrum.

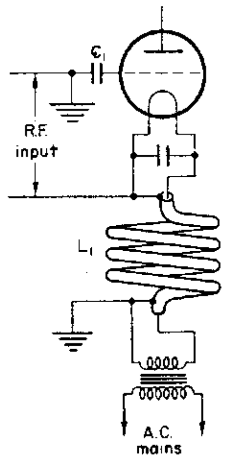


FIG. 4.11 Grounded-grid circuit with cathode choke of low-impedance heavy-current coaxial cable, such as Pyrotenax.

In the circuit arrangement shown in Fig. 4.11, cathode choke L_1 has to carry the heater current as well as the d.c. and r.f. current of the cathode circuit, but the supply transformer is at low r.f. potential. The choke should be made of a very low impedance (approx. 5 Ω) coaxial cable, preferably a heat-resistant type such as 'Pyrotenax'. The problem of obtaining adequate reactance at low frequencies with the avoidance of half-wave resonance at high frequencies, is the same as that for the cathode choke in Fig. 4.10. In consequence, the range of frequencies is somewhat limited, but by suitable design it will cover a 7 to 1 range from 4 MHz to 28 MHz. In spite of this limitation, the very simplicity of the circuit and the absence of spurious resonance paths, makes it quite attractive and it has been used in numerous designs over many years.

The circuit arrangement of Fig. 4.12 shows an improved method of constructing the cathode choke L_1. The inductor element is a short length of 'busbar sandwich', consisting of two flat conductors separated by a strip of insulating material surrounded by a number of ferrite rings. The great advantage is that the ferrite rings increase the inductance to a value suitable for the low frequencies of the h.f. band, yet the length is so short that there is no possibility of a half-wave resonance appearing at the upper frequencies. Furthermore, the ferrite rings are not magnetically saturated by the high level of heater current, because they surround both the go and return leads. It is immediately apparent that cathode chokes of this type are very simple to construct.

The same figure shows the grid connected directly to ground, thereby, as mentioned earlier, eliminating the difficulties associated with grounding by means of capacitors. Isolation of the d.c. grid-cathode path is provided by blocking capacitor C_1. This enables grid-cathode bias to be applied without the possibility of spurious resonances affecting the performance.

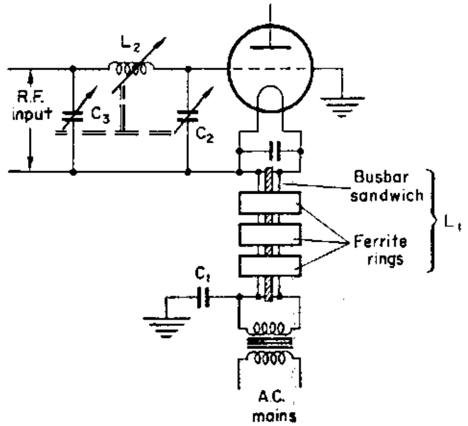


FIG. 4.12 Grounded-grid circuit with cathode choke consisting of a busbar sandwich through ferrite rings.

There are two other factors which must be considered in relation to triodes in grounded-grid circuits. These are the low input capacitance and the low input resistance. The input capacitance of any type of valve is always less in a grounded-grid circuit than it is in a grounded-cathode, because only the cathode-grid capacitance is present. Also, the cathode-grid capacitance of triodes is always less than tetrodes of similar power rating, so the input capacitance of triodes in grounded-grid is considerably less than that of the tetrodes in grounded-cathode.

The effect of low input capacitance is twofold. First, the input r.f. circulating current is relatively low, so the input-circuit components can be smaller and the losses are less. The other effect is a disadvantage at the lower frequencies, particularly in linear applications, where the grid-cathode reactance should be very low at second harmonic frequencies, i.e. the Q factor of the input circuit should be as high as practically possible. With the low input resistance of grounded-grid circuits, typically less than 50 Ω, the low input capacitance means that the Q factor is less than unity at the lower frequencies and still poor at the higher frequencies. So capacitance must be added between the cathode and the grid.

The most convenient type of input circuit to meet these requirements is a Π coupler as shown in Fig. 4.12. Because the circulating current will be relatively

low, even with the terminating capacitor C_2, a variable inductor is quite a practical proposition for the series element.

As the valve input resistance is likely to be rather less than 50 Ω, it is obviously an advantage to consider the coupling from the driving stage to be via a 50 Ω feeder. This means that the impedance change across the Π circuit is quite low, so the shunt components C_2 and C_3 and the series component L_2 can all be ganged together. The circuit then provides a simple one-knob control for obtaining the lowest reflection back along the feeder to the driving stage, by matching the valve input impedance to the 50 Ω feeder. This is not possible at all signal levels, due to the change in input impedance with grid current, but the best possible compromise can be obtained by matching at a mean signal level.

For high-performance, high-power linear amplifiers in the h.f. range, it is recommended that the main features of the input circuit shown in Fig. 4.12 should be used.

5

Cooling Systems

5.1 THE NEED FOR EFFICIENT COOLING SYSTEMS

It will be appreciated that the difference between the total power consumed by a transmitter and the r.f. power at the output terminals is lost within the transmitter itself. It follows that the only way in which this lost power can be dissipated is by raising the temperature of the transmitter components. The essence of any cooling system is to ensure that the temperature rise on all components does not exceed the specified safe limits during operational service, and is sufficient to avoid catastrophic failure under fault conditions.

As the demand for transmitters of greater power output increases, so the cooling problems increase and the need for high overall efficiency becomes more apparent. For example, consider two transmitters designed for 500 kW r.f. power output, one having an overall efficiency of 60% and the other 66·6%. The respective input powers will be 833 kW and 750 kW, so the cooling system for the less efficient equipment must be capable of removing 83 kW more power than that of the more efficient equipment. Also, having removed the heat from the transmitter by means of the cooling medium, there is still the problem of the best means of ultimate disposal. It need hardly be stressed that cooling is one of the major problems of high-power transmitter design, and many designs owe their success to the efficient method of the cooling system employed.

Within the transmitter itself the main bulk of the lost power is dissipated in the final-stage valve, but the other power components can be designed more economically by the judicious use of cooling. In many cases there is a requirement for an 'artificial antenna' load in which to dissipate the r.f. power output for test or lining-up purposes. Where such a load is for use on multi-transmitter stations, it is more convenient for the load cooling system to be a separate entity from the transmitter cooling. But occasions do arise, such as with transmitters operating in parallel, where the r.f. load is a functional part of the transmitter equipment. As such it is better for the cooling system of the load to be integrated with that of the transmitter, thereby forming the major item of the heat-transfer system.

5.2 TYPES OF TRANSMITTER COOLANT

With the earliest valves designed for liquid cooling, an oil of low viscosity was used as a coolant for the anodes. The main reason was that the high resistance of the liquid columns to and from the anode did not increase either the d.c. or r.f. losses. The disadvantages were that oil was not a good medium for heat transfer and a considerable fire risk was involved. Consequently the use of oil as a coolant

was short-lived, and at present the three basic types of coolant used are air, water and 'vapour'.

Although 'vapour' cooling is an expression in general use, a more accurate description would be change-of-state cooling, because it depends on the latent heat energy required to convert water to steam. Colloquially, valves cooled by this method are sometimes called 'kettles', which is probably even more self-explanatory.

Regarding anode cooling, there are no clearly defined power levels for which air, water or vapour are most appropriate. This is best determined by the system proposed by the manufacturer of the valve selected for its electrical properties. Even this is not a clear-cut decision, since many types of valve can be supplied with anode jackets for any of the three systems. Improvements in anode fin structure, and the replacement of glass with ceramics for the insulating envelope, have enabled air cooling to be used for increased power levels. Typically, air cooling is used for transmitters with output power up to about 100 kW.

Taking the other extreme, the development of vapour cooling for very high powers has resulted in such improvements in vapour-cooling techniques that this system is often used for transmitters of less than 100 kW output. Consequently, water cooling is now the exception, rather than, as in the past, the rule. The main application for water cooling is for equipment requiring an artificial antenna load as part of the integrated cooling system. In these circumstances the measurement of output power calorimetrically is often an operational or test requirement.

In all cases of anode cooling by water or vapour, a secondary air system is essential to cool the remainder of the transmitter. This requirement can be the deciding factor in favour of air cooling where a single cooling system is obviously an operational and economic advantage.

Before giving details of individual systems, the following comparison shows typical flows required to dissipate 100 kW.

Air: 8500 ft³/min (20°C temperature rise)
Water: 16 gal/min (20°C temperature rise)
Vapour: 0·6 gal/min of water
(3000 ft³/min of air (approx.) is raised 1°F by 1 kW).

5.3 AIR COOLING

Air flow, pressure and density

The amount of air flow required to cool an equipment depends on the power to be dissipated, the temperature rise permitted on the components and the mean temperature rise of the air stream. The relationship between the air flow, temperature rise and power dissipated depends on the inlet temperature of the air, and is given by the formula

$$\Delta T = \frac{T_K \times W}{164Q}$$

Where ΔT = temperature rise in degrees centigrade
T_K = inlet temperature in degrees Kelvin (°C + 273)
W = power in watts
Q = air flow in cubic feet per minute

In order to show the effect more clearly, the formula has been reproduced in graphical form in Fig. 5.1, in which the power required to raise 1 ft^3/min of air by 1°C is given for inlet temperatures between 0°C and 120°C.

The air pressure required to produce a given air flow depends on the restriction offered to the passage of air by the equipment to be cooled. For a given restriction, the pressure increases approximately as the square of the flow rate, so that if the flow rate is doubled, approximately four times the pressure will be required.

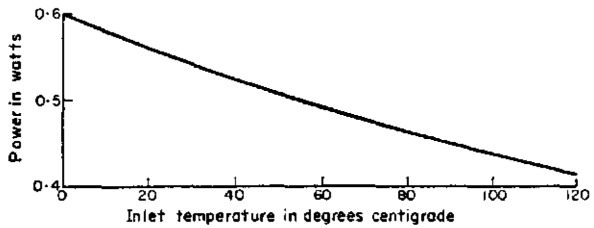


Fig. 5.1 Power required to raise 1 ft^3/min of air by 1°C.

Conversely, if the flow velocity is reduced, the restriction is also reduced, so air ducts should have a large cross-sectional area.

The effectiveness of air as a cooling medium depends on the air density, which is the weight per cubic foot. Therefore, if the density decreases, the rate of air flow must be increased to obtain the same effective cooling. The two main causes of reduced air density are increased temperature and increased altitude, so the

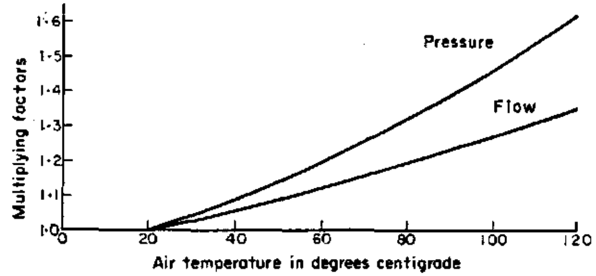


Fig. 5.2 Multiplying factors for pressure and air flow to allow for reduced air density above 20°C.

volume of air flow must be increased for both high-temperature and high-altitude operation. The reduced density with increase in temperature is the reason for the reduced effective cooling with rise of temperature, as shown in Fig. 5.1. As the flow increase is directly proportional to the density decrease, and the pressure increase approximately to the square of the flow increase, the pressure increase is approximately proportional to the square of the density reduction.

Cooling fans deliver a constant air flow in cubic feet per minute irrespective of the air density and their pressure/volume characteristics are specified for operation at ambient temperature (approx. 20°C) and sea-level. In order to select

a fan with suitable characteristics for operation at high temperature and/or high altitudes, correction factors must be applied to the volume and pressure requirements at ambient temperature and sea-level. The correcting factors for inlet temperatures above 20°C are given in Fig. 5.2, and those for altitude above sea-level in Fig. 5.3.

Although a more detailed example is given later in this section, a simple one should clarify the technique. Consider an equipment requiring an air flow of 1000 ft^3/min and a pressure of 2 in. water-gauge for 20°C and at sea-level; but it is to be installed at a site 6000 ft above sea-level where the maximum ambient temperature is 40°C. Taking flow first, Fig. 5.2 gives a factor 1·06 for 40°C and

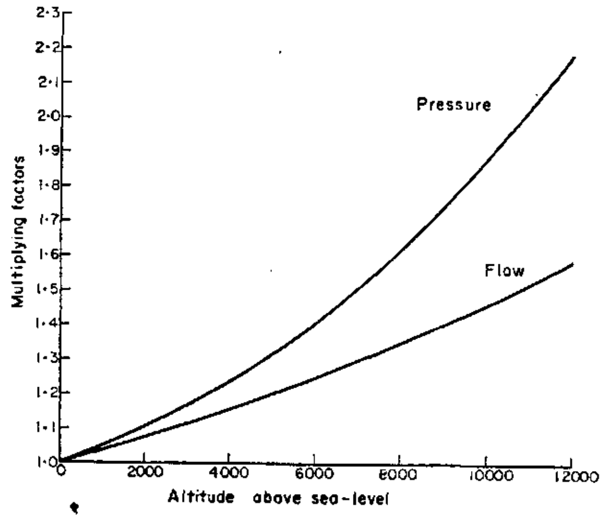


Fig. 5.3 Multiplying factors for pressure and air flow to allow for reduced air density above sea-level.

Fig. 5.3 gives 1·25 for 6000 ft, so the total flow required is $1000 \times 1{\cdot}06 \times 1{\cdot}25 =$ 1325 ft^3/min. Similarly, the factors for pressure are 1·09 and 1·41, so the pressure required is $2 \times 1{\cdot}09 \times 1{\cdot}41 = 3{\cdot}1$ in. water-gauge. This means that a fan must be selected which will deliver 1325 ft^3/min of air against a pressure of 3·1 in. of water at 20°C and sea-level.

Referring to the flow/pressure characteristics of a particular fan shown in Fig. 5.4, it will give 1625 ft^3/min of air against a pressure of 3·1 in. water-gauge. If this fan were to be used, the flow would increase beyond the 1325 ft^3/min required, and the pressure also would increase. In fact, a balance would be achieved at a flow of about 1450 ft^3/min at a pressure of about 3·7 in. water-gauge. This fan would therefore be quite suitable, having a pressure allowance in hand to cater for some increase in air restriction, e.g. the filter being partially blocked with dust particles. It will also be appreciated that the flow was calculated for the worst conditions, so for most of the time the air system will be well within the cooling requirements.

There are several other features of air-cooling systems which must be taken into account.

(1) Any abrupt change in either duct area or direction will cause air turbulence, with accompanying pressure drop—the extent of which is not easy to assess. All changes of duct area or direction should be gradual, by curved sections for change of direction and tapered sections for change of area.

(2) Acoustic noise can be quite objectionable in air-cooling systems, particularly if due regard is not given to size and shape of the air ducts. As a general guide, if an air velocity of 1200 ft/min is not exceeded acoustic noise should not be a problem. It is for this reason that preference should be given to low-speed fans with large inlet and outlet apertures.

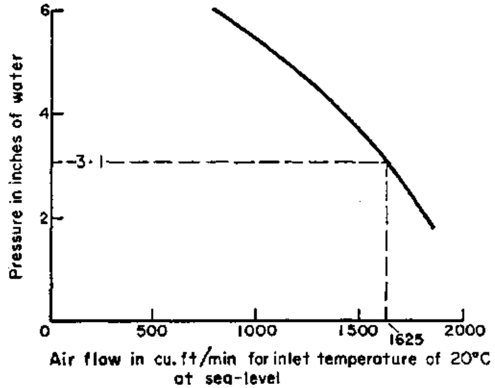


FIG. 5.4 Characteristics of cooling fan.

The air velocity through anode cooling fins is likely to be well in excess of 1200 ft/min, but valve manufacturers have taken considerable care in fin design to ensure low acoustic noise.

(3) The effectiveness of air filters depends on the closeness of the mesh of the filter material. The tighter the mesh the better will be the filtering action, but the greater will be the pressure drop across the filter. To reduce the pressure drop with a given filter material, the area should be as large as possible. To this end, the filter area can be increased by two or three times the duct area by fitting the filter material into a zig-zag formation.

Filters do clog up after a period of time, so some allowance must be made for an increase in pressure drop, and filters should be cleaned regularly and often.

(5) The air inlet and outlet must not be adjacent, otherwise there will be a tendency for the hot air to be recirculated.

(6) It is very important that the air must be circulating before the power supplies are switched on, so the power supplies must be interlocked with the air system in such a way that they will not come on, or conversely be tripped off, if the air flow is less than a predetermined rate.

It is also preferable to fit a two-stage pressure gauge across the air filter. The first stage to give a warning that the filter needs cleaning and the second to trip the power supplies if the pressure drop increases by an excessive amount.

Cooling by air blowing or air suction

Most of the power converted into heat in a high-power transmitter is dissipated at the anode of the final amplifier valve, to which the major portion of any cooling system must be directed. Other items of the equipment can be designed more economically if adequately cooled to maintain their temperature within reasonable limits. There are obvious advantages in using a single cooling system for a complete transmitter. The relative merits of cooling by blowing or by suction must be considered in relation to providing the most efficient method of cooling a complete transmitter with a single system.

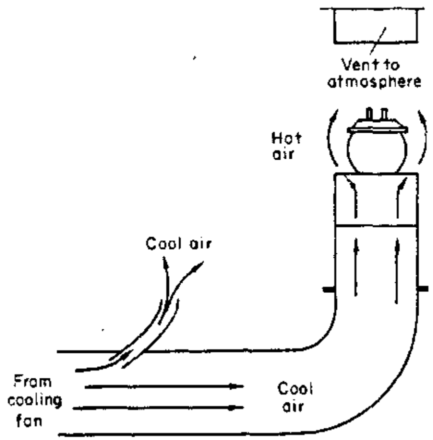


FIG. 5.5 Cooling by blowing.

A simplified diagram of cooling by blowing is shown in Fig. 5.5, for a valve which must be mounted with the anode downwards. The main stream of air is heated by the anode and must be ducted out of the cabinet as directly as possible, because it will be too hot to provide any appreciable cooling for other items. In fact it is likely that it would aggravate the temperature rise on some components. A portion of the cool inlet air must be tapped off the main stream and directed towards the items to be cooled. The pressure inside the whole cabinet will be higher than the outside atmosphere and there will be a tendency for the air to escape through apertures other than the main outlet. This is a good thing from a cleanliness point of view, in that clean air will always be blowing outwards. On the other hand, any doors that are not closed firmly all round will tend to be opened by the internal pressure. (An internal pressure of only 2 in. water-gauge will apply an outward pressure of 0·072 lb/in.², which is a total of 124 lb on a door 6 ft by 2 ft.) This often causes undesirable sparking around doors which are not too well electrically bonded to their frames.

Another disadvantage of the arrangement in Fig. 5.5 is that the hot air from the anode flows directly across the metal–glass or metal–ceramic seals of the valve, which is exactly opposite to the condition recommended by valve manufacturers.

For a valve mounted anode downwards, the alternative method of cooling by suction is shown in Fig. 5.6. It is immediately apparent that the hot air from the anode is ducted directly out of the cabinet via the exhaust fan, and the total volume of air is available for general transmitter cooling. Prior to entering the valve, the air flow can be directed towards likely hot spots by means of baffles and deflector plates. As the general losses are only a small part of the total losses, the temperature of the air entering the valve is not much above ambient, so the air flowing past the valve seals is still relatively cool.

One of the disadvantages of suction cooling is to draw unfiltered air into the cabinet via any aperture, however small. Even so, as the cabinet is at a pressure below atmospheric, any apertures or poorly fitting doors are closed more tightly

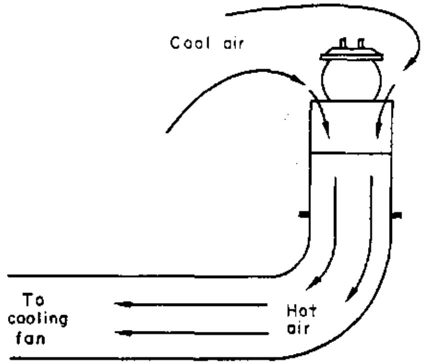


FIG. 5.6 Cooling by suction.

by the external pressure. With a well-designed air system, the pressure drop in the cabinet will be little more than that due to the incoming air filter, so the amount of unclean air entering the cabinet will be a very small proportion of the main volume.

Another point to be considered with suction systems is that the air entering the fan is above ambient by the temperature to which it has been raised in passing through the transmitter. This means that when determining the performance required by a fan from the factors given in Fig. 5.2, the outlet air temperature must be considered, not the ambient temperature. Fans used for suction-cooling must deliver a greater volume of air at a higher pressure than those used for cooling by blowing. It is only of the order of 1·06 times the flow and 1·09 times the pressure for a 20°C rise through the transmitter, but it must be considered.

For valves mounted with the anodes upwards (Fig. 5.7), the relative advantages and disadvantages of air cooling by blowing or by suction are practically the same as for valves with anodes downwards. The main difference is that in the case of blowing the whole cabinet is at the full pressure required to pass the requisite air through the valve. This does mean that all the air is available for general transmitter cooling, but the leakage of air through apertures and loose-fitting doors is aggravated. In fact it can give rise to whistling and other air noise,

which can be most unpleasant. Whether blowing or suction is contemplated, the anode must be surrounded by an insulating tube to ensure that the main air stream does pass through the anode cooling fins. This insulating tube has to be pushed upwards to remove it out of the way for valve changing, but this is not a serious problem.

Some design engineers prefer a balanced air system with a pair of fans, one blowing and the other sucking, both mounted on a common shaft driven by a single motor. This arrangement combines the advantages of suction with the

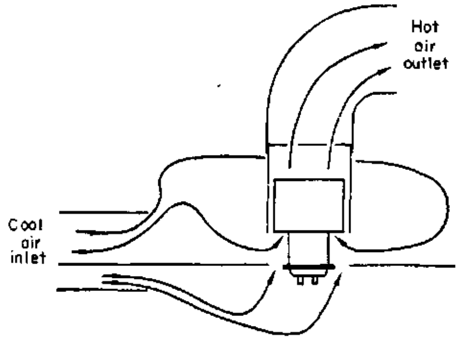


FIG. 5.7 Air cooling for valves with anodes upwards.

ability to direct the blown air more accurately at possible hot spots. The need for such a system depends on the estimated temperature rise in various parts of the transmitter. If the probability of local heating can be reduced by careful initial design, the author considers that a suction system with a simple single fan is the preferred method for cooling transmitters with air.

Typical arrangement for a high-power amplifier with single cooling system

An example of amplifier cooling with a single air system is shown in Fig. 5.8. After passing through the inlet filter, the air first cools the components in the power compartment, after which it separates into two paths. The upper path is via the low-power r.f. stages (if included) the final stage grid/cathode components then passing the valve seals and anode cooling fins to the exhaust fan. The major portion of air flow is in the lower path, cooling the final-stage anode components before passing through the anode cooling fins.

For diagrammatic simplicity, the air inlet and outlet are shown in line with the amplifier. It is often advantageous to duct the air in and out of the back of the amplifier as a means of economizing in floor space and reducing building costs. This does not necessarily mean that the inlet and outlet will be adjacent to one another, because externally the outlet of the exhaust fan can be directed away from the inlet. Underfloor ducts should be avoided, for they add an unnecessary cost and time for installation, particularly if the amplifier is to be fitted into an existing building.

An important point to remember is that it should be possible to change the air filter while the equipment is on power. Filter cleaning is a periodic function which

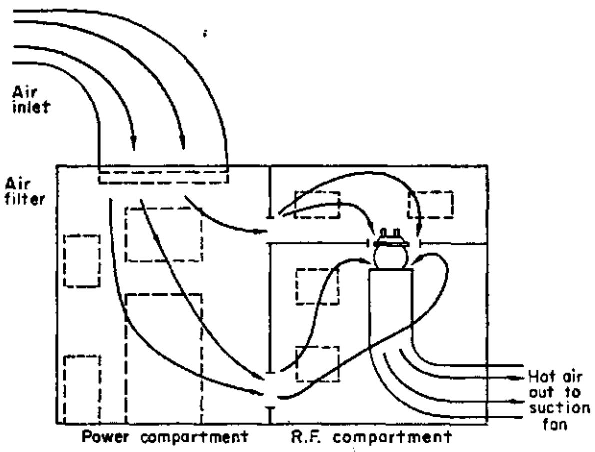


FIG. 5.8 Typical air-cooling system by suction.

can easily be forgotten, and if the filter is not readily replaceable, can result in a considerable loss of traffic.

An example of air flow and air-pressure requirements

Consider the minimum requirements for suction cooling a 30 kW p.e.p. amplifier with an auxiliary rating of 20 kW continuous, of a type which might be installed anywhere in the world.

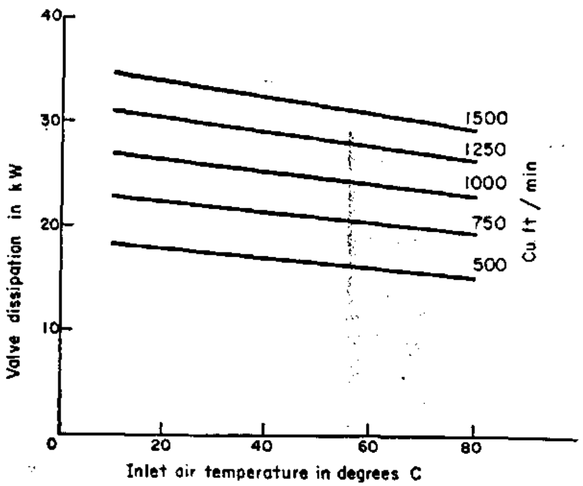


FIG. 5.9 Total valve dissipation relative to air flow and air-inlet temperature.

It was shown in Chapter 2 that the anode dissipation will be highest at the 20 kW rating. A combined allowance for operating in ambient temperature of 40°C, at an altitude of 6000 ft, would be adequate to cover all but the most exceptional cases.

The starting-point is the total valve loss, which is the sum of the anode dissipation, control-grid loss, heater power and screen-grid loss if the valve is a tetrode. Assume this total is 21 kW and the air-cooling characteristics of the valve used are as shown in Figs 5.9 and 5.10. From Fig. 5.9 it will be seen that a total valve dissipation of 21 kW requires a flow of 750 ft^3/min at 40°C inlet temperature, and this flow requires a pressure of 1·1 in. water-gauge, from Fig. 5.10. To this pressure

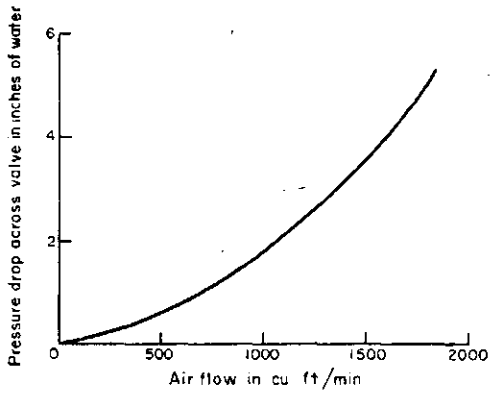


FIG. 5.10 Air flow–air pressure characteristic.

must be added the drop in the ducts, filter and general cabinet cooling. Assume this to be 0·9 in. water-gauge, making a total pressure of 2 in. water-gauge.

Next calculate and/or estimate all the other losses in the amplifier, including transformers, rectifiers, low-power r.f. equipment and final-stage r.f. circuit losses. For the purpose of this example, these losses are assumed to total 4 kW. By calculation from the curve given in Fig. 5.1 from an inlet temperature of 40°C, a power of 4 kW will raise 750 ft^3/min of air by

$$\frac{4000}{750 \times 0{\cdot}522} = 10{\cdot}2°\text{C, say } 10°\text{C}$$

Thus the air entering the valve will be at 50°C for an ambient of 40°C. Checking back on the air characteristics of the valve in Fig. 5.9, 750 ft^3/min of air is just adequate for cooling with an inlet temperature of 50°C.

By calculation from the curve of Fig. 5.1, from an inlet temperature of 50°C, a power of 21 kW will raise 750 ft^3/min of air by

$$\frac{21\,000}{750 \times 0{\cdot}508} = 57{\cdot}7°\text{C, say } 58°\text{C}$$

Thus the air temperature at the valve output and the fan inlet is 108°C (50 + 58). For this temperature at an altitude of 6000 ft, the correcting factors from Fig. 5.2

and Fig. 5.3 are 1·3 × 1·25 for flow and 1·52 × 1·4 for pressure, i.e., a total of 1·625 for flow and 2·13 for pressure. These factors applied to the initial flow of 750 ft^3/min at 2·0 in. water-gauge, give a flow of 1220 ft^3/min at a pressure of 4·3 in. water-gauge. A suitable fan will be one which will give this flow and pressure at *an inlet temperature of 20°C at sea-level.*

Referring to the fan characteristics given in Fig. 5.4, that fan will give 1350 ft^3/min against a head of 4·3 in. water-gauge. Therefore the initial flow would be of the order of 1280 ft^3/min at a pressure of 4·5 in. water-gauge and there would be some allowance for a partially blocked air filter, even under the worst environmental conditions.

If this amplifier were to be air cooled by blowing, the correction factors for altitude would be the same, but the fan-inlet temperature would be reduced from 108°C to 40°C and the fan requirements would be 1000 ft^3/min at 3 in. water-gauge. It would be possible to use a smaller fan, but this must be weighed against the general greater advantages of cooling by suction.

5.4 WATER COOLING

Although the present tendency is to use either air or vapour cooling for valve anodes, the main features of water cooling are discussed in order to give a direct comparison with the other systems.

The main point is that water is a conductor, and anodes are at a high potential with respect to any water supply which may be used, so precautions must be taken to ensure that the d.c. and r.f. losses in the water flowing to and from the anodes are not excessive. This is accomplished by providing a long leakage path by using insulating water-pipes in the form of a column of several turns. Even so, as the flow required is likely to be several gallons per minute, the tube bore cannot be too small, otherwise the pressure restriction could be excessive. The specific resistance of the water should be high, and this depends on the impurity content of the water; ordinary tap water is generally less than 4000 Ω/in^3. and distilled water about 40 000 Ω/in^3.

If tap water is used through 1·125 in. i.d. tubes, 10 ft long, the resistance of the two water columns in parallel will be 240 000 Ω. For a valve with an h.t. supply of 12 kV, the d.c. power lost will be about 600 W. The r.f. losses will be greater because the r.f. power will be dissipated in the top turn or two. Distilled water is recommended in preference to tap water, which means that a closed-circuit system must be used, such as that shown schematically in Fig. 5.11.

The use of tap water has another disadvantage, the water will normally be run to waste. As the flow required is likely to be more than 5 gal/min, this represents a waste consumption of more than 1·3 million gallons per year for a single valve on a single transmitter station operating only 12 h per day. A similar amount of secondary cooling water would be required for a water-to-water heat exchanger, instead of the water-to-air type shown in Fig. 5.11. The problem of wastage would be equally serious with tap water, but if a transmitter is to be installed adjacent to a large supply, such as a river, then a water-to-water heat exchanger is an obvious choice.

The corona shields on top of each hose-column serve two purposes by distributing the r.f. field across a greater portion of each column. The r.f. losses are reduced and the tendency to ionization, at the junction of the metal/insulating

tube, is virtually eliminated by reducing the potential gradient at that point. The corona shields do add capacitance to the anode circuit, which is a disadvantage in the h.f. band but not serious at m.f.

The same metal/insulating tube junction point is particularly prone to corrosion and electrolysis, which can reduce the water resistivity and increase the flow restriction by silting. Protective targets must be fitted adjacent to this point to reduce corrosion and electrolysis and these must be replaced periodically, as they disintegrate. Corrosion and electrolysis can also be set up by different metals in the pipe system. As the valve anodes are made of copper, the whole cooling system should be of copper, to the exclusion of iron, aluminium and brass.

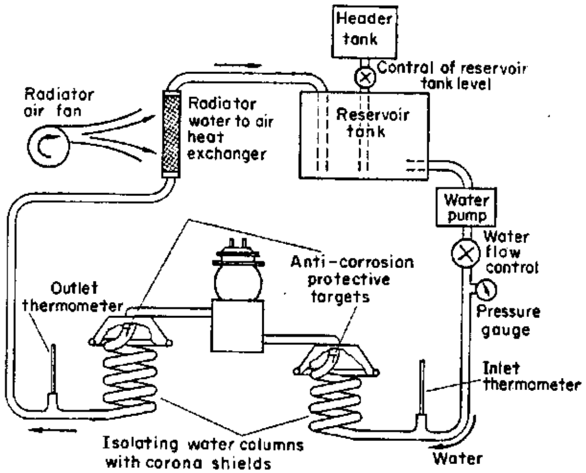


FIG. 5.11 Closed-circuit water cooling system.

If air is absorbed into the water, its cooling efficiency decreases, so splashing should be avoided by feeding water into the reservoir tank well below the water-level.

When equipment is to be installed in cold climates, anti-freeze mixtures can be added to the water, but care must be taken that the type used does not reduce the water resistivity. For this reason ethylene glycol must not be used. Any additive to distilled water reduces the cooling efficiency, so the flow must be increased correspondingly, and, consequently, so must the pressure. This will increase the pressure demand on the circulating water pump, which has to overcome all the restrictions imposed by the pipes, radiator and water head, in addition to that of the valve itself.

Temperature considerations for a water-cooling system

Valve manufacturers specify the cooling requirements in terms of flow/pressure and flow/dissipation for different temperatures of water inlet. Those for flow/dissipation are based on the mean temperature of water at the valve outlet being sufficiently below 100°C to ensure that boiling does not occur at hot spots within the anode jacket. A typical value selected is 70°C.

In a closed-circuit system, ignoring general radiation and convection cooling, the temperature drop across the heat exchanger is bound to be the same as the temperature rise across the valve. With a water-to-air heat exchanger, the effective cooling of the water and air are the same, and a compromise decision on the relative temperature change across valve and exchanger must be made. If the water flow is doubled so that the temperature rise across the valve is reduced from 40°C to 20°C, then the air cooling has to drop the temperature by 20°C instead of 40°C. However, with an upper limit of 70°C at the valve outlet, the decision mainly rests on the ambient temperature of the air into the cooling fan of the heat exchanger.

Consider cooling the valve of a modulated r.f. amplifier, giving an output of 100 kW, suitable for operation at an altitude of 6000 ft and an ambient temperature of 40°C. Assume that the temperature rise across the valve is equal to the temperature drop across the heat exchanger, i.e., the valve-inlet temperature is 55°C and the outlet 70°C. For an anode conversion efficiency of 80%, the anode loss will be 25 kW on carrier and 37·5 kW at 100% modulation. Allowing for grid losses and heater power, the total power to be considered for cooling will be of the order of 45 kW.

A useful formula to remember is that 189 gal/h of water will be raised by 1·0°C by 1·0 kW, i.e., if the water flow is adjusted to 189 gal/h, the power in kilowatts is the same as the temperature rise in degrees centigrade. On this basis, the water flow required to limit the temperature rise across the valve for 45 kW dissipation is given by

$$\frac{189 \times 45}{60 \times 15} = 9{\cdot}45 \text{ gal/min}$$

The equivalent power must be removed from the water in the heat exchanger to reduce the temperature by 15°C from an ambient of 40°C. Referring to Fig. 5.1, 0·522 W will raise 1·0 ft³/min of air by 1·0°C from an inlet temperature of 40°C. Therefore, 45 kW will be dissipated, at sea-level and an ambient of 20°C, with a temperature rise of 15°C by

$$\frac{45\,000}{0{\cdot}522 \times 15} = 5750 \text{ ft}^3\text{/min of air}$$

If under the same conditions the pressure drop across the heat exchanger is 1·5 in. water-gauge, for an ambient of 40°C at an altitude of 6000 ft the fan requirements from Figs 5.2 and 5.3 will be 7700 ft³/min at a pressure of 2·3 in. water-gauge. If the useful area of the radiator is 9 ft² (3 ft × 3 ft), the air velocity will be 850 ft/min, so the acoustic noise should not be objectionable.

It must not be forgotten that the water system cools the valves only and an additional air system is required for cooling other components. For example, assume the total general losses to be 8 kW and the maximum temperature permitted on some components is 70°C, then the *mean* outlet temperature should not exceed 60°C, i.e., a rise of 20°C from an ambient of 40°C. Reference to Fig. 5.1 shows that 770 ft³/min of air will be required. Assume the total pressure drop in the air system is 2 in. water-gauge for the flow and that cooling by blowing and suction are to be considered. For blowing, the multiplying factors for 40°C and 600 ft from Figs 5.2 and 5.3 are 1·325 and 1·55 for flow and pressure, respectively. Thus the fan required is one which will give 1020 ft³/min of air at 3·1 in. water-gauge at 20°C, at sea-level.

For a suction system, because the inlet temperature is 60°C, the flow and pressure factors increase to 1·4 and 1·68, respectively. So the fan required is one which will deliver 1080 ft³/min against a head of 3·4 in. water-gauge, at 20°C at sea-level.

Water-cooled r.f. load

One of the main advantages of water cooling is the ready and accurate means of power measurement, from a knowledge of the temperature rise and water flow rate. A typical water-cooled load for measuring r.f. power is shown in Fig. 5.12.

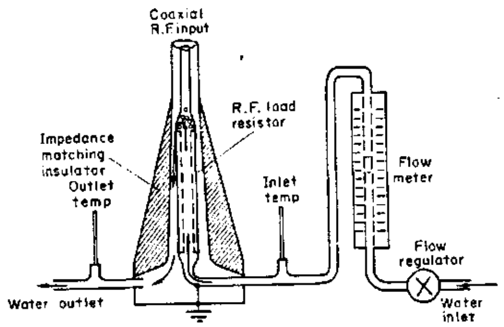


FIG. 5.12 Water-cooled r.f. load.

The actual load element is a ceramic tube, coated on the outside with a thin metallic film of resistance material and having the same resistance as the feeder impedance, usually 50 Ω. Although only rated at 50 W in free air, with suitable water cooling it is capable of dissipating 30 kW of r.f. power continuously. The load resistor is connected directly between the coaxial feeder inner and earth, with the water passing up the inside of the tube and down round the outer. Thus the water is fed in and out at earth potential. The shape of the outer metal case and the insulator are so designed that the load offers a matched impedance over a wide frequency range, typically from l.f. to 200 MHz.

Power measurement is simplified by adjusting the water flow to 189 gal/h, when the temperature rise in degrees centigrade is the power output in kilowatts. It is important that the thermometer bulbs are fully immersed in the water and that water turbulence is avoided, otherwise the thin film will burn out due to local hot spots. A great advantage of this type of load is that it is completely screened and unwanted radiation is negligible.

5.5 VAPOUR COOLING

The advantages of vapour cooling

The basic idea of cooling valves by allowing water to boil in anode jackets is by no means new, for such a system was patented in 1934 [1]. It is only comparatively recently that the advantages of the system have been generally accepted.

This is probably due to the increasing demand for transmitters of higher power, together with valves being designed specifically for vapour cooling.

It is advisable to follow the cooling arrangements recommended by manufacturers of valves for vapour cooling. For this reason most of the information given in this section is extracted from a comprehensive description of vapour cooling given in a booklet issued by Eimac [2].

The main advantage stems from the high level of energy required to convert water to steam vapour. To transform 1 g of water at 100°C to steam requires 540 cal, whereas to raise 1 g of water by 1·0°C requires only 1 cal. The effect of this difference is shown clearly by comparing the amount of water required by the two systems to dissipate the same power.

In a water-cooling system, with a water-to-air heat exchanger, it has been shown that the temperature rise across the valve is typically half the difference between the temperatures of the ambient air and of the maximum permitted at the valve outlet. For an ambient temperature of 20°C and a valve outlet temperature of 70°C, the temperature rise across the valve will be 25°C; the energy required being 25 cal for 1 g of water. Therefore the energy absorbed by converting water at 100°C to steam would be 21·6 times (540 ÷ 25) that required to raise 1 g of water by 25°C. In terms of water flow, 3·15 gal/min (189 gal/h) will be raised 25°C by 25 kW, whereas to absorb 25 kW in converting water to steam requires only 0·146 gal/min. For an ambient temperature of 40°C the difference is even more significant, in that 5·25 g/min will be required in a water system to dissipate 25 kW with a temperature rise limited to 15°C, i.e., the water-cooling system requires thirty-six times the water flow, compared with a vapour-cooling system. In fact the flow is so low that the water circulates due to thermosyphoning action in a well-designed system, and a circulating pump is not required.

Another result of the low water flow is that a short length of small-bore pipe can be used for the inlet water to the anode, with negligible power loss. For example, assuming the resistivity of distilled water is 40 000 Ω/in.^{3}, the resistance of a 2 ft column of water in a $\frac{1}{2}$ in.-bore pipe will be approximately 5 MΩ. With a 12 kV anode supply the d.c. loss will be about 30 W and the r.f. loss will be quite low due to the small bulk involved.

The outlet from the valve being steam vapour, the resistivity of which is many times that of distilled water, quite a large-bore pipe can be used for the valve outlet with negligible power loss. The large corona rings necessary with long water columns are not needed, so the minimum anode capacitance is less with a vapour system. However, electrolytic targets must be used to prevent corrosion and water contamination.

Turning to the heat exchanger, a condenser of any given thermal capacity can be reduced in size if the mean temperature gradient between the cooled liquid and the secondary coolant can be increased. In a water-cooled system the water enters the heat exchanger at 70°C and leaves at about 54°C, the mean temperature being 62°C. With a water-to-air heat exchanger and an air temperature of 40°C, the mean temperature differential is 22°C (62 − 40). In a vapour-cooling system vapour enters at 100°C and water leaves at 100°C, so the mean temperature is also 100°C. The mean temperature differential between steam–water and air is then 78°C (100 − 22) which is rather more than three times that of the water-cooled system. The secondary coolant requirements of the vapour system are between one-third and one-quarter of those for a water system.

For an air-cooled heat exchanger, advantage of this higher temperature gradient can be taken to reduce the size of the condenser and blower equipment. In fact, where the power involved is relatively low, it is often possible to obtain sufficient condensation without a cooling blower, thereby enabling the complete cooling system to be operated in silence, without any rotating machinery.

For very high power, where water is necessary as the secondary coolant, a similar reduction in condenser size is obtained with a vapour-cooling system. Even where equipment is installed adjacent to a ready supply of secondary water, such as a river, there is obviously a lower initial cost. Where tap water is run to waste for the secondary coolant, there is the additional advantage of lower running costs.

A typical vapour-cooling system

A typical vapour-phase cooling system is shown in Fig. 5.13, in which the secondary coolant arrangement has been omitted for simplicity. The diagram is

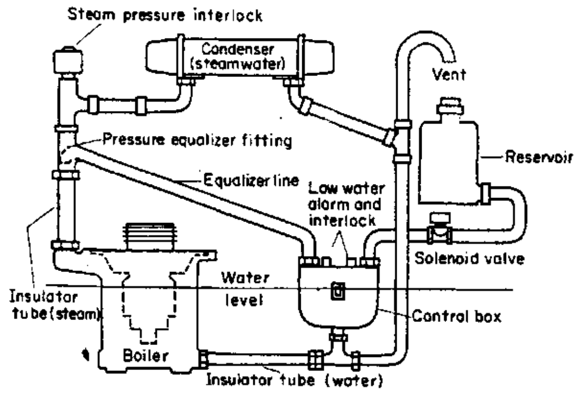


FIG. 5.13 Typical vapour-phase cooling system (Eimac).

largely self-explanatory, but the essential feature of the system is that the level of water in the boiler (valve jacket) must be controlled within fine limits. This function is performed by the control box, in conjunction with the solenoid valve from the reservoir tank, in accordance with the water flow dictated by the valve dissipation under operating conditions. The control box also provides an alarm, to give warning that the water level is approaching predetermined limits, and also an executive interlock which trips the power supplies when the limits are reached.

The pressure-equalizing line is an equally important feature. Without this line, the higher pressure on the vapour side of the system would depress the water level in the boiler, relative to the level in the control box. The relative levels would be determined by the boiling-rate, i.e., the valve dissipation, and the function of the control box would be nullified.

The actual water-level limits within the boiler are determined by the design of the anode fin structure and are specific for each valve type. These limits are predetermined to ensure maximum cooling efficiency consistent with the prevention of hot spots and liability to valve damage. In order to achieve these optimum conditions, a thorough understanding of the laws of heat transfer and thermodynamics is necessary for the design of anodes for vapour cooling. While this is not necessary in the present context, some aspects of these laws are worth considering in order to appreciate the need for maintaining the water level within the designed limits.

Heat-transfer characteristics

The Nukiyama curves shown in Fig. 5.14 present the heat-transfer capability (in watts per square centimetre) of a heated surface submerged in water at various

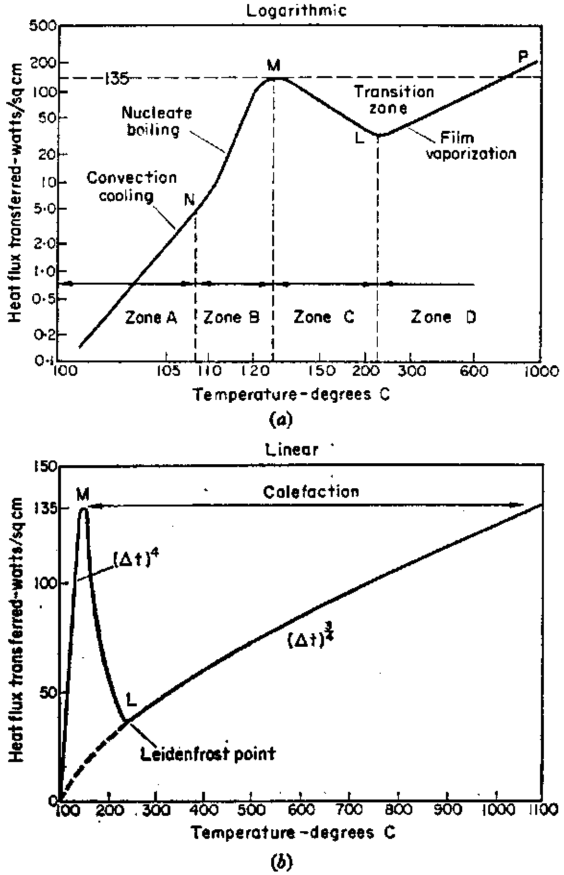


FIG. 5.14 Nukiyama heat-transfer curves.

temperatures. The first portion of the curve, zone A, indicates that from 100°C to about 108°C, heat transfer is a linear function of temperature differential

between the hot surface and the water, reaching a maximum of about 5 W/cm^2 at that temperature, point N. This is the so-called convection-cooling zone. Boiling takes place in the heated water at some point away from the surface.

From 108°C to 125°C, zone B, heat transfer increases as the fourth power of the temperature differential, until at 125°C, point M, it reaches 135 W/cm^2. This zone is characterized by nucleate boiling; i.e., individual bubbles of vapour are formed at the hot surface, break away, and travel upward through the water to the atmosphere.

Above 125°C an unstable portion of the Nukiyama curve is seen, where increasing the temperature of the heated surface actually reduces the unit thermal conductivity. In this area, zone C, the vapour partially insulates the heated surface from the water, until about 225°C is reached on the hot surface. At this point, called the Leidenfrost point, the surface becomes completely covered with a sheath of vapour and all the heat transfer is accomplished through this vapour cover. Thermal conductivity of only 30 W/cm^2 is realized at the Leidenfrost point, which is only about one-quarter of the thermal conductivity realized at point M, 125°C.

From this point through zone D, the 'film vaporization zone', the heat transfer increases with temperature until, at about 1000°C, the value of 135 W/cm^2 is again reached.

The linear plot of the Nukiyama curve indicates that zones A and B are relatively narrow areas and that a heated surface with unlimited heat capacity will tend to pass from zone A to zone D in a short time.

This irreversible super-heating is known as 'calefaction'. For a cylindrical vacuum valve anode the passing into total calefaction would not be tolerable, as any unit heat-transfer density above 135 W/cm^2 would result in temperatures above 1000°C, well above the safe limits for the valve.

In order to avoid the danger regions and at the same time obtain a high density of heat transfer, the surface area of the anode cooling fins is made several times greater than the internal surface area of the anode.

When operating at low dissipation levels, boiling takes place at the root of the fins. Increasing power density causes this boiling area to move out towards the ends of the fins, until at rated dissipation boiling is taking place on the outer half of the fins. Present-day materials and techniques dictate that the outer edges of the anode fins always remain at less than 125°C.

Comparison with other cooling systems

In common with water cooling, vapour-cooling systems for valve anodes require an additional air system for other items of transmitter equipment which require cooling. Both systems also require a secondary cooling system using either air or water as the coolant. Thus both water and vapour systems require more equipment than air-cooling systems. They are consequently more costly, and more maintenance will be required during operational service. Hence the author's preference for wholly air-cooling systems wherever practical and a pointer in favour of high conversion efficiency, particularly at high power levels.

Due to the advantages of vapour systems over water systems shown earlier in this section, vapour systems are preferred for powers too high for air systems, accounting for the decreasing tendency to use water-cooling systems.

REFERENCES

[1] PRIVETT, P. E. 'Improvements in or relating to methods of and means for cooling thermionic valves'. Patent specification No. 423,891, application date Feb. 3, 1934.

[2] *Care and Feeding of Power Grid Tubes.* Eimac division of Varian, pp. 121–138.

6

Power Amplifier Design for H.F. Communications

6.1 OPERATIONAL REQUIREMENTS

The essential features of any communications transmitter are linearity, ready means of frequency changing throughout the h.f. band and reliability.

Linearity

Linearity is dictated by the operational requirement to radiate any type of traffic on a multi-channel basis. In order to fully exploit the transmitter by using as much power as possible on each channel, without objectionable cross-talk or too high a level of out-of-band radiation, the linearity should be of a high order. Multi-channel operation also involves i.s.b. techniques, and it is unlikely that transmitters having p.e.p. in excess of 30 kW will be required. In fact it is probable that satellite links will take over many of the long-range h.f. circuits, so that the future demand will tend towards h.f. powers of the order of 7–10 kW p.e.p.

As valves in classes C and D are non-linear, the remaining choice between classes A and B is biased strongly in favour of class B for high power because of the poor conversion efficiency in class A. If triodes are used, the difficulties associated with high-level neutralizing preclude the use of grounded-cathode circuits, and grounded grid is the only effective solution, with the attendant requirement of a high level of drive power. Tetrodes can be used in grounded-cathode circuits and the drive power required is relatively low, but an additional power supply is necessary for the screen, with associated interlocks. The choice between triodes and tetrodes is mainly a question of the relative availability and cost of a high-level drive or a screen supply. Pentodes fall within the same category as tetrodes but they are not usually available with high power rating.

Frequency changing and reliability

The necessity for changing frequency falls into two categories. In one case the traffic ceases on a particular route and that transmitter is used for another service, which may or may not require a frequency change. In the other category, in order to continue traffic on the same route, a frequency change is necessary due to changing propagation conditions, such as occur around sunrise and sunset. In either category the time taken to change frequency should be as short as possible, for tactical or economic reasons, depending on the application being military or commercial.

In the case of transferring a transmitter to another route or service, in general only one transmitter will be involved at a time. Consequently, providing that the frequency changing process is simple, a manual change by one man can be accomplished in a short time; the more simple the process, the less traffic time will be lost.

Where frequency changes are necessary because of changing propagation conditions, it is likely that a number of routes will be affected at about the same time. On multi-transmitter stations, this means changing frequency on several transmitters almost simultaneously. For manually tuned transmitters it calls for quite a number of operating staff who will only be required about twice each day. If adequate staff are not available, there is a distinct possibility of lost traffic time, not only through the inability to retune several transmitters simultaneously, but also due to the greater liability of human errors when doing things in a hurry. It follows that a degree of unreliability is introduced.

To overcome both the technical and staff difficulties, it is quite common practice on manually tuned multi-transmitter stations to operate two transmitters on different frequencies during the changeover periods. Obviously this arrangement requires a greater number of standby transmitters, but it does show the importance which operating organizations attach to obtaining the lowest possible lost traffic time.

On all counts there is a clear case for tuning transmitters automatically. In its simple form, automatic tuning consists of controls being changed between a number of pre-set positions, previously determined by manual tuning. The first disadvantage of this system is the limited number of frequencies which can be selected and the time taken to set up a frequency for which a pre-set position is not available. The second disadvantage is the inability of pre-set arrangements to cater for final-stage tuning and loading corrections, necessary to compensate for changes in feeder impedance due to changing weather conditions influencing antenna and feeder/transmission line characteristics. During operation, either the power must be reduced in bad weather conditions or periodic tripping is likely, with consequent loss of traffic.

Both these disadvantages can be overcome by automatic self-tuning, in which the final-stage anode/output-circuit controls are always in operation, so that the final-stage valve continues to operate in the optimum condition in spite of changes in feeder impedance.

In common with any electromechanical device, automatic tuning is liable to fail at some time. The more complicated the device, the greater the liability to failure, and the greater the number of devices, the still greater the likelihood of a fault occurring. Automatic self-tuning is necessarily somewhat complex, so for the greatest reliability the number of circuits which have to be self-tuned should be reduced to an absolute minimum.

6.2 THE PRINCIPLE OF SELF-TUNING

In manually tuned equipments it is normal to tune to a minimum d.c. current to the valve anode. This is unsuitable as an indication for self-tuning because the response curve is flat near resonance, and direction-sensitive information is not available. Furthermore, with tetrodes, if the d.c. meter is in the cathode circuit,

as is usual, the resonance indication is even more flat, because the screen current rises to a peak at tune.

A more accurate indication of tune is given by the relative phase of the r.f. voltages on the grid and anode, which are in antiphase at anode circuit resonance. If samples of these two r.f. voltages are compared in a phase discriminator, an error voltage can be obtained in which the sense is dependent on the phase being leading or lagging. This error voltage can be used to drive the tuning motor towards the resonant point. This method was first tried in 1945 [1], but it was not considered to be sufficiently reliable until solid-state devices were available to replace the valves of the original model.

The principle of a phase discriminator can be seen by reference to the vector diagram, Fig. 6.1. Two antiphase r.f. voltages $V_{a'}$ and $V_{a''}$ are derived from the

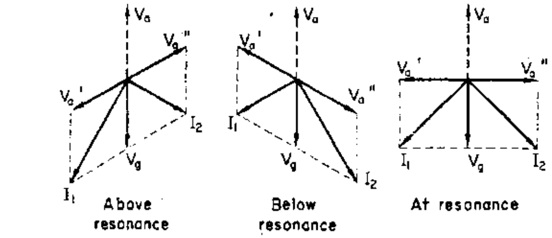


FIG. 6.1 Vector diagrams of the phase relationship in phase discriminator (Fig. 6.2).

anode voltage V_a, and at resonance these voltages are in quadrature with the grid r.f. voltage V_g. Voltages $V_{a'}$ and V_g are applied to one diode, D_1, while voltages $V_{a''}$ and V_g are applied to a second diode, D_2. Vectors I_1 and I_2 represent the amplitudes of the rectified currents in the load resistors of diodes D_1 and D_2, respectively, and at resonance they are equal. Above and below resonance these currents change rapidly in a differential manner and can be combined to give an error signal indicative of the direction in which the anode circuit is off-tune.

Anode-circuit tuning

The main elements of self-tuning an anode circuit by means of a phase discriminator are shown in Fig. 6.2. Capacitors C_1 and C_2 are low-capacitance high-voltage types, pre-set to give approximately the same voltages on the diodes D_1 and D_2. This is not a critical setting, because phase discriminators are insensitive to anode and grid voltage levels over a very wide amplitude range.

This arrangement is effective in the vicinity of resonance where the phase changes very rapidly, but the sensitivity falls off on either side, giving a limited capture range. Because of the rapid increase in sensitivity as resonance is approached, a high degree of stabilizing feedback is required in the servo-system to prevent hunting, thereby further limiting the capture range. It is necessary to provide additional tuning information in order to bring the anode-circuit tuning within the capture range of the phase discriminator. This is relatively coarse tuning information which also serves to prevent mis-tuning on to an harmonic frequency. The method of obtaining this coarse information is described in Section 6.3.

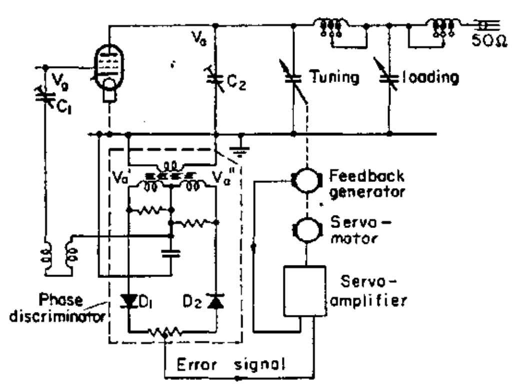


FIG. 6.2 Fine-tuning section of self-tuned amplifier.

Automatic loading

Provided that no grid current is drawn in a linear amplifier, the ratio of anode-to-grid r.f. voltage can be taken as a measure of the anode load impedance and used to provide an error signal to operate the loading control. This linear relationship does not hold if the valve is driven into grid current, so a system based on these two quantities would tend to increase loading when in the grid current region, but would be quite satisfactory at low signal levels.

At high signal levels, the ratio of anode r.f. voltage to d.c. cathode current provides a reasonably true measure of loading, but at low signal levels the relationship is neither sufficiently linear nor accurate for loading information. However, a satisfactory arrangement can be obtained by combining the two systems, as shown in Fig. 6.3, wherein a level discriminator samples the r.f. grid and anode

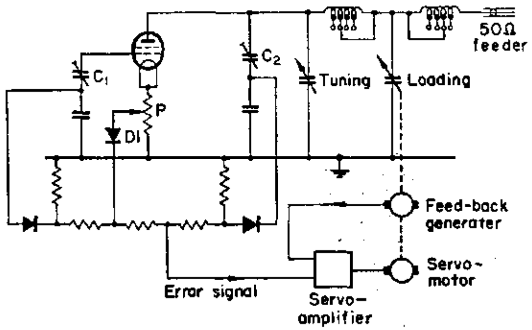


FIG. 6.3 Loading discriminator of self-tuned amplifier.

voltages at low signal levels. At higher levels, where the cathode current is rising faster than the grid voltage, the clamping diode D_1 conducts and the cathode current sample predominates.

A complication arises in the loading sequence due to the fact that the information from the loading-level discriminator is only correct when the anode circuit is in tune. When out of tune, the anode r.f. voltage will be low and the discriminator will tend to unload the valve to increase the anode voltage. As the loading capacitor affects the anode tune to some extent, a situation can arise where, if the circuit is out of tune on the low-frequency side, unloading will further increase the total circuit capacitance. In fact, at the highest tunable frequency the tuning and loading controls could run to the opposite ends of the scale and stay there.

These defects can be overcome by ensuring that the first step is always to bring the anode circuit into tune bv arranging for the output of the tuning-servo feedback generator to mute the input to the loading servo-motor, so that it does not move until the tuning motor stops. The whole tuning and loading sequence then takes place in a series of successively smaller steps, until the correct conditions are reached.

Setting the loading conditions

The actual loading conditions are determined by the required linearity and output power. As a general rule the linearity improves with increased loading, which is decreased conversion efficiency, so it is usual to set the loading to give the required linearity at the best conversion efficiency.

The elements which control the loading are capacitors C_1 and C_2 and potentiometer P, Fig. 6.3. All these controls are independent of frequency, thus a constant loading is obtained over the frequency band in terms of valve output for a given drive voltage. The drive will also be controlled at a constant level throughout the band, so the output at the transmitter terminals will be lowest when the circuit losses are at a maximum. This will normally be on the highest-frequency range and at the lowest frequency of that range, when the anode circuit is a tapped-inductor variable-capacitor arrangement. At the frequency giving the lowest efficiency, the full specified output must be guaranteed by allowing an extra 5% or 10% power to cover such differences as might be given by different valves and different equipments. At other frequencies there will be more power in hand, but not a lot if the circuits are well designed.

Some allowance must also be made for linearity performance, which is usually expressed in terms of intermodulation products (i.p.'s) produced during a two-tone test. Typical specifications call for i.p.'s to be not greater than −36 dB in relation to the level of either of two equal test tones. In order to guarantee i.p.'s of −36 dB, a figure of −40 dB should be obtained in proving tests over the whole band, and especially under the worst conditions. To achieve this figure with reasonable conversion efficiency, it is often an advantage to incorporate a frequency-independent circuit to improve linearity, such as that described in Chapter 3, Section 3.3.

6.3 THE INPUT CIRCUIT AND COARSE TUNING

It was pointed out in Section 6.2 that reliability is improved by reducing the number of automatically tuned circuits, so if only one stage is tuned the optimum reliability will be obtained. This can be achieved with a self-tuned final stage, driven by a system of wideband amplifiers covering the whole frequency spectrum without tuning. The time required for a frequency change is also reduced by this means, because a signal appears at the final-stage input immediately after the frequency-determining source has been changed.

The high-input capacitance of power valves, particularly tetrodes, does present a problem with this arrangement. It means that the impedance of the input line would have to be very low to maintain an impedance match over the whole frequency band. This in turn would mean a low-value terminating resistor and an appreciable power output from the driving amplifier, which is not required by the final-stage valve. For example, the input capacitance of a tetrode to give 10 kW output is about 150 pF and a peak driving voltage of the order of 200 V will be required. The reactance of 150 pF at 27·5 MHz is 39 Ω, which corresponds to a mismatch of 1·4 v.s.w.r. on a 15 Ω line, and a drive power of 1·3 kW would be required.

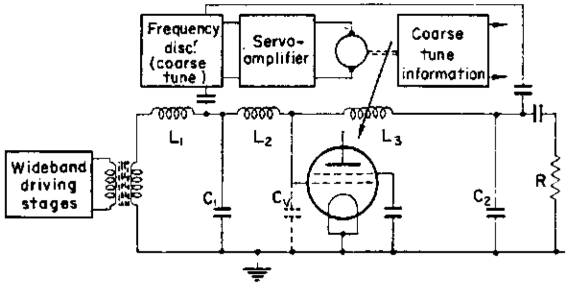


FIG. 6.4 Final amplifier input circuit for self-tuning.

In order to reduce the drive power required, the valve-input capacitance can be partly tuned out, thereby enabling a higher impedance line and termination to be used (with a line impedance of 200 Ω, only 100 W would be required). By tuning the input circuit an additional control is required, but the setting is not at all critical and the tuning can be automatic from information supplied by a frequency discriminator on the input line. Furthermore, the angular position of a shaft, geared down from the driving-motor spindle, can be used to provide coarse tuning and loading information for the anode and output circuits of the final stage. Whatever method is used to supply coarse tuning information, a motor would be required for complete automatic control, therefore the input-tuning motor is not really additional, it is made to perform two functions.

A block diagram of the input circuit for tetrode valves is shown in Fig. 6.4. The input is fed through a wideband transformer to the input line, which consists of series elements L_1, L_2, L_3 and shunt elements C_1, C_V, C_2, and is terminated by resistor R. The capacitance C_V is the input capacitance of the tetrode and is much higher than required for a shunt component of the line, so it is effectively reduced by adjusting the series element L_3.

The frequency discriminator provides a direction-sensitive error signal, which is amplified and used to drive the series inductor L_3 to such a value that the line is matched to the terminating resistor R.

With this system the input-circuit tuning process takes place before the anode and screen supplies are switched on, so the anode and output circuits are set

approximately to the correct settings prior to the application of these supplies. The settings are determined by the input frequency only, no pre-setting is required for any frequency within the band of the transmitter.

For triode valves in a grounded-grid circuit, the cathode load provides the termination for the input line, but this is not present until the anode supply is on. Consequently, when triodes are used the line must be terminated by a resistor, the value of which is equivalent to the cathode-load resistance when in operation. This resistor is automatically switched out of circuit on completion of the input-tuning process, before the anode voltage is applied.

The value of the cathode-load resistance is likely to be less than 50 Ω, so the power required from the driving stage will be much higher than that required for tetrodes. However, the major portion is fed into the anode-cathode load, and being in series with the anode output appears as part of the valve output.

6.4 A COMPLETE SELF-TUNING ARRANGEMENT

The tuning and loading process

A block diagram for automatic tuning of the input, anode and output circuits of a linear amplifier is shown in Fig. 6.5. For simplicity the power supplies, interlocks and drive switching are not shown, but they do come into the automatic frequency-changing process.

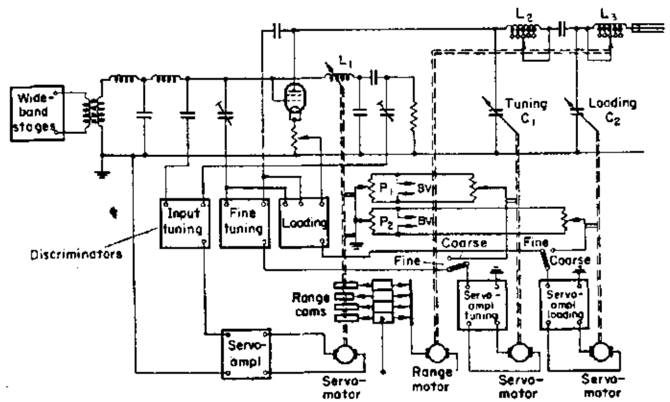


FIG. 6.5 Block diagram of complete automatic system for a self-tuned amplifier.

Probably the best explanation of the system can be given by detailing the sequence of events from the end of traffic on one frequency to traffic commencing on the new frequency.

The only manual operation required in the whole self-tuning system is to select the new frequency in the drive unit, and this can be controlled locally or from a remote position. On changing the frequency-determining switch, three functions occur sequentially. First the d.c. power supply to the final stage is switched off—the anode supply for triodes, but only the screen supply need be removed in the case of tetrodes. Next the frequency changes, followed by the traffic lines being disconnected and replaced by a tuning signal at the new carrier frequency, at a level of about 8 dB below that required for full power output. The tuning signal is essential, because in s.s.b. systems there is no r.f. signal in the absence of traffic and the low level is used to avoid overloading or tripping during tuning.

As the wideband driving stages have not been switched off, the low-level carrier at the new frequency appears on the input line to the final stage. The input-tuning discriminator senses the frequency change and provides an error signal, which is amplified to drive the input-tuning servo-motor and hence the inductor L_1 towards the position for matching the input line. When this motor starts up, the 'coarse–fine' switches are changed from 'fine' to 'coarse', so that the information supplied by potentiometers P_1 and P_2 can be used to set the tuning and loading capacitors C_1 and C_2 to approximately the required tuning position for the new frequency. The position of the potentiometers is determined by the angular position of the input-tuning motor shaft, as are the cams for selecting the appropriate range taps on the anode and output inductors L_2 and L_3 via the range motor.

When the line inductor L_1 reaches a value which matches the input line, the input discriminator ceases to provide an error signal and the input-tuning motor stops. When it stops, the power supplies are applied to the final stage and the 'coarse–fine' switches are changed to 'fine' to enable the anode and output capacitors to be controlled from the fine tuning and loading discriminators. The input-tuning motor is interlocked with the power supplies, so that it cannot start up while these supplies are on. This prevents the possibility of a range switch being changed with power on.

Immediately the power supplies are on, the fine tuning and loading discriminators both provide error signals, but the loading control is muted during fine tuning, so only the tuning system operates until the anode circuit is in tune. When in tune the loading control operates, but in so doing the tuning is affected and it is necessary to re-tune when the loading motor stops. These tuning and loading controls operate sequentially in progressively smaller steps until both are set correctly for the incoming frequency.

When both these motors stop for the first time, an automatic level control operates to adjust the drive input to give a transmitter output at, say, 8 dB below p.e.p. This step is necessary to compensate for changes in the gain-frequency response of the wideband driving stages, when driven by the low-level carrier at a constant amplitude. On completion of the level adjustment, the carrier signal is removed, the incoming lines are restored and traffic can recommence.

The tuning and loading settings reached are appropriate for the feeder impedance at the time, but this is liable to change quite erratically in bad weather conditions. To compensate for any such change, the tuning and loading system remains in operation continuously, to enable the valve to be always working in the optimum condition. By this means trips caused by excessive overloading or underloading are avoided, unless the v.s.w.r. exceeds a limit set by the impedance range which the transmitter can accommodate.

Economic advantages

With this system the traffic time lost depends to some extent on the difference between the two frequencies, but it is always less than a minute and usually

between 30 s and 45 s. The advantage is emphasized in the case of multi-transmitter stations, where each transmitter normally radiates several traffic channels simultaneously. In terms of total channel traffic, the economic advantage is quite impressive.

Due to acreage required for the antenna systems, it is usual for multi-transmitter stations to be sited remote from populated areas, not only because of land availability but also the land cost. For this reason operating organizations often find difficulty in obtaining an adequate complement of personnel of the right calibre. With self-tuned transmitters, a large staff is not required at the main times for changing frequency and a reduction in operating staff means a reduction in supporting staff for canteen facilities, etc. Consequently the use of self-tuned transmitters means an even greater economy than would be expected by the reduction in operating personnel alone.

Antenna selection

On many main traffic routes, the frequencies and antennas are predictable for long periods, although changes may be required two or three times a day. Each frequency can be allocated to a particular antenna for each transmitter, by means of a pre-patching board in the interlocking system associated with a feeder switching matrix. Thus, when one of these frequencies is selected, the appropriate antenna is also selected for that service during the time taken for the amplifier input circuit to tune. This means that when the power supplies are switched on, the correct antenna is already connected to that transmitter and no additional time is required for antenna changing.

The pre-patching equipment can be very simple to operate, so that departures from the regular frequency-antenna combinations can be made at short notice to cover any special demands as they arise.

REFERENCES

[1] *I.E.E. Convention on H.F. Communications.* London, March 25–27, 1963.
[2] Stokes, V. O. *Automatic tuning.* British patent application No. 33964/45.

7

Typical Designs for Broadcasting Applications

7.1 DESIGN FOR 100 kW CARRIER POWER IN THE H.F. BAND

Reasons for circuit arrangement

Transmitters in the h.f. band for long-range broadcasting use high-gain antennas, which require a balanced feeder input. For carrier powers of the order of 100 kW, unbalance-to-balance feeder transformers are practical for a single frequency, but quite unsuitable for the operational frequency changes necessary in the h.f. band. Wideband transformers or baluns are not presently available for this power level, so it follows that the final amplifier anode and output circuits should also be balanced, with two valves in a push–pull circuit.

The input circuit of a push–pull stage must also be balanced, but this does not necessarily mean that all the driving stages must be balanced. With high-gain tetrodes in the final stage, the required drive power is such that wideband unbalance-to-balance transformers can be used. Even with the power absorbed by the resistors necessary to obtain a matched condition, the total power required from the driving amplifier is only between 3 kW and 5·5 kW. Therefore the driving amplifier can be an unbalanced circuit, which is more economical in valves and components and more suitable for frequency changing.

The valve in the driving amplifier can also be a high-gain tetrode, which means that it requires a low level of drive that can be obtained from a wideband amplifier system, right from the frequency-determining source. The net result is an amplifying system with only two tuned stages to cover all the broadcast bands between 3 MHz and 26 MHz, giving a carrier output of 100 kW suitable for high-level amplitude modulation.

By international agreement, broadcasting in the h.f. spectrum is limited to a number of relatively narrow bands, so transmitters for this application need not have continuous frequency coverage. Also, the times and frequencies of the services must be predetermined in order that the listeners can have this information in advance of the programmes being radiated. This means that any range-switching required can be in the form of band selection, to be carried out at specific times, and automatic self-tuning is not so necessary as it is for communication transmitters. With such a system, manual tuning is required at each frequency change, so the changing process should be as simple as possible, with all the necessary controls on the front panel. As an aid to simplicity many of the variable elements can be motor-driven, with the controlling switches on the front panel adjacent to the appropriate indicating meters. Motor drive also simplifies

the design by avoiding the use of complicated mechanical drives, which would be required in many cases to obtain front panel control.

On multi-transmitter stations the pre-programming means that the frequency-changing times can be sequential rather than simultaneous, making it unnecessary to have a large staff for frequency changing only. With a limited staff and manually tuned transmitters, even more emphasis should be directed towards simplifying the frequency-changing process.

The screen supply and modulation

In order to radiate programmes of the high quality required for broadcasting when tetrodes are used in the modulated amplifier, it is essential to modulate the screen as well as the anode. Further, the method of modulating the screen plays an important part in determining the order of quality obtainable in terms of audio-frequency distortion.

There are two main methods normally employed to modulate the screen. The more simple one makes use of the change in screen current which occurs when the anode is amplitude modulated, by feeding the d.c. screen supply through a series resistor, providing self-modulation. However, the screen current is a non-linear function of the applied voltage, to overcome which a high value of series resistor must be used, combined with a high-voltage d.c. supply, resulting in considerable power loss.

In the other method, the audio voltage for modulating the screen is obtained from a tertiary winding on the modulation transformer. By this means a very good performance can be obtained over a limited audio-frequency range, but the distortion can be quite high at the upper and lower frequencies, due to the relative difference in phase between the audio voltages on anode and screen. In addition, the voltage level on the screen is critical, further complicating the design of the modulation transformer with its tertiary winding.

It has been found in practice that the best performance can be obtained with a combination of the two methods, by passing the d.c. screen supply through a tertiary winding and a series resistor of relatively low value. This gives a degree of self-modulation from the series resistor, with an over-riding alternating voltage from the tertiary winding. This combined method is remarkably insensitive to phase and amplitude and gives a low level of distortion over the whole audio-frequency band. In fact, the audio-harmonic distortion measured on the rectified r.f. output of the transmitter is not significantly inferior to that at the modulator output.

As the amplitude of the self-modulating voltage is not critical, the value of the series resistor is not critical for modulation. Therefore the resistor value can be selected to give the required d.c. voltage on the screen from a range of d.c. voltages. It is probable that a d.c. supply within this range will be used for some stage in the modulator, eliminating the need for a special supply for the screen only.

Typical circuit diagram and description

Based on the concepts outlined, a simplified circuit diagram for the final amplifier of a 100 kW broadcast transmitter is given in Fig. 7.1. Only the items associated with the r.f. function are shown, all feed meters, interlocks, safety devices, test facilities, etc., have been excluded.

Spurious oscillations and their prevention

Valves operating in push–pull circuits are liable to generate spurious oscillations, especially when neutralized at the fundamental, because of the greater number of coupling paths provided.

The most usual type of spurious oscillation occurs at a frequency very much higher than the fundamental, mainly due to phase reversal in the coupling path, when the reactance changes from capacitative to inductive. These spurious oscillations can be suppressed, without difficulty, by two methods, both of which are shown in Fig. 7.1.

Consider the case of the leads connecting the screen to earth decoupling capacitors C_{81} and C_{82}, represented on the diagram by L_{31} and L_{32}. To be effective at the fundamental, the reactance of these capacitors must be very low; typical values being 0·001 μF with a reactance of 6 Ω at 25 MHz. At 100 MHz the capacitive reactance is only 1·5 Ω, so if the inductive reactance is greater than 1·5 Ω the reactance of the decoupling path would become positive at 100 MHz and spurious oscillation is probable. The inductance required to give a reactance of 1·5 Ω at 100 MHz is only about 0·01 μH. This type of oscillatory tendency can be effectively eliminated by connecting non-inductive resistors across, and close to, the connecting leads, indicated by R_{51} and R_{52}. The value of resistor is not critical, but the optimum is that which will give a Q factor of 1 with the reactance of the lead inductance at the spurious frequency.

Another form of spurious oscillation, also at a frequency much higher than the fundamental, is produced by high Q resonances within the active inductive and capacitative elements of the fundamental circuits. This type can be prevented by applying damping at high-potential points of the spurious frequency in the form of 'resistive antennas', commonly called 'whiskers'. A resistor is connected directly to each point of spurious high-potential, with the other end looking into space. Examples of this type of cure on the anode-circuit components are shown on the diagram as R_{61}, R_{62} and R_7.

To be most effective the resistors should be long in terms of wavelength at the spurious frequency, and 10 in. long carbon rods are quite usual. The resistor value is not critical, being typically 100 Ω or 200 Ω, but the actual value is often determined by the fundamental power dissipated in each resistor.

Both these methods of spurious frequency suppression are only applicable when the spurious frequency is very much higher than the fundamental. In cases where the frequency difference between the spurious and fundamental is not so great, the problem must be tackled more basically by improving the design to eliminate the possibility of such oscillations occurring.

The cooling system

With any vapour-cooling system for high-power valves, it is necessary to have air-cooling for other transmitter components, and probably for the heat exchanger as well, as pointed out in Chapter 5. In a well-designed layout the vapour and air systems can be combined to give a most economical cooling system.

An example of an arrangement of this type, for the final r.f. amplifier and modulator of a 100 kW broadcast transmitter, is shown in Fig. 7.2, in which the vapour-to-air heat exchanger is mounted on the top of the transmitter cabinet. Note that a suction air system is used and the fan can be an all-weather type, suitable for mounting outside the building.

The filtered air is drawn through the main h.t. rectifier stack, which receives adequate cooling by virtue of the construction of the individual modules and the way in which these are mounted within the stack. It then flows through the modulator and r.f. cabinets in parallel paths, to enter the heat exchanger. The construction of the cabinet is such as to force the air through the grid, cathode and screen-grid components before it passes upwards through the anode circuits and the heat exchanger. The total quantity of heat removed by the 8000 ft^3/min of air flow is approximately 120 kW, resulting in an air-temperature rise of about 5°C from the cabinets and a total rise of 23°C at the outlet of the heat exchanger.

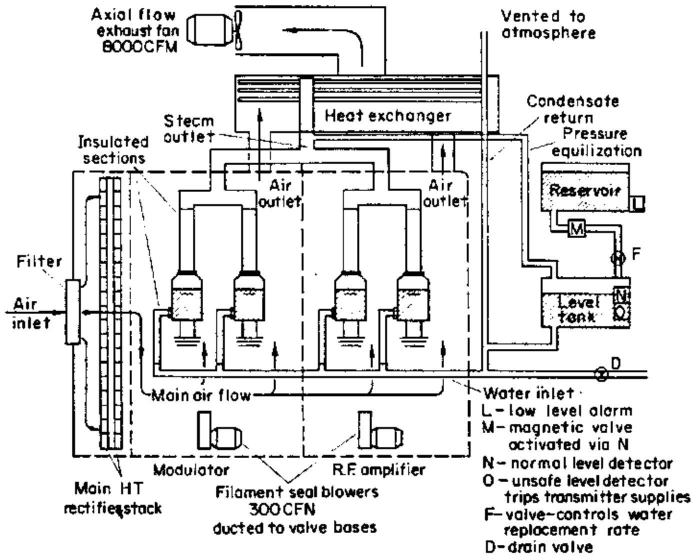


FIG. 7.2 Integrated vapour- and air-cooling system.

The CY 1170 J vapour-cooled tetrodes (English Electric Valve Co.) used in the final stages of both the modulator and r.f. amplifier are exceptional in being designed with the anode mounted upwards, the boiler being an integral part of the valve. This allows the valve socket, with filament, control grid and screen-grid connections, to be fixed, and avoids bringing the vapour outlet through the input circuits. The consequent simplification of these connections in the r.f. amplifier contributes to the easy handling and good efficiency at the highest radio frequencies. To remove a valve it is only necessary to disconnect the fittings for the water inlet and vapour outlet.

A suitable drive equipment

The r.f. input to the final amplifier is a maximum carrier power of 5 kW at any of the operational frequencies, preferably with a low harmonic content. Because of the need for operation on a number of frequencies, it is also desirable that the

The incoming r.f. from the driving stage is stepped up in voltage by the wide-band ferrite-core unbalance-to-balance transformer T_1, thence via d.c. blocking capacitors C_{11} and C_{12}, to the input Π matching circuits, which are partly terminated by resistors R_{11} and R_{12} and partly by the grid loading of the valves. The input capacitor C_2 and inductors L_{11} and L_{12} are in circuit for all ranges, but for the highest frequencies the large input capacitance of the tetrode valves is adequate for the output capacitance of the Π circuits. For lower frequencies, the variable portion of the Π output capacitors C_{31} and C_{32} are first paralleled with the valve capacitance, and at still lower frequencies the fixed portions of these capacitors are also added. By this means, matching is improved by maintaining high Q factors and the design of the inductors L_{11} and L_{12} is simplified. Even so,

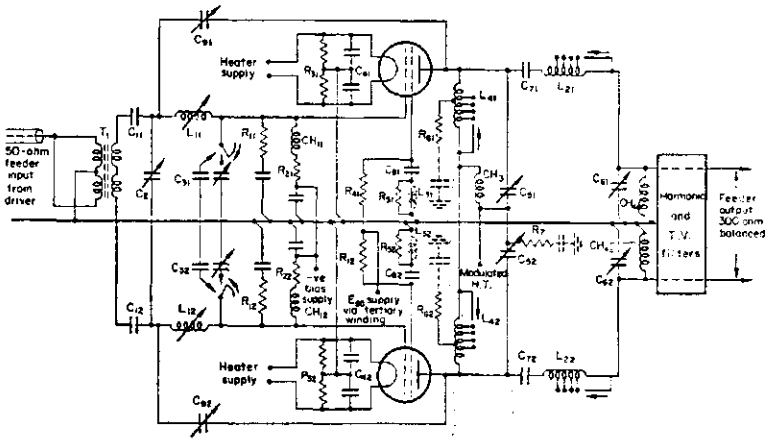


FIG. 7.1 Simplified circuit diagram of 100 kW h.f. modulated amplifier.

the design of these inductors presents quite a problem, because of the very low inductance value required at the highest frequencies, combined with the inductance range necessary to cover the whole frequency band from 3·2 MHz to 26·1 MHz. Although they are shown as single inductors, this is for diagrammatic simplicity only, and in practice a degree of range switching is employed.

The control-grid bias is partly fixed from a negative d.c. supply and partly automatic from the voltage developed across resistors R_{21} and R_{22} by the grid current of the individual valves. The cathodes are maintained effectively at earth potential for both r.f. and d.c., by means of the centre-pointed capacitor–resistor networks C_{41}, R_{31} and C_{42}, R_{32}.

The anode tuning and loading circuit is a balanced Π arrangement, with variable-shunt capacitors and series inductors, having tapping points which are range switched. The input capacitance of the Π circuits is the sum of the valve capacitances and the capacitors C_{51} and C_{52}. Tapped inductors L_{41} and L_{42} are shunted across the valves to reduce their effective capacitance, which would otherwise be too high for resonance with practical inductors in the Π circuit at the higher frequencies. In addition, the junction point between them is at a low r.f. potential, into which to feed the d.c. anode supply. Consequently, the

design of the feed choke CH_3 is simplified because of the low r.f. potential across it. Capacitors C_{71} and C_{72}, in series with the inductors L_{21} and L_{22}, are to isolate the d.c. anode supply from the feeder. At the same time they remove the d.c. voltage from the inductors and the output capacitors C_{61} and C_{62}, easing the voltage clearance problem.

The output capacitors C_{61} and C_{62} are shunted by chokes CH_{41} and CH_{42}, which perform two protective functions. First, they provide a leakage path for static voltages picked up on the antenna, which can build up to quite a high voltage. Second, should a fault occur which causes a flashover effectively across capacitors C_{71} and C_{72}, the d.c. h.t. voltage with an a.c. component would be applied momentarily across the output capacitors C_{61} and C_{62}, most probably destroying them. With chokes CH_{41} and CH_{42} fitted, the capacitors would not be affected by such a fault, which would be cleared in this case by an h.t. overload trip. Because of the very heavy fault current possible through these chokes, they must be made of a conducting material of sufficient current-carrying capacity to withstand this type of fault.

Although the internal anode-to-control grid capacitance of these high-power tetrodes is very small, some neutralizing is necessary to compensate for the unwanted positive feedback produced with such high-gain tetrodes. In this circuit, neutralizing is provided by variable capacitors C_{91} and C_{92} of very small value, between the valve anodes and the input side of the Π input circuit, where the r.f. phase is correct for negative feedback.

Due to the waveforms generated in class C amplifiers, the harmonic content of the valve anodes is relatively high. The single Π anode-output circuit shown in Fig. 7.1 does not attenuate the harmonics to the low level required by international regulations. Also, the maximum permissible harmonic is specified in milliwatts, so the higher the fundamental power the greater the attenuation required. In consequence, with a fundamental power output of 100 kW, the harmonic attenuation must be considerably greater than that provided by the single Π circuit.

As broadcast transmitters operate on specific frequencies, it is quite practical to employ tuned harmonic filters to give adequate attenuation on all the troublesome harmonic frequencies. For transmitters in the h.f. band, such as the one being described, the filters can be range-switched for the various fundamental frequencies in operational service. The most suitable position for these filters is in the output feeder, but it is very important to enclose the filters in a well-earthed and bonded screen. Otherwise it is possible for the filters themselves to radiate a higher harmonic signal than that which would be produced without them. The screen grids are by-passed to earth for r.f. by capacitors C_{81} and C_{82}. The d.c. screen supply is first fed through the tertiary winding of the modulation transformer, then via self-modulating resistors R_{41} and R_{42} for the individual valves.

From the modulation transformer the d.c. anode supply is fed through a common r.f. choke CH_3, then individual inductors L_{41} and L_{42}, which must be changed by taps for different frequencies, for they must offer a very high impedance to the operating frequency, because they are in series across the highest r.f. potential in the whole amplifier.

It should be noted that the centre-point between the Π input capacitors C_{51} and C_{52} is bonded to the h.t. supply, so these capacitors do not have to withstand the d.c. potential.

Spurious oscillations and their prevention

Valves operating in push–pull circuits are liable to generate spurious oscillations, especially when neutralized at the fundamental, because of the greater number of coupling paths provided.

The most usual type of spurious oscillation occurs at a frequency very much higher than the fundamental, mainly due to phase reversal in the coupling path, when the reactance changes from capacitative to inductive. These spurious oscillations can be suppressed, without difficulty, by two methods, both of which are shown in Fig. 7.1.

Consider the case of the leads connecting the screen to earth decoupling capacitors C_{81} and C_{82}, represented on the diagram by L_{31} and L_{32}. To be effective at the fundamental, the reactance of these capacitors must be very low; typical values being 0·001 μF with a reactance of 6 Ω at 25 MHz. At 100 MHz the capacitative reactance is only 1·5 Ω, so if the inductive reactance is greater than 1·5 Ω the reactance of the decoupling path would become positive at 100 MHz and spurious oscillation is probable. The inductance required to give a reactance of 1·5 Ω at 100 MHz is only about 0·01 μH. This type of oscillatory tendency can be effectively eliminated by connecting non-inductive resistors across, and close to, the connecting leads, indicated by R_{51} and R_{52}. The value of resistor is not critical, but the optimum is that which will give a Q factor of 1 with the reactance of the lead inductance at the spurious frequency.

Another form of spurious oscillation, also at a frequency much higher than the fundamental, is produced by high Q resonances within the active inductive and capacitative elements of the fundamental circuits. This type can be prevented by applying damping at high-potential points of the spurious frequency in the form of 'resistive antennas', commonly called 'whiskers'. A resistor is connected directly to each point of spurious high-potential, with the other end looking into space. Examples of this type of cure on the anode-circuit components are shown on the diagram as R_{61}, R_{62} and R_7.

To be most effective the resistors should be long in terms of wavelength at the spurious frequency, and 10 in. long carbon rods are quite usual. The resistor value is not critical, being typically 100 Ω or 200 Ω, but the actual value is often determined by the fundamental power dissipated in each resistor.

Both these methods of spurious frequency suppression are only applicable when the spurious frequency is very much higher than the fundamental. In cases where the frequency difference between the spurious and fundamental is not so great, the problem must be tackled more basically by improving the design to eliminate the possibility of such oscillations occurring.

The cooling system

With any vapour-cooling system for high-power valves, it is necessary to have air-cooling for other transmitter components, and probably for the heat exchanger as well, as pointed out in Chapter 5. In a well-designed layout the vapour and air systems can be combined to give a most economical cooling system.

An example of an arrangement of this type, for the final r.f. amplifier and modulator of a 100 kW broadcast transmitter, is shown in Fig. 7.2, in which the vapour-to-air heat exchanger is mounted on the top of the transmitter cabinet. Note that a suction air system is used and the fan can be an all-weather type, suitable for mounting outside the building.

The filtered air is drawn through the main h.t. rectifier stack, which receives adequate cooling by virtue of the construction of the individual modules and the way in which these are mounted within the stack. It then flows through the modulator and r.f. cabinets in parallel paths, to enter the heat exchanger. The construction of the cabinet is such as to force the air through the grid, cathode and screen-grid components before it passes upwards through the anode circuits and the heat exchanger. The total quantity of heat removed by the 8000 ft³/min of air flow is approximately 120 kW, resulting in an air-temperature rise of about 5°C from the cabinets and a total rise of 23°C at the outlet of the heat exchanger.

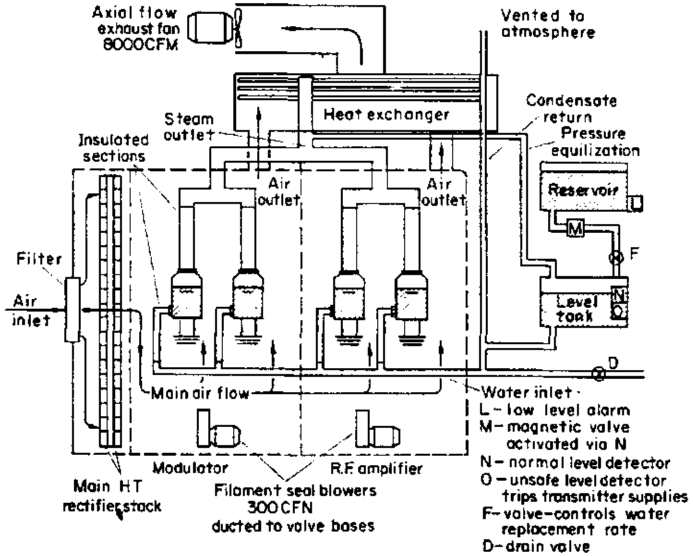


FIG. 7.2 Integrated vapour- and air-cooling system.

The CY 1170 J vapour-cooled tetrodes (English Electric Valve Co.) used in the final stages of both the modulator and r.f. amplifier are exceptional in being designed with the anode mounted upwards, the boiler being an integral part of the valve. This allows the valve socket, with filament, control grid and screen-grid connections, to be fixed, and avoids bringing the vapour outlet through the input circuits. The consequent simplification of these connections in the r.f. amplifier contributes to the easy handling and good efficiency at the highest radio frequencies. To remove a valve it is only necessary to disconnect the fittings for the water inlet and vapour outlet.

A suitable drive equipment

The r.f. input to the final amplifier is a maximum carrier power of 5 kW at any of the operational frequencies, preferably with a low harmonic content. Because of the need for operation on a number of frequencies, it is also desirable that the frequency-changing process should be simple and reliable. All these features are provided by a medium-power h.f. communications transmitter. By using this type of transmitter as the complete drive system of the 100 kW broadcast transmitter there is a considerable saving in development costs. It is only necessary to eliminate the s.s.b. equipment, which is usually in a separate cabinet, and to provide crystal oscillators for the frequencies required.

It would be possible to use automatic tuning for the drive to further speed up frequency changing, but this would need a modification to cater for the variation in drive level required over the frequency band. However, only one stage has to be tuned in the communications transmitter, so the manually tuned version is recommended for this application.

7.2 DESIGN OF A VERY HIGH POWER M.F. TRANSMITTER

Design considerations

It was pointed out in Chapter 3 that the actual power level of transmitters for very high power is dictated by the power output which can be obtained from the limited range of suitable valves. For the most economical design in terms of cost per kilowatt, both initially and during operation, it means that the maximum power output should be obtained with the minimum number of high-power valves. With a single valve capable of delivering an r.f. carrier-power output of 250 kW, combined with the obvious advantages of using the same type of valve in the high-level modulator as in the r.f. stage, Table 7.1 has been compiled to show the reason for selecting a carrier power of 750 kW.

TABLE 7.1

Carrier power, kW	No. of valves in r.f. amplifier	No. of valves in modulator	Total valve complement	Carrier power per valve, kW
250	1	2	3	$\frac{250}{3} = 83{\cdot}3$
500	2	2	4	$\frac{500}{4} = 125$
750	3	2	5	$\frac{750}{5} = 150$

The type of valve used is the VCP 2002 vapour-cooled tetrode, with a maximum anode dissipation of 180 kW. It would not be possible to obtain 250 kW from one of these valves operating in class C without exceeding the permitted anode dissipation at 100% modulation. Thus the 750 kW rating with three valves is only possible with the r.f. valves operating in a 'Tyler high-efficiency' circuit (class D) and a lower a.f. output is required from two valves of the same type in the modulator.

Apart from this consideration, for a specified power output the mains-to-r.f. conversion efficiency given by class D is an important factor in reducing running costs, especially at power levels of this order.

Frequency changing is not an operational requirement at m.f., so the additional tuning circuits associated with the third harmonic resonators of the high-efficiency circuit do not present a routine difficulty. In fact the tuning of the harmonic resonators is not critical, and, due to the self-correcting properties of the system, can be accomplished quite readily from simple instructions.

The antennas used for these m.f. transmitters are monopoles requiring an unbalanced input, so the final amplifier should be single-sided. This is another reason why an odd number of valves (three) can be considered for the r.f. amplifier, where they can be connected in parallel. Again, the operation of valves in parallel is practical at m.f., where the high input and output capacitance of tetrode valves is not so important as at h.f.

Although the positive feedback produced by the low value of internal anode-to-grid capacitance is quite low, there is also likely to be some external feedback which it is almost impossible to avoid with equipments of this size. The net result is that some neutralizing is required. This can be obtained by feeding a sample of the anode voltage back into the grid of the same valves via a wideband transformer to provide the necessary phase reversal for negative feedback. The use of a wideband transformer also has the advantage that the neutralizing is effective at the third harmonic as well as at the fundamental; an important feature with this circuit.

With a fundamental output power of 750 kW from a non-linear amplifier, it will be appreciated that the harmonic content of the anode-output circuit is such that particular precautions must be taken to reduce the harmonic level in the feeder output. The attenuation is provided by a number of tuned harmonic filter circuits in a well-screened compartment adjacent to the outgoing feeder. These filters need not be reset after being adjusted on installation for the operational frequency allocated to the transmitter, for the requirement to change frequency on such an equipment would be most exceptional.

The output feeder has an impedance of 100 Ω and a special design is required for this power rating. Allowing for a v.s.w.r. of 1·5 to 1, the peak voltage on the feeder inner at 100% modulation is 30 000 V, with a maximum r.m.s. current of 130 A under the same conditions. Adequate safety factors must be allowed on these ratings, and for voltage this should be 3 to 1 on a feeder, giving a nominal flashover voltage of 90 000 V between inner and outer. This gives an idea of the constructional problem of the feeder, particularly in equalizing the potential gradient across the insulators supporting the feeder inner. It also serves to show that the components comprising the harmonic filters present quite a problem in voltage and current rating.

Circuit description

A simplified circuit diagram of the final r.f. amplifier of a 750 kW m.f. broadcast transmitter is shown in Fig. 7.3. It is only simplified in relation to the number of components comprising the inductor and capacitor elements indicated, for fundamentally the design is essentially very simple. In terms of wavelength, the lengths of the interstage connections are such that it is not necessary to use a feeder between the drive output and the final stage, thereby eliminating the need for

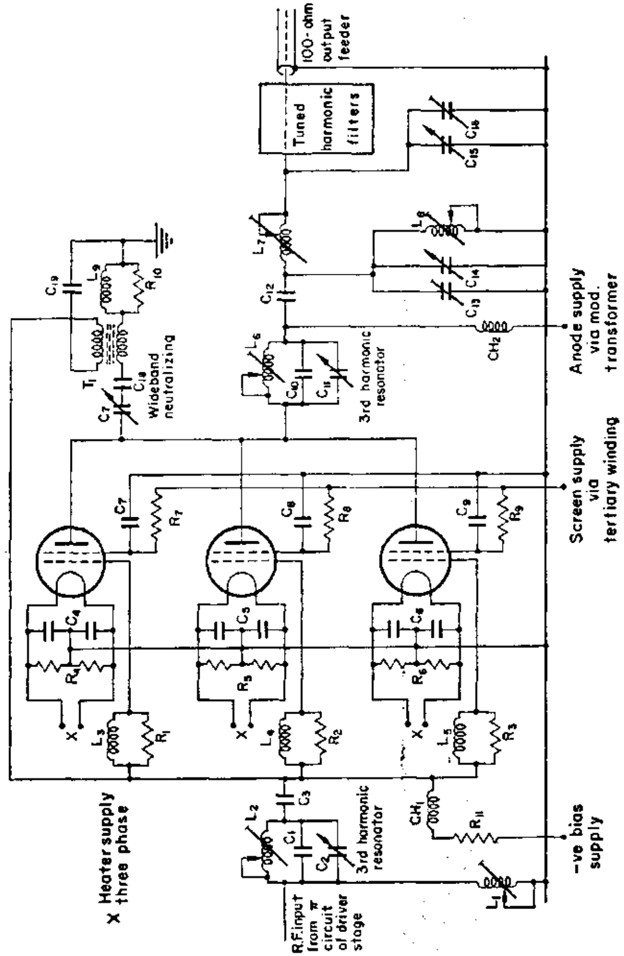


FIG. 7.3 Final amplifier for 750 kW m.f. transmitter.

step-down and step-up circuits. In this case the r.f. drive is fed directly from the Π circuit of the driving amplifier into the final-stage tuned input circuit, which includes a third harmonic resonator circuit $L_2C_1C_2$. It then passes through the d.c. blocking capacitor C_3 to the control grids of the three power valves, via their individual anti-spurious circuits L_3R_1, L_4R_2 and L_5R_3. The valve cathodes are effectively grounded to r.f. and d.c. by the capacitor-resistor networks C_4R_4, C_5R_5 and C_6R_6.

The valve anodes are directly paralleled and feed the Π output circuit via the third harmonic resonator circuit $L_6C_{10}C_{11}$ and the d.c. blocking capacitor C_{12}. These third harmonic resonators are an essential feature of the Tyler high-efficiency system, which was described fully in Chapter 3. It is of passing interest to note that any harmonic resonance in the circuit including the valve capacitance and a Π input capacitor (tuning) is the very thing to be *avoided* in the h.f. linear amplifiers.

The Π output circuit consists of input shunt capacitors C_{13} and C_{14} in parallel with series inductor L_7 and output shunt capacitors $C_{15}C_{16}$. In fact, both sets of capacitors are made up of several units in parallel—for two reasons. First, in order to obtain sufficient capacitance for the low-frequency end of the band, 525 kHz, with the necessary voltage and current rating, a number of standard units can be used instead of a specially designed and very expensive single unit. Second, for frequencies near the upper end of the band, 1625 kHz, some of the capacitors can either be linked out of circuit or not even supplied. The purpose of L_8 is to improve the Q factor of the Π circuit, which would otherwise be very low with this arrangement.

The tuned harmonic filters are connected between the output of the Π circuit and the 100 Ω output feeder. Inductor L_8 also performs the dual function of static leak for the feeder and fault protection for the Π circuit components, exactly as for the 100 kW h.f. transmitter described in Section 7.1.

For neutralizing, capacitor C_{17} is adjusted to give the correct feedback voltage to the grids of the valves, via the wideband phase reversal transformer T_1. Capacitor C_{19} is for d.c. blocking and circuit L_9R_{10} is for anti-spurious purposes.

The screen supply is modulated partly by passing the d.c. through the tertiary winding of the modulation transformer, and partly by the self-modulating action of the individual series resistors R_7, R_8 and R_9. The bias supply is also partly automatic from the common resistor R_{11}, which is in series with the d.c. bias and the grid choke CH_1.

With three valves in the final stage, it is convenient to supply the heaters from a three-phase transformer, so reducing the amplitude of mains noise level from this source. An interesting feature for the reduction of mains noise level is that mains h.t. is supplied from two rectifiers in parallel, each being fed by two three-phase mains transformers with a 30° phase difference between them, giving an effective twelve-phase commutation. This not only means smaller smoothing components, but allows a public supply system of a lower fault capacity to be used and reduces the harmonics produced in the mains supply.

The maximum r.f. drive power required for full output is only 6 kW, which can be obtained from a single tetrode amplifier with only 15 W of drive. At m.f. it is a simple matter to obtain a 15 W drive from a solid-state low-power system, so the whole r.f. system of the 750 kW transmitter requires only four valves, three of the same type and also of the type used in the final modulator stage.

7.3 ECONOMICAL DESIGN FOR A 1·0 kW M.F. TRANSMITTER

The need for an m.f. transmitter of this power

With the ever-increasing demand for higher power in the m.f. band and the liability to interference due to spectrum overcrowding, it might appear that there is no need for m.f. broadcast transmitters of only 1·0 kW carrier power. This opinion is probably influenced by the fact that the majority of radio transmitters are manufactured in industrial areas of densely populated countries, where there is no such need. However, in many parts of the world, especially in the developing countries, the population tends to be grouped in relatively small communities separated from one another by considerable distances. It is true that few communities are outside the long-range coverage of h.f. broadcasts in many languages, but suitable h.f. receivers are rather costly, reception is not renowned for reliability or quality, and much of the programme material lacks local interest. Long-range m.f. broadcasts are better in many respects but they have nothing like the coverage area given by h.f., and the programme material tends to be of the propaganda type.

In these isolated communities there is undoubtedly an increasing demand for high-quality programmes for entertainment and education in their own language. With the development of inexpensive, battery-operated, portable, transistorized receivers, this demand can be realized in each community with a local m.f. broadcast transmitter of about 1·0 kW output.

Transmitter features

The keynotes of design for these transmitters must be simplicity and reliability. In general, the stations will be operated by local personnel, whose knowledge of transmitters may be limited to that gained on a short training course held by the manufacturer. As far as possible, the operational controls should be little more than necessary for switching on and off—but not by a too-liberal use of automation which, due to its complexity, would make fault-finding more difficult.

All components should be operated well below their maximum ratings, to give a high degree of reliability over long periods and reduce the number of spares required on a station. Consumable spares must be minimized, which means using the smallest possible number of valves. The life expectancy of valves can be extended by under-running, so valves should also be operated well below their maximum ratings. Even so, valves do die and when r.f. valves are changed it will be necessary to retune the associated circuits. The tuning process should be very simple, if possible even more simple than that for equipment requiring regular changes of frequency, because retuning will be a rare occurrence.

The cooling system should be simple also, so this is a clear case for cooling the complete transmitter with a single exhaust fan of ample capacity for stations at high altitude and with a high ambient temperature.

The modulation system

Various types of modulation are available at this power level, but anode modulation, ampliphase and Doherty systems all require more than one power valve, and are therefore not considered to meet the requirement for the lowest possible

number of valves, i.e., one. On the score of simplicity and quality of performance, the modulated drive system is preferred, with the final r.f. amplifier in a linear condition. This also has the advantage that the low level of modulated drive can be obtained at m.f. from an all solid-state drive. The only valve in the complete transmitter is in the final r.f. amplifier.

In a class B linear amplifier with a modulated r.f. drive, the anode current is constant at any level of modulation, so the amplifier efficiency increases with the modulation depth. In this respect it behaves in a similar manner to a class A amplifier.

The conversion efficiency of the final amplifier in this arrangement is not so high as for high-level modulation systems, so the power consumption is higher for the same power output. However, this running cost is more than offset by the higher replacement cost of the greater number of valves used in the other systems.

Final amplifier arrangement

For a carrier power of 1·0 kW, the peak power at 100% modulation is 4 kW, so the final r.f. valve must be capable of supplying this peak power when operated in a linear condition. This means that the valve is a much larger type than would normally be expected for a 1·0 kW transmitter, but as the depth of modulation for the average broadcast programme is only 30% or less, for most of the time the valve is being under-run.

It would be reasonable to consider a tetrode for this application on the grounds of the low level of drive which would be required. However, there is a 'zero bias' triode available, type 3XC3000F7 (Eimac), with characteristic features which make it more suitable both operationally and economically. It has an anode dissipation of 3 kW and has been designed for linear operation with no negative bias on the grid. The grid current is quite low and flows during the whole of the positive r.f. cycle, so there is an absence of the non-linearity which normally occurs when the positive grid-swing exceeds the bias voltage. Also, the load produced by the grid current is sufficiently constant over the driving cycle to provide a termination for the drive source without additional resistance loading. As only heater and h.t. supplies are required, which can be applied simultaneously, there is an obvious economic advantage in not having to supply bias and screen voltages, with their associated interlocks.

The method of setting up the triode for linear amplification of a modulated r.f. input signal, in relation to the load line on the constant current characteristics, is the same as for a tetrode (described fully Chapter 3, Section 3.3). When set up for a carrier level output of 1·0 kW, the drive power required is 20 W, which is well within the capability of a solid-state drive at these frequencies.

A simplified circuit diagram of the final amplifier is shown in Fig. 7.4. The drive can be fitted adjacent to the final-stage input, so no intervening coaxial cable is required. The output impedance of the solid-state drive is low, making it necessary to insert the wideband transformer T_1 to step up the voltage to the level required at the grid. The phase-reversing action of the transformer provides a point of correct phase for neutralizing the triode, via capacitor C_1. It is apparent that no input tuning is necessary over the m.f. band, but neutralizing must be adjusted for the operational frequency, and again when a valve is changed.

The valve output is fed to the Π output circuit via the d.c. blocking capacitor

C_2. The Π circuit input capacitor C_3 and output capacitor C_4 consist of a number of fixed units, which can be linked in and out of circuit to give coarse tuning and loading. The variable inductor L_2 also has taps for coarse tuning, while fine tuning is effected by adjusting a damping copper ring near or around the inductor (spade tuning). Tapped inductor L_3 provides the fine loading control necessary to couple to the output feeder within the permissible limits of 1·4 to 1 v.s.w.r.

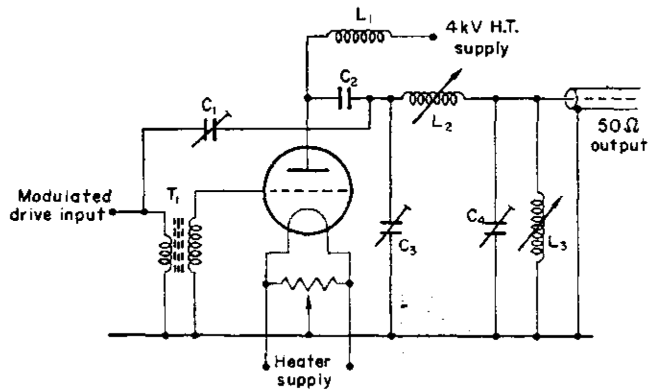


FIG. 7.4 Linear amplifier for 1·0 kW m.f. transmitter with low-level modulation.

This inductor also provides a leakage path for static charges picked up on the antenna and offers fault protection for the Π circuit capacitors. After changing a valve, the only retuning required is a readjustment of the variable inductor L_2.

It is interesting to note that the r.f. harmonics in the feeder are less than 50 mW with the attenuation of the Π circuit only; no harmonic filters are used. This is an indication of the linearity of the 3CX3000F7 triode and a pay-off obtained by using a low-level drive and linear-amplifier combination for this type of equipment. It should also be apparent that the transmitter is eminently suitable for this broadcast application with relatively unskilled operators, and that the maintenance requirements are minimal.

8

An L.F. Transmitter Design

8.1 CHARACTERISTIC FEATURES

Comparison between l.f. and h.f. systems.

In the early days of radio-communications, low-frequency, and very low-frequency transmissions were employed for all long-range circuits, using telegraphy only. With the advent of h.f. communications, such as the Marconi beam system, the pattern changed, for a number of reasons. At h.f., higher keying speeds could be used, even higher than the long-distance cable capability at that time, and the greater bandwidth made it possible to transmit more than one channel simultaneously. This feature went a long way to meet the rapidly increasing demand for a greater number of international communication circuits. In addition, the bandwidth at h.f. was sufficient for voice frequencies, so radio-telephony became practical for long-range broadcasting as well as communications.

Operating in the h.f. band made it possible to use efficient antennas with directional properties, so that the e.r.p. in any required direction could be of the order of ten times greater than the power output at the transmitter terminals. Thus the transmitter power at h.f. could be much lower than at l.f., with a consequent saving in initial and running costs. Bearing in mind also that h.f. antennas themselves are less expensive than l.f. antennas, there is a considerable saving in overall costs by employing h.f. systems.

However, it is a well-known characteristic of low-frequency propagation that ground-wave attenuation and ionospheric reflection are much lower than in the h.f. band. The combination of these two effects means that deep and rapid fading is practically non-existent at l.f., and slow fading is predictable and less severe. Another important feature is that reception is less affected by ionospheric storms, even in the polar regions where h.f. systems are often unreliable.

These characteristics enable low frequencies to be used as a reliable medium for long-range communication circuits and, in the field of navigation particularly, they offer overriding advantages over all other systems. In the form of radio beacons, l.f. systems provide a very accurate means of long-range position determination for both civil and military organizations, especially in the polar regions. Another important advantage is that communication can be established with submerged submarines from remote land stations.

Choice of frequency, bandwidth and power output

In order to assess the most generally useful frequency range and power output, the main factors to be considered are propagation attenuation and antenna efficiency in relation to frequency. Associated considerations are the required distances to be covered, available bandwidth and economic aspects.

At frequencies below 200 kHz, ground-wave propagation is the predominant mode and path attenuation increases with frequency. For the longest transmission path the greatest field strength will be given by operating at the lowest frequency. On the other hand, antenna efficiency decreases with frequency increase, because the antenna length decreases with frequency in terms of wavelength: the optimum radiation efficiency being obtained with an antenna one-quarter wavelength long.

In order to show the combination of these effects, consider a simplified example for a transmission path 3000 km long with frequencies of 20 kHz and 40 kHz. By ground-wave propagation only, a radiated power of 12 kW at 20 kHz will give a signal strength at the receiver of +25 dB with reference to 1 μV/m, whereas at 40 kHz the same radiated power will give +16 dB on 1 μV/m, i.e., the remote signal will be 9 dB higher at 20 kHz. The antenna efficiencies will be of the order of 12% at 20 kHz and 36% at 40 kHz, so transmitters of the same power would radiate three times more power at 40 kHz, i.e., +5 dB. For the same power at the transmitter terminals, the remote signal strength at 40 kHz would be 4 dB (9 − 5) lower than at 20 kHz.

The Q factor of an antenna depends on the efficiency, so the proportional bandwidth also depends on efficiency. In the example, this means that the proportional bandwidth at 40 kHz is three times that at 20 kHz, and as the carrier frequency is twice that at 20 kHz, the bandwidth, in hertz, will be six times that at 20 kHz. Typical Q factors for antennas at 20 kHz are of the order of 200, giving a bandwidth of ±50 Hz at the 3 dB points. On the above basis, the bandwidth at 40 kHz will be ±300 Hz at the 3 dB points.

The available bandwidth determines the number of frequency-shift telegraphy channels that can be radiated simultaneously either in f.d.m. or t.d.m. Allowing for spacing between channels, it means that three telegraph channels can be radiated simultaneously at 40 kHz for a single channel at 20 kHz, using the same frequency shift.

In order to make the maximum use of the total available bandwidth in the l.f. and v.l.f. spectrums, there are obvious advantages in being able to operate a number of channels on each assigned frequency. Combining this advantage with the fact that the remote signal strength is only about 4 dB lower at the higher frequency (in the typical example) for the same r.f. power at the transmitter terminals, a carrier frequency of 40 kHz is preferred to 20 kHz for general applications.

Even at frequencies in the region of 40 kHz, there may be a need for a total bandwidth greater than the limitation imposed by the antenna. The most convenient means of increasing the bandwidth is to apply resistive loading to the antenna circuit. This reduces the radiated power and it may mean that the damping element has to be cooled, e.g. by blowing. The alternative is to erect a bigger antenna, which would be very costly and might not even be practical.

There are some applications where the lowest possible carrier frequency has overriding advantages, such as for communicating with submerged submarines, but these are limited and multi-channel operation is not the prime requirement.

Considering the upper end of the low-frequency band, say 200 kHz, antenna efficiency will be about twice that at 40 kHz, and, the frequency being five times,

the antenna bandwidth will be about ten times that at 40 kHz, i.e., ±3000 Hz. Thus the Q factor of the antenna does not limit the bandwidth at the upper end of the l.f. band and speech transmission is practicable.

The attenuation of the propagation path is much higher at these upper frequencies, but the fading effect produced by multi-path propagation is not too serious, so this portion of the band can be used with advantage for shorter-range circuits.

The actual upper frequency limit is governed by international radio regulations, which allocate 160–255 kHz to broadcasting in some regions. The most appropriate frequency range for communications in the l.f. band is considered to be 40–160 kHz.

Turning to the question of transmitter power, the range to be covered and/or the remote signal strength will be greater for higher power, but so will be the initial and running costs. Consequently the actual power level depends on the relative importance of these factors, and a transmitter power of 100 kW is considered to be a satisfactory compromise.

A discussion of the relative merits of l.f. and v.l.f. would not be complete without mentioning the antenna structure. Even to obtain the order of antenna efficiency quoted for the lower end of the v.l.f. band, the height and area covered by the antenna will be much greater than in the case of l.f. Not only will the initial and maintenance costs of v.l.f. antennas be higher, but the required acreage will also be greater. These are additional factors in favour of l.f. operation, especially where the available land space is limited.

Multi-channel operation and frequency stability

In order to achieve multi-channel operation within the limited bandwidth available, particularly at the lower end of the l.f. spectrum, the frequency shift and channel separation must be low. This necessitates a high order of frequency stability to avoid cross-modulation between adjacent channels, with consequent degradation of service.

For standard time-signals and accurate information for navigation, the frequency stability should be even higher than that necessary for multi-channel operation.

A convenient method of obtaining high stability is by frequency division from the output of an h.f. synthesizer system, driven by a master oscillator; the actual stability being that of the master oscillator. Short-term stabilities of the order of 1 part in 10^{11} (0·0000004 Hz at 40 kHz) can be obtained from master oscillators based on a rubidium-gas cell or caesium beam for navigation and time-signals. For general communication purposes, less expensive oscillators with a stability of about 1 part in 10^{8} (0·0004 Hz at 40 kHz) are suitable.

An additional advantage of deriving the carrier frequency by l.f. synthesizer methods is that a standard equipment can be used as the basis for both h.f. and l.f. drives, reducing the number of equipment types for organizations employing both h.f. and l.f. communication circuits.

In order to avoid restricting the type of modulation used in multi-channel operation, it is necessary to use linear amplification. Therefore, to offer the greatest operational flexibility all amplifiers after the final mixing process must be linear.

8.2 CIRCUIT ARRANGEMENT

The output circuit

The impedance of antennas in the l.f. band is such that they are not suitable for connecting directly to transmitter output circuits, and the actual impedance at any particular frequency depends on the individual antenna structure. A network is necessary to convert the antenna impedance to one which is suitable for matching to the transmitter output circuit. The matching network is normally called the antenna tuning unit (a.t.u.) and is mounted underneath the antenna at a position convenient for the downlead.

Due to the high Q factors of l.f. antennas and associated a.t.u.'s at the lower frequencies in the band, the peak voltage and r.m.s. current in these circuits is much higher than in the anode circuit of the final amplifier. The actual values depend on the circuit constants of each antenna, but for 100 kW output it is not unusual for the peak voltage to be 170 kV, with a circulating current of 170 A r.m.s. Particular attention must be given to avoid corona discharge, to provide weather protection for the a.t.u. and to prevent access by personnel when the equipment is in operation.

The antenna input is an unbalanced arrangement, and as the a.t.u. is likely to be some distance from the transmitter building, it is convenient to match the antenna to a 50 Ω coaxial feeder for connecting to the transmitter output. This is a very convenient arrangement, for it means that a standard transmitter can be used with a 50 Ω output and fully tested into a 50 Ω dummy load. Each a.t.u. is then adjusted on site to match the antenna impedance to the 50 Ω feeder at the operating frequency, which would be necessary in any case because of the individual characteristics of each antenna.

With an unbalanced output it follows that the final amplifier should be single-sided, together with the other amplifiers in the linear chain.

The final amplifier

The transmitter about to be described was initially designed to meet a specification calling for a simple cooling system, which indicated air-cooling with a single exhaust fan. It was considered that there was no single air-cooled valve available which would give the required linearity at 100 kW p.e.p. output, without grid current and with the adequate margins necessary for a reliable design. Consequently, two valves were used for the final amplifier, connected in parallel to comply with the need for a single-sided circuit. Even if the initial valves supplied were a matched pair, it would be unlikely for all the valves used throughout the life of the transmitter to be matched, so provision had to be made for separate bias and r.f. input level controls for each valve, in order to obtain optimum linearity. The method of setting up valves for linear operation was fully described in Chapter 2, Section 2.5, but in this case the output is the sum of that from each valve, but by operating in parallel the linearity is slightly degraded.

The circuit arrangement of the final amplifier is shown in Fig. 8.1, the valves used being tetrodes type 4CX35000C. The output from the driving amplifier is fed into the wideband input transformer T_1 and loaded with resistor R_1 across the primary winding. Loading on the primary is necessary in this case to maintain the linearity performance of the driving stage, with and without r.f. feedback on

the final amplifier, which alters the characteristics of the input circuit. It also means that the input transformer has a lower throughput power.

The secondary of the input transformer is fitted with tapping points, in order to feed each valve with the appropriate r.f. level to obtain the correct operating conditions with the signal applied. The bias levels are set to give the same static anode feed to each valve in the no-signal condition. The r.f. input to the valves is fed via the d.c. blocking capacitors C_2C_3 and the anti-spurious networks L_2R_2 and L_3R_3.

It is interesting to note that anti-spurious networks L_6R_6 and L_7R_7 are also fitted in the anode circuit, and L_8R_8 and L_9R_9 in the screen circuit, which might

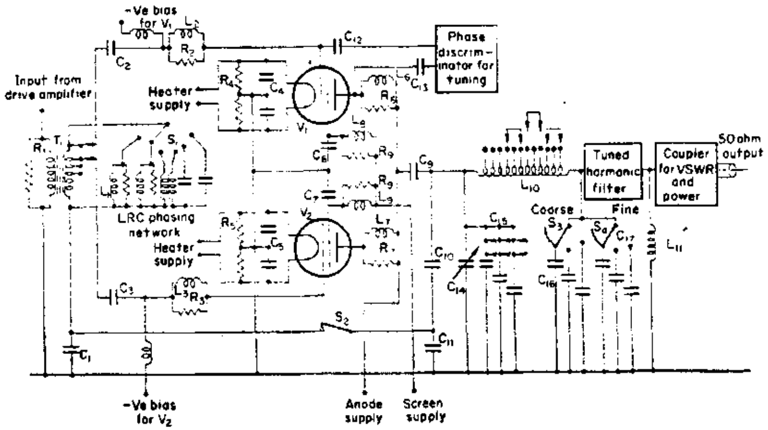


FIG. 8.1 Circuit arrangement of final amplifier for 100 kW p.e.p. in the l.f. band.

appear to be a 'belt and braces' approach. To some extent this is true, but it does not mean that all are essential to avoid spurious oscillation. In one respect valves in parallel are similar to valves requiring high-level neutralizing, in that additional paths are available to provide the appropriate positive feedback for the generation of spurious oscillations. Such oscillations tend to appear at frequencies between 50 MHz and 120 MHz. For transmitters in the l.f. band, the difference in frequency is so great that every probable spurious path can be fitted with an effective damping network for these high frequencies, without having any detrimental effect at the operating frequencies.

The anode-output circuit is a Π arrangement fed via d.c. blocking capacitor C_9. The input capacitance is made up of a variable capacitor C_{14}, and a number of fixed capacitors shown as C_{15}, which are linked in circuit before power is applied, depending on the frequency of operation. The Π output capacitance consists of two sets of fixed capacitors in parallel, the larger set of values, C_{16}, being selected by the coarse loading switch S_3, and the lower values, C_{17}, for fine loading are selected by switch S_4.

The inductor of the Π circuit, L_{10}, has a number of tapping points and links, but contains no continuously variable section. In order to provide the large number of tapping points necessary, this inductor is made of copper tube for simplicity, as distinct from the normal approach of litz wire and a variometer. It will be appreciated that this inductor is quite large to cover 40 kHz, so to save space it is mounted around the air-duct immediately above the final-amplifier valves.

The capacitance range of the tuning capacitor C_{14} is such a small portion of the total anode-circuit capacitance, that the tuning is too flat for minimum cathode current to give a clear indication of resonance. If the tapping points and links are set correctly, the possible amount off-tune cannot be such as to produce a detrimental width of ellipse on the load line, but the operators want to verify that the circuit is in tune. Hence the provision of a phase discriminator unit as a means of indicating anode-circuit resonance even when r.f. feedback is on.

With a circuit of this nature, containing a single variable capacitor of limited range combined with a large number of fixed capacitors with link connections and inductor tapping points, a necessary complement is a chart showing which links and taps are required for any frequency within the band, for both tuning and loading. It is equally important that the links and taps are clearly marked so that they can be readily identified from the chart information. By means of the charts supplied with this transmitter a typical time to change frequency is 30 min.

Frequency changing is not a normal operational requirement in the l.f. band but occasions do arise when a frequency change has to be made. It is also an advantage to be able to demonstrate the performance on a number of frequencies to potential customers, without too much delay.

Although the tetrode type 4CX35000C has good linear characteristics, it is not possible to obtain 50 kW p.e.p. per valve, with i.p.'s as low as –35 dB on the two-tone test, when operating with two valves in parallel. This has been overcome by using about 12 dB of r.f. feedback.

A sample of the anode r.f. voltage is obtained from the capacitance potentiometer $C_{10}C_{11}$, and fed via switch S_1 into the secondary winding of the input transformer at the opposite end to that feeding the valve grids. At this point the phase is approximately correct for negative feedback, being opposite in phase to the grid voltage, by virtue of the earth point along the winding being determined by the relative values of the input capacitance of the valves and that of capacitor C_9, which are in series across the secondary winding. Fine control of the phase of the feedback voltage is given by selecting the most appropriate arm of the *L-R-C* phasing network by means of selector switch S_1. In fact the operation of switch S_2 compensates for the high input capacitance of the two valves in parallel, by adjusting the tune of the input circuit.

Switch S_2 in the feedback circuit is to enable feedback to be applied after the main supplies are switched on, thereby avoiding disturbing transients set up by switching surges. This switch is linked with a switch in the low-level drive, which controls the drive attenuation in such a way that the power output is the same with feedback on or off.

The r.f. feedback reduces the harmonic content at the valve output, as well as reducing the distortion, by cleaning up the waveform. Nevertheless, the combined effect of r.f. feedback and the harmonic attenuation provided by the Π anode-output circuit is insufficient to reduce the level of harmonics in the output feeder to the 100 mW maximum permitted by international regulations (–60 dB relative to fundamental for an output power of 100 kW). Additional attenuation at the lower-order harmonics for frequencies is given by the tuned

filter in the output feeder, immediately after the Π circuit and mounted in a well-screened compartment. On the antenna side of the filter inductor L_{11} provides a leakage path for static picked up on the antenna, and the output coupler supplies information for measurements of v.s.w.r. and power output.

Power gain and drive required

In any amplifying stage with negative feedback applied, the stage gain is reduced by the amount of feedback, and the driving power required is increased by the same amount.

The final amplifier in this equipment has a gain of 25 dB with 12 dB of feedback on, so a drive power of 300 W (100 kW – 25 dB) is required for the full output of 100 kW p.e.p. Without feedback the stage gain is, obviously, 37 dB, for which condition the drive required for full output is only 20 W. This exemplifies the need for a fixed loading on the driving amplifier by means of a resistor (R_1), to reduce the change in load line which would be given by the 15 to 1 difference between the feedback on and off conditions. It is worth noting that the 20 W drive necessary without feedback is due to circuit losses only, because the final-stage valves do not run into grid current.

The valve in the driving amplifier is a pentode type 5CX1500A, having sufficient gain to enable the total output required by the loading resistor and the final-stage input to be obtained without running into grid current. In fact this is possible with 3 dB of negative feedback, which is applied in the same manner as on the final stage, except that the feedback switch is omitted. Even this small amount of feedback gives an appreciable improvement in stage linearity.

With feedback, the effective gain of this stage, including the loss in the loading resistor, is 18 dB. To obtain the 300 W drive level required for a transmitter output of 100 kW, the drive required by the penultimate amplifier is 5 W (300 W – 18 dB). Thus the overall gain of the two stages is 43 dB, which represents a power gain of 20 000 to 1.

Considering the two-stage gain without feedback, it would be 58 dB, a power gain of 600 000 to 1 from an input power of 170 mW. A gain of this magnitude from a two-stage power amplifier would be liable to 'round-the-loop' positive feedback, which would degrade linearity, even if it did not cause self-oscillation at the fundamental frequency. This could be overcome by resistive loading, but negative feedback is preferable because it performs a dual function.

The input power of 5 W to the penultimate stage can readily be obtained from a solid-state linear amplifier in the l.f. band, so the complete 100 kW transmitter contains only two valve amplifier stages using three valves.

9

Transmitters in Parallel

9.1 THE NEED FOR PARALLEL OPERATION

To obtain a given output power by paralleling two lower-power transmitters, usually involves a greater cost for equipment than that for a single transmitter of the same output, hence there must be valid reasons for the need to parallel. The most obvious of these is to obtain more power then can be given by a single transmitter, and falls into three categories in terms of power level.

At lower- to medium-power levels the most likely reason is to extend the capability of an existing transmitter by adding another of the same power. This may be either to increase the service range, or to increase the number of channels on a particular service, thereby requiring more total power to maintain the same power per channel to cover the same range. In either case it is usually on a permanent basis, and it is more economical to add a transmitter of the same power than it would be to buy a new transmitter of higher power.

At medium- to high-power levels, the need tends to be restricted to relatively short periods to cover special programmes or to counteract poor propagation conditions. Most of the time the two transmitters are used for separate services. Thus the keynote of this arrangement is a flexibility which could not be obtained with a single transmitter of higher power.

For very high-power levels the economic aspect changes, because the probable sales of such transmitter-output-combining equipments are likely to be very small. In consequence, the high development cost would have to be recovered on only two or three equipments, and the selling price would be higher than two transmitters of half the power.

However, the main reason for parallel operations at all power levels is reliability. By using a method of paralleling, whereby one transmitter continues to operate if the other fails, unbroken service continues, but at a lower power. With most paralleling arrangements and only one transmitter in operation, the radiated power is reduced to one-quarter of the combined output, which is 6 dB down. A reduction in signal strength of 6 dB at a receiver would hardly be noticed, particularly with automatic volume control (a.v.c.) except possibly in the fringe areas of reception. In consequence, the loss of one of a paralleled pair of transmitters does not incur any break in service, so reliability is greatly improved by operating two transmitters in parallel.

9.2 REQUIREMENTS FOR PARALLEL OPERATION

It is essential that both transmitters are driven at exactly the same frequency, which means using a common drive source. As each transmitter is normally

supplied with its own drive, in order to continue the reliability theme arrangements should be made to use either drive with automatic changeover facilities to cover the possibility of one failing. Preferably, each drive should be used on alternate days, to ensure that both are in a service condition. For such an arrangement, the power output of each drive must be sufficient to drive the two transmitters and to cover any attenuation there may be in the combining and splitting networks. In some cases this will mean the addition of a low-gain amplifier, or preferably two for reliability.

For optimum conditions, the output of the two transmitters must be the same in both phase and amplitude. To achieve these conditions, there must be means of indicating the relative phase and amplitude at each output, in association with arrangements for controlling them within each transmitter, normally at the input.

Correct phase relationship is probably the more important indication, because any differences cannot be detected on the normal power meters on transmitters. Means of indicating phase are essential and these may take the form of phase discriminators, or, more simply, amplitude detectors, in such a position that they give an indication of relative phase.

It is worth noting that when tuning transmitters to an amplitude indication, such as minimum anode current, the trough or peak is fairly flat around the resonant point, but the phase change is very rapid. Consequently there is liable to be quite a difference in phase between the outputs of two identical transmitters, even when tuned as accurately as possible by the same person. At the same time, this very feature enables phase adjustments to be made by trimming one or more tuned circuits, without any noticeable mistune as regards amplitude. However, it is considered preferable to tune the transmitters conventionally and to correct the relative phase by means of special circuits between the drive and each transmitter input.

An indication of relative output amplitude can be given by means of simple detector circuits, as accuracy is not so important as phase. In the case of class B linear transmitters, the level can be controlled by the attenuators normally fitted at the input. This is not possible with class C transmitters, because the automatic biasing arrangement is designed to make the output independent of changes in drive level. With these transmitters, the output level should be controlled by the coupling on the final amplifier.

The method of combining the output of the two transmitters depends on the power level and the frequency band. At medium-, high- and very high-power levels in the m.f., l.f. and v.l.f. bands, where frequency changes are not an operational feature, combining networks consist of a number of capacitors and inductors adjusted for one frequency, with a load to take the out-of-balance power.

A similar arrangement can be used in the h.f. band for the same power levels, by switching between pre-set positions on the capacitors and inductors for changing frequency. This is suitable for broadcast transmitters where the operational frequencies are known and limited in number.

For medium- and high-power h.f. communication transmitters it is preferable to feed each transmitter into a separate antenna, so that the field patterns are combined to give a gain in the desired direction. The phasing is not so simple with this arrangement but it does offer the facility of changing the beam direction by altering the relative phase between the two outputs.

For low- to lower-medium power transmitters in the h.f. band, as well as in the m.f. and l.f. bands, the most satisfactory arrangement is a system of wideband ferrite-cored transformers for both input- and output-combining. The upper power limit for this system is determined by practical design problems concerning heat dissipation in the output transformers and by the v.s.w.r.

9.3 PARALLELING BY MEANS OF A CAPACITOR-INDUCTOR NETWORK

A block diagram of the main units in a paralleling arrangement for the transmitters containing class C or class D amplifiers is shown in Fig. 9.1. Note that

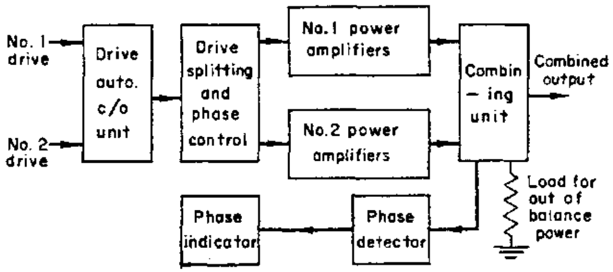


FIG. 9.1 Block diagram for operating two transmitters in parallel by means of a capacitor-inductor combining network.

there is no input-level control, the output power level being adjusted by the output-stage coupling. This arrangement is suitable for all power levels in the v.l.f., l.f., m.f. and h.f. bands.

The drive automatic changeover unit

This is normally a switch, the position of which is controlled by the output of the drives. Solid-state switching is more simple to control automatically, but whatever type is used considerable attention must be given to prevent a leakage from the standby drive into the operating drive, otherwise there will be a beat on the output at the difference frequency between the two.

Drive splitting and phasing networks

A useful form of drive splitting and phase-controlling system is shown in Fig. 9.2, where a 75 Ω input line is connected directly to two 150 Ω networks, terminated with a 150 Ω resistor in each transmitter.

To explain the operating principle, consider each branch as two Π networks in series. If they are identical, any mismatch produced by the first is cancelled out by the reciprocal action of the second, and the phase delay is twice that of each Π circuit. As the reactance of the component value approaches 150 Ω, each Π circuit becomes nearer to a quarter-wave network, with a phase delay approaching 90°, giving a total delay of up to 180°. In a practical application, the four inductors are of the same value and the phase delay can be made adjustable by using a four-gang variable capacitor in each branch.

With the reactance of the capacitors being adjusted between 100 Ω and 320 Ω at any one frequency, a phase control of about 90° can be obtained by each branch, as shown in Fig. 9.3. This phase change holds good for inductors having reactances between 120 Ω and 150 Ω, so any one set of inductors will enable the same

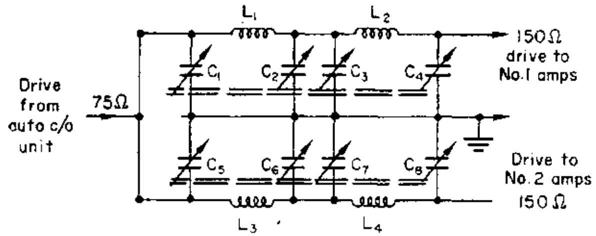


FIG. 9.2 Drive splitting and phasing network.

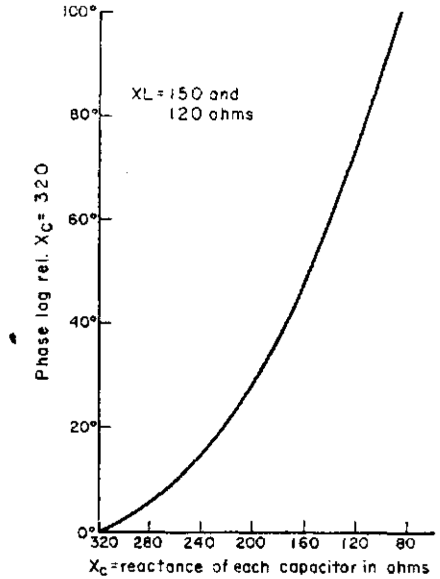


FIG. 9.3 Phase change produced by each double Π section of the phasing network.

phase change to be obtained over a frequency range of 1·25 to 1. Over the same reactance range of the capacitor arms, the v.s.w.r. produced at the 75 Ω input is shown in Fig. 9.4 to be 1·4 or less, and the maximum amplitude change at the output is less than 0·3 dB.

In operation, it would be most unusual to require a phase adjustment of anything like 90° in either transmitter, so by using four-gang capacitors of 50–500 pF each, the circuit is suitable for a wide range of frequencies, as shown in Fig. 9.5, although fixed capacitors must be added at some frequencies. Different inductors are also required, as each value will only cover a frequency range of 1·25 to 1.

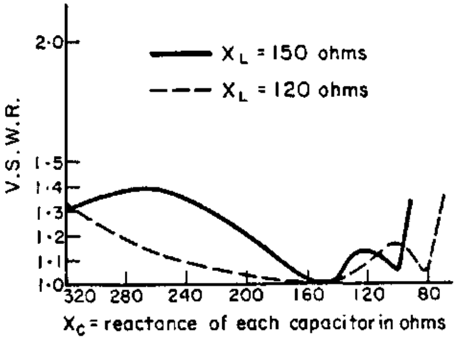


FIG. 9.4 V.S.W.R. at input of phasing network with each output matched into 150 Ω.

Frequency in MHz	0·1	1·0	10	30
Arrangment of each capacitor arm	5000 PF	500 PF		50 PF
Capitance range in pF	5050 to 5500	550 to 1000	50 to 160	25 to 45
Reactance range in ohms	320 to 295	290 to 160	320 to 100	215 to 120
Phase control on each transmitter	0° to 8° = 8°	4° to 48° = 44°	0° to 89° = 89°	24° to 74° = 50°

FIG. 9.5 Phase control available at various frequencies with each variable capacitor covering 50–500 pF.

Output combining with a bridged 'T' network

The circuit diagram of a bridged "T" network for paralleling the output of two transmitters is shown in Fig. 9.6. It was devised by Bartlett in 1951 [1, 2], and is a very simple method, containing only two capacitors, two inductors and a load resistor.

When the outputs from the two transmitters are of the same amplitude and in phase at the input to the network, there is no potential difference across the L_1R_1 branch, so no current is flowing in it and no power is dissipated in the load resistor R_1. The two outputs pass through capacitors C_1 and C_2, respectively, and are paralleled at the input to the outgoing feeder, therefore the power output in the feeder is twice that of each transmitter. The series capacitative reactance of C_1 and C_2 is neutralized by the reactance of inductor L_1 shunted across the feeder, so that each transmitter is correctly terminated by a resistive feeder.

If one transmitter fails, the other continues to operate into a matched load, with half the power being dissipated in resistor R_1 and half being transferred to the output feeder. Thus, the power in the feeder with only one transmitter on is one-quarter of the output with both on, i.e., 6 dB down.

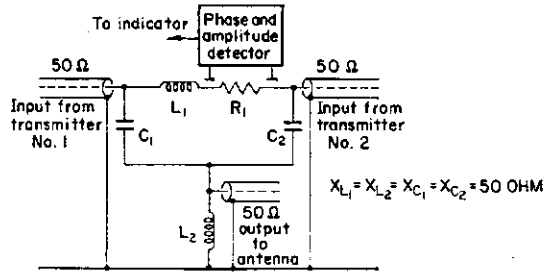


FIG. 9.6 Bridged 'T' paralleling network.

The function of the network components under both conditions can be seen by reference to Figs 9.7 and 9.8. With both transmitters on, in the balanced

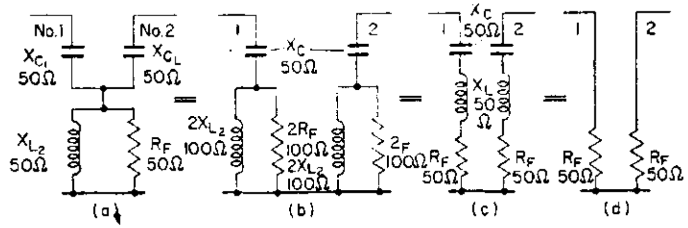


FIG. 9.7 Breakdown of the bridged 'T' network in the balanced condition.

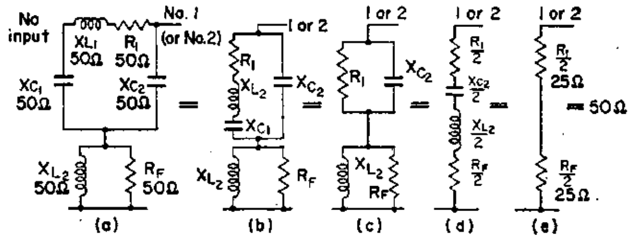


FIG. 9.8 Breakdown of the bridged 'T' network with one transmitter off.

condition, branch L_1R_1 is not carrying any current and can be redrawn as Fig. 9.7(*a*). This can be redrawn as Fig. 9.7(*b*), which is the exact equivalent of Fig. 9.7(*a*). By converting the two parallel sections $2L_2$ and $2R_F$ to series components, as in Fig. 9.7(*c*), the exact equivalence is retained. In this arrangement the series capacitative and inductive reactances cancel out, giving a pure resistive loading on each transmitter of 50 Ω. This proves that in the parallel and balanced condition each transmitter is correctly matched and all the power appears in the output feeder.

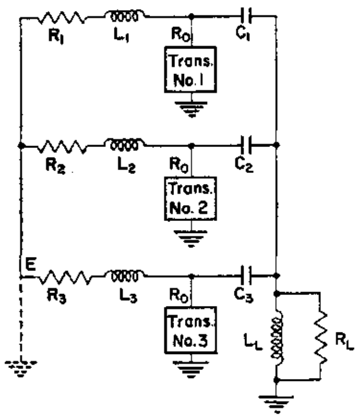
If the outputs of the two transmitters are not of the same amplitude, or not in phase, there is current in the L_1R_1 branch and power will be dissipated in the load R_1. The remaining power is passed to the common output feeder. A device which detects the power in the load R_1 gives a means of indicating the combined amplitude and phase error. By successively trimming the phase and amplitude controls to minimum reading on the indicating meter, the desired condition of balance is obtained.

The function of the network when one transmitter fails can be seen by reference to Fig. 9.8, where Fig. 9.6 has been reproduced as Fig. 9.8(*a*). In Fig. 9.8(*b*) the components have been rearranged to show more clearly that the reactances of L_1 and C_1 cancel out, to give the circuit of Fig. 9.8(*c*). The parallel components of Fig. 9.8(*c*) have been converted to their series equivalent in Fig. 9.8(*d*), which shows that the reactive components cancel out, giving the equivalent circuit, Fig. 9.8(*e*). This shows that the power of the one remaining transmitter is divided equally into load R_1 and feeder R_F, which are effectively half their resistance value, giving a total matching load of 50 Ω.

The failure of either transmitter does not affect the loading of the other transmitter and one-quarter of the total power is radiated. With a v.s.w.r. on the feeder, the ratio of power radiated to power loss depends on the type of mismatch, but transmitter loading is unaffected.

More than two transmitters in parallel

It is not at all unusual for the output of more than two transmitters to be connected in parallel by means of a bridged 'T' network, and an arrangement



R_0 = Transmitter output impedance
$R_L = R_0$ = feeder impedance
$R_1 = R_2 = R_3$ $L_1 = L_2 = L_3$
$C_1 = C_2 = C_3$

Item	Ohmic value in terms of R_0	
	3 Transmitters	N Transmitters
R_1	2/3	(N−1)/N
XL_1	$\sqrt{2}/3$	$\sqrt{(N-1)}/N$
XL_L	$1/\sqrt{2}$	$1/\sqrt{(N-1)}$
XC_1	$-\sqrt{2}$	$-\sqrt{(N-1)}$

FIG. 9.9 Bridged 'T' paralleling network for more than two transmitters.

of three in parallel is often used for unattended transmitters in relatively remote areas.

The circuit arrangement is shown in Fig. 9.9 for three transmitters, but any number may be paralleled in this way, so the resistance and reactance values are given for three and N transmitters.

When all transmitter outputs are of equal amplitude and in phase, it can be shown by an analysis similar to that given in Fig. 9.7, that each transmitter is correctly terminated. In this case the circuit at point E is effectively at earth potential.

An analysis of the conditions when one transmitter is off, by the method shown in Fig. 9.8, will show that the remaining transmitters are correctly terminated and the amount by which the total output power is reduced will depend on the number of transmitters involved. In the general case of N transmitters, the power dissipated in the resistors R_1 will be $1/N$ of the remaining output and the power output $(N-1)^2/N^2$. Therefore, with one transmitter failing out of a total of three, the power lost is one-third of the remaining output, which $2^2/3^2$ of the output of three transmitters, i.e., 3·5 dB down.

9.4 PARALLELING MEDIUM- AND HIGH-POWER H.F. TRANSMITTERS BY COMBINING THE RADIATED FIELD PATTERN

The basic arrangement for operating transmitters in parallel in the h.f. band is shown in Fig. 9.10, but different applications call for some circuit modifications.

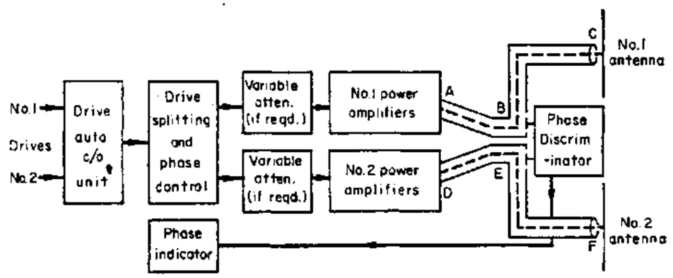


FIG. 9.10 Block diagram for operating two medium- or high-power h.f. transmitters in parallel by combining the field pattern.

The input circuit

For broadcast applications with non-linear amplifiers, it is likely that parallel operation will be on a permanent basis. Consequently the drive changeover unit, drive splitting and phase control can be identical with those described in Section 9.3, but the variable attenuators shown in Fig. 9.10 are not required, because amplitude control is at the output of the final amplifiers. If a 75 Ω cable is normally the input to the power amplifiers, the termination could be changed to suit the 150 Ω output of the phasing networks. However, it is preferable to feed the output of the phasing networks directly into wideband step-down transformers from 150 Ω to 75 Ω and to feed the amplifiers via a 75 Ω cable.

For communication applications with linear amplifiers within this power range, it is more likely that parallel operation will be required only occasionally. In these cases the input circuit should be rearranged as shown in Fig. 9.11, with switches to by-pass the automatic drive changeover, phasing units and step-down transformer. It will be seen that the step-down transformer is necessary to provide a match between the output of the phasing units and the input of the

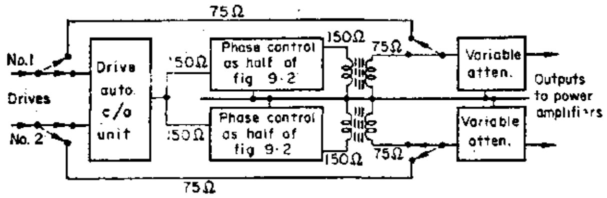


FIG. 9.11 Input arrangement when paralleling is required on a temporary basis.

variable attenuators. By using this type of circuit, the transmitters can be used either in parallel or separately for different services, the variable attenuators being necessary for both applications.

There are, obviously, other suitable input circuits for these applications, such as those described in Section 9.5, so the arrangement should be considered as a typical example to meet the necessary conditions.

The output circuit

The output arrangement shown in Fig. 9.10 is also typical, because the same principle applies if the output feeders are twin wire instead of concentric cable. The feeder-switching arrangements normally associated with h.f. transmitting stations have also been omitted for diagrammatic simplicity.

There are two main features associated with this arrangement. First, for parallel operation, the electrical length of the feeders from each amplifier to its antenna, should be as near identical as possible, so that the phase delay is similar. In this respect, the more important sections are BC and EF, respectively, because the phase comparison is made at points B and E, as representative of the relative phase of the input at each antenna. Sections AB and DE should also be of the same electrical length, to avoid too much phase-correcting of the inputs to the amplifiers. However, these sections are not so easy to control as regards length, because in all probability they will include the feeder-switching matrix of the station.

The second feature is the need to bring the output feeders close together at some point such as BE, in order to avoid long r.f. input leads to the phase discriminator, with the consequent liability of errors in the phase comparison.

The ultimate proof that the relative phasing is correct can only be determined by measuring the radiated field pattern at some distance from the station. This type of measurement is the only reliable method of checking the required phase relationship for a given direction if beam-swinging by phase adjustment is a feature of the application, particularly where a number of frequencies are involved.

9.5 PARALLELING TRANSMITTERS AT LOW- AND LOWER-MEDIUM POWER LEVELS

At these power levels, the applications are mainly in the h.f. band; there may be a few requirements at m.f., but it is highly improbable that there is any application at l.f. or v.l.f. Nevertheless, the system described in this section is suitable for all these bands, and utilizes wideband ferrite-cored transformers for output combining. The use of wideband transformers is particularly suitable for the h.f. band, by virtue of the fact that the output-combining requires no adjustment or component change for any frequency in the h.f. spectrum. A block diagram of the basic arrangement is shown in Fig. 9.12.

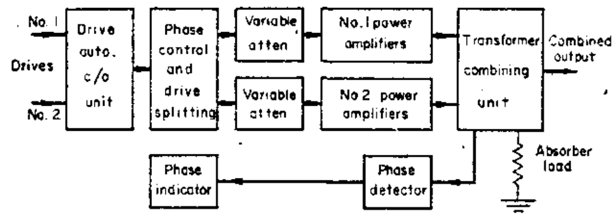


FIG. 9.12 Block diagram of paralleling by means of wideband transformers.

Paralleling with two amplifiers in push–pull

This type of operation, with the signals in the two amplifiers in antiphase, is indicated by using a single output-combining transformer, with the outputs of the two amplifiers fed into the opposite ends of the primary winding. It follows that the input to the amplifiers must also be in antiphase, and one method of achieving this condition is shown in Fig. 9.13.

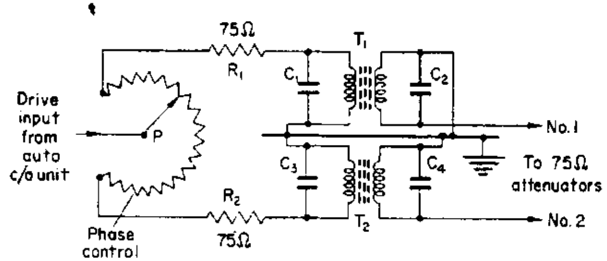


FIG. 9.13 Phasing and drive splitting for push–pull operation.

There are some interesting features in this type of input circuit. By using two 1 to 1 transformers, T_1 and T_2, with the secondary windings cross-connected to give the antiphase outputs, a better match is possible than that which would be given by a single transformer. Capacitors C_1, C_2, C_3 and C_4 are fitted to give a more linear frequency response. The 75 Ω resistors R_1 and R_2, each in series with one of the 75 Ω primary windings, are necessary to give a match with the 75 Ω input cable, by providing two 150 Ω parallel paths. From this, it is apparent that the voltage on the primary and secondary of each transformer is half the input voltage. Potentiometer P is a non-inductive resistor of low ohmic value which provides a very simple method of changing the relative phase of the two transmitters with one control. The phase change possible by this method varies over the h.f. band, and is rather limited at the lower frequencies. However, the range of phase control is adequate for low-power transmitters which are likely to contain only one or two stages.

A method of combining the transmitter outputs with a single transformer is shown in Fig. 9.14. As on the input circuit, capacitors C_1, C_2, C_3 and C_4 are fitted to improve the frequency response. When the two inputs to the transformer are of the same amplitude and in antiphase, all the power is transferred to the 50 Ω secondary, and hence to the output feeder. Any departure from this condition

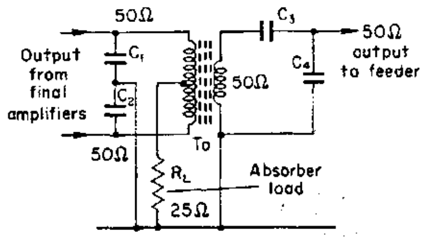


FIG. 9.14 Transformer combining unit for push–pull operation.

results in an out-of-balance power in the absorber load R_L. An amplitude detector across this load gives an indication of amplitude unbalance and incorrect phasing. A minimum on the detector meter indicates correct phase relationship and a zero reading indicates amplitude balance.

With one transmitter off, the antenna load is transformed to appear as 25 Ω across one-half of the primary winding. Half the power is transferred to the output feeder, the other half being dissipated in the absorber load, and the remaining transmitter is still terminated correctly with a 50 Ω load. Any voltage induced in the inactive primary is in phase with the voltage in the other half; as such it is effectively in shunt with the absorber load and so the input terminal of the inactive winding is effectively at earth potential.

One of the limitations with this system is the difficulty in designing the combining transformer. An accurate balance between the two halves of the primary, and the correct transfer ratio between each half and the secondary, is almost impossible to achieve over the whole h.f. band. Consequently there is inevitably a rather high v.s.w.r. reflected on to each transmitter output, and some power is lost in the absorber load. However, it is suitable for a limited frequency coverage—with some reduction in reflected v.s.w.r. and power loss—by means of improving the match with appropriate loading capacitors and inductors, as described in Chapter 10.

Paralleling with two transmitters in push–push

In view of the limitations of the push–pull system, it has been found in practice that better conditions can be provided by operating the two amplifiers in phase

(push–push). Although this entails an additional transformer in the output-combining network, the input circuit is further simplified and improved.

The arrangement of the input circuit is shown in Fig. 9.15. The potentiometer P for phase control and series resistors R_1 and R_2, for matching the input cable,

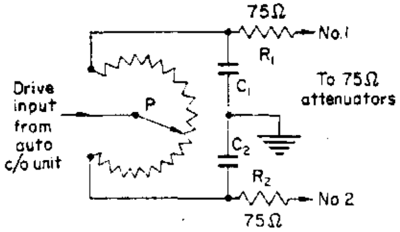


FIG. 9.15 Phasing and drive splitting for push–push operation.

are exactly the same as in the push–pull case, and only half of the input voltage is available at each amplifier input. The difference is that the two in-phase inputs can be fed directly into the amplifiers, no transformers being required to obtain an antiphase output from the single input. Thus the match on the input cable is also improved over the frequency band, due to reducing the number of components.

The output combining circuit for push–push operation is shown in Fig. 9.16. Capacitors C_1, C_2, C_3, C_4, C_5 and C_6 with inductors L_1 and L_2 are provided to

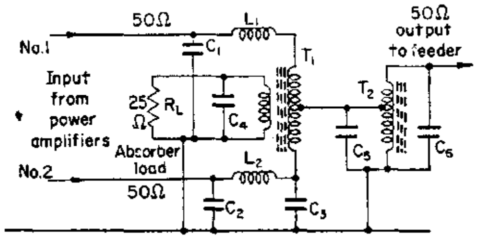


FIG. 9.16 Transformer combining unit for push–push operation.

improve the frequency response of the network. In this case, transformer T_1 is only necessary for the single amplifier condition. In parallel operation the flux induced by the two in-phase inputs cancel one another in the primary, so the two inputs are connected directly together at the centre-point output. The only design problem with this transformer for parallel operation, is that the balance between the two halves of the primary should be as near perfect as possible. Auto-transformer T_2 provides a 25 Ω-to-50 Ω ratio for matching the feeder to the transmitters, via the centre-point of T_1. Therefore the balancing and matching functions are separated, and in consequence more readily achievable.

With only one transmitter on, the power is divided equally between the output feeder and the 25 Ω absorber load, provided that the ratio of T_1 is 1 to 1 in this condition. If loads of other resistance value, such as 50 Ω, are more convenient, the ratio of transformer T_1 should be designed accordingly, but the power distribution will remain the same. There is negligible output at the terminal of the unused transmitter in either case with single operation.

Parallel operation by means of transformer combining with in-phase amplifiers is particularly appropriate for use with wideband h.f. amplifiers, because the phase delay in amplifiers of the same type is likely to be very similar. Therefore the simple phase-adjusting system shown in Fig. 9.15 gives sufficient phase control by its differential action.

For tuned amplifiers, where a greater range of phase control is likely to be required, the circuit arrangement shown in Fig. 9.2 would be more suitable. This also applies if parallel operation is required in the m.f. band, because the available phase change of the potentiometer system depends on its electrical length in terms of wavelength, which obviously decreases with frequency decrease.

REFERENCES

[1] BARTLETT, H. 'The parallel operation of broadcast transmitters'. British Patent No. 743473 (1951).

[2] MORCOM, W. J. 'The operation of transmitters in parallel'. *Sound and Vision Broadcasting*, **2**, No. 1 (spring 1961).

Part 2

Medium and Low Power

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*).

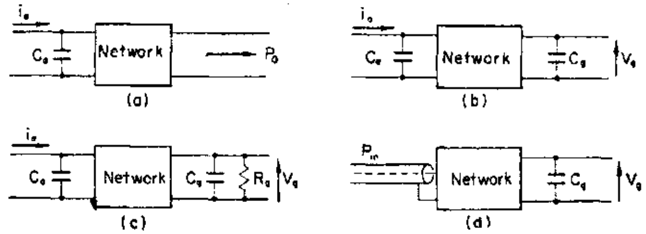


FIG. 10.1 Working conditions for a wideband amplifier. (*a*) At the output. (*b*), (*c*) At the interstage coupling. (*d*) At the input.

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]

$$P_o \max = \frac{\pi}{2} \cdot i_a^2 \cdot XC_a \qquad (1)$$

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

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

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]

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

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

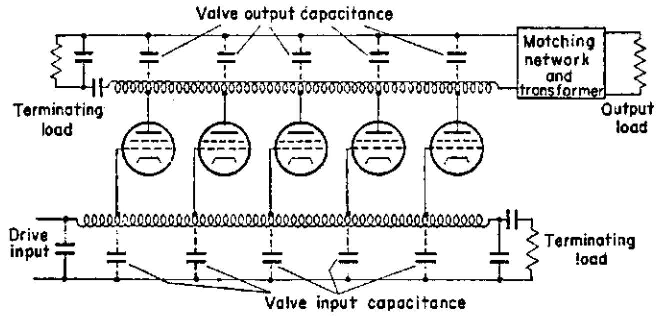


FIG. 10.2 Simplified diagram of a distributed amplifier.

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 [illegible]ract maximum power over a wide band from a given set of valves; the [illegible] provided that it [illegible]w, 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).

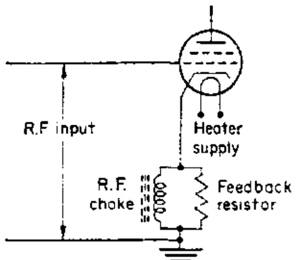


FIG. 10.3 Circuit for r.f. cathode feedback.

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

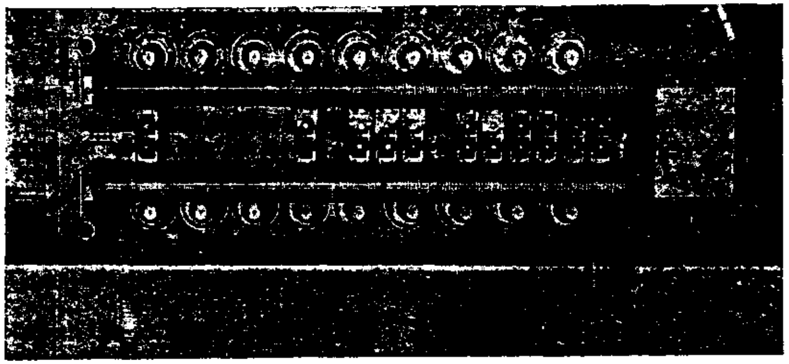


FIG. 10.4 Anode delay lines showing the tapering sections.

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

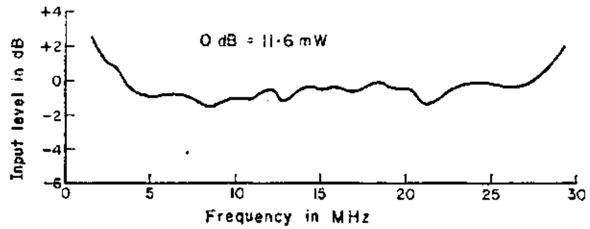


FIG. 10.5 Frequency response of two-stage distributed amplifier, in terms of drive level required for 1 kW output.

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.

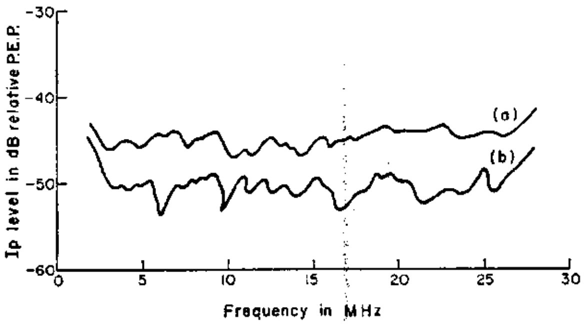


FIG. 10.6 Intermodulation products between two closely spaced tones with reference to p.e.p. (*a*) 1 kW p.e.p. (*b*) 500 W p.e.p.

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 change-over 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

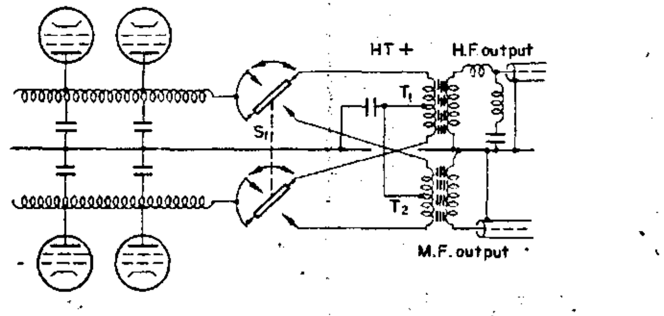


FIG. 10.7 Output circuit arrangement for h.f. and m.f. version of distributed amplifier.

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 ship-borne 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^2R 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.

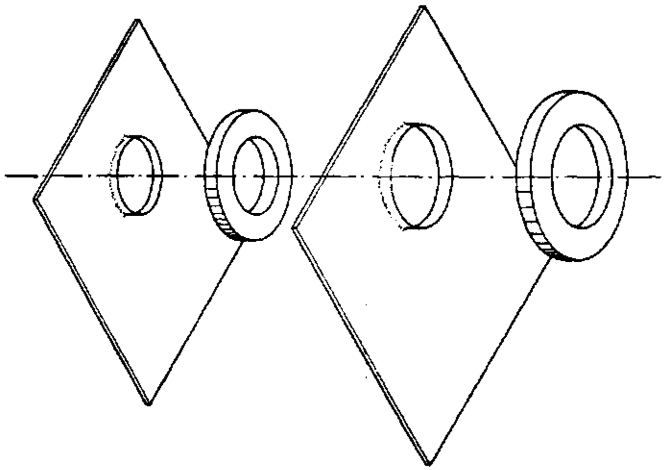


FIG. 10.8 Method of ferrite core assembly for 40 kW h.f. transformer.

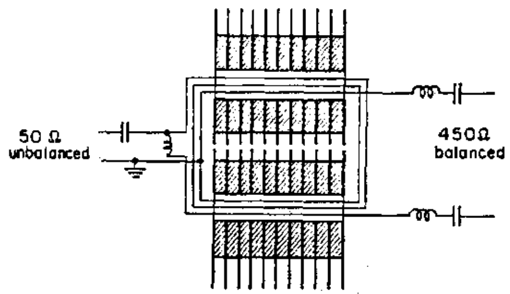


FIG. 10.9 Circuit arrangement of a 40 kW wideband transformer.

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.

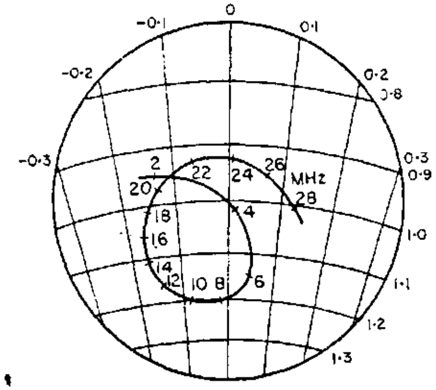


FIG. 10.10 Smith's chart presentation of input matching of a 40 kW wideband transformer.

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.

REFERENCES

[1] GREEN, E. 'Amplitude-frequency characteristics of ladder networks'. *Marconi Rev.* (1957), Monograph, p. 31.
[2] WHEELER, H. A. 'Wideband amplifiers for television' *Proc. Inst. Radio Engrs.*, **27**, 429 (1939).
[3] *Reference Data for Radio Engineers.* 4th edition, p. 187. International Telephone and Telegraph Co., New York (1956).
[4] WEINBERG, L. 'Network design by use of modern synthesis technique and tables'. Technical Memoranda Nos. 427 and 434 Research Lab., Hughes Aircraft Co.
[5] PERCIVAL, W. S. British Patent No. 460562 (1936).
[6] GINZTON, E. L., HEWLETT, W. R., JASBERG, J. H. and NOE, J. D. 'Distributed amplification'. *Proc. Inst. Radio Engrs.*, **36**, 956 (1948).
[7] HORTON, W. H., JASBERG, J. H., and NOE, J. D. 'Distributed amplifiers: practical considerations and experimental results'. *Proc. Inst. Radio Engrs.*, **38**, 748 (1950).
[8] CORMAK, A. 'Distributed amplification'. *Electronic Engng.*, 24, 144 (1952).
[9] ESPLEY, D. C. British Patent No. 623175 (1944).
[10] SOSIN, B. M. British Patent No. 846633 (1958).
[11] MAURICE, D. and MINNS, R. H. 'Very wide band radio-frequency transformers'. *Wireless Engr.*, **24**, 168, 209 (1947).
[12] 'On the use of ferrites in wide-band h.f. transformers'. *Point to Point Telecomm.*, **1**, 22 (1957).
[13] ELSEY, G. and JACKSON, N. C. O. 'A new approach to the design of ferroxcube cores for wide-band h.f. transformers'. *Mullard tech. Commun.*, **5**, 182 (1960).

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

17·5 V. As the intermediate amplifier will not normally run into grid current, the incoming voltage can be stepped up with a wideband transformer, shown as T_1 in Fig. 11.1. The terminating resistor R_1, connected across the secondary, is a much higher value than 75 Ω, so the peak voltage made available for driving the valve is increased. The amount by which the voltage can be stepped up is limited—mainly by the input capacitance of the valve in relation to the operating frequency, but also by the transformer design.

Input capacitative reactance must be of the order of 1·7 times the value of the terminating resistor if the v.s.w.r. on the drive cable is not to exceed 1·2 to 1. If the resistor on the secondary has a value of 600 Ω, the upper frequency of

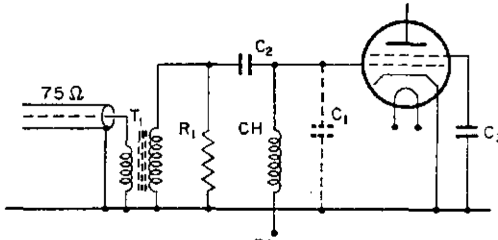


FIG. 11.1 Input circuit using a wideband step-up transformer.

operation is reached when the reactance of the valve input capacitance is about 1000 Ω.

For 600 Ω to be the correct termination on the transformer secondary from a 75 Ω cable, the voltage step-up ratio is $\sqrt{(600/75)} = 2\sqrt{2}$. The design of a transformer with this ratio is not a problem at a throughput level of 2 W, and by its use the available peak voltage at the grid is raised from 17·5 V to 50 V.

The valve used for intermediate amplifiers will be of a comparatively low power rating and not large, so the input capacitance is likely to be less than 50 pF, and is often nearer to 20 pF. The reactance of 50 pF at 3 MHz is 1060 Ω, so the input circuit arrangement of Fig. 11.1 will be suitable for the v.l.f., l.f., and m.f. bands, with a transformer having a step-up ratio of $2\sqrt{2}$ and a terminating resistor of 600 Ω on the secondary.

In order to take advantage of a similar voltage step-up for frequencies above 3 MHz, consideration should first be given to selecting a type of valve which has a low input capacitance, consistent with other characteristics being satisfactory. For a final-stage drive power (including loading) of up to 300 W, tetrode type 4CX250B (Eimac) is suitable, having an average input capacitance of 16 pF. The reactance of 16 pF at 10 MHz is 1000 Ω, so the input circuit of Fig. 11.1 will be satisfactory up to 10 MHz when using a 4CX250B tetrode.

For frequencies above 10 MHz the valve input capacitance could be tuned out for each frequency used, but bearing in mind the operational need to change frequency in the h.f. band, this is not a good solution. But, a partial tuning arrangement is quite practical, using the input circuit shown in Fig. 11.2. Values of inductance are selected by a switch, S_1, so that in combination with the valve input capacitance C_1, each inductor will cover a range of frequencies at which the reactance is 1000 Ω or more. In the majority of cases it will be found that the

two inductors L_1 and L_2 will be sufficient to cover frequencies in the h.f. band at which the reactance of the input capacitance would otherwise be less than 1000 Ω.

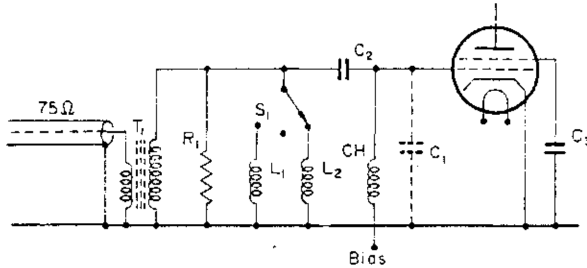


FIG. 11.2 Partial tuning of the input circuit.

For intermediate stages in linear amplifier chains, it is preferable to avoid the use of r.f. feedback as a means of improving linearity, because the input level is reduced by the amount of feedback employed. Consequently, the available output is also reduced by the same amount. If an improvement in linearity is required, the circuit arrangement shown in Fig. 3.1, Chapter 3, should be used. This is a form of envelope feedback which does not reduce the drive level appreciably.

11.3 INTERSTAGE COUPLING WITH A Π CIRCUIT

The most suitable method of interstage coupling when driving a final amplifier in a grounded-cathode arrangement is a Π circuit as shown in Fig. 11.3. It is essentially simple for tuning and loading and the final-stage input capacitance forms part of the shunt capacitance of the Π circuit, so it does not have to be tuned out by a separate control. This means that the value of the terminating

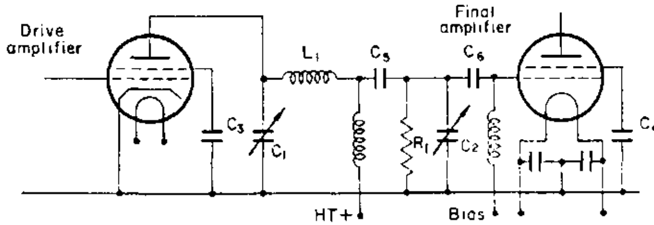


FIG. 11.3 Interstage coupling with a Π circuit.

resistor R_1 is determined only by the driving voltage required in relation to the power available from the driving amplifier.

For linear operation without grid current, apart from circuit losses, the resistor R_1 provides a constant load over the whole frequency range to be covered, so the driving amplifier loading conditions are the same for all frequencies.

For linear operation with grid current, although the grid current is not likely to be high, the load it produces must be only about one-quarter of that produced by the loading resistor R_1, otherwise peak flattening will occur with a consequent reduction in linearity. The peak value of the grid current should be kept low, otherwise considerable power will be required from the drive for good linearity.

Grid current is normal with class C operation and it is preferable to use some automatic bias on the final stage as a means of regulating the drive voltage and maintaining a constant load on the driving amplifier. Under these circumstances the load resistor R_1 is not necessary, although a resistor of high value is sometimes fitted as a further stabilizing device.

In all cases of the final stage taking grid current, there is an advantage in using a Π circuit for interstage coupling, when the final stage is in a grounded-cathode arrangement; the advantage is greater in linear amplifiers. Due to the phase reversal in the Π circuit, the *driven* negative-going voltage on the driver anode produces the positive-going driving voltage on the grid of the final stage. Without this phase reversal, such as would be produced by direct capacitative coupling, the driving power would be obtained from the stored energy in the resonant circuit. As such, the amount of peak flattening would depend on the Q factor of the tuned circuit and would give poor linearity with low Q factors.

11.4 INTERSTAGE CAPACITATIVE COUPLING

While capacitative coupling is unsuitable for driving a grounded-cathode stage, it gives the correct phase relationship for driving a grounded-grid stage because the driving half cycle is negative going. The circuit arrangement is shown in Fig. 11.4, where shunt components L_1C_1 tune the anode circuit of the driving

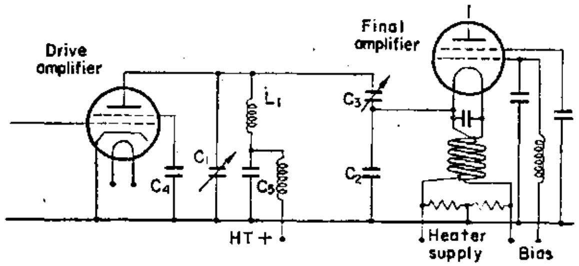


FIG. 11.4 Capacitative interstage coupling.

amplifier, and the capacitor C_3, of capacitance potentiometer C_3C_2, provides the loading control. No additional loading resistor is needed because the main load is provided by the anode–cathode current in the cathode circuit. This load is practically constant over the driving cycle and is usually much greater than that produced by the grid current. This means that peak flattening is very small, so the circuit is very suitable for linear amplification.

There are a number of other points which should be noted in connection with this arrangement. First, the driving power required is very much greater than that needed for a grounded-cathode amplifier, but it is not all wasted because

the power in the anode–cathode circuit is in series with the output power, and appears as throughput in the output circuit.

Secondly, even at minimum setting of C_3 the capacitance of the potentiometer is added to the output capacitance of the driving amplifier, reducing the frequency to which the driving amplifier will tune. This is particularly the case with amplification over a wide band of frequencies, where capacitor C_2 must always provide a low-impedance path to harmonic frequencies. In consequence, the value of C_3 is governed by the value of C_2 required at the low-frequency end of the band, so C_3 tends to be higher than the value required for the higher frequencies. This effect can be reduced by making C_2 either variable or changed in steps over the frequency range.

Thirdly, in the circuit shown, the final amplifier valve is a tetrode, which gives a high order of linear performance in a grounded-grid circuit. In grounded-grid circuits the first and second points are equally applicable to triodes which also give good linearity, but the triodes do not require a screen supply. So, it is more usual to employ triodes if the grounded-grid arrangement is used.

11.5 INTERSTAGE COUPLING WITH A QUARTER-WAVE NETWORK

One of the main characteristics of a line one-quarter of a wavelength long, is that its impedance $Z_0 = \sqrt{(R_1 R_2)}$, where R_1 is the effective resistance at the input, when the line is terminated with R_2. At any particular frequency a network of lumped capacitance and inductance elements in a Π circuit arrangement can behave as a quarter-wave line, if the shunt and series components all have the same reactance. The value of this reactance is then the impedance of the quarter-wave network of lumped elements.

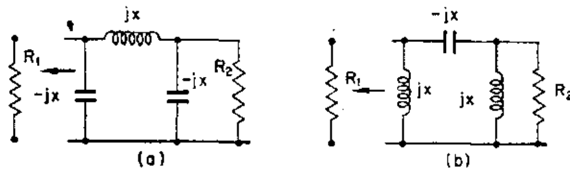


FIG. 11.5 Two types of quarter-wave network.

Two such networks are shown in Fig. 11.5(a) and (b), wherein $jx = -jx = Z_0$. In both cases the network will look like R_1 at the input when terminated by R_2, which may be of any value, i.e., $\sqrt{R_1} = Z_0/\sqrt{R_2}$, or $R_1 = Z_0{}^2/R_2$. By reducing R_2 to half value the input resistance is increased from R_1 to $2R_1$.

Now consider a requirement for a constant voltage at the output for a change in terminating resistor R_2 to half value. The power in R_2 will increase from V^2/R_2 to $V^2/0{\cdot}5R_2$ for the lower value and the power at the input of the network will also be increased by the same ratio. If I_0 is the current in the effective input resistance R_1, the power at the input will be $I_0{}^2/R_1$ for R_2 on the output. If I_1 is the input current when the output termination is $0{\cdot}5R_2$, the power will be $I_1{}^2/2R_1$, because R_1 will have increased to $2R_1$. But the power at the input will also be twice the original power, so $I_0{}^2/R_1 = 2I_1{}^2/2R_1$ and $I_1 = I_0$. It can be seen that irrespective of the value of the terminating resistor R_2 ***a quarter-wave network will provide a constant voltage output from a constant current input.***

A constant voltage is exactly that required for driving a linear amplifier when the effective input resistance changes over the driving cycle, such as occurs in a grounded-cathode stage running into grid current.

Reference to the characteristics of any tetrode will show that for a given driving voltage the anode current is substantially constant over a very wide range of anode voltage. In fact, this is so unless the anode voltage approaches the same as, or is less than, the screen voltage. By using a quarter-wave network for interstage coupling from a driving tetrode, peak flattening due to grid-current loading can be virtually eliminated. It follows that this system is particularly suitable for interstage coupling in linear amplifier chains.

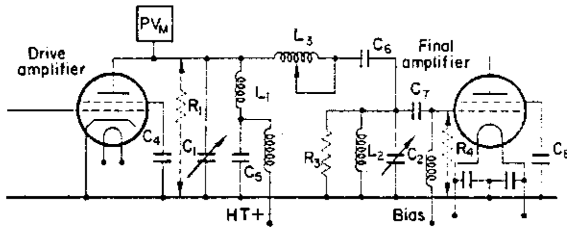


FIG. 11.6 Practical arrangement for interstage coupling with a quarter-wave network.

In giving consideration to a practical circuit arrangement, (*a*) or (*b*) in Fig. 11.5 are equally effective, but there is a slight preference for (*a*) because of its better attenuation to harmonic frequencies which may be produced in the driving amplifier. This leads to the practical configuration shown in Fig. 11.6.

The first point to consider is the impedance of the network Z_0 from the driving requirements of the final amplifier, to give its full output, and the operational load line of the drive amplifier to give the peak power of the final-stage drive.

The effective input resistance R_4 is taken as the peak grid voltage divided by the peak grid current. The added resistor R_3 is to limit the change in load-line resistance of the drive caused by the change between zero and peak grid current. Typically, R_3 is about equal to R_4, although the value is quite arbitrary and much higher values can be used effectively. The value used for the terminating resistance R_2 is given by resistances R_3 and R_4 in parallel.

The r.m.s. power required for the peak load is given by $PV^2/2R_2$, which enables the load line of the driving amplifier to be selected for peak linear output. As linear amplification is being considered, the value R_1 is twice the value of the slope resistance for all practical purposes. Therefore the required Z_0 of the network is given by $\sqrt{(R_1 R_2)}$ and the three arms of the network must be set to this numerical value. Drive anode circuit $L_1 C_1$ (including the valve output capacitance) is set to equal $-jx$, L_3 is set to jx and $L_2 C_2$ (including the final amplifier input capacitance) is set to $-jx$.

The necessity to adjust three controls for each frequency, combined with the fact that unorthodox tuning procedures make it unsuitable for automatic self-tuning, is the main reason why this arrangement is not used extensively in the

h.f. band. However, the effective Q factor of the network is unity, so the settings are not critical, and automatic frequency changing between pre-selected settings is quite a practical proposition. Its use has been proven when driving linear amplifiers with output power up to 30 kW, with triodes in grounded-cathode circuits, where the peak grid current was about one-third of the peak anode current.

In order to obtain the correct values for pre-setting the circuits, the series element L_3 must first be calibrated in terms of reactance and frequency. This can best be done by disconnecting it from the circuit at the points where it connects to the shunt circuits, and resonating the whole series arm with known values of capacitance. This is a once-only operation, subsequently the inductor is set either to logged positions or to a calibration chart.

With L set for a particular frequency, the bias is removed from the final stage (without h.t. of course), thereby providing an effective short across $L_2 C_2$, so that L_3 is in shunt with $L_1 C_1$. With a low-level signal into the drive amplifier, circuit $L_1 C_1$ is resonated with L_3, as indicated by a *maximum* reading on the peak voltmeter, making $L_1 C_1$ equal to $-jx$. Then, with final-stage bias on and the signal level below that required to give final-stage grid current, circuit $L_2 C_2$ is adjusted to series resonate with L_3, as indicated by a *minimum* reading on the peak voltmeter. This means that $L_2 C_2$ is also equal to $-jx$. Once set for a frequency, no attempt should be made to trim any of the component arms on full power. The circuit properties can only be degraded by such action.

11.6 TYPICAL EXAMPLES OF QUARTER-WAVE NETWORK CONDITIONS

Consider a final amplifier requiring a drive of 225 peak volts and 0·5 A peak grid current, which give a value of 450 Ω (225/0·5) for R_4 and an r.m.s. power of 56·25 W. If a value of 550 Ω is used for R_3, then R_2 will be 247·5 Ω, say 250, at full signal level, and a total power of 100 W will be required from the drive amplifier.

Tetrode type 4CX250B is capable of giving 100 W as a linear amplifier and the linearity is good, so reference should be made to the characteristics shown in Fig. 11.7 for this application.

With an h.t. voltage of 1400 and a static feed of 0·1 A, operating on load line A, to 0·5 A at 400 V, will give a maximum output condition of 100 W, by the approximate method

$$\frac{(0{\cdot}5 - 0{\cdot}1) \times 1000}{4} = 100 \text{ W}$$

The slope resistance of the load line is 2500 Ω (1000/0·4), giving the effective r.f. load across the input of the network R_1 of 5000 Ω. The required Z_0 of the network is $\sqrt{(R_1 R_2)} = \sqrt{(5000 \times 250)} = 1118\ \Omega$, say 1100 Ω. This is the value to which the *jx* and *−jx* arms should be set by the method described, for every operational frequency.

Now consider the low-level condition when there is no grid current in the final amplifier. Resistance R_4 does not exist and the total load R_2 is given by the 550 Ω of R_3. From $Z_0 = \sqrt{(R_1 R_2)}$, $R_1 = Z_0{}^2/R_2 = 2200\ \Omega$ and the slope of the load line changes to 1100 Ω, shown as *B* on Fig. 11.7. Obviously, the excursion is not that given by the full extent of the load line, but it enables the ratio of anode current change to be obtained by reference to the 0·4 A curve. For the same grid voltage of –10 V, the anode current on line A is 0·4 A and on line B it is 0·41 A, which is an increase of 2·5%.

Via the quarter-wave network, this means that the regulation of the final amplifier driving voltage is only 2·5% from no grid current to peak grid current.

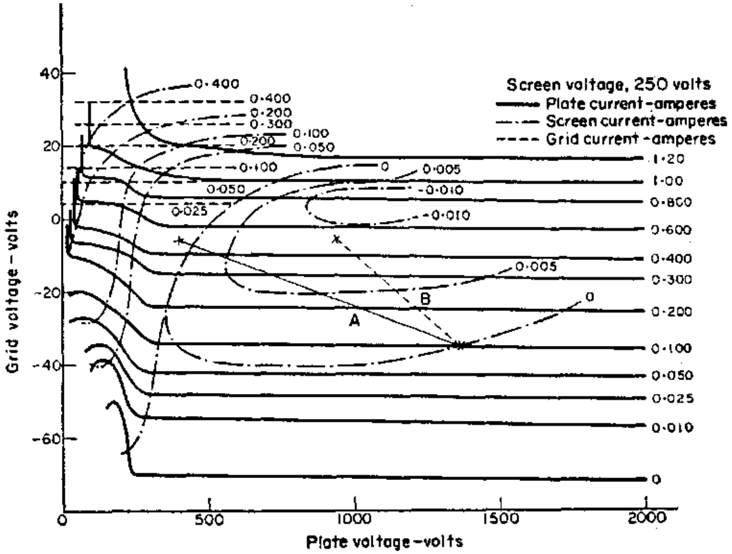


FIG. 11.7 Constant-current characteristics of tetrode type 4CX250B (Eimac).

In order to obtain a regulation of 2·5% with other forms of interstage coupling from a tetrode driving stage, the shunt resistor would have to be so low that at least 1 kW of total drive power would be required; an impossible consideration to supply an actual driver power of only 56·25 W r.m.s.

Referring again to Fig. 11.7, the power output from the driving amplifier is given with a grid excursion from –36 V to –6 V, i.e., 30 V. In the case considered as an example, with a 2 W drive available, the required 30 V grid swing can be obtained with a transformer ratio of 1·75 to 1, terminated on the secondary with 225 Ω. As the input capacitance of the 4CX250B is 16 pF (average), its reactance is 360 Ω at 28 MHz. With a shunt resistor of 225 Ω, it means that the input circuit shown in Fig. 11.1 will be satisfactory up to 28 MHz, without any shunt inductance needed for higher values of shunt resistance as shown in Fig. 11.2.

12

Amateur Transmitters

12.1 POWER CONSIDERATIONS

A feature of h.f. transmitting equipment for use by amateurs, is the regulation which imposes a limit on the maximum power that may be used. This limit is defined as the maximum peak r.f. power given by a d.s.b. amplitude modulated system (A3) with 150 W d.c. into a final r.f. amplifier operating at 66% efficiency.

With a conversion efficiency of 66% for a d.c. power input of 150 W, the r.f. feeder carrier output is 100 W and the d.c. input is unaffected by anode modulation; the increased output power being supplied by the modulator. At 100% modulation, the power in each of the two sidebands is one-quarter of the carrier power, which is half the carrier amplitude per sideband. This power and amplitude relationship is shown in Fig. 12.1(*a*). It will be seen that with a carrier power of 100 W, each half-amplitude sideband is 25 W and the peak amplitude is twice the carrier amplitude, which is 400 W p.e.p.

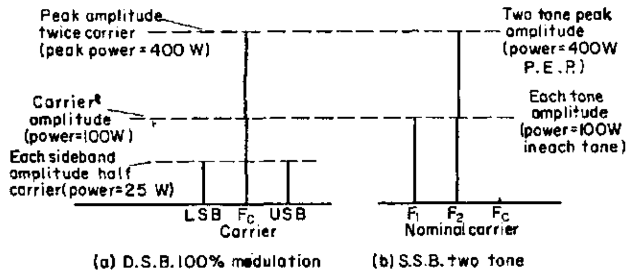


FIG. 12.1 Comparison between the spectral components in d.s.b. and s.s.b. to give the same p.e.p.

Peak power relationship between d.s.b. and s.s.b.

With an upper power limit in terms of p.e.p., there is an obvious advantage in employing a system by which as much as possible of the total power is used to convey the intelligence, e.g. by using an s.s.b. system.

In a d.s.b. system all the intelligence is conveyed in either of the two sidebands, each of which is only one-sixteenth of the peak power, as shown in Fig. 12.1(*a*) (25 W out of a total of 400 W). With an s.s.b. system, after d.s.b. modulation at a low-power level the carrier and one sideband are filtered out so that only one sideband is radiated at a peak power equal to the total peak power. The peak intelligence power radiated with s.s.b. is sixteen times that with d.s.b. (In order to recover the intelligence, it must be remembered that the carrier must be re-inserted at the receiver at the correct frequency.)

To check the performance of s.s.b. systems, it is usual to apply two equal amplitude tones to one sideband, so that the peak power is reached every time the radio frequencies corresponding to the tones are in phase, i.e., at the difference frequency of the tones. With two-tone modulation applied, the power relationship with s.s.b. is shown in Fig. 12.1(*b*) and compared with d.s.b. in Fig. 12.1(*a*).

The dynamic condition of the two systems is shown in Fig. 12.2(*a*) and (*b*) for the same p.e.p. as would be seen by displaying the r.f. waveforms on an oscilloscope. The two-tone method of checking is only a test condition, and in practice the whole of the peak power is available for the single-channel operation normally used.

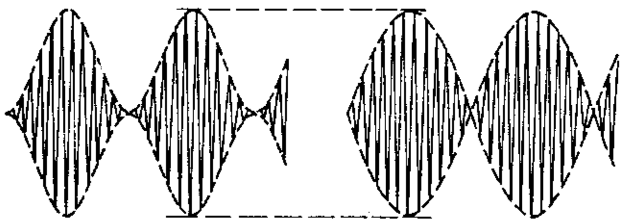


FIG. 12.2 Comparison between d.s.b. and s.s.b. r.f. waveforms as seen on an oscilloscope.

Most amateurs started with very low power with such good results that a p.e.p. of 400 W on s.s.b. seems unnecessarily high, and few use as much as 200 W p.e.p. Even at this level, the power is eight times the power of 25 W in one sideband of a d.s.b. system which has a total p.e.p. of 400 W.

It might be argued that the two sidebands of d.s.b. system add up at the receiver to give a higher signal strength, but this is not always the case. In propagation conditions, appropriate for selective fading, the path length travelled by the two sidebands is different, and the difference varies. The result is a relative change of phase between the two sidebands at the receiver, so they add and subtract in a random manner, causing fading of the signal. This cannot occur with only one sideband, and accounts for the comparative absence of this type of fading with s.s.b. operation, resulting in a more steady signal strength.

Mean power and peak power

The relationship between mean and peak power conditions on s.s.b. is given in Table 12.1 for single frequency and the two-tone test condition for amplifiers operating in class B. These assume sinusoidal waveforms and perfectly linear valve characteristics, but they are sufficiently accurate for practical purposes when used with valves suitable for linear r.f. amplification. If greater accuracy is required, reference should be made to the method described in Chapter 2.

The formulae relate to single-valve operation and the following definitions apply for PI_A and PV_A.

PI_A = the anode current excursion from zero to the peak current at peak signal (the static feed at zero signal level is ignored by this approximation);

PV_A = the anode voltage excursion from the h.t. voltage to the anode voltage at peak anode current.

TABLE 12.1

	Single frequency	Two-tone test
I_A, d.c.	$\frac{PI_A}{\pi}$	$\frac{2PI_A}{\pi^2}$
Power output, watts r.m.s.	$\frac{PV_A \cdot PI_A}{4}$	$\frac{PV_A \cdot PI_A}{8}$
Power input, watts, d.c.	$\frac{PI_A \cdot \text{h.t.v.}}{\pi}$	$\frac{2PI_A \cdot \text{h.t.v.}}{\pi^2}$
Anode dissipation, watts	$PI_A\left(\frac{\text{h.t.v.}}{\pi} - \frac{PV_A}{4}\right)$	$PI_A\left(\frac{2\text{h.t.v.}}{\pi^2} - \frac{PV_A}{8}\right)$
Conversion efficiency % (d.c. to r.m.s. output)	$\frac{25 \cdot \pi \cdot PV_A}{\text{h.t.v.}}$	$\frac{6{\cdot}25 \cdot \pi \cdot PV_A}{\text{h.t.v.}}$

First, note that the mean power to peak power ratio is 1 to 2. The next step is to obtain the required operating load line on the characteristics of a particular valve type by means of the formulae. As an example, assume that the required output is 200 W p.e.p. On the two-tone test the mean power output will be 100 W (50 W per frequency) and the valve will need to have an anode-dissipation capability of about 100 W. A tetrode is the obvious choice because of the low-power drive required. Tetrode type QV08-100 (Mullard) allows an anode dissipation of 100 W and the characteristics are quite linear, as seen in Fig. 12.3. In addition, it is an inexpensive valve.

At peak anode current in the region of 1·0–1·5 A, it is seen that the characteristics are substantially linear down to V_A of 120 V, so with an h.t. voltage of 800 V the PV_A swing will be 680 V. Allowing 5 W for circuit losses and any departure of individual valves from the typical characteristics, 210 W p.e.p. will be required and a mean power of 105 W.

Then $PV_A = PI_A/8 = 105$, and as $PV_A = 680$ V, $PI_A = 8 \times 105/680 = 1{\cdot}235$ A. Thus

I_A d.c. $= 2PI_A/\pi^2 = 0{\cdot}25$ A
Power input, d.c. $= 0{\cdot}25 \times 800 = 200$ W
Power output, r.m.s. $= 105$ W
Anode dissipation $= 95$ W

Even in the two-tone condition the anode dissipation is below the rated maximum of 100 W and the output is 210 W p.e.p.

The static anode feed is selected to give good linearity, provided that the anode dissipation is not exceeded. Typically, for a valve of this type it will be about 0·1 A, giving a zero-signal dissipation of 80 W. This means that the range of d.c. input power is from 80 W with no signal, to 200 W on two-tone. With normal speech the average modulation depth is about 20–30%, so the *average* d.c. input will be of the order of 100–110 W, with signal peaks of 210 W p.e.p.

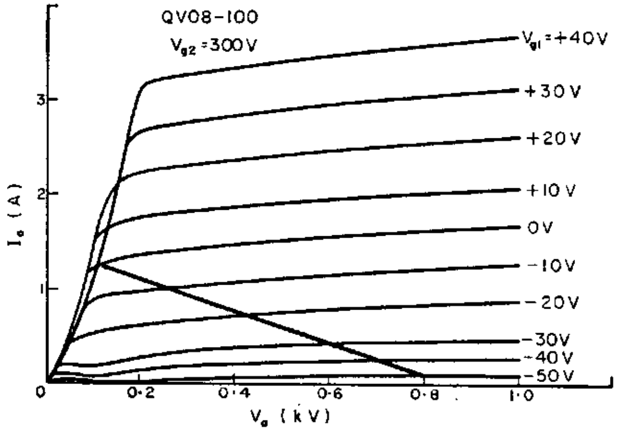


FIG. 12.3 Characteristics of tetrode type QV08-100 with load line for 210 W p.e.p. output.

Compare this with the d.c. input power for 400 p.e.p. output with a d.s.b. system at 100% modulation. The a.f. power required to modulate the 150 W input to the r.f. amplifier is 75 W, and the modulator efficiency can be taken as 50%. Thus the d.c. input to the modulator is 150 W, giving a total d.c. input of 300 W for 400 W p.e.p. output.

With an average modulation depth of 20–30% on speech, the d.c. input to the modulator is about 90–100 W (assuming a static feed of 0·1 A for the two modulator valves). The d.c. into the r.f. amplifiers remains the same, at 150 W, so the *average* total d.c. input is 240–250 W, for a peak power in each sideband of only 25 W.

Drive power for s.s.b. operation

Referring to the load line on the valve characteristics shown in Fig. 12.3, the bias is −50 V and a grid voltage of −2 V is needed to give I_A of 1·235 A at an anode voltage of 120 V. The required grid swing is 48 V and there will be no grid current, so the valve itself will require no drive power.

However, the 75 Ω cable must be matched correctly, and if terminated directly with a 75 Ω resistor a drive power of 15 W is required.

In cases where an amateur wishes to add an amplifier to increase the output power, terminating the feeder with a 75 Ω resistor directly across the valve input

is the obvious choice (Fig. 12.4). It ensures that the feeder is correctly matched at all frequencies and provides sufficient damping on the input to give a very stable operating condition without neutralizing. Note that all the drive power is dissipated in the terminating resistor.

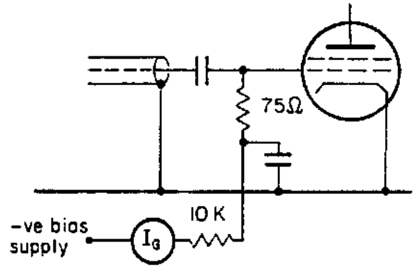


FIG. 12.4 Input circuit arrangement when high-power drive is available.

If only a low-level drive is available, a voltage step-up transformer must be used, and this can be either a wideband arrangement with a ferrite core, or separate transformers with tuned secondaries. In either case, the feeder must be terminated by a high-value resistor on the secondary, the actual value depending on the step-up ratio. If a preferred value of resistor is used, say 6800 Ω, the required transformer ratio is 9·5 to 1; $\sqrt{(6800/75)}$. Neglecting circuit losses, this means that the necessary drive power will be only 0·17 W p.e.p.

12.2 TYPICAL CIRCUIT DIAGRAM FOR AN R.F. AMPLIFIER FOR 200 W P.E.P.

A typical circuit arrangement for a 200 W p.e.p. amplifier with a tetrode valve is shown in Fig. 12.5. To cover the five amateur bands between 3·5 MHz and

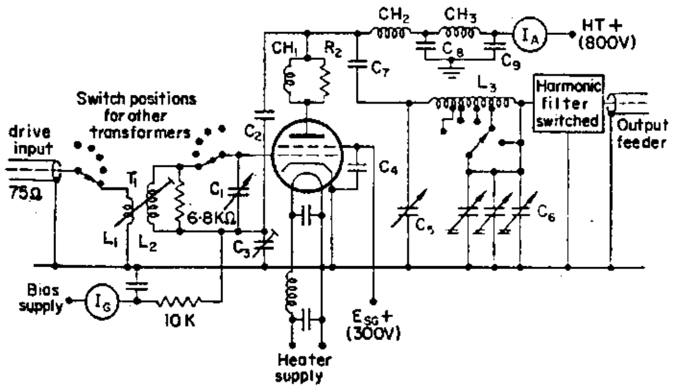


FIG. 12.5 Typical circuit diagram for r.f. amplifier.

28·7 MHz, the output circuit is switched in ranges, but separate transformers are used for each band on the input circuit.

The input circuit

It is theoretically possible to use a single input transformer with tapping points for both primary and secondary to cover the five ranges. However, practical difficulties arise in obtaining adequate mutual inductance between primary and secondary to produce a good match on the incoming feeder. Furthermore, separate transformers are an advantage if the available drive power differs between bands, enabling individual terminating resistors and different step-up ratios to be used for each band.

Wideband transformers are a special case, so consideration is limited to individual band transformers with tuned secondaries, although only one transformer is shown on the diagram, with a secondary resistor of 6·8 kΩ.

The drive is fed into the primary L_1 of the step-up transformer T_1, the secondary of which, L_2, is tuned by the variable capacitor C_1.

The negative bias is fed to the grid via a low-reading grid-current milliammeter and a 10 kΩ resistor, which is to prevent the grid current rising too rapidly. Note that there is no grid choke. If a choke were to be fitted at the high potential end it would ruin the feeder termination, and if fitted lower down it would tend to cause instability. For full output the peak grid voltage approaches zero (typically −2 V), so a reading on the meter shows that this voltage is being exceeded at peaks, and gives an indication of the required peak drive for full power output.

Neutralizing and stabilizing arrangements

The low-potential end of the transformer secondary is connected to earth via capacitor C_3, which is in series with the input capacitance of the valve across the winding. This provides a point on the input circuit where the r.f. voltage is in antiphase with the driving voltage, which is the correct phase for neutralizing directly from the valve anode, via capacitor C_2. Although the interelectrode capacitance between anode and control grid is quite low with this valve type (less than 0·9 pF), neutralizing is necessary to reduce the positive feedback, which is ruinous to linearity, even if not required to prevent self-oscillation at the fundamental frequency.

If capacitor C_2 were the variable element of the neutralizing path, it would have to be a very low-capacitance type with a high voltage rating and would be quite tricky to adjust, because neither end is earthed. By making C_3 the variable element these objections are overcome, and the variable capacitor is much less expensive.

The network consisting of choke CH_1 and resistor R_2 is an anti-spurious device to prevent self-oscillation at much higher frequencies. It is usually made by winding two or three turns round a 1 W resistor but insulated from it. The inductance of the choke must be quite low, otherwise the fundamental voltage across the network will produce excessive dissipation in the resistor. It may be necessary to use a resistor of a higher dissipation rating to avoid excessive heating, or even burn out. The optimum value of resistor is that which will give a network Q factor of unity, i.e., the resistance being equal to the reactance of the choke at the most likely spurious frequency. In this arrangement the spurious frequency can be considered as being in the region of 90 MHz.

It should be noted that choke CH_1 has to carry the r.f. current through the anode-screen-earth capacitance of the valve, which is about 15 pF including strays. The reactance of 15 pF at 28·7 MHz is 370 Ω. As the peak anode voltage is 680 V, the r.m.s. current through the choke will be about 0·75 A on the two-tone test at 28·7 MHz, but less at lower frequencies. The choke also has to carry the d.c. current of 0·25 A under the same conditions.

An important feature for the prevention of a spurious oscillation in the region of 90 MHz is the method of connecting the capacitor C_4, which is an r.f. by-pass between screen grid and cathode. The capacitance value is not critical, 0·002 μF being typical, but the capacitor must be a low-inductance type and the connecting leads must be as short and wide as possible, in order to reduce the inductance of the by-pass path. In this type of oscillation the screen behaves as an anode, hence the importance of reducing the inductance of the by-pass path.

An alternative approach is to make C_4 variable and to tune it to series resonance with the lead inductance at the spurious frequency, providing zero reactance between screen and cathode. Once set, this capacitor need not be readjusted, because this circuit is unaffected by switching between the various frequency bands.

The anode-output circuit

The anode-output circuit is a simple Π arrangement. Capacitor C_5 is the input shunt arm for tuning, tapped inductor L_3 is the series element and capacitor C_6 is the output shunt arm for controlling the valve loading. Capacitor C_6 consists of three variable capacitors in parallel ganged together, because coupling directly from a Π circuit into a feeder requires a very large capacitance for 3·5 MHz. Due to the large capacitance of C_6 necessary at the lower frequencies, it is preferable to use a 75 Ω output feeder, for which the actual value will be about two-thirds of that required for a 50 Ω feeder.

As in normal amateur operation there is only a single channel, good linearity is not required to avoid interchannel cross-talk. It is necessary for good quality, and especially to reduce harmonic radiation. A single Π circuit does not give adequate harmonic attenuation, hence the tuned harmonic filter in the output feeder, if necessary band-switched with the input and anode circuits. Particular attention should be given to harmonics falling within television bands, because T.V. interference is frequently—but often erroneously—attributed to amateurs. For this band it is not difficult to achieve an attenuation of about 70 dB with a relatively simple filter.

The chokes CH_2 and CH_3 in combination with capacitors C_8 and C_9 form a filter network to keep the r.f. out of the anode d.c. meter and the h.t. feed line. Capacitor C_7 is a blocking capacitor to isolate the d.c. anode supply from the Π circuit, enabling lower-voltage components to be used. However, it does mean that choke CH_2 is difficult to design, because the full r.f. voltage at the valve anode is applied across it at all operating frequencies. Consequently, considerable care must be taken to avoid resonance within the choke, usually by using a choke wound in sections of different sizes. It should not be forgotten that blocking capacitor C_7 has to carry the total r.f. current into the Π circuit. This will be of the order of 5 A r.m.s. at 28 MHz for a 200 W p.e.p. amplifier on the two-tone test.

12.3 DESIGN OF THE INPUT CIRCUIT

Features of the input circuit

The main problem in the design of the input transformer is to obtain both the correct voltage step-up ratio and a good match for the 75 Ω incoming drive cable, at each of the five frequency bands between 3·5 MHz and 28·7 MHz. In this respect the relatively narrow frequency range in each band is an advantage, because it enables a mid-band setting to be suitable for each band.

Using the circuit shown in Fig. 12.5 the coupling between primary and secondary must be very tight to obtain a resistive termination on the incoming feeder. Here a ferrite core is an advantage, but quite a good match can be obtained with an air-core transformer.

The square root of the primary–secondary resistance ratio determines the required step-up voltage ratio, which is also given approximately by the primary–secondary turns ratio. The number of primary turns for any range is approximately the number of secondary turns divided by the square root of the resistance ratio, but the optimum primary turns must be determined experimentally.

In order to calibrate the transformers, the grid-current meter is connected directly to earth instead of via the negative bias. This is a test condition by which the valve itself and the grid-current meter behave as a peak voltmeter, enabling the circuit to be set up on all frequencies, with a signal generator at the input instead of the drive. The only stipulation is that the signal generator has a 75 Ω output.

Note that the input capacitance C_V of the valve is in series with capacitor C_3 (which is the lower end of the neutralizing capacitance potentiometer C_2C_3) across the secondary. As the value of C_3 is likely to be at least twenty times the input capacitance C_V, for practical purposes it can be considered that the valve capacitance is directly across the secondary.

Input transformer design and calculation

Numerous transformer configurations can be considered, but there are certain features which are applicable to all designs.

(*a*) The turns must be readily adjustable and tapping points may be required. This means the use of bare wire and imposes a minimum limit on wire gauge and spacing between turns.

(*b*) The lead lengths from the transformer, including the switches, should be as short as possible. This indicates that the transformer must not be too large either in length or diameter, imposing an upper limit on wire gauge and turn spacing.

(*c*) In order to limit the transformer size, it is an advantage to use a high-value tuning capacitor. On the other hand, if the maximum capacitance is high, the minimum capacitance will also be high, so very low inductance values will be required for the high-frequency ranges and adjustment will be very critical.

(*d*) The whole input circuit must be well screened from the anode circuit because a tuned-grid tuned-anode arrangement is very prone to self-oscillation with very little coupling between them.

The overall size of the transformer can be reduced by using a ferrite core with the winding directly on to the ferrite. However, this tends to make the primary

adjustment very critical, especially for the higher-frequency ranges, so an air-cored transformer is used in the example shown below.

Consider a design to cover the frequency bands of 3·5–3·8 MHz, 7·0–7·1 MHz, 14·0–14·4 MHz, 21·0–21·4 MHz and 28·0–28·7 MHz, with a variable capacitor of 300 pF maximum and 30 pF minimum.

The input capacitance of the tetrode type CV08-100 is 30 pF, so the maximum capacitance available is 330 pF to tune to the lowest frequency of 3·5 MHz. With an allowance for tuning, consider a capacitance of 300 pF for determining the maximum inductance of the transformer secondary.

For the highest frequency of 28·7 MHz, the minimum possible capacitance is the sum of the capacitor minimum and the valve capacitance, i.e., $30 + 30 = 60$ pF. With allowance for tuning, work out the inductance required for the smallest secondary, on the basis of tuning with 70 pF.

Assuming the secondary is shunted by a resistor of 6·8 kΩ, for the input to match a 75 Ω feeder, the transformer ratio is $\sqrt{(6800/75)} = 9{\cdot}5$ to 1. As the turns ratio between secondary and primary is approximately the same as the voltage ratio, the primary turns will be at about one-tenth of the secondary turns *on all ranges*.

Bearing in mind the previously mentioned features (*a*), (*b*) and (*c*), for this example it is considered that a reasonable compromise is given by winding the lowest-frequency transformer on a 1·25 in. diameter former, with 18-gauge tinned-copper wire at twelve turns per inch. By reference to Appendix V, or by some other method, it will be found that the required maximum secondary inductance of 6·9 μH is given by twenty turns, giving a basic transformer size of approximately 1·5 in. diameter, 1·75 in. long. A similar former would be suitable for the 7 MHz band, or tapping points on the 3·5 MHz transformer may be used.

The next step is to work out the inductance required for the other frequency bands, starting with the highest frequency.

The conditions for the lower- and upper-frequency limiting bands are given in Table 12.2, together with the conditions for the intermediate bands, selected on the basis of a linear frequency/Q factor characteristic. For frequency bands above 7 MHz it will be necessary to use formers of smaller diameter, otherwise the primary will be a small fraction of one turn, and adequate mutual coupling between primary and secondary will be impossible. For these transformers of small diameter, reference should again be made to Appendix V.

TABLE 12.2

Frequency band, MHz	3·5–3·8	28–28·7	7–7·1	14–14·4	21–21·4
Frequency for calculation, MHz	3·5	28·7	7	14	21
Secondary, pF	300	70	162	100	80
Secondary reactance, Ω	145	80	140	114	95
Q factor	46·7	85	48·6	60	71·5
Variable capacitance, pF	270	40	132	70	50
Secondary inductance, μH	6·9	0·44	3·18	1·3	0·72
Secondary turns	20	—	11	—	4
Primary turns (approx.)	2	—	1·125	—	—

Matching the input transformer to the feeder

This adjustment is made with the aid f a signal generator having a 75 Ω output of about 1 V. A low-reading r.f. voltmeter will also be required if one is not fitted to the instrument. Set the frequency required, check the output voltage into a 75 Ω resistor at the end of a 75 Ω cable just long enough to connect (subsequently) into the socket for the drive input, call this voltage V_i. With the grid circuit arranged for testing and the heater, but no other supplies, on, remove the test resistor and connect the signal generator output into the drive socket, leaving the peak voltmeter connected.

Tune the secondary to a maximum indication on the grid-current meter. If the transformer is correctly matched from the 6·8 kΩ secondary to 75 Ω primary, the input voltage V_i will be the same as with the test resistor, and the grid-current meter will read about 0·95 mA (0·95 mA in 10 kΩ = 9·5 V). If voltage V_i is higher or lower than that when feeding the test resistor, the primary for that frequency band must be adjusted in steps until the correct impedance match is obtained. Tapping points might be used with advantage.

On completion of the matching process, it is obviously necessary to remove the earth connection from the grid meter and restore the bias circuit to normal.

12.4 DESIGN OF THE ANODE-OUTPUT CIRCUIT

Description of the Π network

The Π network shown in Fig. 12.5 is a very convenient arrangement for coupling the valve output to the feeder. The two variable capacitors C_5 and C_6 control the tuning and coupling, respectively, and band changing is provided by switching taps on inductor L_3. A simplified version of this circuit is shown in Fig. 12.6. By means of the network, the feeder impedance R_F is converted to a

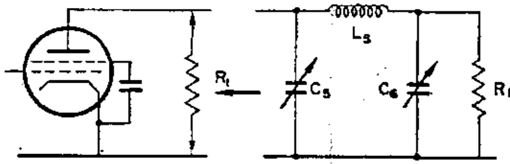


FIG. 12.6 Basic Π circuit for anode-output coupling.

non-reactive load R_l across the valve, the value of which is selected to give the required operating conditions on the valve. The feeder impedance need not be purely resistive, the circuit will give a non-reactive load on the valve with a complex feeder impedance, but a v.s.w.r. limit is usually imposed of about 2 to 1.

It is important to note that the value of R_l is *not* the resistance of the valve load line, but is derived from the output power and the peak r.f. voltage excursion ($R_l = PV^2/2$ W). In the case of a single valve operating in class B, R_l *is almost exactly twice the load-line resistance.*

A more detailed explanation of the function of the Π circuit is given in Fig. 12.7, with appropriate formulae in terms of capacitative and inductive reactance.

Obviously, the actual values of capacitance and inductance must be calculated for the operating frequency. Working back from the feeder, shunt components R_F and C_6 are converted to their equivalent series components r_2 and C_5. An inductor, L_{32}, having a reactance of the same numerical value as the reactance of C_5 is connected in series with r_2 and C_5, leaving r_2 only. The next step is to provide a series inductor and shunt capacitors of appropriate reactances to convert resistor r_2 to a pure resistance R_l across the valve, of the correct resistance value. The ratio R_l/r_2 is the starting-point from which the reactance of the series inductor L_{31} can be calculated. Series components r_2 and L_{31} are then converted to their equivalent shunt components R_l and L_p. Finally, shunt capacitor C_5 is added, the reactance of which is numerically equal to the reactance of L_p,

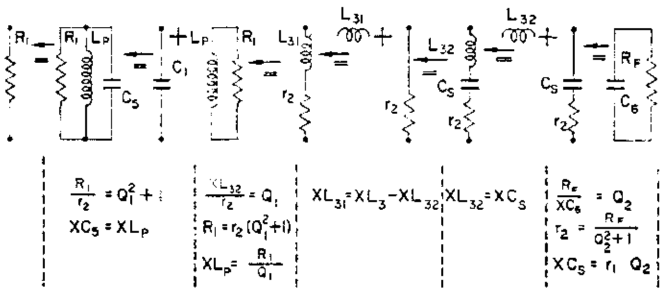


FIG. 12.7 Breakdown of the Π circuit.

thereby tuning out the reactance and leaving the resistive load R_l on the valve anode.

Two points are worth noting while considering the breakdown of the Π circuit. First, the division of inductor L_3 into two sections L_{31} and L_{32} is a device for calculation only. In practice, the point of division is determined by the values of C_5 and C_6.

Second, at this division point the circuit is purely resistive of value r_2. This means that the second part of the network, including the feeder impedance, looks like the pure resistance r_2. Consequently the r.f. current in this resistor is the same as the current in the second part of the inductor, so it is the same as in the whole inductor. Also, the output power effectively appears in this resistor, so it is a simple matter to determine the r.f. current in the series inductor. For example, if the value of r_2 is calculated to be 4 Ω and the mean power output is 100 W, the current in the series inductor L_5 is 5 A r.m.s. ($I = \sqrt{(100/4)} = \sqrt{25} = 5$). This means that the gauge of wire used for the inductor must be sufficiently large to carry this current without excessive loss and consequent overheating. For 5 A r.m.s. at 28 MHz it would be advisable to use tinned-copper wire of not less than 10 gauge (0·125 in. diameter approx.).

Typical design for an amplifier giving 200 W p.e.p.

In Section 12.1 it was shown that tetrode type QV08-100 is suitable for use as a linear amplifier with an output power of 210 W p.e.p.

From the characteristics given in Fig. 12.3 at a screen voltage of 300 V and a d.c. anode supply of 800 V, the selected load line operated from 0·1 A at 800 V to 1·235 A at 120 V. Thus the anode voltage swing was 680 V and the load-line resistance 600 Ω (680/1·135). However, the effective r.f. load resistor R_1 is obtained from the power output of 210 W and the peak anode voltage excursion, i.e., $R_1 = 680^2/2(210\ \text{W}) = 1100\ \Omega$.

Having determined R_1, the next step is to work out the component values for the highest-frequency band, in conjunction with the probable values required for the lowest-frequency band. It is necessary to consider the two extreme frequencies together, because of the conflicting requirements. At the lowest frequency the output capacitance C_6 (Fig. 12.5) has not only to cover the value required by a resistive feeder, but also that due to an inductive type of mismatch; hence a large value is required, and in consequence C_5 must also be relatively large. If C_5 maximum is large, its minimum will be greater than it would be if its maximum value were lower. As the frequency range to be covered is 8·2 to 1, a capacitance value of 300 pF maximum for C_5 is a reasonable compromise.

Example of determination of component values

By using the formulae given in Fig. 12.7, component values for a 210 W p.e.p. amplifier covering the five amateur bands from 3·5 MHz to 28·7 MHz are shown in Table 12.3. The basic information in the calculations is given below.

(1) The valve considered was tetrode type QV08-100, with the dynamic load line as drawn on Fig. 12.3, giving an effective r.f. resistance R_1 of 1100 Ω.

(2) The valve output capacitance is normally 13 pF, but allowing for strays on the anode circuit, the minimum was considered to be 25 pF.

(3) The variable tuning capacitor had a range of 30–300 pF, so the minimum capacitance total was 55 pF. With some allowance for tuning, 70 pF was considered to be the resonant capacitance at 28·7 MHz.

(4) With the same variable capacitor, the maximum possible total capacitance was 325 pF, so 300 pF was taken as that required for resonance at 3·5 MHz.

(5) Having determined the limiting conditions for the two extreme frequencies, the values selected for the intermediate bands was on the basis of a linear Q factor/frequency characteristic.

In Table 12.3 it will be noted that the r.f. current in the Π circuit inductor has been calculated in order to determine the size of wire with which to wind the solenoid. Although tinned-copper wire has a high r.f. resistance compared with bare copper, at the current values indicated 0·125 in. o.d. tinned copper wire can be used without undue temperature rise.

The turns shown are based on an inductor of 2·5 in. diameter, wound with 0·125 in. tinned-copper wire at four turns per inch. Preferably, the inductor former should be of the skeleton type, in order to simplify the tapping point arrangement and to provide better cooling.

It should be pointed out that the v.s.w.r. of 2 to 1 was selected to show its effect on the value of the output capacitor, particularly at lower frequencies. If such a condition should arise in practice, it would be far more advantageous to improve the feeder match.

Having calculated the component values, it is just as important to confirm that they are realized in the amplifier. A very useful method is to obtain a capacitance calibration for one of the variable capacitors, such as that used for tuning the Π

TABLE 12.3

Frequency band, MHz	3·5–3·8	28–28·7	7–7·1	14–14·4	21–21·4
Frequency for calculation, MHz	3·5	28·7	7·0	14·0	21·0
Π input capacitance (tuning), pF	300	70	171	134·5	82·5
Effective series resistance, r_2, Ω	20·2	5·86	16·0	10·9	7·6
Series inductance, μH	8·25	0·55	3·70	1·56	0·86
Π output capacitance (loading) pF, for $R_F = 75\ \Omega$	1000	255	583	416	300
Range of output capacitance for 2:1 v.s.w.r., pF	530–1470	198–312	350–816	300–532	224–378
Current in inductor at 100 W r.m.s. (200 W. p.e.p.) amps, r.m.s.	4·125	2·23	2·5	3·03	3·63
Inductor turns (approx.)	17	2·25	8·5	5	3

circuit. With the aid of a grid dip oscillator this enables the inductors to be set to the required values and the valve output capacitance to be measured *in situ*, etc., when setting up the circuit. Furthermore, it provides a means of checking that the tuning and loading conditions calculated are achieved in practice.

The method of setting the tapping points on the Π inductor is indicated by Fig. 12.8. The output-loading capacitor is short circuited and the inductor taps

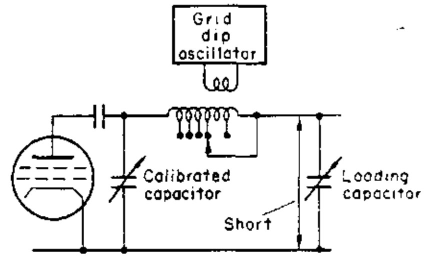


FIG. 12.8 Method of calibrating inductor for Π circuit.

are adjusted to give the required values to resonate with the known capacitance of the tuning capacitor. Resonance is indicated in the grid dip oscillator. This method enables all the strays to be taken into account when setting the tapping points, without the amplifier being switched on.

12.5 SEND–RECEIVE SWITCHING

A feature of amateur operation is the need for a quick changeover between send and receive. Normally this is accomplished by means of two relays, one of which changes the antenna between transmitter and receiver, while the other disconnects the transmitter screen supply when receiving. Even with the screen supply removed, the white noise (wideband thermal noise) produced by the transmitter valve and input circuits is relatively high compared with the received signal level. Also, the attenuation of the antenna changeover relay is not infinite, so some of

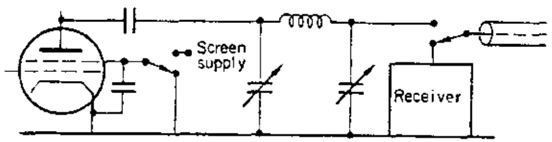


FIG. 12.9 Send–receive switching with relays in the receive position.

the transmitter noise appears on the antenna in the receive position. The result is a background of 'mush' on the received signal, which is often attributed to atmospherics picked up by the antenna.

This breakthrough of transmitter noise into the antenna can be attenuated to a negligible level by using a back contact on the screen supply relay, to earth the screen during reception, as shown in Fig. 12.9. It has been found in practice that the earthed screen produces a remarkable reduction in background 'mush', enabling very low-level signals to be received with clarity.

13

Solid-State Amplifiers

13.1 THE PRESENT STATE OF THE ART

The progress in solid-state devices continues unabated, but at the present time their application to r.f. power amplification is limited to low-power levels. In parallel with this progress, manufacturers of solid-state devices issue application reports at the earliest possible opportunity, in order to keep prospective customers up to date with their products. In consequence, it is considered that a comprehensive review of low-power solid-state amplifiers would not only be superfluous, but most probably out of date by the time this book is published. Therefore, consideration is restricted to certain similarities and differences between transistors and valves when used as r.f. power amplifiers.

Transistors are low-impedance devices with characteristics which make them particularly suitable for on–off applications. As such, they are more appropriate for non-linear amplifiers. Another characteristic feature of transistors is their susceptibility to temperature changes, tne performance falling with rising temperature. This is a further point in favour of non-linear amplification with transistors, which is more efficient than linear amplification, so the power dissipated within the transistor is less for the same power output, and the temperature rise is less.

It follows that as solid-state devices with higher-power rating become available—which undoubtedly they will—the increase in r.f. power output from transistor amplifiers will be most rapid in applications requiring non-linear amplification. It also follows that there is every incentive towards high conversion efficiency if the maximum possible output is to be obtained. For example, a transistor having a dissipation rating of 10 W will give an r.f. output of 40 W at 80% efficiency, but 90 W at 90% efficiency, which is an *increase* of 125% in power output for an efficiency improvement of only 10%.

While considering internal dissipation, transistors have a definite advantage over valves in that no heater supply is required. This will be even more advantageous for transistors having a higher power rating.

13.2 NOTES ON NON-LINEAR TRANSISTOR AMPLIFIERS

Conversion efficiency

Operation of the active device in class C is the most usual configuration for non-linear amplifiers, giving a reasonably high conversion efficiency. The basic circuit for a class C transistor amplifier is shown in Fig. 13.1, in which Π circuits are used for input and output because of the low impedances involved. With silicon transistors, an external bias is not normally required because bias is provided automatically by the internal voltage V_{be}.

With this simple circuit, a transistor amplifier is not conducive to high conversion efficiency. In order to use the transistor type efficiently, the cut-off

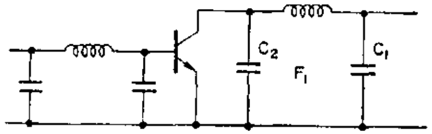


FIG. 13.1 Basic circuit for transistor amplifier, class C.

frequency F_T will not usually be much higher than the operating frequency F_1, and the effect cannot be neglected. The effect is a delay between the drive waveform V_b and the output waveform $V_c(1)$, as shown in Fig. 13.2.

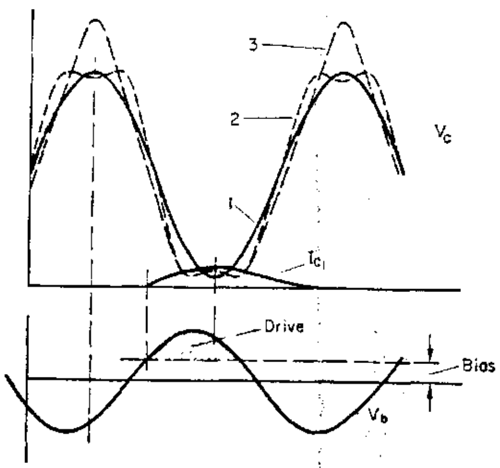


FIG. 13.2 Input and output waveforms of transistor amplifiers.

When transistors are driven to saturation, another difficulty arises due to the storage charge causing a spread in the length of the current pulses, as shown by I_{c1}. It is obvious that current flowing on either side of the voltage trough will cause poor efficiency.

In order to improve the conversion efficiency, consider the application of the principle used by Tyler [1] for high-efficiency valve amplifiers (class D), as shown in Fig. 13.3. It is seen that the output contains an additional circuit which is resonant at the third harmonic of the fundamental frequency. When tuned correctly the collector waveform will be shown by $V_c(2)$, Fig. 13.2, giving an obvious improvement in efficiency compared with the class C waveform $V_c(1)$.

This arrangement differs from the high-efficiency valve amplifier, in that pulse broadening of the input waveform is not necessary (for an explanation of the function of the third harmonic circuit as a means of improving efficiency, refer to Chapter 3, Section 3.5).

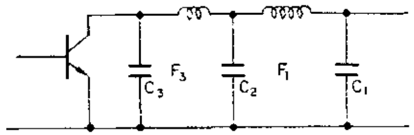


FIG. 13.3 Circuit for high-efficiency operation.

With transistor amplifiers, as the frequency increases the need for the third harmonic resonator diminishes. As a result of the low working impedance, stray series inductance (lead length) and relatively small collector capacitance (Fig. 13.4), the collector waveform produced is as shown by $V_c(3)$, Fig. 13.2. This waveform has the flat trough necessary for high conversion efficiency. The peak overswing does not appear to have a detrimental effect on the transistor, even when the nominal breakdown voltage is exceeded, provided that the operating

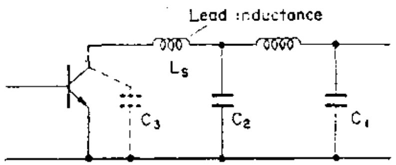


FIG. 13.4 Effective output circuit at higher frequencies.

frequency is high. The result is that many transistors designed for high-frequency operation can produce more output power with safety when operating at higher frequencies. This appears contrary to straightforward reasoning, and accounts for conversion efficiencies higher than expected from class C operation.

Other circuits have been devised for obtaining conversion efficiencies which approach 100%, based on rectangular waveforms produced by switching transistors on and off. Practical difficulties do arise with such arrangements, not the least being the correct timing for switching on and off, but it is considered that this line of attack will be used as a means of obtaining higher r.f. power from non-linear transistor amplifiers

Parasitic oscillations

Transistor amplifiers are subject to a type of parasitic oscillation not normally present in valve amplifiers. These are parametric oscillations due to the non-linear characteristic of the collector-base capacitance. Consider the frequency spectrum shown in Fig. 13.5, in which the fundamental operating frequency is F_1. If the collector-base capacitance is sufficiently non-linear, with high-amplitude signals, an output will be produced parametrically at two other frequencies

$(F_1 - F_x)$ and $(F_1 + F_x)$, in the presence of power at the frequency F_x. The ratio of powers will be proportional to the frequency ratio. If the output circuit is tuned to accept the difference frequency $(F_1 - F_x)$, then the input impedance is

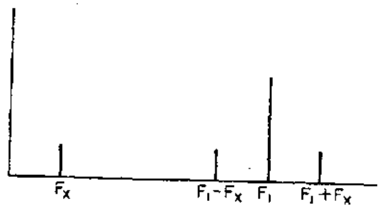


FIG. 13.5 Spectrum showing low-frequency parametric oscillation due to non-linear collector-base capacitance.

negative at frequency F_x and parasitic oscillation will be present, without the application of power at frequency F_x, if the circuit is not damped sufficiently at this low frequency, F_x.

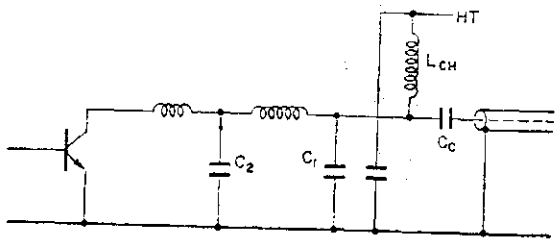


FIG. 13.6 Practical circuit for producing spurious parametric oscillation.

Consider the practical circuit shown in Fig. 13.6, where the h.t. supply is fed via the choke L and the output is coupled to the feeder via capacitor C_c. If the choke L_{CH} and capacitor C_c are not sufficiently large, and the resultant circuit

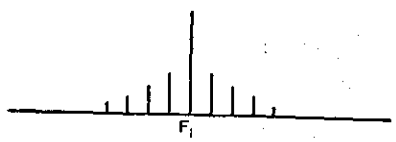


FIG. 13.7 Spectrum of frequencies produced by the circuit shown in Fig. 13.6.

not sufficiently lossy at both low r.f. and audio frequencies, parametric oscillations will be produced on either side of the fundamental frequency F_1. The frequency spectrum produced by this type of oscillation is shown in Fig. 13.7.

13.3 LINEAR AMPLIFICATION WITH TRANSISTORS

Transistor linearity compared with valves

Transistors are often considered to be suspect for linear amplification and inferior to valves, mainly due to comparisons being made on a non-realistic basis, particularly regarding power level.

For the low-power stages of linear transmitters, linearity is the overriding consideration and conversion efficiency is unimportant. Consequently, it is quite normal to use valves in class A with a rated dissipation of about 2 W for a few milliwatts output. Yet when transistors are used for a similar application, the same order of linearity is expected from models having a dissipation similar to the output. A more practical comparison can be made by using a non-linearity factor (n.l.f.), which is defined as the ratio of amplifier dissipation to p.e.p. output, for a given distortion level in terms of intermodulation products (i.p.'s). On this basis the following typical example shows that transistors can be superior to valves as linear amplifiers.

Consider a requirement for 2 W p.e.p. output from 2 MHz to 30 MHz with i.p.'s of -50 dB, from valves and transistors of suitable rating.

Valve n.l.f. = 30, excluding heater power.
Valve n.l.f. = 40, including heater power.
Transistor n.l.f. = 25.

Another point is that no useful purpose is achieved by comparing the theoretical r.f. linearity of valves having a 3/2 power law with that of transistors having an exponential law, because valves are voltage driven, whereas transistors are current driven.

Non-linearity in transistors

One of the most important factors contributing to non-linearity in transistors is the variation of cut-off frequency F_T in relation to the V_c/I_c characteristics. Before discussing this further, it might be advisable to explain the term cut-off frequency. The gain of a transistor is given by the ratio of the collector current

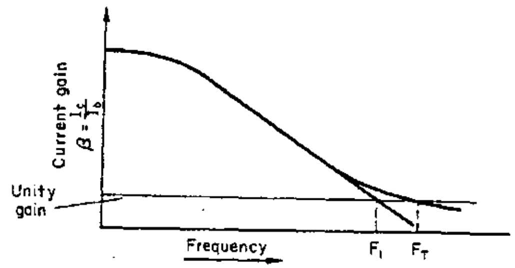


FIG. 13.8 Gain/frequency response for constant V_c.

I_c to the base current I_b, and is called β ($\beta = I_c/I_b$). With transistors the current gain decreases with rising frequency in the form of the curve shown in Fig. 13.8. The frequency at which the current gain falls to unity ($I_c = I_b$) is known as the cut-off frequency, F_T.

It should be noted that over an appreciable frequency range, the gain β drops linearly with frequency increase. If this straight portion is continued, the unity gain condition is reached at a frequency F_1, which is somewhat lower in frequency than F_T.

Frequency F_1 is often used by design engineers instead of F_T, because it enables the gain at frequencies along the straight portion to be estimated more readily.

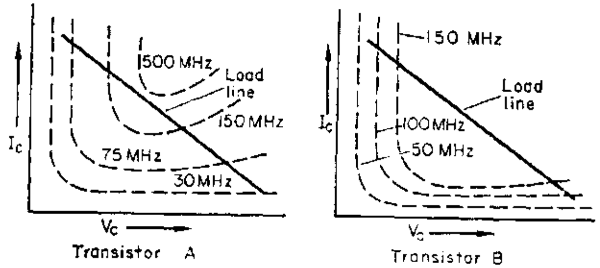


FIG. 13.9 Typical contours of cut-off frequency F_T relative to voltage V_c, for two transistors.

The variation of cut-off frequency in relation to the V_c/I_c characteristics for two different transistors is shown in Fig. 13.9, with typical load lines added. It is clear that the gain of an amplifier varies with current and voltage swing, thus giving rise to distortion. The effect is more pronounced at higher operating frequencies, so for linear amplification it is necessary to use transistors having cut-off frequencies very much higher than the required operating frequency.

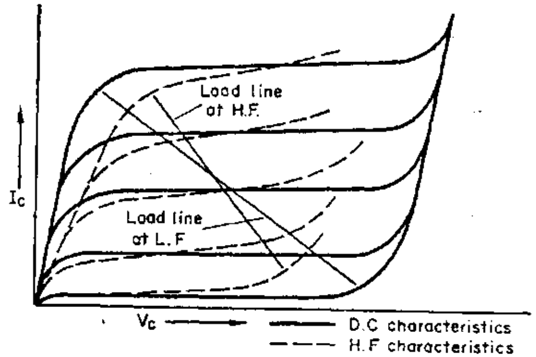


FIG. 13.10 Effect of increasing frequency on transistor I_c/V_c characteristics.

For this reason transistors having the characteristics similar to B in Fig. 13.9 are preferred to those with A characteristics.

Another way of showing this effect is given in Fig. 13.10, in which the solid line characteristics are at d.c. as given in the published data sheets. However, as the frequency increases, the effect of the cut-off frequency, i.e., gain reduction, causes the characteristics to tilt, as shown in broken lines. The result is a reduction

in voltage and current excursions at higher frequencies, with a consequent increase in distortion. The change in optimum load line is also shown.

Class B transistor amplifiers

It is quite practical to operate linear r.f. amplifiers with a single valve in class B. This is possible because of the high resistance of the load line, particularly with tetrodes, enabling circuits of high Q factor to be used to supply energy during the un-driven half-cycles.

On the other hand, transistors have very low-resistance load lines, typically 3 Ω for a 75 W transistor, so circuits of high Q factor are completely impractical. For linear r.f. amplifiers, transistors in class B must be operated in push–pull pairs in a similar manner to class B valves in a.f. amplifiers.

In order to obtain good linearity with class B amplifiers in push–pull and low Q factor circuits, it is important that the bias is set accurately. If the bias is too high, cross-over distortion will result; whereas if it is too low, the cross-over distortion will occur to a lesser extent, but excessive dissipation is highly probable.

Secondary breakdown

Maintaining the bias correctly under operating conditions presents quite a problem, because the required bias point changes with temperature, which in turn changes with operating power level. In fact, the static feed increases with rising temperature for a fixed-bias voltage, so the dissipation increases and the transistor can be destroyed by secondary breakdown. This is the term applied to an effect which causes silicon transistors to break down under d.c. conditions, at a dissipation level much lower than that which can be safely maintained under

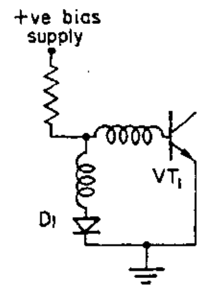


FIG. 13.11 Forward-biased diode for bias compensation.

r.f. conditions. A practical example will show the serious nature of this effect. Consider that a transistor, having a rated secondary breakdown at 8 W, is biased to dissipate 5 W with no r.f. signal. On application of r.f. the dissipation rises to 20 W and the silicon will get hot, but the transistor does not break down. On removal of the r.f. signal, the bias point will have changed and a d.c. dissipation of 10 W is likely. This is above the rated d.c. dissipation, so the transistor will be destroyed by secondary breakdown.

The simplest method of bias compensation is to use a forward-biased diode which is held at the same temperature as the transistor, and connected as shown in Fig. 13.11. The diode D_1 should be mounted adjacent to the transistor VT_1 on the same heat sink, so that the two silicon chips are at approximately the same temperature and have the same temperature characteristic. This method is more satisfactory if the diode and transistor are both on the same silicon chip.

An alternative method of compensation is to use a constant-current bias source together with a diode-connected transistor, by some arrangement such as shown in Fig. 13.12. This is more satisfactory for higher power. Some new power transistors have a diode incorporated on the same chip, but of a very small rating so that a d.c. amplifier must be incorporated.

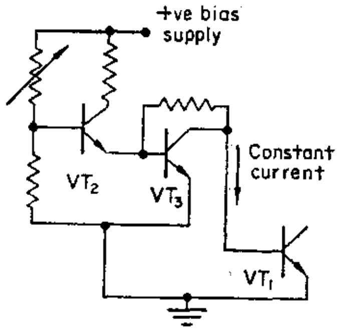


FIG. 13.12 Constant-current method of bias compensation.

Linear wideband r.f. amplifier, 2–30 MHz

Because transistors are low-impedance devices, they are more appropriate to wideband amplification than they are to narrow band tuned arrangements. However, in considering a bandwidth of several octaves, such as 2–30 MHz, some gain/frequency compensation must be applied to give a substantially flat frequency response, because of the change in gain (β) of transistors over the

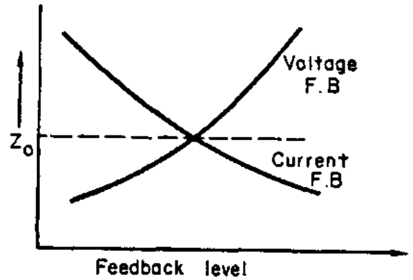


FIG. 13.13 Change in input impedance with increasing feedback level.

operating frequency band. Compensation can best be obtained with feedback, because feedback also improves linearity. Feedback also changes the impedance of transistors, thereby providing mismatch, with consequent degradation of linearity. It so happens that with increasing feedback level, voltage feedback increases the output impedance, but current feedback reduces it. By applying

voltage and current feedback in parallel, it is possible to obtain a substantially constant output impedance over a wide range of frequencies, even though the feedback level changes (Fig. 13.13).

Wideband circuit arrangement

A basic circuit arrangement for a class B wideband transistor amplifier is shown in Fig. 13.14. It will be seen that this incorporates a bias supply arranged to compensate for changes in bias point with rising temperature, together with both voltage and current feedback. This arrangement is quite satisfactory as a linear amplifier from 2 MHz to 30 MHz, with a suitable choice of component values. If more gain is required by two such stages in cascade, it is preferable to

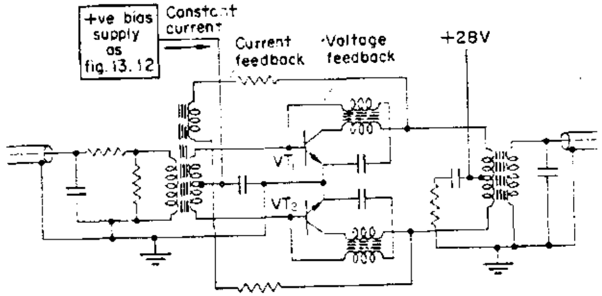


FIG. 13.14 Single-stage wideband class B r.f. amplifier.

avoid the use of a feeder by using transformer matching directly into the second-stage input.

With the very low impedance of transistors, the impedance of quite short leads can be troublesome and transformers must be designed with the lowest possible leakage inductance. Even so, as the power level is raised it becomes increasingly difficult to couple the two sides of the class B stage sufficiently tightly together. For higher powers it may be necessary to revert to partial tuning to overcome the side-to-side coupling problem, and effectively reduce the bandwidth. Experimentally it has been found that an improvement in linearity in terms of i.p.'s can be obtained with amplifiers slightly mistuned. This is due to cancellation, and is not consistent at all signal levels. It is not a suitable method to use for a production equipment.

REFERENCE

[1] TYLER, V. J. 'A new high-efficiency high-power amplifier'. *Marconi Rev.*, 21 3rd quarter (1958).

Appendix I

A Graphical Method of Harmonic Analysis

If often happens that a quick harmonic analysis is required in circumstances where speed is more important than a high order of accuracy. The following method is a ready means of determining the amplitude and phase relationships in a complex waveform, provided that the levels of the fifth and higher orders are relatively low. The accuracy obtainable depends upon the plotting accuracy and the relative levels of harmonics and fundamental, with higher harmonic levels giving a greater accuracy.

Before proceeding with the description however, it should be mentioned that the method is not new [1]. It was devised by J. Harrison many years ago and has since appeared in several textbooks, one of which is Castle's *Manual of Practical Mathematics*, published as long ago as 1920. However, a recent sampling (admittedly a small one) taken among young engineers and students seems to indicate that it has fallen into disuse—indeed, it was quite unknown to them.

DESCRIPTION OF THE METHOD

Perhaps the simplest way of describing the process of waveform analysis by this method is by practical illustration. Figure AI.1 is a plot of a complex waveform, the main constituents being fundamental, second harmonic and third

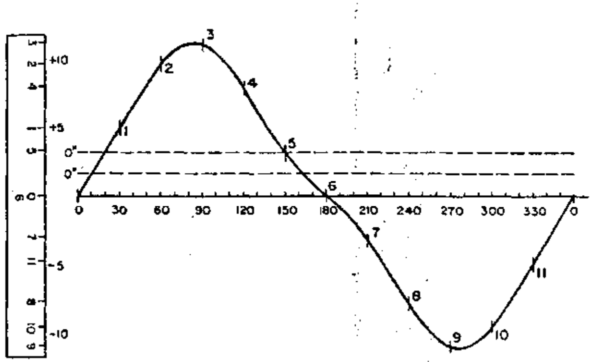


Fig. AI.1 A complex waveform with second and third harmonic content. The points identified numerically every 30° on the waveform are projected to determine the graduations on the paper strip.

harmonic. Points 0–11 are marked on the waveform at 30° intervals: 0 at 0°, 1 at 30°, 2 at 60°, etc. On a separate strip of paper, also shown in Fig. AI.1, the amplitude of points 0–11 are marked with respect to zero level. In practice, the paper strip is moved along in 30° steps, to obtain the best accuracy.

This marked strip is used to plot curves A, B and C in shown Fig. AI.2 and described later, of the constituent components of the waveform, by algebraic subtraction and addition of the amplitudes of the various points in accordance with Table AI.1.

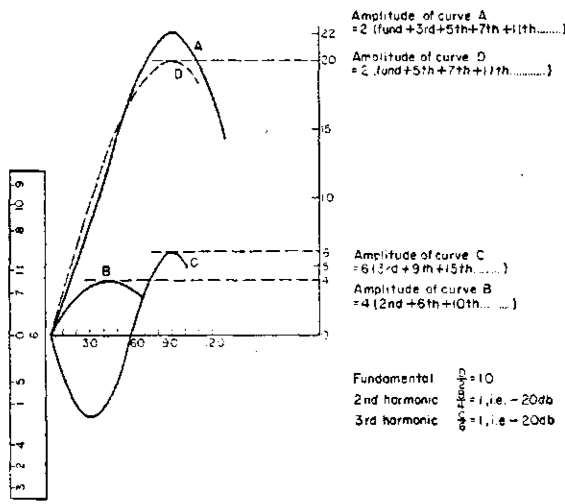


Fig. AI.2 The constituent components of Fig. AI.1 waveform plotted from the information given in Table AI.1. *Note:* For clarity subdivisions are not shown, but 10 on the linear amplitude scale represents twenty small divisions.

In Table AI.1 mathematical expressions representing component parts of a Fourier expansion are given for the three curves. As can be seen from these expressions, half the amplitude of curve A gives the peak amplitude of the fundamental, third harmonic, ignoring the fifth, seventh, etc. One-quarter of the amplitude of curve B gives the peak amplitude of the second harmonic, ignoring

Table AI.1

θ	A 2 (fund. + 3rd + 5th)	B 4 (2nd + 6th + 10th)	C 6 (3rd + 9th + 15th)
0°	0–6	(0–3) + (6–9)	(0–2) + (4–6) + (8–10)
30°	1–7	(1–4) + (7–10)	(1–3) + (5–7) + (9–11)
60°	2–8	(2–5) + (8–11)	(2–4) + (6–8) + (10–0)
90°	3–9	(3–6) + (9–0)	(3–5) + (7–9) + (11–1)
120°	4–10	(4–7) + (10–1)	(4–6) + (8–10) + (0–2)

the sixth, tenth, etc. One-sixth of the amplitude of curve C gives the peak amplitude of the third harmonic, ignoring the ninth, fifteenth, etc.

The peak amplitude of the fundamental is obtained by subtracting one-third of the amplitude of curve C from curve A, at every 10° point, giving curve D, which is twice the fundamental amplitude.

The relative phase of the three components of the original waveform is obvious from Fig. AI.2.

Because the peak amplitudes of curves A and C are coincident at 90° in the example, both relative phase and amplitude could be obtained without drawing curve D, but in the general case it is necessary to plot curve D for an accurate assessment.

PLOTTING THE CURVES

To plot the curves, the strip is reversed and points marked at 30° intervals, as shown in Fig. AI.3. The method of plotting the points for curves A, B and C is shown in Fig. AI.3 (*A*), (*B*) and (*C*), respectively. Points S.T.U. and V are points for curve A, with points L, M and N for curve B, and points G and H for curve C, all at 30° intervals.

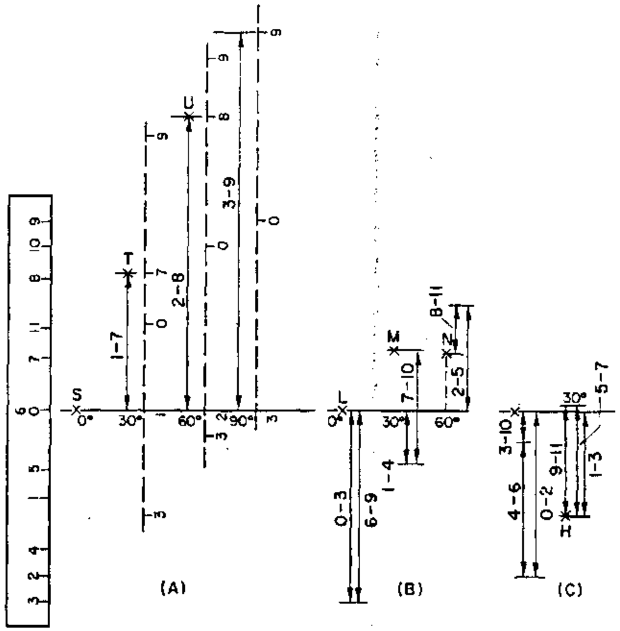


Fig. AI.3 The method of marking points at 30° intervals for curves A, B and C, in accordance with Table AI.1 and using the paper strip inverted.

Figure AI.3 is used to give an explanation of the method of using the inverted strip. The actual curves should be plotted as shown in Fig. AI.2, in order that the relative phase of harmonics and fundamentals can be determined.

From these points a rough approximation of curve A could be drawn, but they are quite inadequate to draw B and C.

Additional points can be obtained at 30° intervals, starting at 10° and 20°, by using two more paper strips. On these strips, points 0–11 are marked, as before, but with reference to new zero lines, 0′ and 0″ (see Fig. AI.1). Points on the 0′ strip are marked 0 at 10°, 1 at 40°, 2 at 70°, etc., giving points for curves A, B and C when inverted and transferred to Fig. AI.2, at 10°, 40°, etc.

Similarly, the other strip 0″, will give points at 20°, 50°, 80°, etc., when marked with reference to level 0″. It is seen that 0′ is the baseline drawn at the level of the complex waveform at 10°, and 0″ is the level at 20°.

ACCURACY

For a second harmonic level of 20 dB and the scale used (see Fig. AI.2, note), the limits of measurement are about half of one small division in eight small divisions (curve B), giving an accuracy of approximately 0·5 dB. Also, the limit of measurement of about one small division corresponds to a harmonic level of −40 dB (one-tenth of curve B or C). At this harmonic level half of one small division represents a tolerance of ±3 dB.

Table AI.2 gives the amplitude of curve D at every 10°, showing that in the example the plot of the fundamental does not depart from a sine wave by more than 2%.

Table AI.2

θ	Level from Fig. AI.2	Rationalized level	$\sin\theta$ level
0°	0	0	0
10°	3·5	0·175	0·1736
20°	6·7	0·335	0·342
30°	10·0	0·5	0·5
40°	12·6	0·63	0·6428
50°	15·0	0·75	0·766
60°	17·0	0·85	0·866
70°	18·5	0·925	0·9397
80°	19·5	0·975	0·9848
90°	20·0	1·0	1·0

It should be pointed out that these limits of accuracy assume that the plot of the original waveform is in itself correct within fine limits. In practical cases where the method of obtaining the original waveform does not enable the plot to be very accurate, the result will not be within such fine limits. For example, when the waveform is obtained from an oscilloscope, photographically or by tracing, the initial error is of the order of 5%. Thus the results obtained will be less accurate by this amount, giving total tolerances of approximately ±1 dB for harmonics 20 dB down and ±5 dB for harmonics 40 dB down.

APPLICATIONS

It is obvious that there are many applications for this method of harmonic analysis, particularly in making early assessments, where quick answers are of more immediate importance than extreme accuracy.

One example is the determination of the harmonic content to be expected from valves and transistors under various operating conditions, based on the published characteristics.

Another example is the measurement of the harmonic content in a coaxial feeder at v.h.f., obtained from a plot of the feeder voltage along a slotted line. In an application of this nature several precautions are necessary, but with a relatively high harmonic content and low v.s.w.r., the results obtained can be substantially correct.

REFERENCE

[1] STOKES, V. O. 'A graphical method of harmonic analysis'. *Wireless World*, 121–123 (March 1966).

Appendix II

The Self-Inductance of Single Straight Conductors of Circular Cross-Section

Although the size of conductors used for interconnections in high-power r.f. circuits is mainly dependent upon current-carrying capacity, in some instances it may be advantageous to select a size of conductor to provide some inductance in a connection which is part of an r.f. circuit. Such a connection may permit reduction in the size of an inductor or other variable element. In some cases the

size of conductor is chosen to reduce the lead inductance in a connection to a component. The extent of reduction possible with typical conductor sizes can be seen from Fig. AII.1. These curves are also useful in estimating the likely total inductance of connections, in order to determine the range of variable element necessary to cover a specified frequency range.

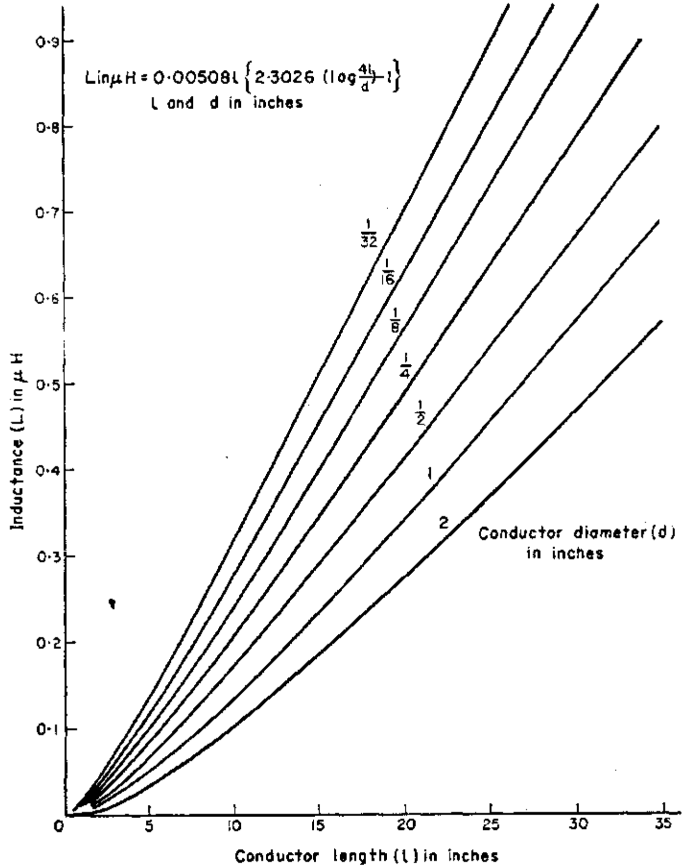


FIG. AII.1 Self-inductance of single straight conductors of circular cross-section.

Appendix III

Self- and Mutual Inductance of Turns of Large Diameter

For the large-diameter conductors necessary to carry the r.f. current and the low values of inductance required in the high-frequency ranges, general formulae for the inductance of multi-turn solenoids are insufficiently accurate. At the same time, these inductors are quite expensive items to manufacture, so any design should be substantially 'right first time'. However, inductors can be designed

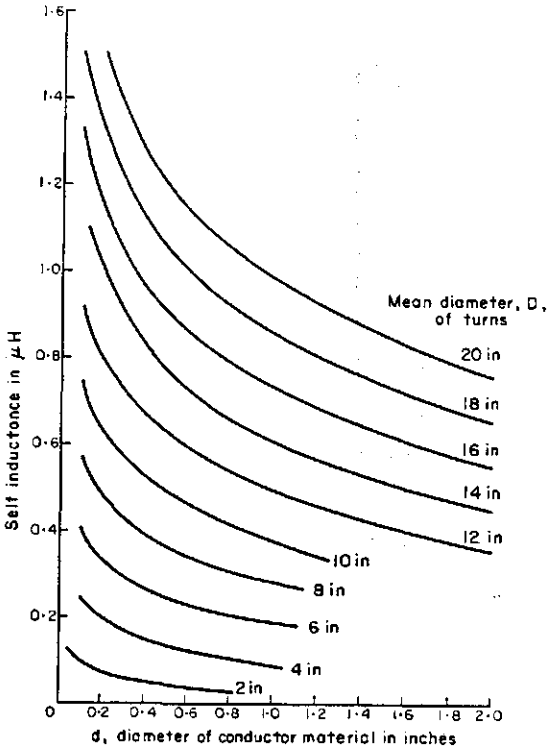


FIG. AIII.1 Self-inductance of single circular turns of circular cross-section material.

with sufficient accuracy based on the inductance value of each turn and the mutual inductances between the various turns. It is for this reason that Figs AIII.1 and AIII.2 have been reproduced, showing the self- and mutual inductance, respectively, of single turns with dimensions typical of those used for inductors in high-power applications.

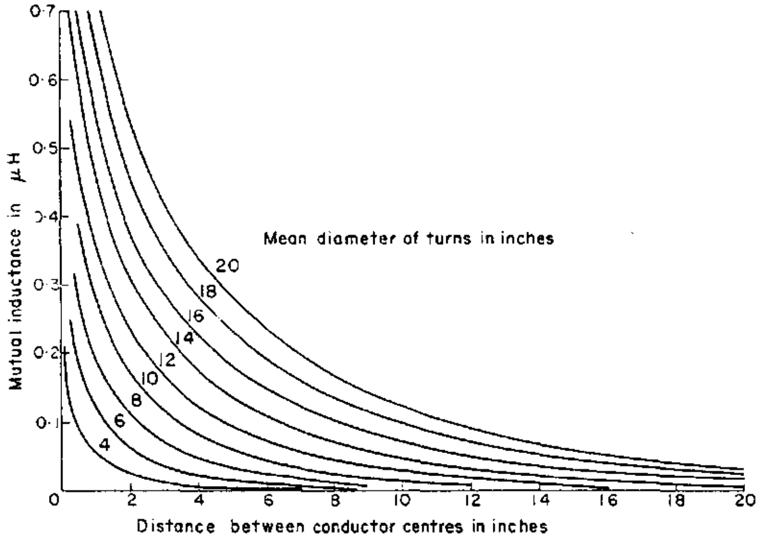


FIG. AIII.2 Mutual inductance of equal coaxial turns at various spacings.

The method of calculation based on Figs AIII.1 and AIII.2 is given below.

(1) With the material diameter, d, and the mean turn diameter, D, known, read off the self-inductance, L, of each turn from Fig. AIII.1.

(2) Assume a spacing between turns of some value, probably determined by voltage clearance, and read off Fig. AIII.2 the mutual inductance between adjacent turns M_1, alternate turns M_2, every third turn M_3, etc.

(3) If N is the total number of turns, then the total inductance is given by;

$$N \cdot L + (N-1)(2M_1) + (N-2)(2M_2) + (N-3)(2M_3) + \cdots +$$

As an example, consider a four-turn inductor, 10 in. mean diameter of 1·0 in. diameter material and 1·8 in. between the centres of adjacent turns. From Figs AIII.1 and AIII.2, $L = 0{\cdot}375\ \mu$H; M_1 (at 1·8 in.) $= 0{\cdot}19\ \mu$H; M_2 (at 3·6 in.) $= 0{\cdot}095\ \mu$H; M_3 (at 5·4 in.) $= 0{\cdot}055\ \mu$H; and total inductance $L = 4(0{\cdot}375) + 3(0{\cdot}38) + 2(0{\cdot}19) + 1(0{\cdot}11) = 3{\cdot}13\ \mu$H.

One of the advantages of the inductance being presented in this form is that it provides a ready means of determining the effect of either a change in turn diameter or a change in material diameter. An application of this type can be useful in controlling the angular position of inductor tapping points.

Appendix IV

Voltage Flashover

The voltage which can appear between two electrodes before flashover occurs is dependent on both the spacing between them and the diameter of the individual electrodes. In high-power applications the most usual concern is with parallel circular conductors. For this type of conductor Fig. AIV.1 shows the voltage at which flashover will occur between electrodes of various diameters and spacing at 0°C at sea-level, with correcting factors for higher temperatures and altitudes.

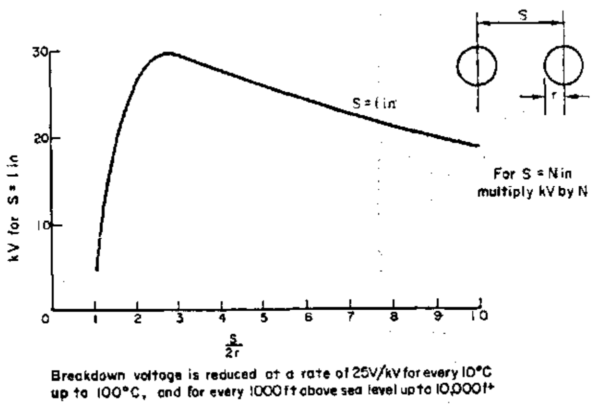


FIG. AIV.1 Flashover voltage between parallel circular conductors at 0°C at sea-level.

It is interesting to consider the effect of selecting ratios of $S/2r$ on either side of that giving the maximum voltage at 2·5. If a ratio of say, 2, is selected, as the voltage approaches breakdown surface ionization is likely to occur. This effectively reduces the ratio by a small amount, but the characteristic is very steep, so that any ionization is invariably followed by breakdown. On the other hand, if ionization occurs when $S/2r$ is greater than 2·5, the effective reduction of the ratio increases the breakdown voltage, and instead of breakdown it is more likely that the ionization will cease.

For the voltage breakdown between a circular conductor and a flat plate, reference should be made to Fig. AIV.2. First, consider an image of the circular conductor at a distance S equal to twice the distance between the centre of the conductor and the surface of the flat plate. Next, determine the breakdown

voltage which would occur between the conductor and its image from Fig. AIV.1. Then the effective breakdown voltage between the circular conductor and the plate is half the voltage which would produce flashover between the conductor and its image.

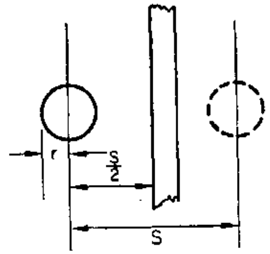


FIG. AIV.2 Flashover voltage between a circular conductor, or a flat plate with a radiused edge, and a flat plate.

Appendix V

Inductance of Single-Layer Solenoids

The curves given in Fig. AV.1 provide a ready means of calculating the inductance of single-layer inductors of constant diameter and pitch. The method is particularly applicable to solenoids of many turns of up to 2 in. mean diameter, so they will normally be wound with wire not exceeding 0·125 in. diameter. Consequently they are appropriate for low-power inductors and chokes.

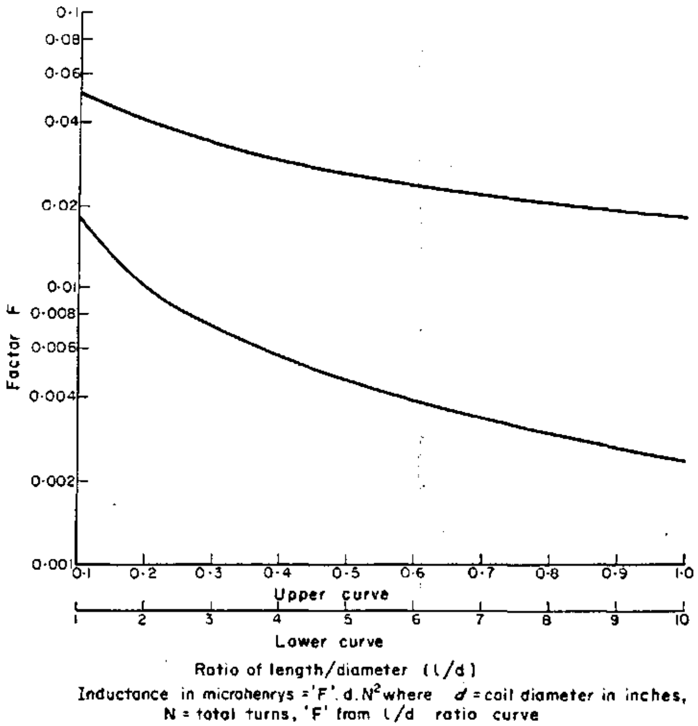


FIG. AV.1 Inductance of single-layer solenoids.

For inductors of a small number of turns, wound with gauges greater than 0·125 in. diameter on large mean diameters, reference should be made to Appendix III.

Appendix VI

S.I. Units

The Eleventh Conférence Générale des Poids et Mesures in 1960 adopted the rationalized and coherent system of units linked to the Giorgi units, and this has been accepted by the Electrotechnical Commission (I.E.C.) and the International Organization for Standardization (I.S.O.). British industry, through the Industry Standards Committee of the B.S.I., has opted to go straight to the use of S.I. units when changing to a metric system for the United Kingdom in the 1970s.

The important feature of the S.I. system is that it is a coherent system of units which covers the majority of quantities needed in all technologies and is based on six fundamental units. A coherent system is one where the product or quotient of any two unit quantities in the system is the unit of the resultant quantity; a simple example of this is that unit area is the result of unit length multiplied by unit width.

BASIC UNITS

Quantity	Name	Symbol
Length	metre	m
Mass	kilogramme	kg
Time	second	s
Current	ampere	A
Temperature	kelvin	K
Luminous intensity	candela	cd

Note: 1K = 1°C, and 0K = −273°C.

Most radio engineers are familiar with the basic units, but some of the derived units may be difficult to assimilate. One of these is the new unit of force, the newton (N) which is expressed as kilogramme metre/s² ($N = kg\ m/s^2$). It is a coherent unit, in that the force applied to unit mass produces unit acceleration and is independent of gravitational acceleration (g). Some rethinking will also be necessary regarding the joule (J), which becomes the sole unit for work, energy and quantity of heat. It is expressed in newton metres (N m).

DERIVED UNITS

Quantity	Name	Symbol
Force	newton	$N = kg\ m/s^2$
Work, energy, heat	joule	$J = N\ m$
Power	watt	$W = J/s$
Charge	coulomb	$C = A\ s$
Potential	volt	$V = W/A$
Capacitance	farad	$F = A\ s/V$
Resistance	ohm	$\Omega = V/A$
Frequency	hertz	$Hz = s^{-1}$
Magnetic flux	weber	$Wb = V\ s$
Flux density	tesla	$T = Wb/m^2$
Inductance	henry	$H = V\ s/A$
Luminous flux	lumen	$lm = cd\ sr$
Illumination	lux	$lx = lm/m^2$

Definitions

A *newton* is that force which gives a mass of one kilogramme an acceleration of one metre per second squared.

A *joule* is the work done when the point of application of a force of one newton is displaced through one metre in the direction of the force.

A *watt* is one joule per second.

A *coulomb* is the quantity of electricity transported in one second by a current of one ampere.

A *volt* is the p.d. between two points of a conducting wire carrying a constant current of one ampere when the power dissipated between these points is one watt.

A *farad* is the capacitance between the plates of a capacitor when a quantity of one coulomb of electricity produces a p.d. between the plates of one volt.

An *ohm* is the resistance between two points of a conductor when a constant p.d. of one volt produces a current of one ampere.

A *hertz* is the frequency of a periodic phenomenon of which the periodic time is one second.

A *weber* is the flux which, linking a circuit of one turn, produces in it an e.m.f. of one volt as it is uniformly reduced in one second (1 tesla $= 10^4$ gauss).

A *henry* is the inductance of a closed circuit which produces an e.m.f. of one volt when the current varies uniformly at one ampere per second.

Density is kilogramme per metre cubed (kg/m^3).

Pressure is newton per square metre (N/m^2).

Further information on S.I. units is given in the following publications.

The Use of S.I. Units, PD5868. British Standards Institution, London.

ANDERTON, P. and BIGG, P. H. *Changing to the Metric System*, 3rd edn H.M.S.O., London (1969).

Index