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Radio Transmitters

R.F. Power Amplification

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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 I

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.

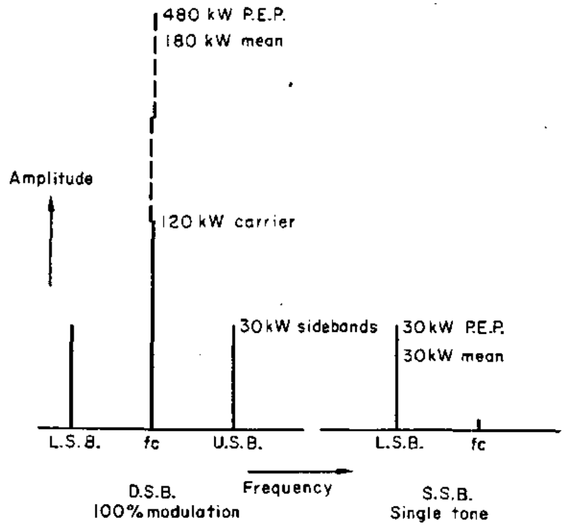

FIG. 1.1 Comparison between radio-frequency spectrum with d.s.b. and s.s.b.

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

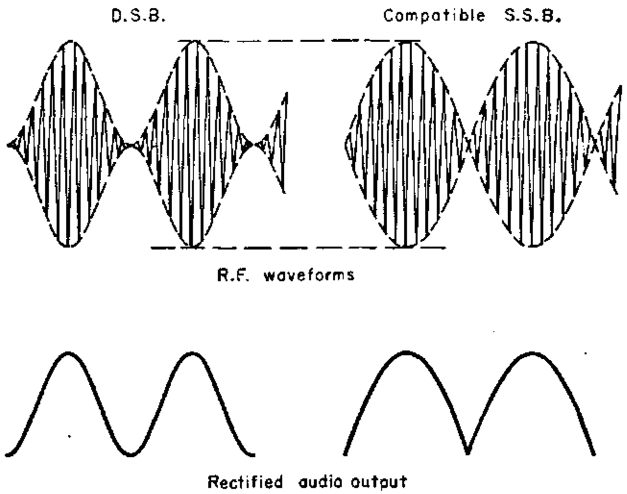

FIG. 1.2 Waveforms produced by d.s.b. and compatible s.s.b. systems.

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*

	Triodes				Tetrodes			
Class	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.

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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.

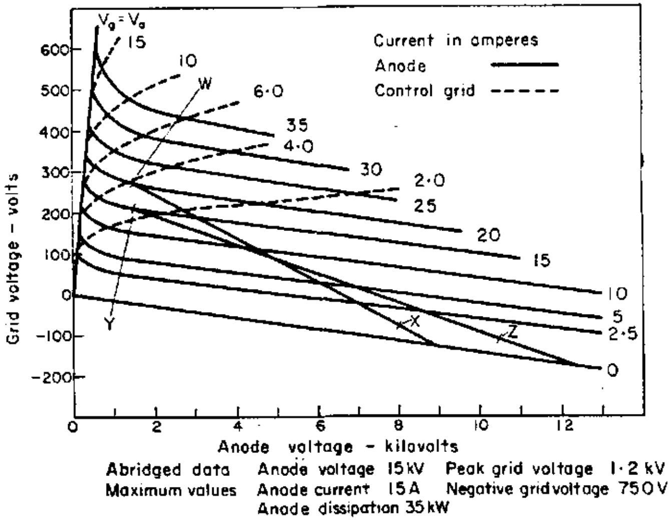

FIG. 2.1 Load lines on constant-current characteristics of triode valves type BR 1161 (English Electric Valve Co.).

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

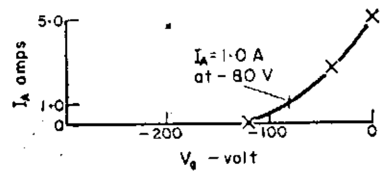

FIG. 2.2 Interpolation of I_A from V_g/I_A graph.

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

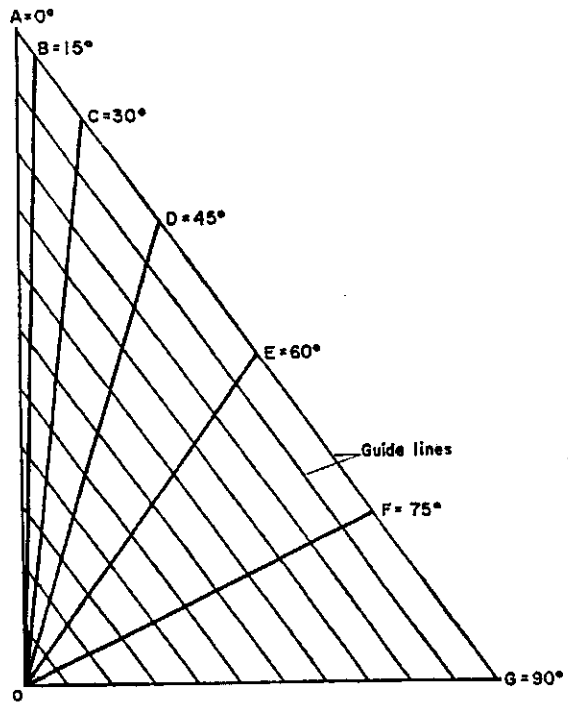

FIG. 2.3 Tube performance computor (Eimac).

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.

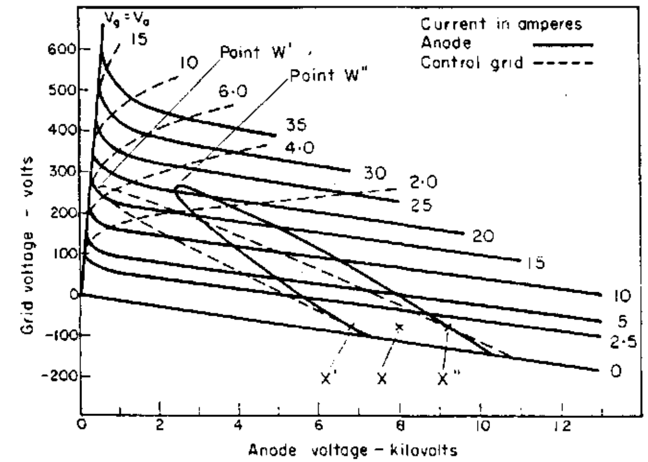

FIG. 2.4 Elliptical load lines on characteristics of triode type BR 1161.

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

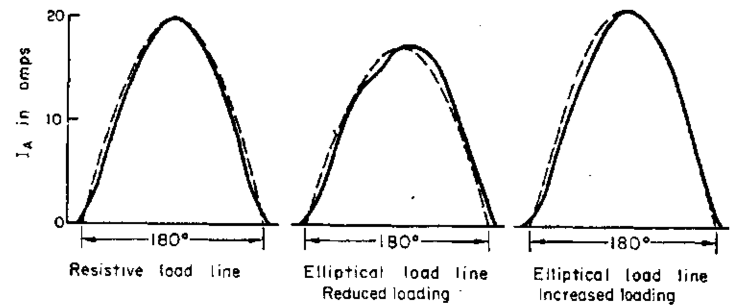

FIG. 2.5 Effect of feeder mismatch on anode current waveform of triode.

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.

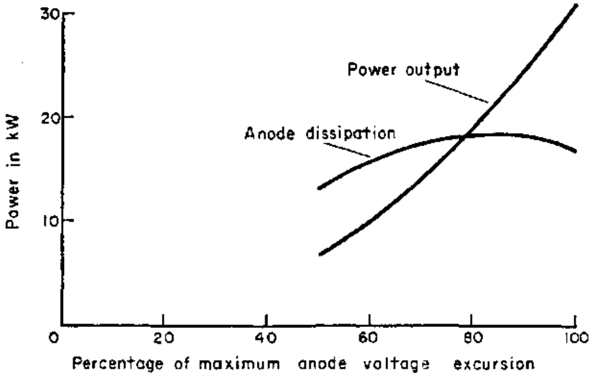

FIG. 2.6 Anode dissipation and power output in relation to anode voltage excursion.

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.

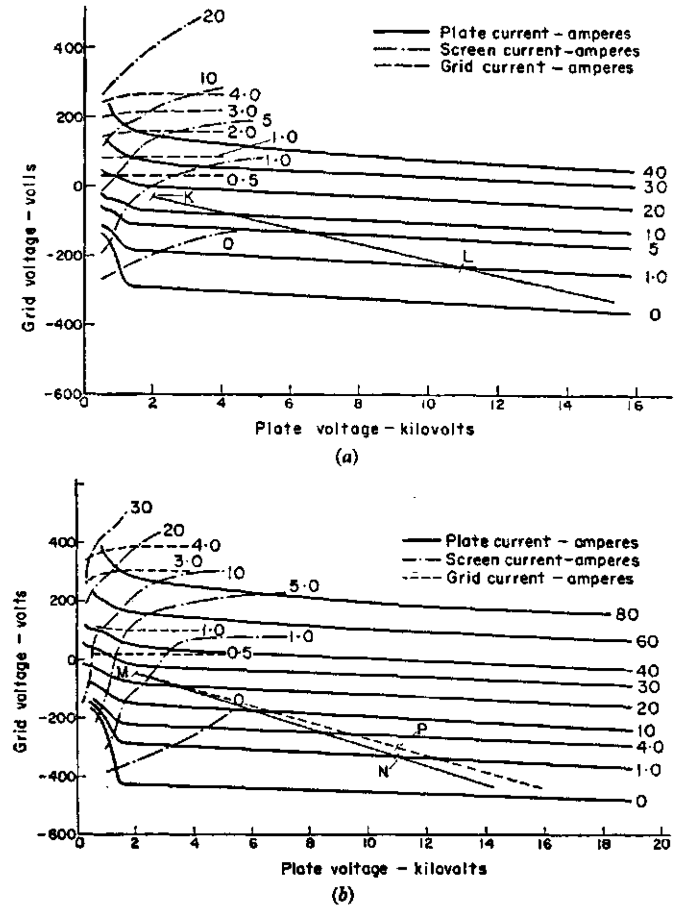

Fig. 2.7 Load lines on constant-current characteristics of tetrode valve type 4C X 35 000 C (Eimac). (a) Screen voltage = 1000 V. (b) Screen voltage = 1500 V.

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

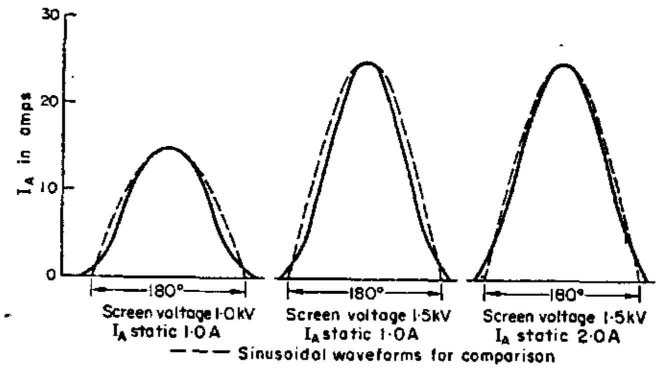

Fig. 2.8 Effect of changing screen voltage and static anode feed on anode current waveform of tetrode.

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.

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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

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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

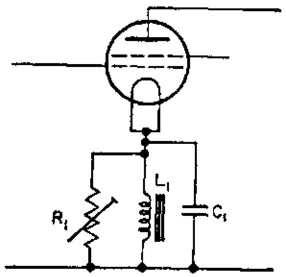


FIG. 3.1 Circuit for improving tetrode linearity.

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

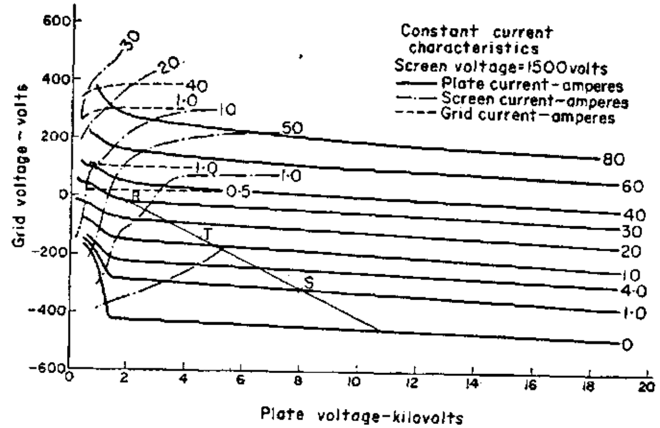


FIG. 3.2 Load line on constant-current characteristics of tetrode valve for low-level modulation.

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 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.

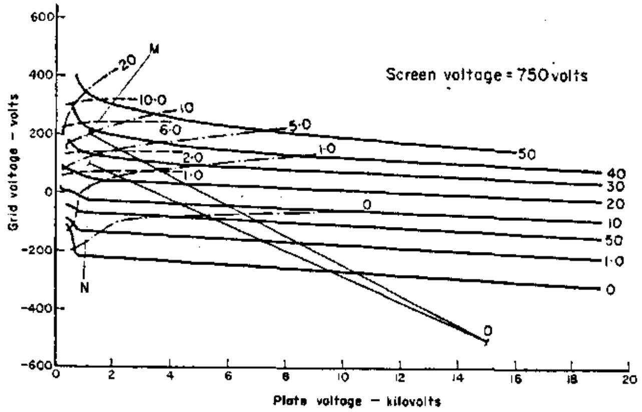


FIG. 3.3 Load lines on constant-current characteristics of tetrode valve for anode modulation.

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

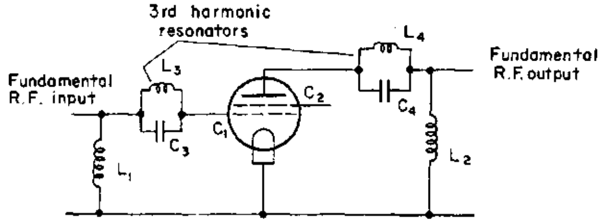


FIG. 3.4 Basic circuit arrangement for class D operation.

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

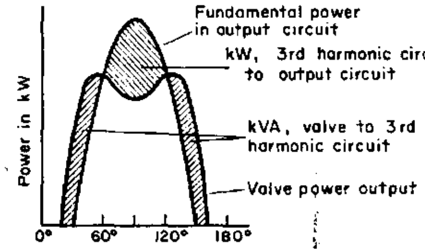


FIG. 3.5 Energy transfer in class D operation.

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:

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

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

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

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

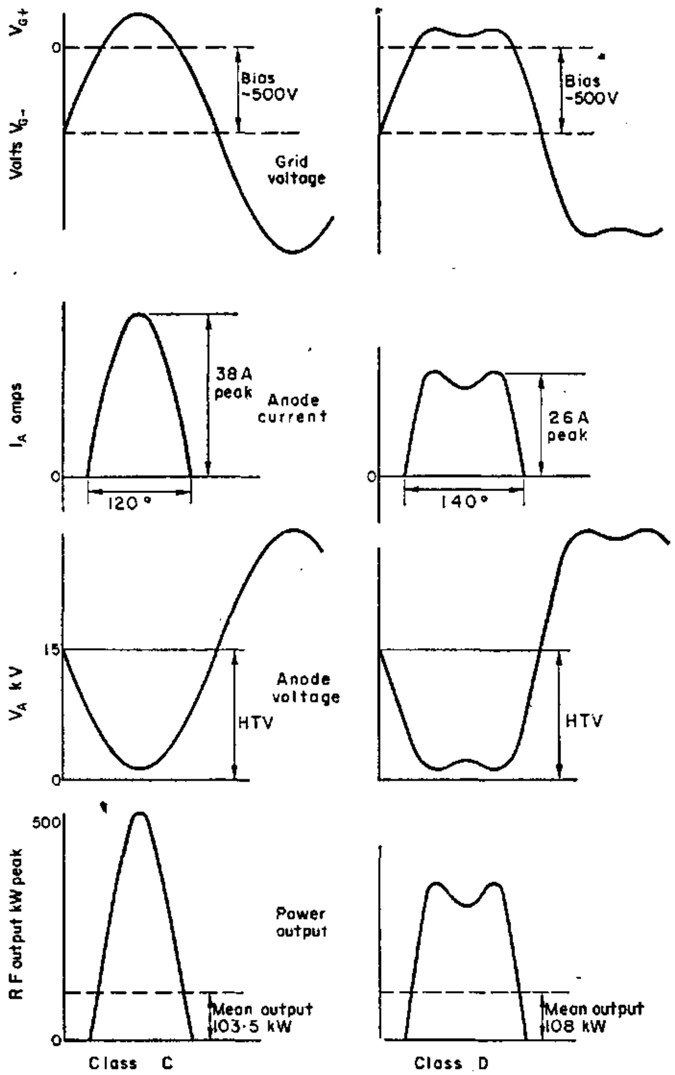


FIG. 3.6 Comparison between class C and class D waveforms.

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

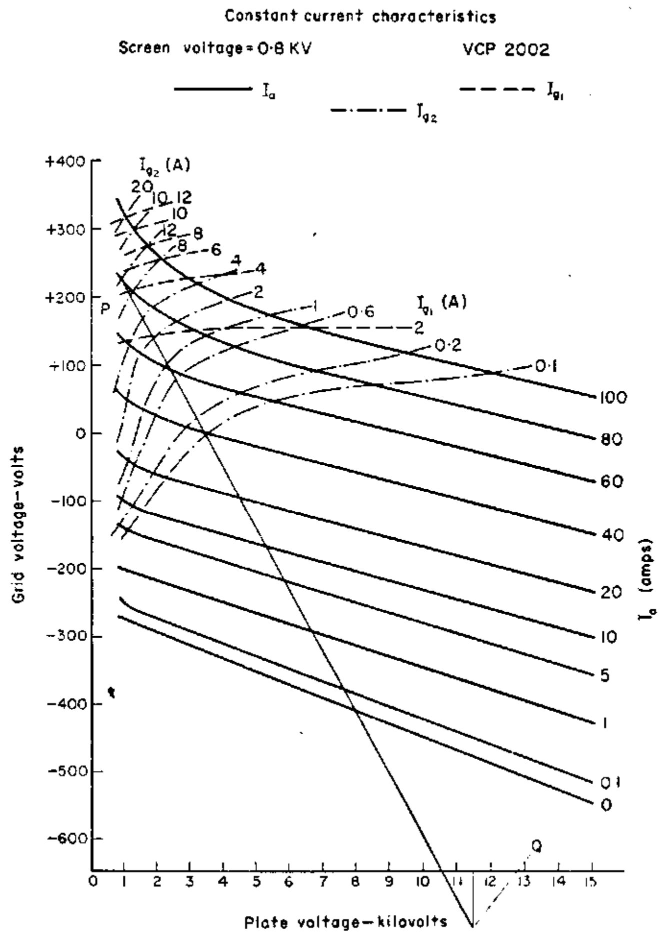


FIG. 3.7 Load line on constant-current characteristics of tetrode for very high power at high efficiency in class D. Valve type VCP 2002 (English Electric Valve Co.).

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_1C_1 and a series-tuned output circuit L_2C_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.

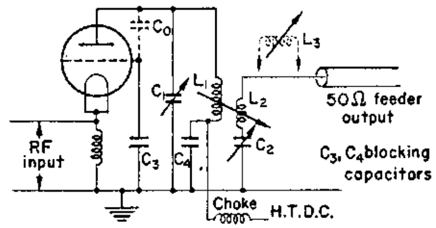


FIG. 4.1 Tuned-anode tuned-output circuits with magnetic coupling for grounded-grid triode.

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

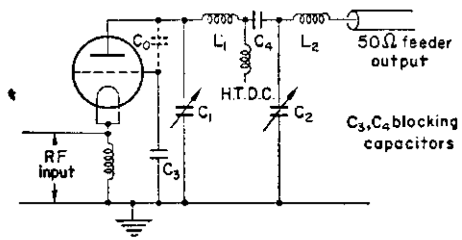


FIG. 4.2 Basic ΠL circuit for grounded-grid triode.

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

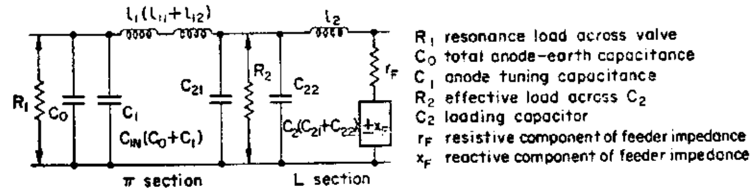
An alternative arrangement of a ΠL circuit is obtained by eliminating the variable capacitor C_1, and making inductor L_1 variable for anode tuning. While this has some technical advantages, variable inductors for high power are not only difficult to construct but are more liable to give trouble in service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

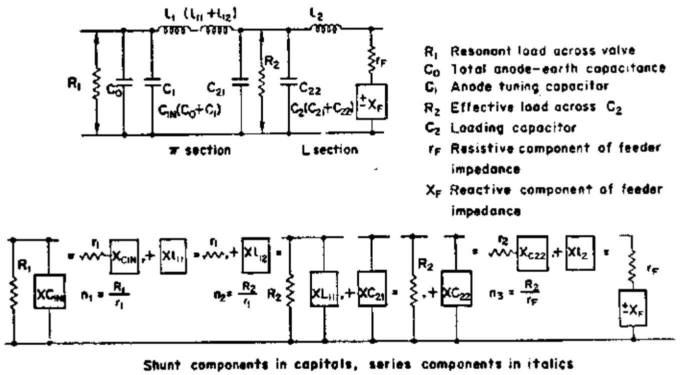
4.3 A METHOD OF CALCULATING COMPONENT VALUES FOR ΠL CIRCUITS

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:

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

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

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

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

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
$\sqrt{(n_1-1)} = \dfrac{R_1}{XC_{IN}}$	10·92	3·28
n_1	120	11·76
$r_1 = \dfrac{R_1}{n_1}$	4·83 Ω	49·32 Ω
$xC_{IN} = r_1\sqrt{(n_1-1)}$	−52·74 Ω	−161·77 Ω
$xl_{11} = -xC_{IN}$	52·74 Ω	161·77 Ω
R_2 (selected)	170 Ω	170 Ω
$n_2 = \dfrac{R_2}{r_1}$	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
$XC_{21} = \dfrac{R_2}{\sqrt{(n_2-1)}}$	−29·06 Ω	−108 Ω
r_F	50 Ω	50 Ω
$n_3 = \dfrac{R_2}{r_F}$	3·4	3·4
$\sqrt{(n_3-1)}$	1·53	1·53
$XC_{22} = \dfrac{R_2}{\sqrt{(n_3-1)}}$	−111·11 Ω	−111·11 Ω
XC_2 (total) $= \dfrac{XC_{21} - XC_{22}}{XC_{21} + XC_{22}}$	−23·06 Ω	−54·8 Ω
$xl_2 = r_F\sqrt{(n_3-1)}$	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 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.

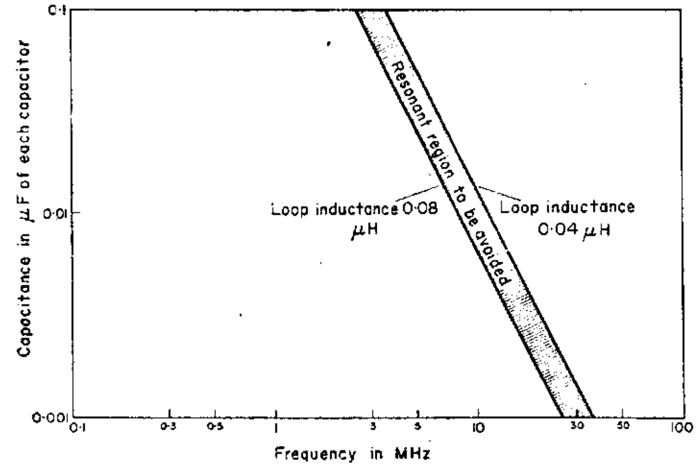


FIG. 4.4 Probable resonance of capacitors in parallel.

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

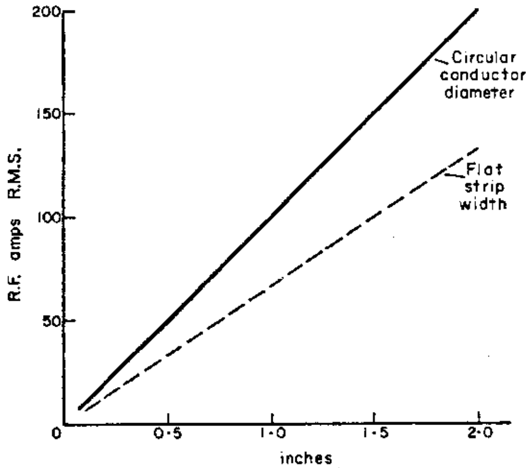


FIG. 4.5 R.F. current-carrying capacity of conductors for frequencies below 30 MHz.

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.

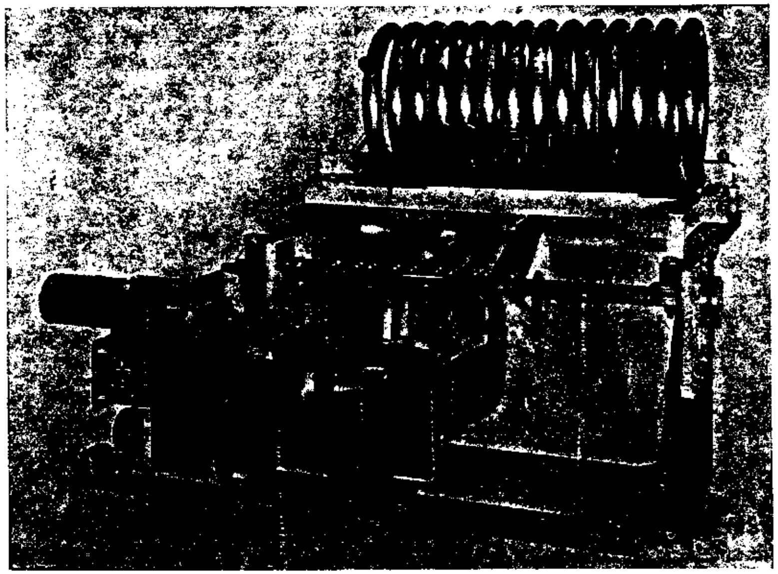


FIG. 4.6 Example of tapped-inductor variable-capacitor anode circuit.

(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.

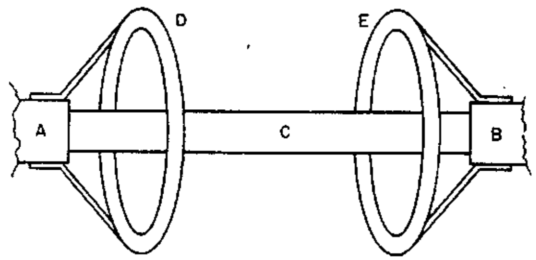


FIG. 4.7 The use of corona rings to increase breakdown voltage.

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

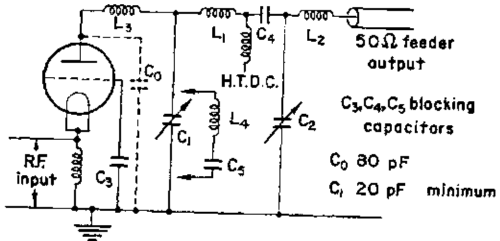


FIG. 4.8 Modified ΠL circuit arrangement for high power.

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.

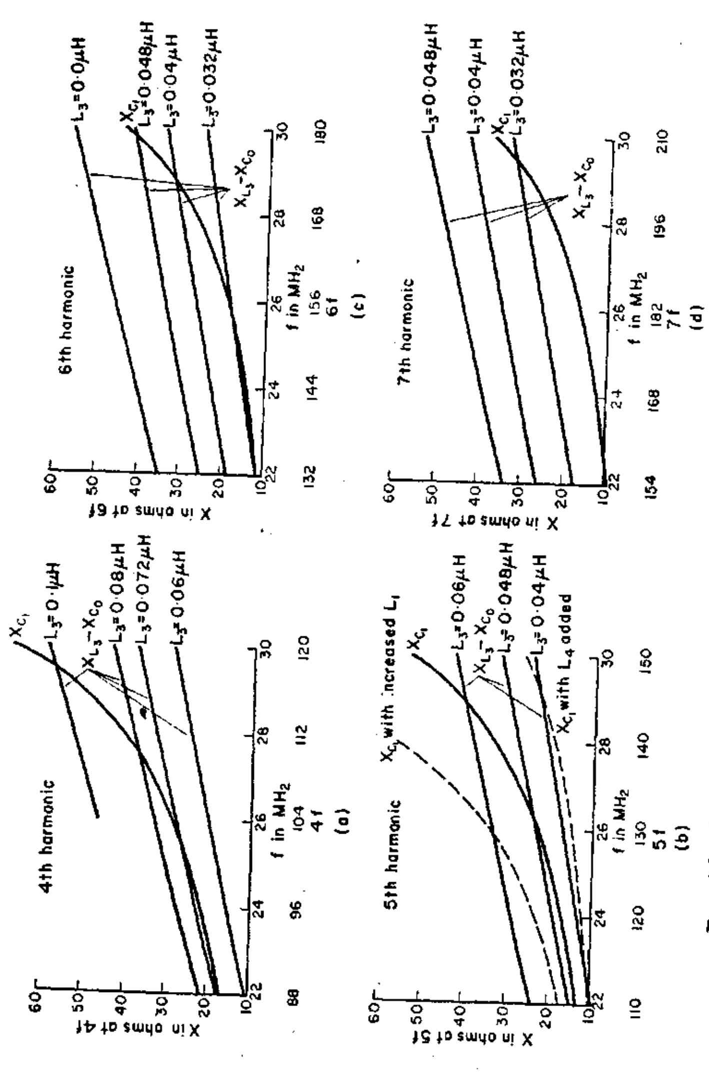


FIG. 4.9 Graphs showing liability for harmonic accentuation in anode circuits.

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.

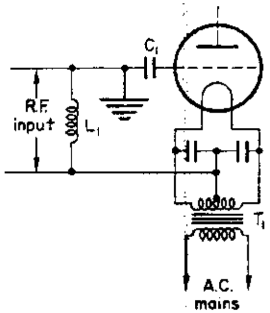


FIG. 4.10 Grounded-grid circuit with low-capacitance heater transformer secondary at r.f. potential.

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.

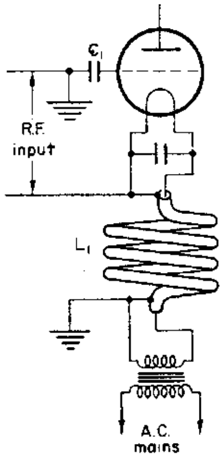


FIG. 4.11 Grounded-grid circuit with cathode choke of low-impedance heavy-current coaxial cable, such as Pyrotenax.

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.

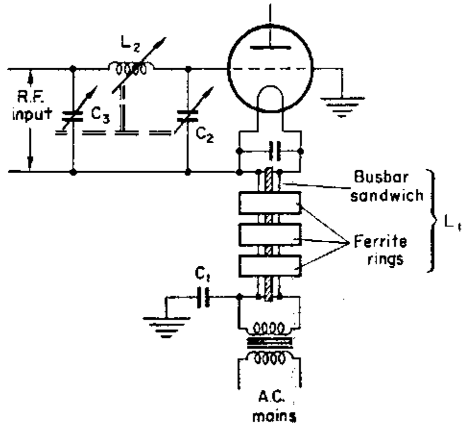


FIG. 4.12 Grounded-grid circuit with cathode choke consisting of a busbar sandwich through ferrite rings.

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 ft^3/min (20°C temperature rise)
Water: 16 gal/min (20°C temperature rise)
Vapour: 0·6 gal/min of water
(3000 ft^3/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.

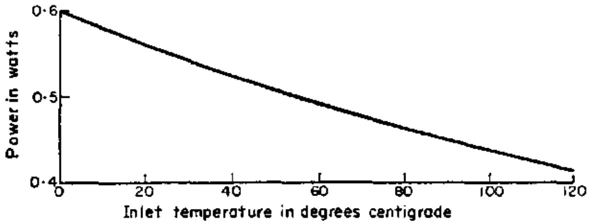


Fig. 5.1 Power required to raise 1 ft^3/min of air by 1°C.

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

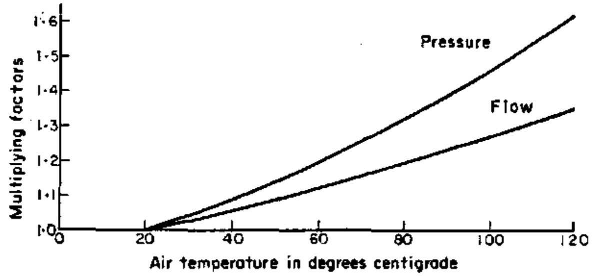


Fig. 5.2 Multiplying factors for pressure and air flow to allow for reduced air density above 20°C.

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

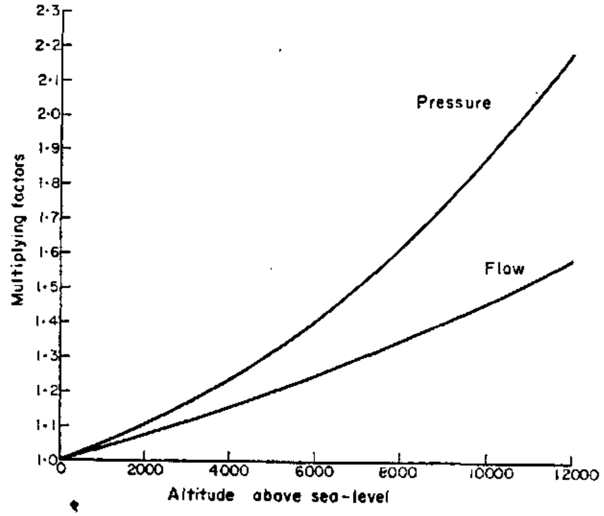


Fig. 5.3 Multiplying factors for pressure and air flow to allow for reduced air density above sea-level.

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.

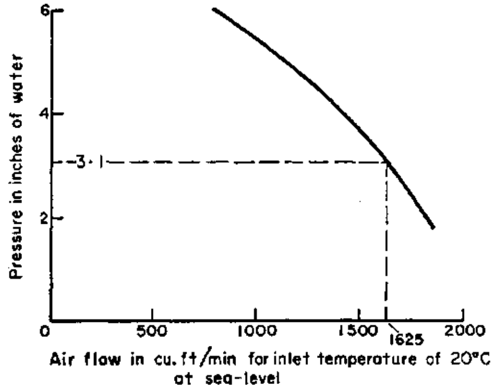


FIG. 5.4 Characteristics of cooling fan.

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.

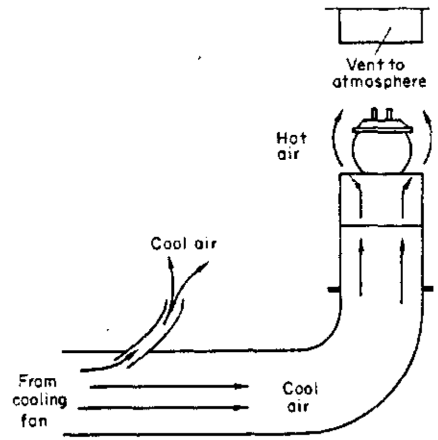


FIG. 5.5 Cooling by blowing.

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 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.

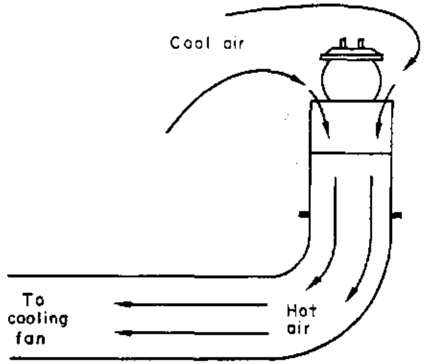


FIG. 5.6 Cooling by suction.

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 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.

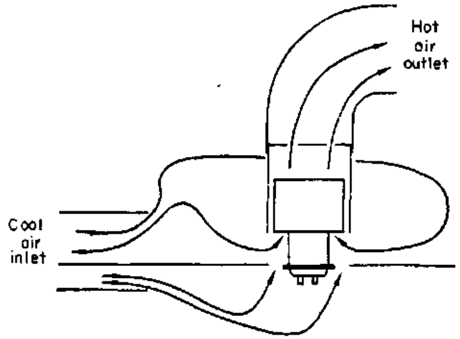


FIG. 5.7 Air cooling for valves with anodes upwards.

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

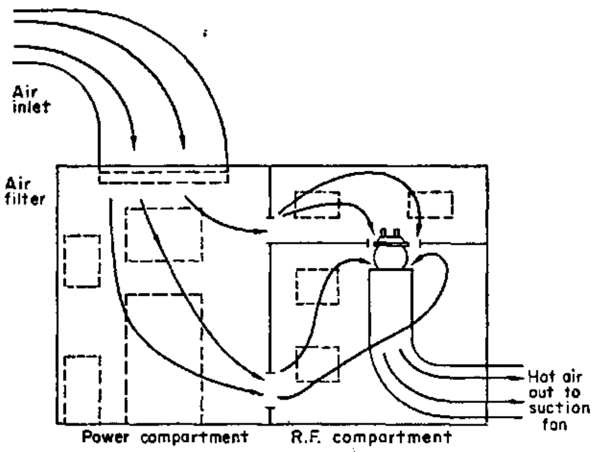


FIG. 5.8 Typical air-cooling system by suction.

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.

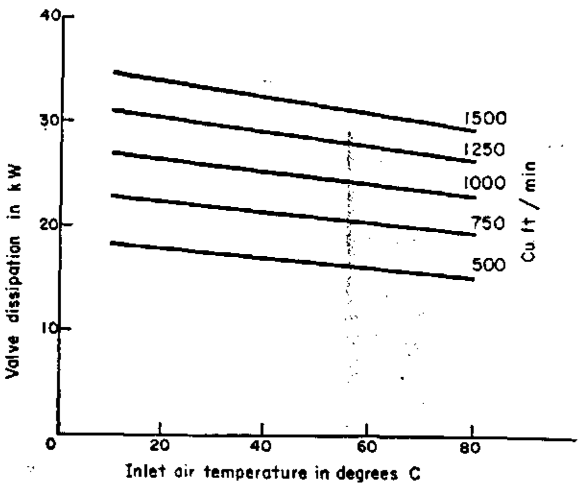


FIG. 5.9 Total valve dissipation relative to air flow and air-inlet temperature.

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

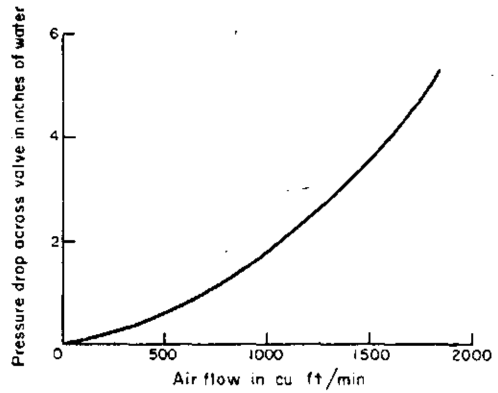


FIG. 5.10 Air flow–air pressure characteristic.

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 ft³/min at 2·0 in. water-gauge, give a flow of 1220 ft³/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 ft³/min against a head of 4·3 in. water-gauge. Therefore the initial flow would be of the order of 1280 ft³/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 ft³/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/in³. and distilled water about 40 000 Ω/in³.

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.

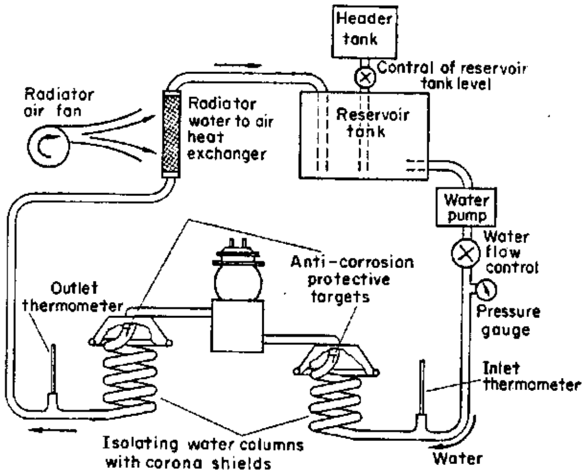


FIG. 5.11 Closed-circuit water cooling system.

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.

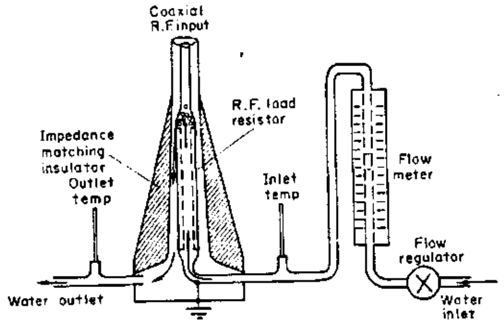


FIG. 5.12 Water-cooled r.f. load.

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

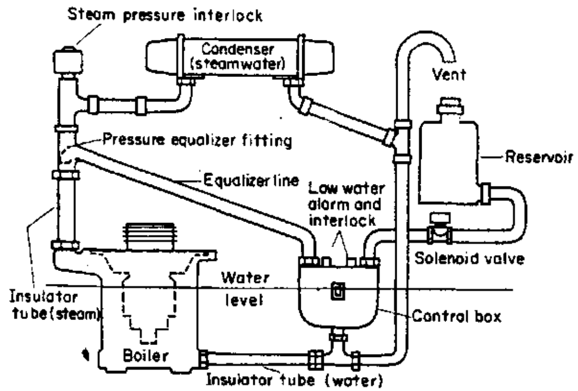


FIG. 5.13 Typical vapour-phase cooling system (Eimac).

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

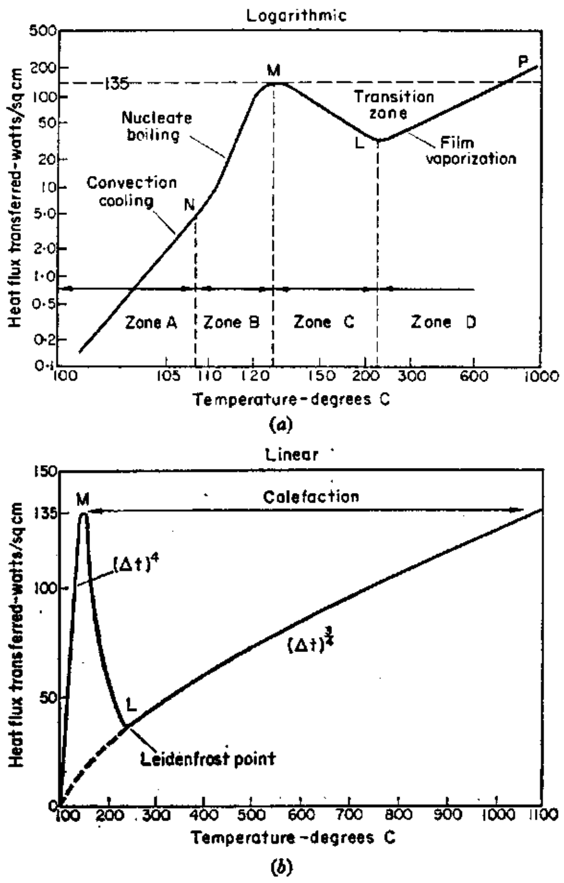


FIG. 5.14 Nukiyama heat-transfer curves.

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

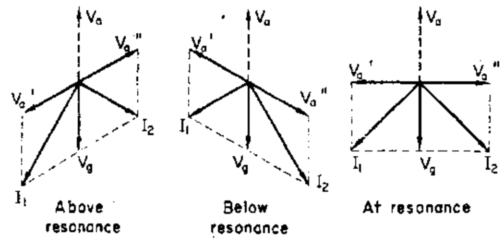


FIG. 6.1 Vector diagrams of the phase relationship in phase discriminator (Fig. 6.2).

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.

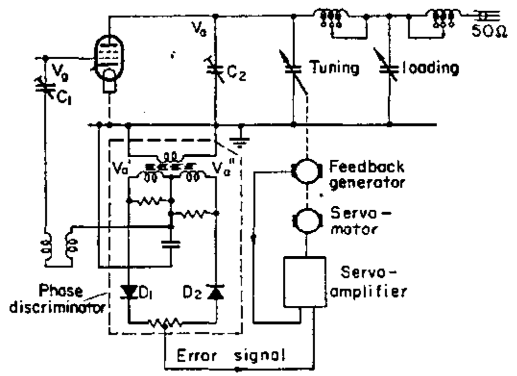


FIG. 6.2 Fine-tuning section of self-tuned amplifier.

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

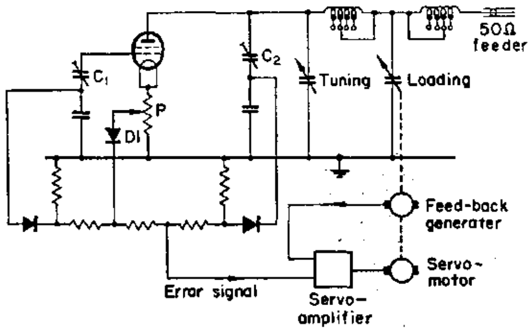


FIG. 6.3 Loading discriminator of self-tuned amplifier.

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.

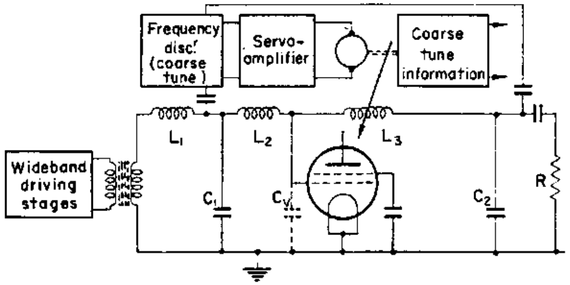


FIG. 6.4 Final amplifier input circuit for self-tuning.

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.

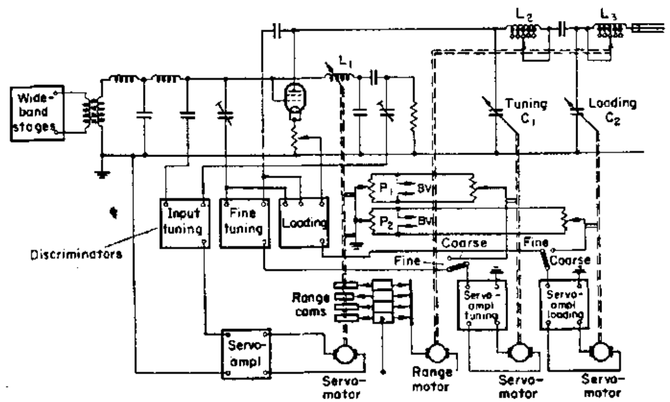


FIG. 6.5 Block diagram of complete automatic system for a self-tuned amplifier.

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.