7

Typical Designs for Broadcasting Applications

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

Reasons for circuit arrangement

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

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

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

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

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

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

The screen supply and modulation

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

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

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

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

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

Typical circuit diagram and description

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

Spurious oscillations and their prevention

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

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

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

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

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

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

The cooling system

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

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

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

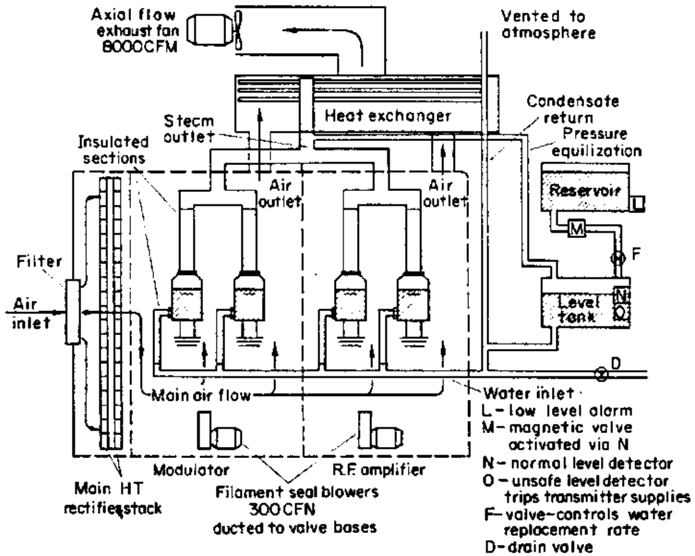

Fig. 7.2 Integrated vapour- and air-cooling system.

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

A suitable drive equipment

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

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

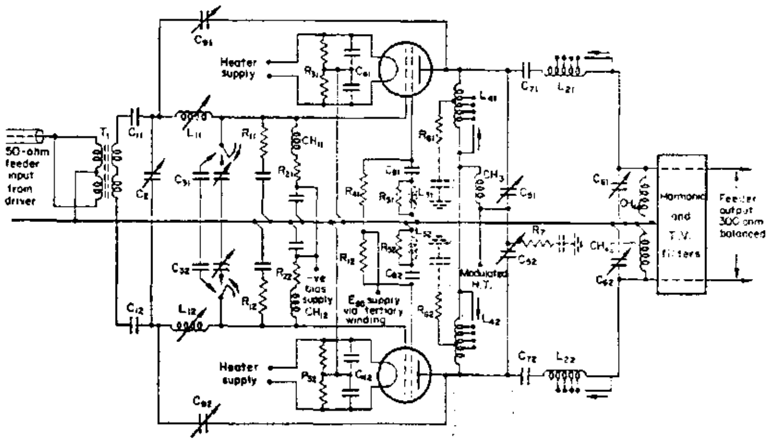

FIG. 7.1 Simplified circuit diagram of 100 kW h.f. modulated amplifier.

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

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

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

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

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

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

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

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

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

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

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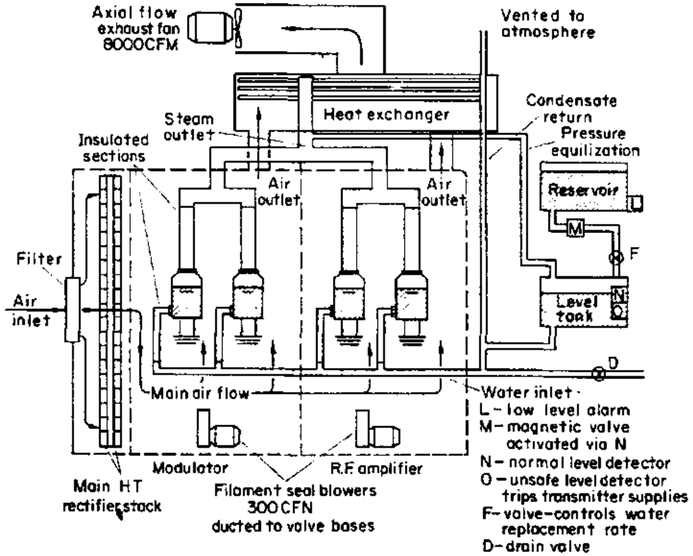

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A suitable drive equipment

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

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

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

Design considerations

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

TABLE 7.1

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

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

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

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

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

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

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

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

Circuit description

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

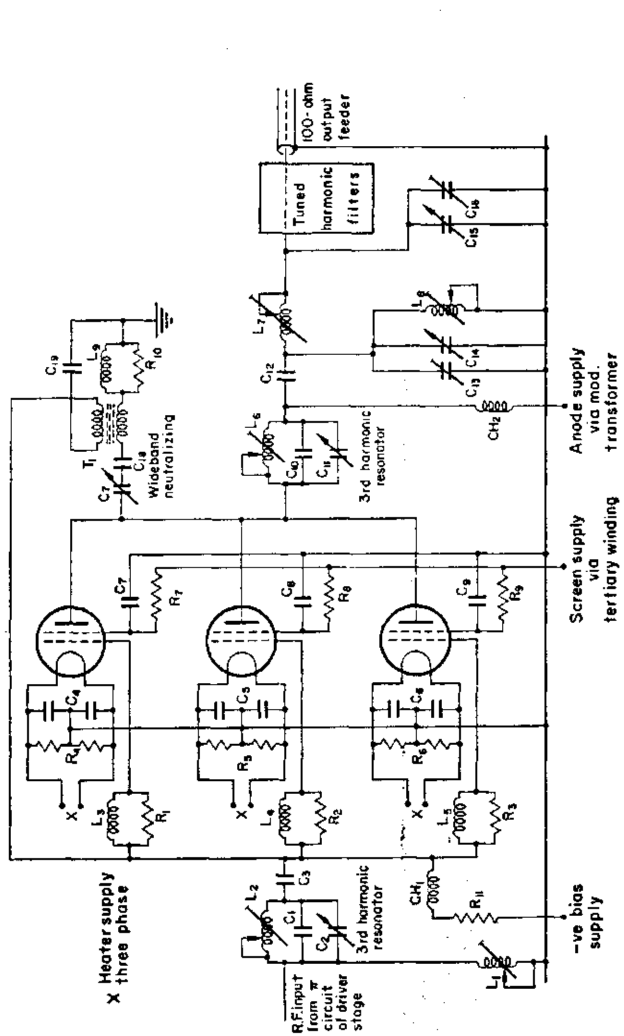


FIG. 7.3 Final amplifier for 750 kW m.f. transmitter.

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

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

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

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

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

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

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

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

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

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

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

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

Transmitter features

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

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

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

The modulation system

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

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

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

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

Final amplifier arrangement

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

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

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

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

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

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

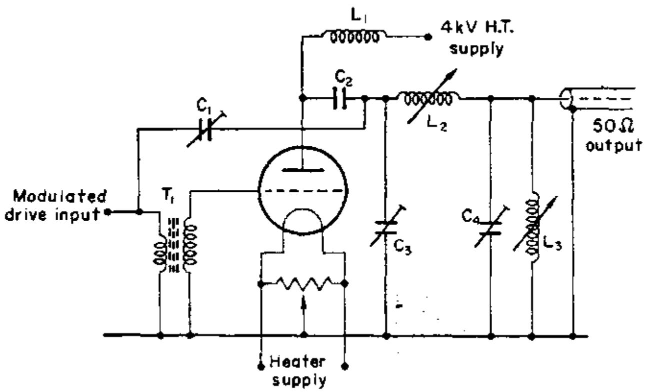


FIG. 7.4 Linear amplifier for 1·0 kW m.f. transmitter with low-level modulation.

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

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

8

An L.F. Transmitter Design

8.1 CHARACTERISTIC FEATURES

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

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

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

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

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

Choice of frequency, bandwidth and power output

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

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

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

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

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

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

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

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

Considering the upper end of the low-frequency band, say 200 kHz, antenna efficiency will be about twice that at 40 kHz, and, the frequency being five times, the antenna bandwidth will be about ten times that at 40 kHz, i.e., ±3000 Hz. Thus the Q factor of the antenna does not limit the bandwidth at the upper end of the l.f. band and speech transmission is practicable.

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

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

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

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

Multi-channel operation and frequency stability

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

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

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

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

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

8.2 CIRCUIT ARRANGEMENT

The output circuit

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

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

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

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

The final amplifier

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

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

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

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

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

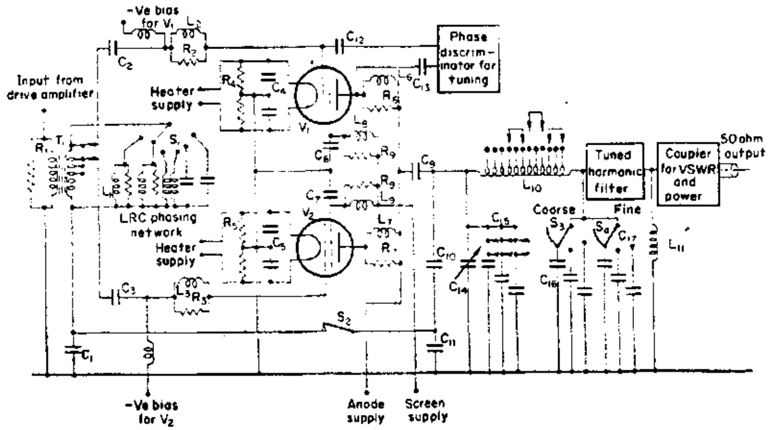


FIG. 8.1 Circuit arrangement of final amplifier for 100 kW p.e.p. in the l.f. band.

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

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

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

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

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

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

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

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

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

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

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

Power gain and drive required

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

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

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

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

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

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

9

Transmitters in Parallel

9.1 THE NEED FOR PARALLEL OPERATION

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

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

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

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

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

9.2 REQUIREMENTS FOR PARALLEL OPERATION

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

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

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

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

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

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

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

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

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

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

9.3 PARALLELING BY MEANS OF A CAPACITOR-INDUCTOR NETWORK

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

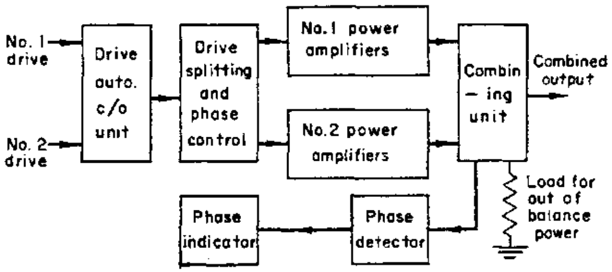


FIG. 9.1 Block diagram for operating two transmitters in parallel by means of a capacitor–inductor combining network.

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

The drive automatic changeover unit

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

Drive splitting and phasing networks

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

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

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

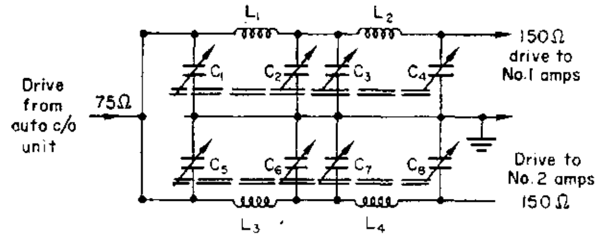


FIG. 9.2 Drive splitting and phasing network.

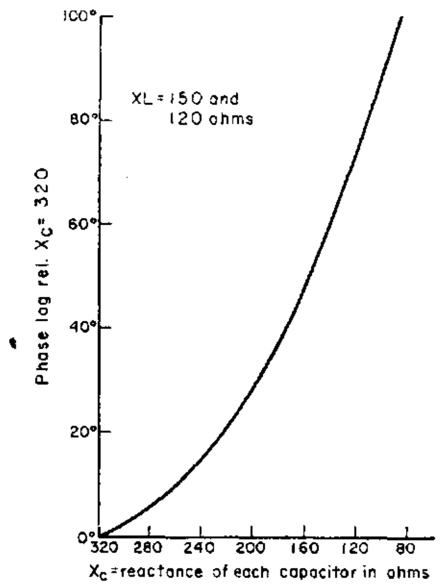


FIG. 9.3 Phase change produced by each double Π section of the phasing network.

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

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

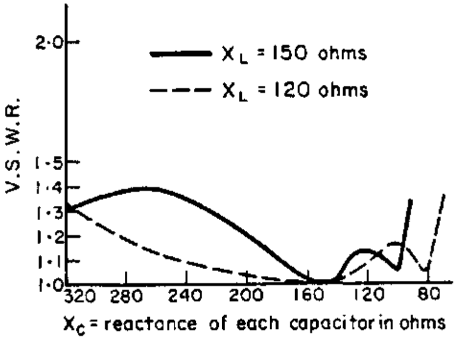


FIG. 9.4 V.S.W.R. at input of phasing network with each output matched into 150 Ω.

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

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

Output combining with a bridged 'T' network

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

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

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

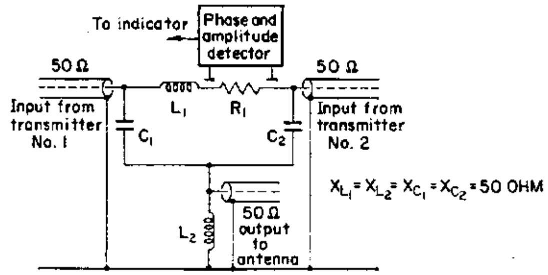


FIG. 9.6 Bridged 'T' paralleling network.

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

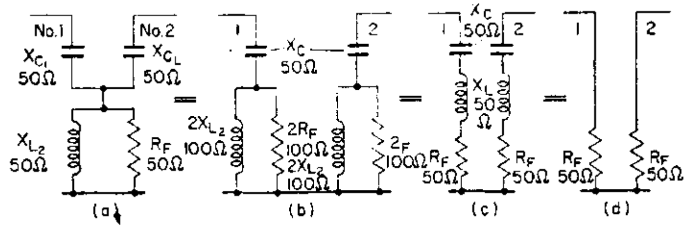


FIG. 9.7 Breakdown of the bridged 'T' network in the balanced condition.

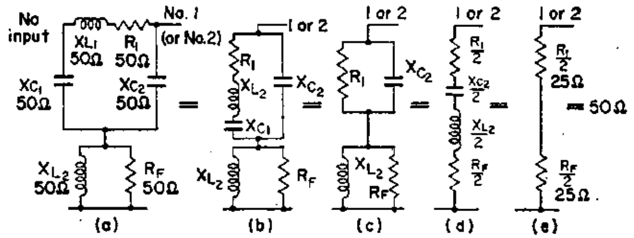


FIG. 9.8 Breakdown of the bridged 'T' network with one transmitter off.

condition, branch $L_1 R_1$ is not carrying any current and can be redrawn as Fig. 9.7(*a*). This can be redrawn as Fig. 9.7(*b*), which is the exact equivalent of Fig. 9.7(*a*). By converting the two parallel sections $2L_2$ and $2R_F$ to series components, as in Fig. 9.7(*c*), the exact equivalence is retained. In this arrangement the series capacitative and inductive reactances cancel out, giving a pure resistive loading on each transmitter of 50 Ω. This proves that in the parallel and balanced condition each transmitter is correctly matched and all the power appears in the output feeder.

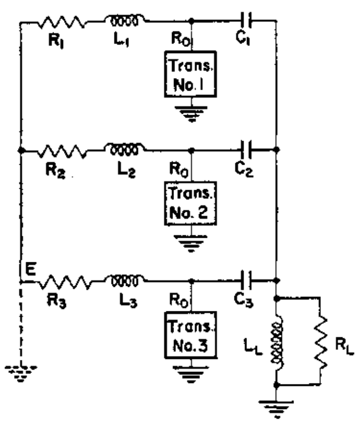
If the outputs of the two transmitters are not of the same amplitude, or not in phase, there is current in the $L_1 R_1$ branch and power will be dissipated in the load R_1. The remaining power is passed to the common output feeder. A device which detects the power in the load R_1 gives a means of indicating the combined amplitude and phase error. By successively trimming the phase and amplitude controls to minimum reading on the indicating meter, the desired condition of balance is obtained.

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

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

More than two transmitters in parallel

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



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

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

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

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

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

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

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

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

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

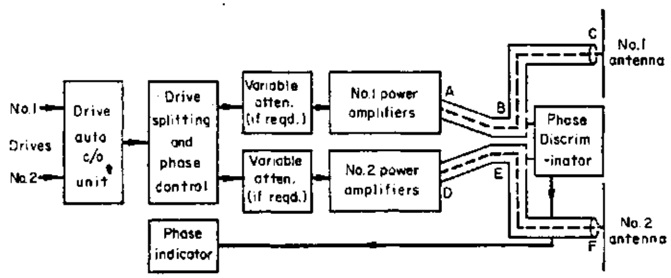


FIG. 9.10 Block diagram for operating two medium- or high-power h.f. transmitters in parallel by combining the field pattern.

The input circuit

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

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

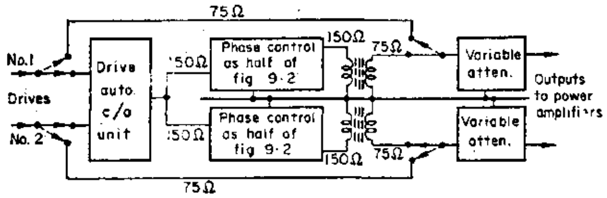


FIG. 9.11 Input arrangement when paralleling is required on a temporary basis.

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

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

The output circuit

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

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

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

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

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

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

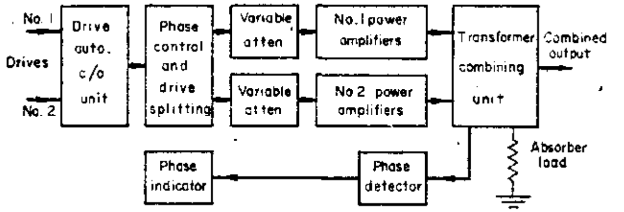


FIG. 9.12 Block diagram of paralleling by means of wideband transformers.

Paralleling with two amplifiers in push–pull

This type of operation, with the signals in the two amplifiers in antiphase, is indicated by using a single output-combining transformer, with the outputs of the two amplifiers fed into the opposite ends of the primary winding. It follows that the input to the amplifiers must also be in antiphase, and one method of achieving this condition is shown in Fig. 9.13.

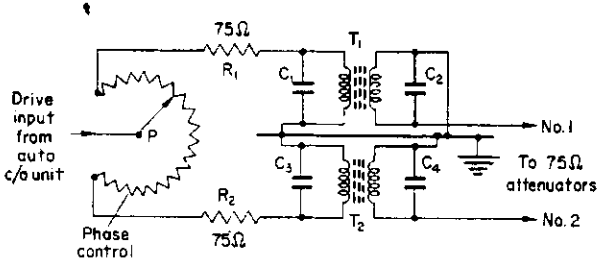


FIG. 9.13 Phasing and drive splitting for push–pull operation.

There are some interesting features in this type of input circuit. By using two 1 to 1 transformers, T_1 and T_2, with the secondary windings cross-connected to give the antiphase outputs, a better match is possible than that which would be given by a single transformer. Capacitors C_1, C_2, C_3 and C_4 are fitted to give a more linear frequency response. The 75 Ω resistors R_1 and R_2, each in series with one of the 75 Ω primary windings, are necessary to give a match with the 75 Ω input cable, by providing two 150 Ω parallel paths. From this, it is apparent that the voltage on the primary and secondary of each transformer is half the input voltage. Potentiometer P is a non-inductive resistor of low ohmic value which provides a very simple method of changing the relative phase of the two transmitters with one control. The phase change possible by this method varies over the h.f. band, and is rather limited at the lower frequencies. However, the range of phase control is adequate for low-power transmitters which are likely to contain only one or two stages.

A method of combining the transmitter outputs with a single transformer is shown in Fig. 9.14. As on the input circuit, capacitors C_1, C_2, C_3 and C_4 are fitted to improve the frequency response. When the two inputs to the transformer are of the same amplitude and in antiphase, all the power is transferred to the 50 Ω secondary, and hence to the output feeder. Any departure from this condition

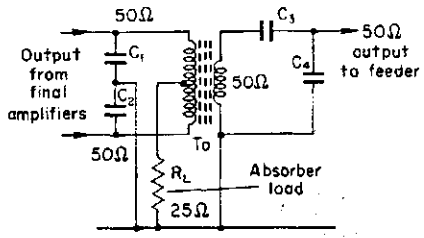


FIG. 9.14 Transformer combining unit for push–pull operation.

results in an out-of-balance power in the absorber load R_L. An amplitude detector across this load gives an indication of amplitude unbalance and incorrect phasing. A minimum on the detector meter indicates correct phase relationship and a zero reading indicates amplitude balance.

With one transmitter off, the antenna load is transformed to appear as 25 Ω across one-half of the primary winding. Half the power is transferred to the output feeder, the other half being dissipated in the absorber load, and the remaining transmitter is still terminated correctly with a 50 Ω load. Any voltage induced in the inactive primary is in phase with the voltage in the other half; as such it is effectively in shunt with the absorber load and so the input terminal of the inactive winding is effectively at earth potential.

One of the limitations with this system is the difficulty in designing the combining transformer. An accurate balance between the two halves of the primary, and the correct transfer ratio between each half and the secondary, is almost impossible to achieve over the whole h.f. band. Consequently there is inevitably a rather high v.s.w.r. reflected on to each transmitter output, and some power is lost in the absorber load. However, it is suitable for a limited frequency coverage—with some reduction in reflected v.s.w.r. and power loss—by means of improving the match with appropriate loading capacitors and inductors, as described in Chapter 10.

Paralleling with two transmitters in push–push

In view of the limitations of the push–pull system, it has been found in practice that better conditions can be provided by operating the two amplifiers in phase

(push–push). Although this entails an additional transformer in the output-combining network, the input circuit is further simplified and improved.

The arrangement of the input circuit is shown in Fig. 9.15. The potentiometer P for phase control and series resistors R_1 and R_2, for matching the input cable,

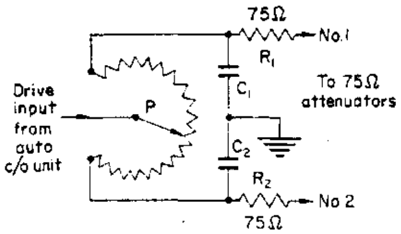


FIG. 9.15 Phasing and drive splitting for push–push operation.

are exactly the same as in the push–pull case, and only half of the input voltage is available at each amplifier input. The difference is that the two in-phase inputs can be fed directly into the amplifiers, no transformers being required to obtain an antiphase output from the single input. Thus the match on the input cable is also improved over the frequency band, due to reducing the number of components.

The output combining circuit for push–push operation is shown in Fig. 9.16. Capacitors C_1, C_2, C_3, C_4, C_5 and C_6 with inductors L_1 and L_2 are provided to

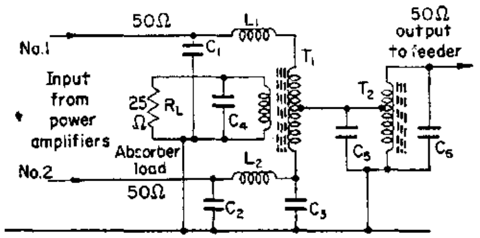


FIG. 9.16 Transformer combining unit for push–push operation.

improve the frequency response of the network. In this case, transformer T_1 is only necessary for the single amplifier condition. In parallel operation the flux induced by the two in-phase inputs cancel one another in the primary, so the two inputs are connected directly together at the centre-point output. The only design problem with this transformer for parallel operation, is that the balance between the two halves of the primary should be as near perfect as possible. Auto-transformer T_2 provides a 25 Ω-to-50 Ω ratio for matching the feeder to the transmitters, via the centre-point of T_1. Therefore the balancing and matching functions are separated, and in consequence more readily achievable.

With only one transmitter on, the power is divided equally between the output feeder and the 25 Ω absorber load, provided that the ratio of T_1 is 1 to 1 in this condition. If loads of other resistance value, such as 50 Ω, are more convenient, the ratio of transformer T_1 should be designed accordingly, but the power distribution will remain the same. There is negligible output at the terminal of the unused transmitter in either case with single operation.

Parallel operation by means of transformer combining with in-phase amplifiers is particularly appropriate for use with wideband h.f. amplifiers, because the phase delay in amplifiers of the same type is likely to be very similar. Therefore the simple phase-adjusting system shown in Fig. 9.15 gives sufficient phase control by its differential action.

For tuned amplifiers, where a greater range of phase control is likely to be required, the circuit arrangement shown in Fig. 9.2 would be more suitable. This also applies if parallel operation is required in the m.f. band, because the available phase change of the potentiometer system depends on its electrical length in terms of wavelength, which obviously decreases with frequency decrease.

REFERENCES

[1] BARTLETT, H. 'The parallel operation of broadcast transmitters'. British Patent No. 743473 (1951).

[2] MORCOM, W. J. 'The operation of transmitters in parallel'. *Sound and Vision Broadcasting*, 2, No. 1 (spring 1961).

Part 2

Medium and Low Power

10

Power Amplification Using Wideband Techniques

10.1 THE CASE FOR WIDEBAND CIRCUITS

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

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

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

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

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

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

10.2 GENERAL PRINCIPLES OF WIDEBAND AMPLIFIERS

Limiting factors

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

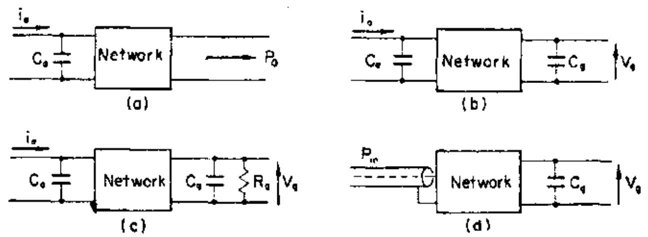


FIG. 10.1 Working conditions for a wideband amplifier. (*a*) At the output. (*b*), (*c*) At the interstage coupling. (*d*) At the input.

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

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

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

$$P_o \max = \frac{\pi}{2} \cdot i_a^2 \cdot XC_a \qquad (1)$$

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

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

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

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

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

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

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

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

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

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

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

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

Distributed amplifiers

In a distributed amplifier [5, 6, 7, 8] (see Fig. 10.2), the valve capacitances are not charged simultaneously. By placing valves along an artificial transmission line, the same instantaneous power will charge the valve capacitances in succession. To produce a given driving voltage across any of the valve-input capacitances, the same amount of power is required. In a simple amplifier, once this power has served the purpose of producing a driving voltage it is dissipated in a termination. In a distributed amplifier, however, it is used to drive the following valve. This process can be repeated over and over again so that an unlimited number of valves (subject to losses) can be driven by the same power that is required to drive one valve.

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

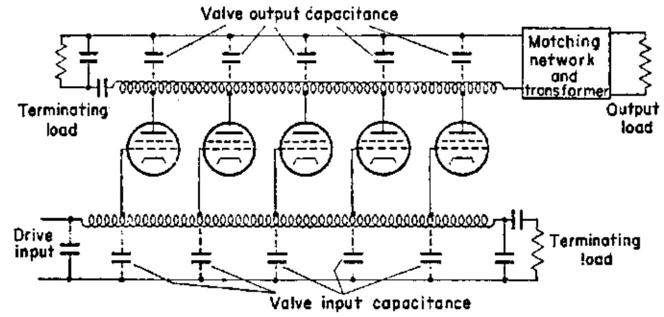


FIG. 10.2 Simplified diagram of a distributed amplifier.

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

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

The problem of power output per valve, and efficiency, is most important in transmitter applications. In an amplifier with a uniform anode line, the number of valves and the h.t. voltage are adjusted to permit optimum operation of individual valves, but half of the available power is lost. In the tapered-line amplifier, valves work inefficiently, but the addition of power is complete. In an amplifier with a uniform anode line for a bandwidth of about 30 MHz, only a few modern low-capacitance valves are required for optimum operation. By this is meant a condition under which the valves operate with a sufficiently high-impedance load line for the anode-voltage swing developed to be comparable with the h.t voltage applied. This optimum has, of course, no relation to the optimum gain per stage of a low-power distributed amplifier. Here, the first consideration is

...ract maximum power over a wide band from a given set of valves; the ... provided ... w, is of secondary importance.

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

$$P_0 = mX'C_a\left(n - \frac{n_1}{2}\right)n_1 \cdot i_a{}^2 \tag{4}$$

as the frequency tends to zero,

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

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

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

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

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

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

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

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

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

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

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

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

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

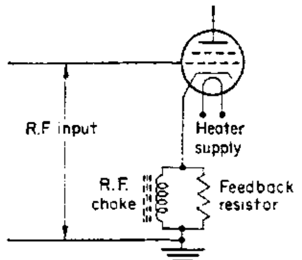


FIG. 10.3 Circuit for r.f. cathode feedback.

10.3 A 1 kW WIDEBAND TRANSMITTER, 2–28 MHz

The final amplifier

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

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

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

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

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

Design and construction of anode and grid lines

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

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

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

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

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

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

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

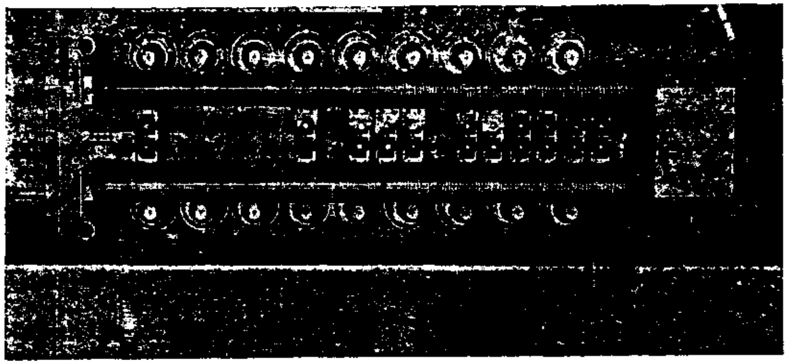


FIG. 10.4 Anode delay lines showing the tapering sections.

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

The penultimate amplifier

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

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

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

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

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

10.4 PERFORMANCE OF 1 kW WIDEBAND TRANSMITTER

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

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

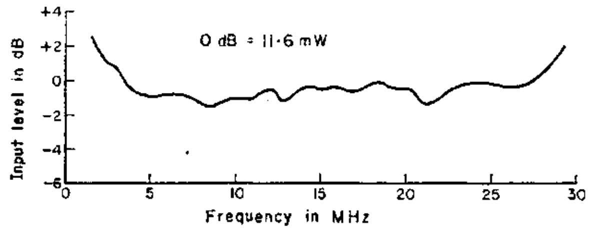


FIG. 10.5 Frequency response of two-stage distributed amplifier, in terms of drive level required for 1 kW output.

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

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

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

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

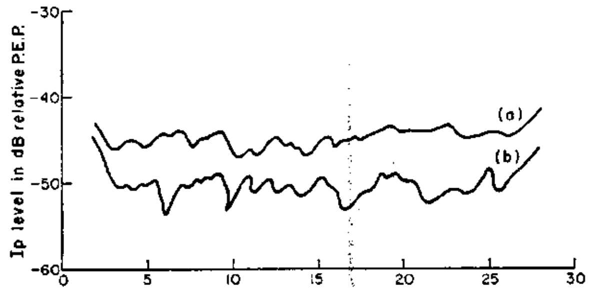


FIG. 10.6 Intermodulation products between two closely spaced tones with reference to p.e.p. (*a*) 1 kW p.e.p. (*b*) 500 W p.e.p.

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

10.5 MULTI-FREQUENCY OPERATION

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

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

TABLE 10.1

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

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

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

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

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

10.6 FREQUENCY EXTENSION TO COVER THE M.F. BAND

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

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

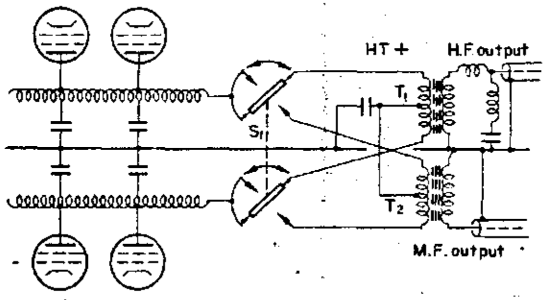


Fig. 10.7 Output circuit arrangement for h.f. and m.f. version of distributed amplifier.

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

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

10.7 WIDEBAND TRANSFORMERS

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

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

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

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

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

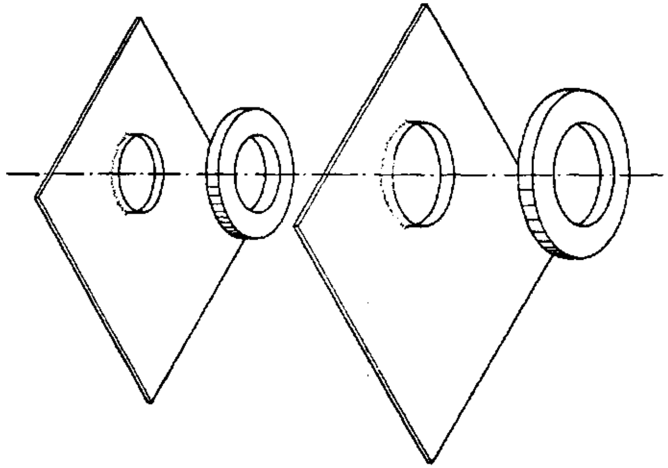


FIG. 10.8 Method of ferrite core assembly for 40 kW h.f. transformer.

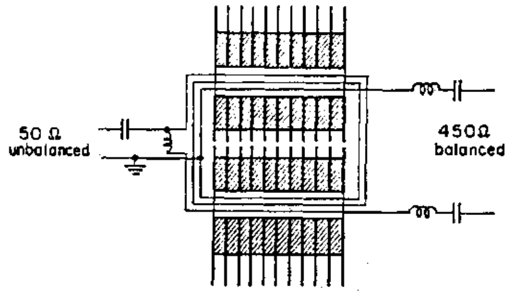


FIG. 10.9 Circuit arrangement of a 40 kW wideband transformer.

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

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

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

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

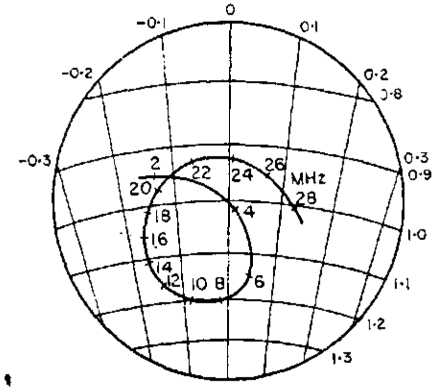


FIG. 10.10 Smith's chart presentation of input matching of a 40 kW wideband transformer.

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

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

REFERENCES

[1] GREEN, E. 'Amplitude-frequency characteristics of ladder networks'. *Marconi Rev.* (1957), Monograph, p. 31.

[2] WHEELER, H. A. 'Wideband amplifiers for television' *Proc. Inst. Radio Engrs.*, **27**, 429 (1939).

[3] *Reference Data for Radio Engineers*. 4th edition, p. 187. International Telephone and Telegraph Co., New York (1956).

[4] WEINBERG, L. 'Network design by use of modern synthesis technique and tables'. Technical Memoranda Nos. 427 and 434 Research Lab., Hughes Aircraft Co.

[5] PERCIVAL, W. S. British Patent No. 460562 (1936).

[6] GINZTON, E. L., HEWLETT, W. R., JASBERG, J. H. and NOE, J. D. 'Distributed amplification'. *Proc. Inst. Radio Engrs.*, **36**, 956 (1948).

[7] HORTON, W. H., JASBERG, J. H., and NOE, J. D. 'Distributed amplifiers: practical considerations and experimental results'. *Proc. Inst. Radio Engrs.*, **38**, 748 (1950).

[8] CORMAK, A. 'Distributed amplification'. *Electronic Engng.*, 24, 144 (1952).

[9] ESPLEY, D. C. British Patent No. 623175 (1944).

[10] SOSIN, B. M. British Patent No. 846633 (1958).

[11] MAURICE, D. and MINNS, R. H. 'Very wide band radio-frequency transformers'. *Wireless Engr.*, **24**, 168, 209 (1947).

[12] 'On the use of ferrites in wide-band h.f. transformers'. *Point to Point Telecomm.*, **1**, 22 (1957).

[13] ELSEY, G. and JACKSON, N. C. O. 'A new approach to the design of ferroxcube cores for wide-band h.f. transformers'. *Mullard tech. Commun.*, **5**, 182 (1960).

11

Intermediate-Stage Amplifiers

11.1 GENERAL CONSIDERATIONS

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

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

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

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

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

11.2 THE INPUT CIRCUIT

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

17·5 V. As the intermediate amplifier will not normally run into grid current, the incoming voltage can be stepped up with a wideband transformer, shown as T_1 in Fig. 11.1. The terminating resistor R_1, connected across the secondary, is a much higher value than 75 Ω, so the peak voltage made available for driving the valve is increased. The amount by which the voltage can be stepped up is limited—mainly by the input capacitance of the valve in relation to the operating frequency, but also by the transformer design.

Input capacitative reactance must be of the order of 1·7 times the value of the terminating resistor if the v.s.w.r. on the drive cable is not to exceed 1·2 to 1. If the resistor on the secondary has a value of 600 Ω, the upper frequency of

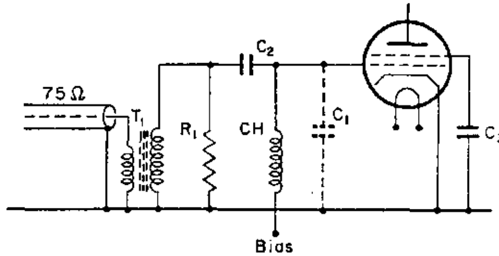


Fig. 11.1 Input circuit using a wideband step-up transformer.

operation is reached when the reactance of the valve input capacitance is about 1000 Ω.

For 600 Ω to be the correct termination on the transformer secondary from a 75 Ω cable, the voltage step-up ratio is $\sqrt{(600/75)} = 2\sqrt{2}$. The design of a transformer with this ratio is not a problem at a throughput level of 2 W, and by its use the available peak voltage at the grid is raised from 17·5 V to 50 V.

The valve used for intermediate amplifiers will be of a comparatively low power rating and not large, so the input capacitance is likely to be less than 50 pF, and is often nearer to 20 pF. The reactance of 50 pF at 3 MHz is 1060 Ω, so the input circuit arrangement of Fig. 11.1 will be suitable for the v.l.f., l.f., and m.f. bands, with a transformer having a step-up ratio of $2\sqrt{2}$ and a terminating resistor of 600 Ω on the secondary.

In order to take advantage of a similar voltage step-up for frequencies above 3 MHz, consideration should first be given to selecting a type of valve which has a low input capacitance, consistent with other characteristics being satisfactory. For a final-stage drive power (including loading) of up to 300 W, tetrode type 4CX250B (Eimac) is suitable, having an average input capacitance of 16 pF. The reactance of 16 pF at 10 MHz is 1000 Ω, so the input circuit of Fig. 11.1 will be satisfactory up to 10 MHz when using a 4CX250B tetrode.

For frequencies above 10 MHz the valve input capacitance could be tuned out for each frequency used, but bearing in mind the operational need to change frequency in the h.f. band, this is not a good solution. But, a partial tuning arrangement is quite practical, using the input circuit shown in Fig. 11.2. Values of inductance are selected by a switch, S_1, so that in combination with the valve input capacitance C_1, each inductor will cover a range of frequencies at which the reactance is 1000 Ω or more. In the majority of cases it will be found that the

two inductors L_1 and L_2 will be sufficient to cover frequencies in the h.f. band at which the reactance of the input capacitance would otherwise be less than 1000 Ω.

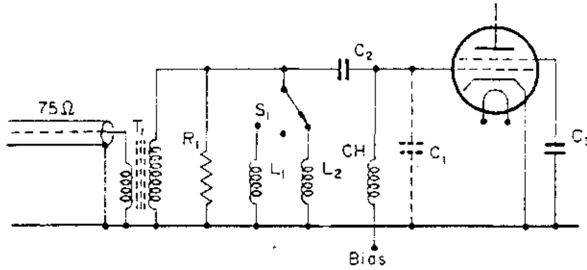


Fig. 11.2 Partial tuning of the input circuit.

For intermediate stages in linear amplifier chains, it is preferable to avoid the use of r.f. feedback as a means of improving linearity, because the input level is reduced by the amount of feedback employed. Consequently, the available output is also reduced by the same amount. If an improvement in linearity is required, the circuit arrangement shown in Fig. 3.1, Chapter 3, should be used. This is a form of envelope feedback which does not reduce the drive level appreciably.

11.3 INTERSTAGE COUPLING WITH A Π CIRCUIT

The most suitable method of interstage coupling when driving a final amplifier in a grounded-cathode arrangement is a Π circuit as shown in Fig. 11.3. It is essentially simple for tuning and loading and the final-stage input capacitance forms part of the shunt capacitance of the Π circuit, so it does not have to be tuned out by a separate control. This means that the value of the terminating

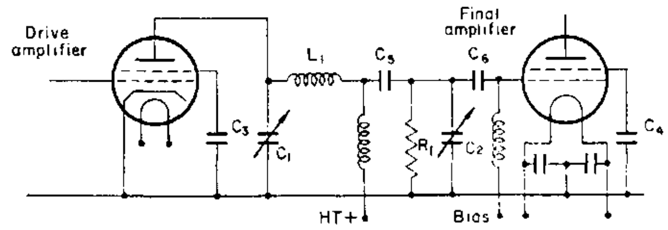


Fig. 11.3 Interstage coupling with a Π circuit.

resistor R_1 is determined only by the driving voltage required in relation to the power available from the driving amplifier.

For linear operation without grid current, apart from circuit losses, the resistor R_1 provides a constant load over the whole frequency range to be covered, so the driving amplifier loading conditions are the same for all frequencies.

For linear operation with grid current, although the grid current is not likely to be high, the load it produces must be only about one-quarter of that produced by the loading resistor R_1, otherwise peak flattening will occur with a consequent reduction in linearity. The peak value of the grid current should be kept low, otherwise considerable power will be required from the drive for good linearity.

Grid current is normal with class C operation and it is preferable to use some automatic bias on the final stage as a means of regulating the drive voltage and maintaining a constant load on the driving amplifier. Under these circumstances the load resistor R_1 is not necessary, although a resistor of high value is sometimes fitted as a further stabilizing device.

In all cases of the final stage taking grid current, there is an advantage in using a Π circuit for interstage coupling, when the final stage is in a grounded-cathode arrangement; the advantage is greater in linear amplifiers. Due to the phase reversal in the Π circuit, the *driven* negative-going voltage on the driver anode produces the positive-going driving voltage on the grid of the final stage. Without this phase reversal, such as would be produced by direct capacitative coupling, the driving power would be obtained from the stored energy in the resonant circuit. As such, the amount of peak flattening would depend on the Q factor of the tuned circuit and would give poor linearity with low Q factors.

11.4 INTERSTAGE CAPACITATIVE COUPLING

While capacitative coupling is unsuitable for driving a grounded-cathode stage, it gives the correct phase relationship for driving a grounded-grid stage because the driving half cycle is negative going. The circuit arrangement is shown in Fig. 11.4, where shunt components L_1C_1 tune the anode circuit of the driving

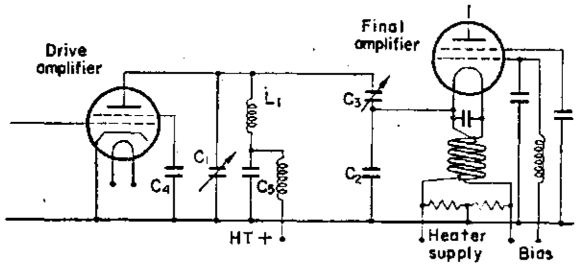


FIG. 11.4 Capacitative interstage coupling.

amplifier, and the capacitor C_3, of capacitance potentiometer C_3C_2, provides the loading control. No additional loading resistor is needed because the main load is provided by the anode–cathode current in the cathode circuit. This load is practically constant over the driving cycle and is usually much greater than that produced by the grid current. This means that peak flattening is very small, so the circuit is very suitable for linear amplification.

There are a number of other points which should be noted in connection with this arrangement. First, the driving power required is very much greater than that needed for a grounded-cathode amplifier, but it is not all wasted because

the power in the anode–cathode circuit is in series with the output power, and appears as throughput in the output circuit.

Secondly, even at minimum setting of C_3 the capacitance of the potentiometer is added to the output capacitance of the driving amplifier, reducing the frequency to which the driving amplifier will tune. This is particularly the case with amplification over a wide band of frequencies, where capacitor C_2 must always provide a low-impedance path to harmonic frequencies. In consequence, the value of C_3 is governed by the value of C_2 required at the low-frequency end of the band, so C_3 tends to be higher than the value required for the higher frequencies. This effect can be reduced by making C_2 either variable or changed in steps over the frequency range.

Thirdly, in the circuit shown, the final amplifier valve is a tetrode, which gives a high order of linear performance in a grounded-grid circuit. In grounded-grid circuits the first and second points are equally applicable to triodes which also give good linearity, but the triodes do not require a screen supply. So, it is more usual to employ triodes if the grounded-grid arrangement is used.

11.5 INTERSTAGE COUPLING WITH A QUARTER-WAVE NETWORK

One of the main characteristics of a line one-quarter of a wavelength long, is that its impedance $Z_0 = \sqrt{(R_1 R_2)}$, where R_1 is the effective resistance at the input, when the line is terminated with R_2. At any particular frequency a network of lumped capacitance and inductance elements in a Π circuit arrangement can behave as a quarter-wave line, if the shunt and series components all have the same reactance. The value of this reactance is then the impedance of the quarter-wave network of lumped elements.

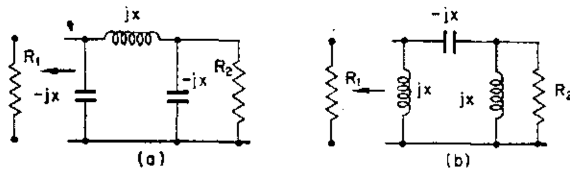


FIG. 11.5 Two types of quarter-wave network.

Two such networks are shown in Fig. 11.5(a) and (b), wherein $jx = -jx = Z_0$. In both cases the network will look like R_1 at the input when terminated by R_2, which may be of any value, i.e., $\sqrt{R_1} = Z_0/\sqrt{R_2}$, or $R_1 = Z_0^2/R_2$. By reducing R_2 to half value the input resistance is increased from R_1 to $2R_1$.

Now consider a requirement for a constant voltage at the output for a change in terminating resistor R_2 to half value. The power in R_2 will increase from V^2/R_2 to $V^2/0{\cdot}5R_2$ for the lower value and the power at the input of the network will also be increased by the same ratio. If I_0 is the current in the effective input resistance R_1, the power at the input will be I_0^2/R_1 for R_2 on the output. If I_1 is the input current when the output termination is $0{\cdot}5R_2$, the power will be $I_1^2/2R_1$, because R_1 will have increased to $2R_1$. But the power at the input will also be twice the original power, so $I_0^2/R_1 = 2I_1^2/2R_1$ and $I_1 = I_0$. It can be seen that irrespective of the value of the terminating resistor R_2 ***a quarter-wave network will provide a constant voltage output from a constant current input.***

A constant voltage is exactly that required for driving a linear amplifier when the effective input resistance changes over the driving cycle, such as occurs in a grounded-cathode stage running into grid current.

Reference to the characteristics of any tetrode will show that for a given driving voltage the anode current is substantially constant over a very wide range of anode voltage. In fact, this is so unless the anode voltage approaches the same as, or is less than, the screen voltage. By using a quarter-wave network for interstage coupling from a driving tetrode, peak flattening due to grid-current loading can be virtually eliminated. It follows that this system is particularly suitable for interstage coupling in linear amplifier chains.

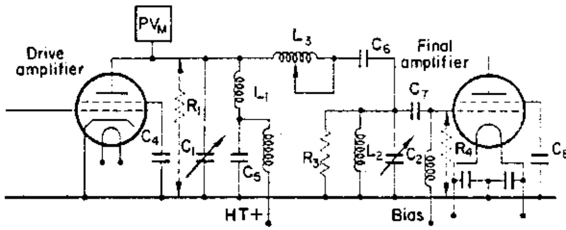


FIG. 11.6 Practical arrangement for interstage coupling with a quarter-wave network.

In giving consideration to a practical circuit arrangement, (*a*) or (*b*) in Fig. 11.5 are equally effective, but there is a slight preference for (*a*) because of its better attenuation to harmonic frequencies which may be produced in the driving amplifier. This leads to the practical configuration shown in Fig. 11.6.

The first point to consider is the impedance of the network Z_0 from the driving requirements of the final amplifier, to give its full output, and the operational load line of the drive amplifier to give the peak power of the final-stage drive.

The effective input resistance R_4 is taken as the peak grid voltage divided by the peak grid current. The added resistor R_3 is to limit the change in load-line resistance of the drive caused by the change between zero and peak grid current. Typically, R_3 is about equal to R_4, although the value is quite arbitrary and much higher values can be used effectively. The value used for the terminating resistance R_2 is given by resistances R_3 and R_4 in parallel.

The r.m.s. power required for the peak load is given by $PV^2/2R_2$, which enables the load line of the driving amplifier to be selected for peak linear output. As linear amplification is being considered, the value R_1 is twice the value of the slope resistance for all practical purposes. Therefore the required Z_0 of the network is given by $\sqrt{(R_1 R_2)}$ and the three arms of the network must be set to this numerical value. Drive anode circuit $L_1 C_1$ (including the valve output capacitance) is set to equal $-jx$, L_3 is set to jx and $L_2 C_2$ (including the final amplifier input capacitance) is set to $-jx$.

The necessity to adjust three controls for each frequency, combined with the fact that unorthodox tuning procedures make it unsuitable for automatic self-tuning, is the main reason why this arrangement is not used extensively in the

h.f. band. However, the effective Q factor of the network is unity, so the settings are not critical, and automatic frequency changing between pre-selected settings is quite a practical proposition. Its use has been proven when driving linear amplifiers with output power up to 30 kW, with triodes in grounded-cathode circuits, where the peak grid current was about one-third of the peak anode current.

In order to obtain the correct values for pre-setting the circuits, the series element L_3 must first be calibrated in terms of reactance and frequency. This can best be done by disconnecting it from the circuit at the points where it connects to the shunt circuits, and resonating the whole series arm with known values of capacitance. This is a once-only operation, subsequently the inductor is set either to logged positions or to a calibration chart.

With L set for a particular frequency, the bias is removed from the final stage (without h.t. of course), thereby providing an effective short across $L_2 C_2$, so that L_3 is in shunt with $L_1 C_1$. With a low-level signal into the drive amplifier, circuit $L_1 C_1$ is resonated with L_3, as indicated by a *maximum* reading on the peak voltmeter, making $L_1 C_1$ equal to $-jx$. Then, with final-stage bias on and the signal level below that required to give final-stage grid current, circuit $L_2 C_2$ is adjusted to series resonate with L_3, as indicated by a *minimum* reading on the peak voltmeter. This means that $L_2 C_2$ is also equal to $-jx$. Once set for a frequency, no attempt should be made to trim any of the component arms on full power. The circuit properties can only be degraded by such action.

11.6 TYPICAL EXAMPLES OF QUARTER-WAVE NETWORK CONDITIONS

Consider a final amplifier requiring a drive of 225 peak volts and 0·5 A peak grid current, which give a value of 450 Ω (225/0·5) for R_4 and an r.m.s. power of 56·25 W. If a value of 550 Ω is used for R_3, then R_2 will be 247·5 Ω, say 250, at full signal level, and a total power of 100 W will be required from the drive amplifier.

Tetrode type 4CX250B is capable of giving 100 W as a linear amplifier and the linearity is good, so reference should be made to the characteristics shown in Fig. 11.7 for this application.

With an h.t. voltage of 1400 and a static feed of 0·1 A, operating on load line A, to 0·5 A at 400 V, will give a maximum output condition of 100 W, by the approximate method

$$\frac{(0{\cdot}5 - 0{\cdot}1) \times 1000}{4} = 100 \text{ W}$$

The slope resistance of the load line is 2500 Ω (1000/0·4), giving the effective r.f. load across the input of the network R_1 of 5000 Ω. The required Z_0 of the network is $\sqrt{(R_1 R_2)} = \sqrt{(5000 \times 250)} = 1118$ Ω, say 1100 Ω. This is the value to which the jx and $-jx$ arms should be set by the method described, for every operational frequency.

Now consider the low-level condition when there is no grid current in the final amplifier. Resistance R_4 does not exist and the total load R_2 is given by the 550 Ω of R_3. From $Z_0 = \sqrt{(R_1 R_2)}$, $R_1 = Z_0{}^2/R_2 = 2200$ Ω and the slope of the load line changes to 1100 Ω, shown as B on Fig. 11.7. Obviously, the excursion is not that given by the full extent of the load line, but it enables the ratio of anode current change to be obtained by reference to the 0·4 A curve. For the same grid voltage of −10 V, the anode current on line A is 0·4 A and on line B it is 0·41 A, which is an increase of 2·5%.

Via the quarter-wave network, this means that the regulation of the final amplifier driving voltage is only 2·5% from no grid current to peak grid current.

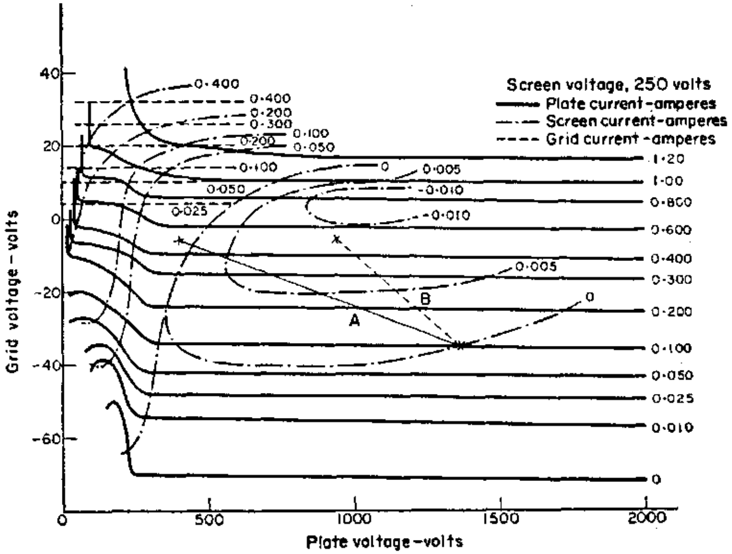


FIG. 11.7 Constant-current characteristics of tetrode type 4CX250B (Eimac).

In order to obtain a regulation of 2·5% with other forms of interstage coupling from a tetrode driving stage, the shunt resistor would have to be so low that at least 1 kW of total drive power would be required; an impossible consideration to supply an actual driver power of only 56·25 W r.m.s.

Referring again to Fig. 11.7, the power output from the driving amplifier is given with a grid excursion from −36 V to −6 V, i.e., 30 V. In the case considered as an example, with a 2 W drive available, the required 30 V grid swing can be obtained with a transformer ratio of 1·75 to 1, terminated on the secondary with 225 Ω. As the input capacitance of the 4CX250B is 16 pF (average), its reactance is 360 Ω at 28 MHz. With a shunt resistor of 225 Ω, it means that the input circuit shown in Fig. 11.1 will be satisfactory up to 28 MHz, without any shunt inductance needed for higher values of shunt resistance as shown in Fig. 11.2.

12

Amateur Transmitters

12.1 POWER CONSIDERATIONS

A feature of h.f. transmitting equipment for use by amateurs, is the regulation which imposes a limit on the maximum power that may be used. This limit is defined as the maximum peak r.f. power given by a d.s.b. amplitude modulated system (A3) with 150 W d.c. into a final r.f. amplifier operating at 66% efficiency.

With a conversion efficiency of 66% for a d.c. power input of 150 W, the r.f. feeder carrier output is 100 W and the d.c. input is unaffected by anode modulation; the increased output power being supplied by the modulator. At 100% modulation, the power in each of the two sidebands is one-quarter of the carrier power, which is half the carrier amplitude per sideband. This power and amplitude relationship is shown in Fig. 12.1(*a*). It will be seen that with a carrier power of 100 W, each half-amplitude sideband is 25 W and the peak amplitude is twice the carrier amplitude, which is 400 W p.e.p.

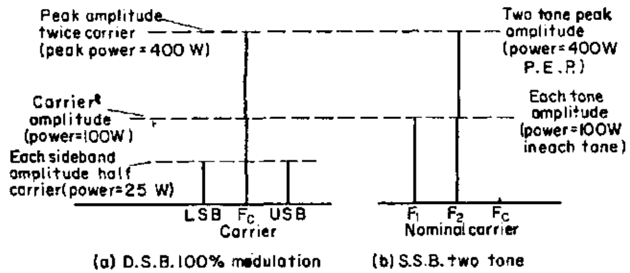


FIG. 12.1 Comparison between the spectral components in d.s.b. and s.s.b. to give the same p.e.p.

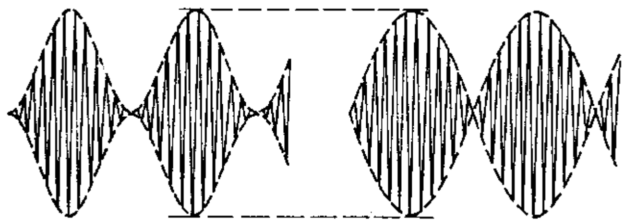
Peak power relationship between d.s.b. and s.s.b.

With an upper power limit in terms of p.e.p., there is an obvious advantage in employing a system by which as much as possible of the total power is used to convey the intelligence, e.g. by using an s.s.b. system.

In a d.s.b. system all the intelligence is conveyed in either of the two sidebands, each of which is only one-sixteenth of the peak power, as shown in Fig. 12.1(*a*) (25 W out of a total of 400 W). With an s.s.b. system, after d.s.b. modulation at a low-power level the carrier and one sideband are filtered out so that only one sideband is radiated at a peak power equal to the total peak power. The peak intelligence power radiated with s.s.b. is sixteen times that with d.s.b. (In order to recover the intelligence, it must be remembered that the carrier must be re-inserted at the receiver at the correct frequency.)

To check the performance of s.s.b. systems, it is usual to apply two equal amplitude tones to one sideband, so that the peak power is reached every time the radio frequencies corresponding to the tones are in phase, i.e., at the difference frequency of the tones. With two-tone modulation applied, the power relationship with s.s.b. is shown in Fig. 12.1(*b*) and compared with d.s.b. in Fig. 12.1(*a*).

The dynamic condition of the two systems is shown in Fig. 12.2(*a*) and (*b*) for the same p.e.p. as would be seen by displaying the r.f. waveforms on an



(a) D.S.B 100% modulation (b) S.S.B. two tone

FIG. 12.2 Comparison between d.s.b. and s.s.b. r.f. waveforms as seen on an oscilloscope.

oscilloscope. The two-tone method of checking is only a test condition, and in practice the whole of the peak power is available for the single-channel operation normally used.

Most amateurs started with very low power with such good results that a p.e.p. of 400 W on s.s.b. seems unnecessarily high, and few use as much as 200 W p.e.p. Even at this level, the power is eight times the power of 25 W in one sideband of a d.s.b. system which has a total p.e.p. of 400 W.

It might be argued that the two sidebands of d.s.b. system add up at the receiver to give a higher signal strength, but this is not always the case. In propagation conditions, appropriate for selective fading, the path length travelled by the two sidebands is different, and the difference varies. The result is a relative change of phase between the two sidebands at the receiver, so they add and subtract in a random manner, causing fading of the signal. This cannot occur with only one sideband, and accounts for the comparative absence of this type of fading with s.s.b. operation, resulting in a more steady signal strength.

Mean power and peak power

The relationship between mean and peak power conditions on s.s.b. is given in Table 12.1 for single frequency and the two-tone test condition for amplifiers operating in class B. These assume sinusoidal waveforms and perfectly linear valve characteristics, but they are sufficiently accurate for practical purposes when used with valves suitable for linear r.f. amplification. If greater accuracy is required, reference should be made to the method described in Chapter 2.

The formulae relate to single-valve operation and the following definitions apply for PI_A and PV_A.

PI_A = the anode current excursion from zero to the peak current at peak signal (the static feed at zero signal level is ignored by this approximation);

PV_A = the anode voltage excursion from the h.t. voltage to the anode voltage at peak anode current.

TABLE 12.1

	Single frequency	Two-tone test
I_A, d.c.	$\frac{PI_A}{\pi}$	$\frac{2PI_A}{\pi^2}$
Power output, watts r.m.s.	$\frac{PV_A \cdot PI_A}{4}$	$\frac{PV_A \cdot PI_A}{8}$
Power input, watts, d.c.	$\frac{PI_A \cdot \text{h.t.v.}}{\pi}$	$\frac{2PI_A \cdot \text{h.t.v.}}{\pi^2}$
Anode dissipation, watts	$PI_A\left(\frac{\text{h.t.v.}}{\pi} - \frac{PV_A}{4}\right)$	$PI_A\left(\frac{2\text{h.t.v.}}{\pi^2} - \frac{PV_A}{8}\right)$
Conversion efficiency % (d.c. to r.m.s. output)	$\frac{25 \cdot \pi \cdot PV_A}{\text{h.t.v.}}$	$\frac{6{\cdot}25 \cdot \pi \cdot PV_A}{\text{h.t.v.}}$

First, note that the mean power to peak power ratio is 1 to 2. The next step is to obtain the required operating load line on the characteristics of a particular valve type by means of the formulae. As an example, assume that the required output is 200 W p.e.p. On the two-tone test the mean power output will be 100 W (50 W per frequency) and the valve will need to have an anode-dissipation capability of about 100 W. A tetrode is the obvious choice because of the low-power drive required. Tetrode type QV08-100 (Mullard) allows an anode dissipation of 100 W and the characteristics are quite linear, as seen in Fig. 12.3. In addition, it is an inexpensive valve.

At peak anode current in the region of 1·0–1·5 A, it is seen that the characteristics are substantially linear down to V_A of 120 V, so with an h.t. voltage of 800 V the PV_A swing will be 680 V. Allowing 5 W for circuit losses and any departure of individual valves from the typical characteristics, 210 W p.e.p. will be required and a mean power of 105 W.

Then $PV_A = PI_A/8 = 105$, and as $PV_A = 680$ V, $PI_A = 8 \times 105/680 = 1{\cdot}235$ A. Thus

I_A d.c. $= 2PI_A/\pi^2 = 0{\cdot}25$ A
Power input, d.c. $= 0{\cdot}25 \times 800 = 200$ W
Power output, r.m.s. $= 105$ W
Anode dissipation $= 95$ W

Even in the two-tone condition the anode dissipation is below the rated maximum of 100 W and the output is 210 W p.e.p.

The static anode feed is selected to give good linearity, provided that the anode dissipation is not exceeded. Typically, for a valve of this type it will be about 0·1 A, giving a zero-signal dissipation of 80 W. This means that the range of d.c. input power is from 80 W with no signal, to 200 W on two-tone. With normal speech the average modulation depth is about 20–30%, so the *average* d.c. input will be of the order of 100–110 W, with signal peaks of 210 W p.e.p.

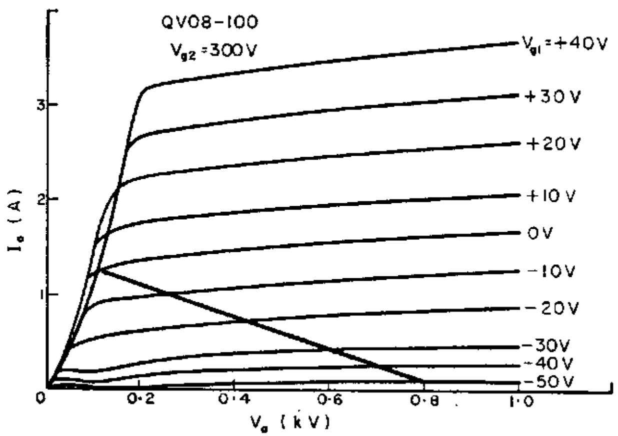


FIG. 12.3 Characteristics of tetrode type QV08-100 with load line for 210 W p.e.p. output.

Compare this with the d.c. input power for 400 p.e.p. output with a d.s.b. system at 100% modulation. The a.f. power required to modulate the 150 W input to the r.f. amplifier is 75 W, and the modulator efficiency can be taken as 50%. Thus the d.c. input to the modulator is 150 W, giving a total d.c. input of 300 W for 400 W p.e.p. output.

With an average modulation depth of 20–30% on speech, the d.c. input to the modulator is about 90–100 W (assuming a static feed of 0·1 A for the two modulator valves). The d.c. into the r.f. amplifiers remains the same, at 150 W, so the *average* total d.c. input is 240–250 W, for a peak power in each sideband of only 25 W.

Drive power for s.s.b. operation

Referring to the load line on the valve characteristics shown in Fig. 12.3, the bias is −50 V and a grid voltage of −2 V is needed to give I_A of 1·235 A at an anode voltage of 120 V. The required grid swing is 48 V and there will be no grid current, so the valve itself will require no drive power.

However, the 75 Ω cable must be matched correctly, and if terminated directly with a 75 Ω resistor a drive power of 15 W is required.

In cases where an amateur wishes to add an amplifier to increase the output power, terminating the feeder with a 75 Ω resistor directly across the valve input

is the obvious choice (Fig. 12.4). It ensures that the feeder is correctly matched at all frequencies and provides sufficient damping on the input to give a very stable operating condition without neutralizing. Note that all the drive power is dissipated in the terminating resistor.

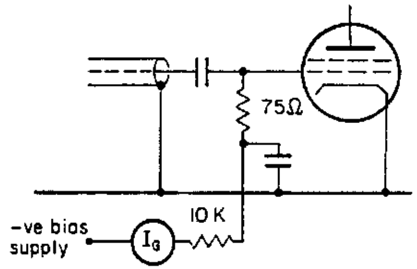


FIG. 12.4 Input circuit arrangement when high-power drive is available.

If only a low-level drive is available, a voltage step-up transformer must be used, and this can be either a wideband arrangement with a ferrite core, or separate transformers with tuned secondaries. In either case, the feeder must be terminated by a high-value resistor on the secondary, the actual value depending on the step-up ratio. If a preferred value of resistor is used, say 6800 Ω, the required transformer ratio is 9·5 to 1; $\sqrt{(6800/75)}$. Neglecting circuit losses, this means that the necessary drive power will be only 0·17 W p.e.p.

12.2 TYPICAL CIRCUIT DIAGRAM FOR AN R.F. AMPLIFIER FOR 200 W P.E.P.

A typical circuit arrangement for a 200 W p.e.p. amplifier with a tetrode valve is shown in Fig. 12.5. To cover the five amateur bands between 3·5 MHz and

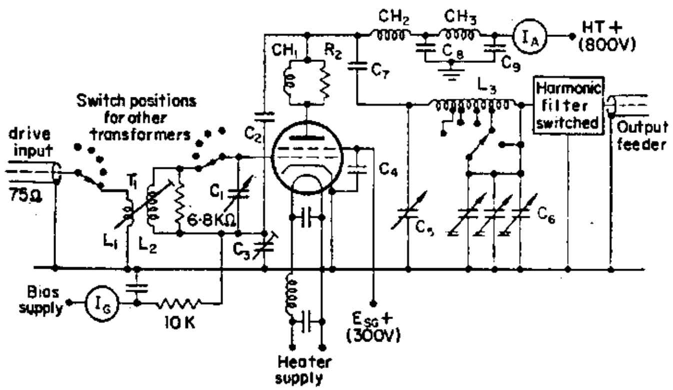


FIG. 12.5 Typical circuit diagram for r.f. amplifier.

28·7 MHz, the output circuit is switched in ranges, but separate transformers are used for each band on the input circuit.

The input circuit

It is theoretically possible to use a single input transformer with tapping points for both primary and secondary to cover the five ranges. However, practical difficulties arise in obtaining adequate mutual inductance between primary and secondary to produce a good match on the incoming feeder. Furthermore, separate transformers are an advantage if the available drive power differs between bands, enabling individual terminating resistors and different step-up ratios to be used for each band.

Wideband transformers are a special case, so consideration is limited to individual band transformers with tuned secondaries, although only one transformer is shown on the diagram, with a secondary resistor of 6·8 kΩ.

The drive is fed into the primary L_1 of the step-up transformer T_1, the secondary of which, L_2, is tuned by the variable capacitor C_1.

The negative bias is fed to the grid via a low-reading grid-current milliammeter and a 10 kΩ resistor, which is to prevent the grid current rising too rapidly. Note that there is no grid choke. If a choke were to be fitted at the high potential end it would ruin the feeder termination, and if fitted lower down it would tend to cause instability. For full output the peak grid voltage approaches zero (typically −2 V), so a reading on the meter shows that this voltage is being exceeded at peaks, and gives an indication of the required peak drive for full power output.

Neutralizing and stabilizing arrangements

The low-potential end of the transformer secondary is connected to earth via capacitor C_3, which is in series with the input capacitance of the valve across the winding. This provides a point on the input circuit where the r.f. voltage is in antiphase with the driving voltage, which is the correct phase for neutralizing directly from the valve anode, via capacitor C_2. Although the interelectrode capacitance between anode and control grid is quite low with this valve type (less than 0·9 pF), neutralizing is necessary to reduce the positive feedback, which is ruinous to linearity, even if not required to prevent self-oscillation at the fundamental frequency.

If capacitor C_2 were the variable element of the neutralizing path, it would have to be a very low-capacitance type with a high voltage rating and would be quite tricky to adjust, because neither end is earthed. By making C_3 the variable element these objections are overcome, and the variable capacitor is much less expensive.

The network consisting of choke CH_1 and resistor R_2 is an anti-spurious device to prevent self-oscillation at much higher frequencies. It is usually made by winding two or three turns round a 1 W resistor but insulated from it. The inductance of the choke must be quite low, otherwise the fundamental voltage across the network will produce excessive dissipation in the resistor. It may be necessary to use a resistor of a higher dissipation rating to avoid excessive heating, or even burn out. The optimum value of resistor is that which will give a network Q factor of unity, i.e., the resistance being equal to the reactance of the choke at the most likely spurious frequency. In this arrangement the spurious frequency can be considered as being in the region of 90 MHz.

It should be noted that choke CH_1 has to carry the r.f. current through the anode-screen-earth capacitance of the valve, which is about 15 pF including strays. The reactance of 15 pF at 28·7 MHz is 370 Ω. As the peak anode voltage is 680 V, the r.m.s. current through the choke will be about 0·75 A on the two-tone test at 28·7 MHz, but less at lower frequencies. The choke also has to carry the d.c. current of 0·25 A under the same conditions.

An important feature for the prevention of a spurious oscillation in the region of 90 MHz is the method of connecting the capacitor C_4, which is an r.f. by-pass between screen grid and cathode. The capacitance value is not critical, 0·002 μF being typical, but the capacitor must be a low-inductance type and the connecting leads must be as short and wide as possible, in order to reduce the inductance of the by-pass path. In this type of oscillation the screen behaves as an anode, hence the importance of reducing the inductance of the by-pass path.

An alternative approach is to make C_4 variable and to tune it to series resonance with the lead inductance at the spurious frequency, providing zero reactance between screen and cathode. Once set, this capacitor need not be readjusted, because this circuit is unaffected by switching between the various frequency bands.

The anode-output circuit

The anode-output circuit is a simple Π arrangement. Capacitor C_5 is the input shunt arm for tuning, tapped inductor L_3 is the series element and capacitor C_6 is the output shunt arm for controlling the valve loading. Capacitor C_6 consists of three variable capacitors in parallel ganged together, because coupling directly from a Π circuit into a feeder requires a very large capacitance for 3·5 MHz. Due to the large capacitance of C_6 necessary at the lower frequencies, it is preferable to use a 75 Ω output feeder, for which the actual value will be about two-thirds of that required for a 50 Ω feeder.

As in normal amateur operation there is only a single channel, good linearity is not required to avoid interchannel cross-talk. It is necessary for good quality, and especially to reduce harmonic radiation. A single Π circuit does not give adequate harmonic attenuation, hence the tuned harmonic filter in the output feeder, if necessary band-switched with the input and anode circuits. Particular attention should be given to harmonics falling within television bands, because T.V. interference is frequently—but often erroneously—attributed to amateurs. For this band it is not difficult to achieve an attenuation of about 70 dB with a relatively simple filter.

The chokes CH_2 and CH_3 in combination with capacitors C_8 and C_9 form a filter network to keep the r.f. out of the anode d.c. meter and the h.t. feed line. Capacitor C_7 is a blocking capacitor to isolate the d.c. anode supply from the Π circuit, enabling lower-voltage components to be used. However, it does mean that choke CH_2 is difficult to design, because the full r.f. voltage at the valve anode is applied across it at all operating frequencies. Consequently, considerable care must be taken to avoid resonance within the choke, usually by using a choke wound in sections of different sizes. It should not be forgotten that blocking capacitor C_7 has to carry the total r.f. current into the Π circuit. This will be of the order of 5 A r.m.s. at 28 MHz for a 200 W p.e.p. amplifier on the two-tone test.

12.3 DESIGN OF THE INPUT CIRCUIT

Features of the input circuit

The main problem in the design of the input transformer is to obtain both the correct voltage step-up ratio and a good match for the 75 Ω incoming drive cable, at each of the five frequency bands between 3·5 MHz and 28·7 MHz. In this respect the relatively narrow frequency range in each band is an advantage, because it enables a mid-band setting to be suitable for each band.

Using the circuit shown in Fig. 12.5 the coupling between primary and secondary must be very tight to obtain a resistive termination on the incoming feeder. Here a ferrite core is an advantage, but quite a good match can be obtained with an air-core transformer.

The square root of the primary–secondary resistance ratio determines the required step-up voltage ratio, which is also given approximately by the primary–secondary turns ratio. The number of primary turns for any range is approximately the number of secondary turns divided by the square root of the resistance ratio, but the optimum primary turns must be determined experimentally.

In order to calibrate the transformers, the grid-current meter is connected directly to earth instead of via the negative bias. This is a test condition by which the valve itself and the grid-current meter behave as a peak voltmeter, enabling the circuit to be set up on all frequencies, with a signal generator at the input instead of the drive. The only stipulation is that the signal generator has a 75 Ω output.

Note that the input capacitance C_V of the valve is in series with capacitor C_3 (which is the lower end of the neutralizing capacitance potentiometer C_2C_3) across the secondary. As the value of C_3 is likely to be at least twenty times the input capacitance C_V, for practical purposes it can be considered that the valve capacitance is directly across the secondary.

Input transformer design and calculation

Numerous transformer configurations can be considered, but there are certain features which are applicable to all designs.

(*a*) The turns must be readily adjustable and tapping points may be required. This means the use of bare wire and imposes a minimum limit on wire gauge and spacing between turns.

(*b*) The lead lengths from the transformer, including the switches, should be as short as possible. This indicates that the transformer must not be too large either in length or diameter, imposing an upper limit on wire gauge and turn spacing.

(*c*) In order to limit the transformer size, it is an advantage to use a high-value tuning capacitor. On the other hand, if the maximum capacitance is high, the minimum capacitance will also be high, so very low inductance values will be required for the high-frequency ranges and adjustment will be very critical.

(*d*) The whole input circuit must be well screened from the anode circuit because a tuned-grid tuned-anode arrangement is very prone to self-oscillation with very little coupling between them.

The overall size of the transformer can be reduced by using a ferrite core with the winding directly on to the ferrite. However, this tends to make the primary adjustment very critical, especially for the higher-frequency ranges, so an air-cored transformer is used in the example shown below.

Consider a design to cover the frequency bands of 3·5–3·8 MHz, 7·0–7·1 MHz, 14·0–14·4 MHz, 21·0–21·4 MHz and 28·0–28·7 MHz, with a variable capacitor of 300 pF maximum and 30 pF minimum.

The input capacitance of the tetrode type CV08-100 is 30 pF, so the maximum capacitance available is 330 pF to tune to the lowest frequency of 3·5 MHz. With an allowance for tuning, consider a capacitance of 300 pF for determining the maximum inductance of the transformer secondary.

For the highest frequency of 28·7 MHz, the minimum possible capacitance is the sum of the capacitor minimum and the valve capacitance, i.e., $30 + 30 = 60$ pF. With allowance for tuning, work out the inductance required for the smallest secondary, on the basis of tuning with 70 pF.

Assuming the secondary is shunted by a resistor of 6·8 kΩ, for the input to match a 75 Ω feeder, the transformer ratio is $\sqrt{(6800/75)} = 9{\cdot}5$ to 1. As the turns ratio between secondary and primary is approximately the same as the voltage ratio, the primary turns will be at about one-tenth of the secondary turns *on all ranges*.

Bearing in mind the previously mentioned features (*a*), (*b*) and (*c*), for this example it is considered that a reasonable compromise is given by winding the lowest-frequency transformer on a 1·25 in. diameter former, with 18-gauge tinned-copper wire at twelve turns per inch. By reference to Appendix V, or by some other method, it will be found that the required maximum secondary inductance of 6·9 μH is given by twenty turns, giving a basic transformer size of approximately 1·3 in. diameter, 1·75 in. long. A similar former would be suitable for the 7 MHz band, or tapping points on the 3·5 MHz transformer may be used.

The next step is to work out the inductance required for the other frequency bands, starting with the highest frequency.

The conditions for the lower- and upper-frequency limiting bands are given in Table 12.2, together with the conditions for the intermediate bands, selected on the basis of a linear frequency/Q factor characteristic. For frequency bands above 7 MHz it will be necessary to use formers of smaller diameter, otherwise the primary will be a small fraction of one turn, and adequate mutual coupling between primary and secondary will be impossible. For these transformers of small diameter, reference should again be made to Appendix V.

TABLE 12.2

Frequency band, MHz	3·5–3·8	28–28·7	7–7·1	14–14·4	21–21·4
Frequency for calculation, MHz	3·5	28·7	7	14	21
Secondary, pF	300	70	162	100	80
Secondary reactance, Ω	145	80	140	114	95
Q factor	46·7	85	48·6	60	71·5
Variable capacitance, pF	270	40	132	70	50
Secondary inductance, μH	6·9	0·44	3·18	1·3	0·72
Secondary turns	20	—	11	—	4
Primary turns (approx.)	2	—	1·125	—	—

Matching the input transformer to the feeder

This adjustment is made with the aid f a signal generator having a 75 Ω output of about 1 V. A low-reading r.f. voltmeter will also be required if one is not fitted to the instrument. Set the frequency required, check the output voltage into a 75 Ω resistor at the end of a 75 Ω cable just long enough to connect (subsequently) into the socket for the drive input, call this voltage V_1. With the grid circuit arranged for testing and the heater, but no other supplies, on, remove the test resistor and connect the signal generator output into the drive socket, leaving the peak voltmeter connected.

Tune the secondary to a maximum indication on the grid-current meter. If the transformer is correctly matched from the 6·8 kΩ secondary to 75 Ω primary, the input voltage V_1 will be the same as with the test resistor, and the grid-current meter will read about 0·95 mA (0·95 mA in 10 kΩ = 9·5 V). If voltage V_1 is higher or lower than that when feeding the test resistor, the primary for that frequency band must be adjusted in steps until the correct impedance match is obtained. Tapping points might be used with advantage.

On completion of the matching process, it is obviously necessary to remove the earth connection from the grid meter and restore the bias circuit to normal.

12.4 DESIGN OF THE ANODE-OUTPUT CIRCUIT

Description of the Π network

The Π network shown in Fig. 12.5 is a very convenient arrangement for coupling the valve output to the feeder. The two variable capacitors C_5 and C_6 control the tuning and coupling, respectively, and band changing is provided by switching taps on inductor L_3. A simplified version of this circuit is shown in Fig. 12.6. By means of the network, the feeder impedance R_F is converted to a

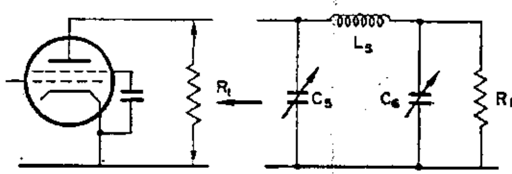


FIG. 12.6 Basic Π circuit for anode-output coupling.

non-reactive load R_1 across the valve, the value of which is selected to give the required operating conditions on the valve. The feeder impedance need not be purely resistive, the circuit will give a non-reactive load on the valve with a complex feeder impedance, but a v.s.w.r. limit is usually imposed of about 2 to 1.

It is important to note that the value of R_1 is *not* the resistance of the valve load line, but is derived from the output power and the peak r.f. voltage excursion ($R_1 = PV^2/2$ W). In the case of a single valve operating in class B, R_1 *is almost exactly twice the load-line resistance.*

A more detailed explanation of the function of the Π circuit is given in Fig. 12.7, with appropriate formulae in terms of capacitative and inductive reactance.

Obviously, the actual values of capacitance and inductance must be calculated for the operating frequency. Working back from the feeder, shunt components R_F and C_6 are converted to their equivalent series components r_2 and C_5. An inductor, L_{32}, having a reactance of the same numerical value as the reactance of C_5 is connected in series with r_2 and C_5, leaving r_2 only. The next step is to provide a series inductor and shunt capacitors of appropriate reactances to convert resistor r_2 to a pure resistance R_1 across the valve, of the correct resistance value. The ratio R_1/r_2 is the starting-point from which the reactance of the series inductor L_{31} can be calculated. Series components r_2 and L_{31} are then converted to their equivalent shunt components R_1 and L_p. Finally, shunt capacitor C_5 is added, the reactance of which is numerically equal to the reactance of L_p,

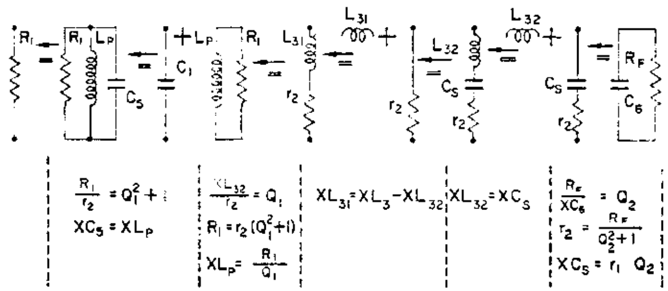


FIG. 12.7 Breakdown of the Π circuit.

thereby tuning out the reactance and leaving the resistive load R_1 on the valve anode.

Two points are worth noting while considering the breakdown of the Π circuit. First, the division of inductor L_3 into two sections L_{31} and L_{32} is a device for calculation only. In practice, the point of division is determined by the values of C_5 and C_6.

Second, at this division point the circuit is purely resistive of value r_2. This means that the second part of the network, including the feeder impedance, looks like the pure resistance r_2. Consequently the r.f. current in this resistor is the same as the current in the second part of the inductor, so it is the same as in the whole inductor. Also, the output power effectively appears in this resistor, so it is a simple matter to determine the r.f. current in the series inductor. For example, if the value of r_2 is calculated to be 4 Ω and the mean power output is 100 W, the current in the series inductor L_5 is 5 A r.m.s. ($I = \sqrt{(100/4)} = \sqrt{25} = 5$). This means that the gauge of wire used for the inductor must be sufficiently large to carry this current without excessive loss and consequent overheating. For 5 A r.m.s. at 28 MHz it would be advisable to use tinned-copper wire of not less than 10 gauge (0·125 in. diameter approx.).

Typical design for an amplifier giving 200 W p.e.p.

In Section 12.1 it was shown that tetrode type QV08-100 is suitable for use as a linear amplifier with an output power of 210 W p.e.p.

From the characteristics given in Fig. 12.3 at a screen voltage of 300 V and a d.c. anode supply of 800 V, the selected load line operated from 0·1 A at 800 V to 1·235 A at 120 V. Thus the anode voltage swing was 680 V and the load-line resistance 600 Ω (680/1·135). However, the effective r.f. load resistor R_1 is obtained from the power output of 210 W and the peak anode voltage excursion, i.e., $R_1 = 680^2/2(210\ \text{W}) = 1100\ \Omega$.

Having determined R_1, the next step is to work out the component values for the highest-frequency band, in conjunction with the probable values required for the lowest-frequency band. It is necessary to consider the two extreme frequencies together, because of the conflicting requirements. At the lowest frequency the output capacitance C_6 (Fig. 12.5) has not only to cover the value required by a resistive feeder, but also that due to an inductive type of mismatch; hence a large value is required, and in consequence C_5 must also be relatively large. If C_5 maximum is large, its minimum will be greater than it would be if its maximum value were lower. As the frequency range to be covered is 8·2 to 1, a capacitance value of 300 pF maximum for C_5 is a reasonable compromise.

Example of determination of component values

By using the formulae given in Fig. 12.7, component values for a 210 W p.e.p. amplifier covering the five amateur bands from 3·5 MHz to 28·7 MHz are shown in Table 12.3. The basic information in the calculations is given below.

(1) The valve considered was tetrode type QV08-100, with the dynamic load line as drawn on Fig. 12.3, giving an effective r.f. resistance R_1 of 1100 Ω.

(2) The valve output capacitance is normally 13 pF, but allowing for strays on the anode circuit, the minimum was considered to be 25 pF.

(3) The variable tuning capacitor had a range of 30–300 pF, so the minimum capacitance total was 55 pF. With some allowance for tuning, 70 pF was considered to be the resonant capacitance at 28·7 MHz.

(4) With the same variable capacitor, the maximum possible total capacitance was 325 pF, so 300 pF was taken as that required for resonance at 3·5 MHz.

(5) Having determined the limiting conditions for the two extreme frequencies, the values selected for the intermediate bands was on the basis of a linear Q factor/frequency characteristic.

In Table 12.3 it will be noted that the r.f. current in the Π circuit inductor has been calculated in order to determine the size of wire with which to wind the solenoid. Although tinned-copper wire has a high r.f. resistance compared with bare copper, at the current values indicated 0·125 in. o.d. tinned copper wire can be used without undue temperature rise.

The turns shown are based on an inductor of 2·5 in. diameter, wound with 0·125 in. tinned-copper wire at four turns per inch. Preferably, the inductor former should be of the skeleton type, in order to simplify the tapping point arrangement and to provide better cooling.

It should be pointed out that the v.s.w.r. of 2 to 1 was selected to show its effect on the value of the output capacitor, particularly at lower frequencies. If such a condition should arise in practice, it would be far more advantageous to improve the feeder match.

Having calculated the component values, it is just as important to confirm that they are realized in the amplifier. A very useful method is to obtain a capacitance calibration for one of the variable capacitors, such as that used for tuning the Π circuit. With the aid of a grid dip oscillator this enables the inductors to be set to the required values and the valve output capacitance to be measured *in situ*, etc., when setting up the circuit. Furthermore, it provides a means of checking that the tuning and loading conditions calculated are achieved in practice.

TABLE 12.3

Frequency band, MHz	3·5–3·8	28–28·7	7–7·1	14–14·4	21–21·4
Frequency for calculation, MHz	3·5	28·7	7·0	14·0	21·0
Π input capacitance (tuning), pF	300	70	171	134·5	82·5
Effective series resistance, r_2, Ω	20·2	5·86	16·0	10·9	7·6
Series inductance, μH	8·25	0·55	3·70	1·56	0·86
Π output capacitance (loading) pF, for $R_F = 75\ \Omega$	1000	255	583	416	300
Range of output capacitance for 2:1 v.s.w.r., pF	530–1470	198–312	350–816	300–532	224–378
Current in inductor at 100 W r.m.s. (200 W. p.e.p.) amps, r.m.s.	4·125	2·23	2·5	3·03	3·63
Inductor turns (approx.)	17	2·25	8·5	5	3

The method of setting the tapping points on the Π inductor is indicated by Fig. 12.8. The output-loading capacitor is short circuited and the inductor taps

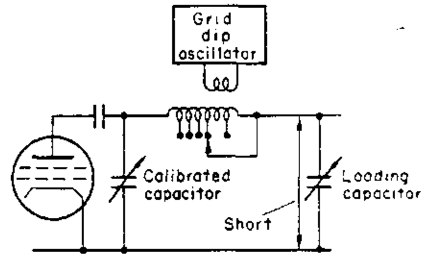


FIG. 12.8 Method of calibrating inductor for Π circuit.

are adjusted to give the required values to resonate with the known capacitance of the tuning capacitor. Resonance is indicated in the grid dip oscillator. This method enables all the strays to be taken into account when setting the tapping points, without the amplifier being switched on.

12.5 SEND-RECEIVE SWITCHING

A feature of amateur operation is the need for a quick changeover between send and receive. Normally this is accomplished by means of two relays, one of which changes the antenna between transmitter and receiver, while the other disconnects the transmitter screen supply when receiving. Even with the screen supply removed, the white noise (wideband thermal noise) produced by the transmitter valve and input circuits is relatively high compared with the received signal level. Also, the attenuation of the antenna changeover relay is not infinite, so some of the transmitter noise appears on the antenna in the receive position. The result is a background of 'mush' on the received signal, which is often attributed to atmospherics picked up by the antenna.

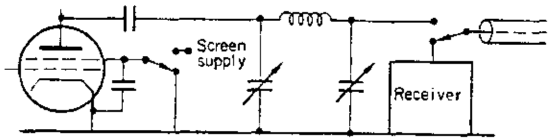


FIG. 12.9 Send–receive switching with relays in the receive position.

This breakthrough of transmitter noise into the antenna can be attenuated to a negligible level by using a back contact on the screen supply relay, to earth the screen during reception, as shown in Fig. 12.9. It has been found in practice that the earthed screen produces a remarkable reduction in background 'mush', enabling very low-level signals to be received with clarity.

13

Solid-State Amplifiers

13.1 THE PRESENT STATE OF THE ART

The progress in solid-state devices continues unabated, but at the present time their application to r.f. power amplification is limited to low-power levels. In parallel with this progress, manufacturers of solid-state devices issue application reports at the earliest possible opportunity, in order to keep prospective customers up to date with their products. In consequence, it is considered that a comprehensive review of low-power solid-state amplifiers would not only be superfluous, but most probably out of date by the time this book is published. Therefore, consideration is restricted to certain similarities and differences between transistors and valves when used as r.f. power amplifiers.

Transistors are low-impedance devices with characteristics which make them particularly suitable for on–off applications. As such, they are more appropriate for non-linear amplifiers. Another characteristic feature of transistors is their susceptibility to temperature changes, tne performance falling with rising temperature. This is a further point in favour of non-linear amplification with transistors, which is more efficient than linear amplification, so the power dissipated within the transistor is less for the same power output, and the temperature rise is less.

It follows that as solid-state devices with higher-power rating become available—which undoubtedly they will—the increase in r.f. power output from transistor amplifiers will be most rapid in applications requiring non-linear amplification. It also follows that there is every incentive towards high conversion efficiency if the maximum possible output is to be obtained. For example, a transistor having a dissipation rating of 10 W will give an r.f. output of 40 W at 80% efficiency, but 90 W at 90% efficiency, which is an *increase* of 125% in power output for an efficiency improvement of only 10%.

While considering internal dissipation, transistors have a definite advantage over valves in that no heater supply is required. This will be even more advantageous for transistors having a higher power rating.

13.2 NOTES ON NON-LINEAR TRANSISTOR AMPLIFIERS

Conversion efficiency

Operation of the active device in class C is the most usual configuration for non-linear amplifiers, giving a reasonably high conversion efficiency. The basic circuit for a class C transistor amplifier is shown in Fig. 13.1, in which Π circuits are used for input and output because of the low impedances involved. With silicon transistors, an external bias is not normally required because bias is provided automatically by the internal voltage V_{be}.

With this simple circuit, a transistor amplifier is not conducive to high conversion efficiency. In order to use the transistor type efficiently, the cut-off

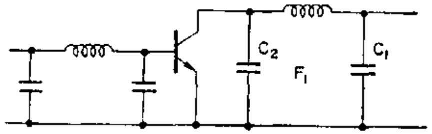


FIG. 13.1 Basic circuit for transistor amplifier, class C.

frequency F_T will not usually be much higher than the operating frequency F_1, and the effect cannot be neglected. The effect is a delay between the drive waveform V_b and the output waveform $V_c(1)$, as shown in Fig. 13.2.

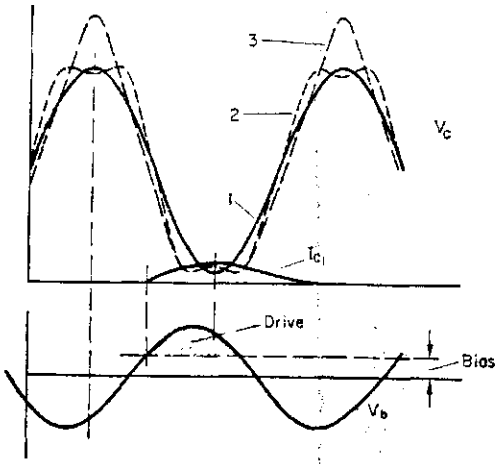


FIG. 13.2 Input and output waveforms of transistor amplifiers.

When transistors are driven to saturation, another difficulty arises due to the storage charge causing a spread in the length of the current pulses, as shown by I_{c1}. It is obvious that current flowing on either side of the voltage trough will cause poor efficiency.

In order to improve the conversion efficiency, consider the application of the principle used by Tyler [1] for high-efficiency valve amplifiers (class D), as shown in Fig. 13.3. It is seen that the output contains an additional circuit which is resonant at the third harmonic of the fundamental frequency. When tuned correctly the collector waveform will be shown by $V_c(2)$, Fig. 13.2, giving an obvious improvement in efficiency compared with the class C waveform $V_c(1)$.

This arrangement differs from the high-efficiency valve amplifier, in that pulse broadening of the input waveform is not necessary (for an explanation of the function of the third harmonic circuit as a means of improving efficiency, refer to Chapter 3, Section 3.5).

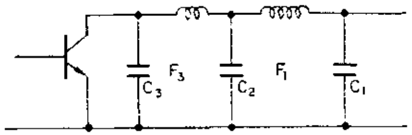


FIG. 13.3 Circuit for high-efficiency operation.

With transistor amplifiers, as the frequency increases the need for the third harmonic resonator diminishes. As a result of the low working impedance, stray series inductance (lead length) and relatively small collector capacitance (Fig. 13.4), the collector waveform produced is as shown by $V_c(3)$, Fig. 13.2. This waveform has the flat trough necessary for high conversion efficiency. The peak overswing does not appear to have a detrimental effect on the transistor, even when the nominal breakdown voltage is exceeded, provided that the operating

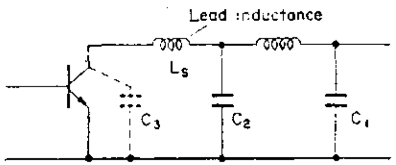


FIG. 13.4 Effective output circuit at higher frequencies.

frequency is high. The result is that many transistors designed for high-frequency operation can produce more output power with safety when operating at higher frequencies. This appears contrary to straightforward reasoning, and accounts for conversion efficiencies higher than expected from class C operation.

Other circuits have been devised for obtaining conversion efficiencies which approach 100%, based on rectangular waveforms produced by switching transistors on and off. Practical difficulties do arise with such arrangements, not the least being the correct timing for switching on and off, but it is considered that this line of attack will be used as a means of obtaining higher r.f. power from non-linear transistor amplifiers

Parasitic oscillations

Transistor amplifiers are subject to a type of parasitic oscillation not normally present in valve amplifiers. These are parametric oscillations due to the non-linear characteristic of the collector-base capacitance. Consider the frequency spectrum shown in Fig. 13.5, in which the fundamental operating frequency is F_1. If the collector-base capacitance is sufficiently non-linear, with high-amplitude signals, an output will be produced parametrically at two other frequencies

$(F_1 - F_x)$ and $(F_1 + F_x)$, in the presence of power at the frequency F_x. The ratio of powers will be proportional to the frequency ratio. If the output circuit is tuned to accept the difference frequency $(F_1 - F_x)$, then the input impedance is

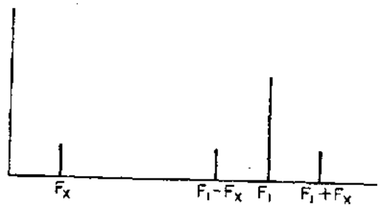


FIG. 13.5 Spectrum showing low-frequency parametric oscillation due to non-linear collector-base capacitance.

negative at frequency F_x and parasitic oscillation will be present, without the application of power at frequency F_x, if the circuit is not damped sufficiently at this low frequency, F_x.

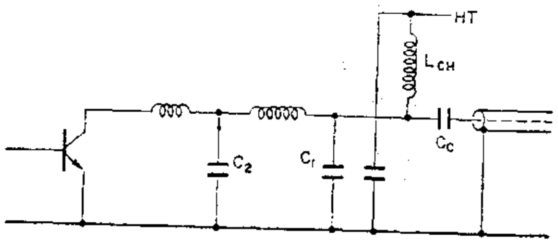


FIG. 13.6 Practical circuit for producing spurious parametric oscillation.

Consider the practical circuit shown in Fig. 13.6, where the h.t. supply is fed via the choke L and the output is coupled to the feeder via capacitor C_c. If the choke L_{CH} and capacitor C_c are not sufficiently large, and the resultant circuit

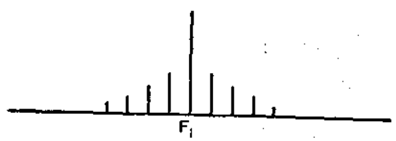


FIG. 13.7 Spectrum of frequencies produced by the circuit shown in Fig. 13.6.

not sufficiently lossy at both low r.f. and audio frequencies, parametric oscillations will be produced on either side of the fundamental frequency F_1. The frequency spectrum produced by this type of oscillation is shown in Fig. 13.7.

13.3 LINEAR AMPLIFICATION WITH TRANSISTORS

Transistor linearity compared with valves

Transistors are often considered to be suspect for linear amplification and inferior to valves, mainly due to comparisons being made on a non-realistic basis, particularly regarding power level.

For the low-power stages of linear transmitters, linearity is the overriding consideration and conversion efficiency is unimportant. Consequently, it is quite normal to use valves in class A with a rated dissipation of about 2 W for a few milliwatts output. Yet when transistors are used for a similar application, the same order of linearity is expected from models having a dissipation similar to the output. A more practical comparison can be made by using a non-linearity factor (n.l.f.), which is defined as the ratio of amplifier dissipation to p.e.p. output, for a given distortion level in terms of intermodulation products (i.p.'s). On this basis the following typical example shows that transistors can be superior to valves as linear amplifiers.

Consider a requirement for 2 W p.e.p. output from 2 MHz to 30 MHz with i.p.'s of -50 dB, from valves and transistors of suitable rating.

Valve n.l.f. = 30, excluding heater power.
Valve n.l.f. = 40, including heater power.
Transistor n.l.f. = 25.

Another point is that no useful purpose is achieved by comparing the theoretical r.f. linearity of valves having a 3/2 power law with that of transistors having an exponential law, because valves are voltage driven, whereas transistors are current driven.

Non-linearity in transistors

One of the most important factors contributing to non-linearity in transistors is the variation of cut-off frequency F_T in relation to the V_c/I_c characteristics. Before discussing this further, it might be advisable to explain the term cut-off frequency. The gain of a transistor is given by the ratio of the collector current

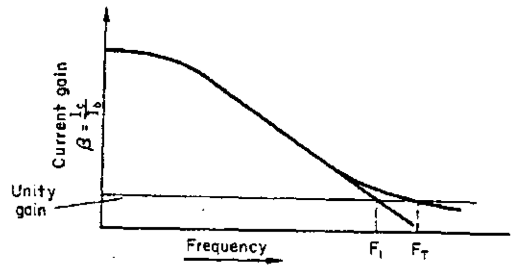


FIG. 13.8 Gain/frequency response for constant V_c.

I_c to the base current I_b, and is called β ($\beta = I_c/I_b$). With transistors the current gain decreases with rising frequency in the form of the curve shown in Fig. 13.8. The frequency at which the current gain falls to unity ($I_c = I_b$) is known as the cut-off frequency, F_T.

It should be noted that over an appreciable frequency range, the gain β drops linearly with frequency increase. If this straight portion is continued, the unity gain condition is reached at a frequency F_1, which is somewhat lower in frequency than F_T.

Frequency F_1 is often used by design engineers instead of F_T, because it enables the gain at frequencies along the straight portion to be estimated more readily.

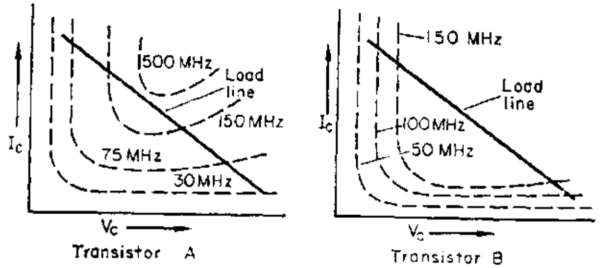


FIG. 13.9 Typical contours of cut-off frequency F_T relative to voltage V_c, for two transistors.

The variation of cut-off frequency in relation to the V_c/I_c characteristics for two different transistors is shown in Fig. 13.9, with typical load lines added. It is clear that the gain of an amplifier varies with current and voltage swing, thus giving rise to distortion. The effect is more pronounced at higher operating frequencies, so for linear amplification it is necessary to use transistors having cut-off frequencies very much higher than the required operating frequency.

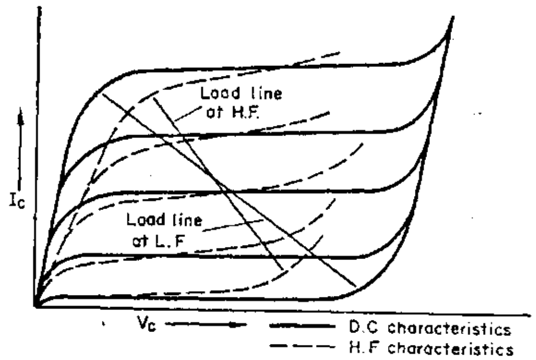


FIG. 13.10 Effect of increasing frequency on transistor I_c/V_c characteristics.

For this reason transistors having the characteristics similar to B in Fig. 13.9 are preferred to those with A characteristics.

Another way of showing this effect is given in Fig. 13.10, in which the solid line characteristics are at d.c. as given in the published data sheets. However, as the frequency increases, the effect of the cut-off frequency, i.e., gain reduction, causes the characteristics to tilt, as shown in broken lines. The result is a reduction

in voltage and current excursions at higher frequencies, with a consequent increase in distortion. The change in optimum load line is also shown.

Class B transistor amplifiers

It is quite practical to operate linear r.f. amplifiers with a single valve in class B. This is possible because of the high resistance of the load line, particularly with tetrodes, enabling circuits of high Q factor to be used to supply energy during the un-driven half-cycles.

On the other hand, transistors have very low-resistance load lines, typically 3 Ω for a 75 W transistor, so circuits of high Q factor are completely impractical. For linear r.f. amplifiers, transistors in class B must be operated in push-pull pairs in a similar manner to class B valves in a.f. amplifiers.

In order to obtain good linearity with class B amplifiers in push-pull and low Q factor circuits, it is important that the bias is set accurately. If the bias is too high, cross-over distortion will result; whereas if it is too low, the cross-over distortion will occur to a lesser extent, but excessive dissipation is highly probable.

Secondary breakdown

Maintaining the bias correctly under operating conditions presents quite a problem, because the required bias point changes with temperature, which in turn changes with operating power level. In fact, the static feed increases with rising temperature for a fixed-bias voltage, so the dissipation increases and the transistor can be destroyed by secondary breakdown. This is the term applied to an effect which causes silicon transistors to break down under d.c. conditions, at a dissipation level much lower than that which can be safely maintained under

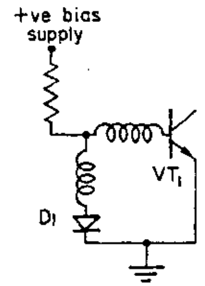


FIG. 13.11 Forward-biased diode for bias compensation.

r.f. conditions. A practical example will show the serious nature of this effect. Consider that a transistor, having a rated secondary breakdown at 8 W, is biased to dissipate 5 W with no r.f. signal. On application of r.f. the dissipation rises to 20 W and the silicon will get hot, but the transistor does not break down. On removal of the r.f. signal, the bias point will have changed and a d.c. dissipation of 10 W is likely. This is above the rated d.c. dissipation, so the transistor will be destroyed by secondary breakdown.

The simplest method of bias compensation is to use a forward-biased diode which is held at the same temperature as the transistor, and connected as shown in Fig. 13.11. The diode D_1 should be mounted adjacent to the transistor VT_1 on the same heat sink, so that the two silicon chips are at approximately the same temperature and have the same temperature characteristic. This method is more satisfactory if the diode and transistor are both on the same silicon chip.

An alternative method of compensation is to use a constant-current bias source together with a diode-connected transistor, by some arrangement such as shown in Fig. 13.12. This is more satisfactory for higher power. Some new power transistors have a diode incorporated on the same chip, but of a very small rating so that a d.c. amplifier must be incorporated.

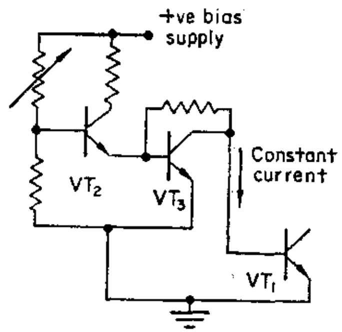


Fig. 13.12 Constant-current method of bias compensation.

Linear wideband r.f. amplifier, 2–30 MHz

Because transistors are low-impedance devices, they are more appropriate to wideband amplification than they are to narrow band tuned arrangements. However, in considering a bandwidth of several octaves, such as 2–30 MHz, some gain/frequency compensation must be applied to give a substantially flat frequency response, because of the change in gain (β) of transistors over the

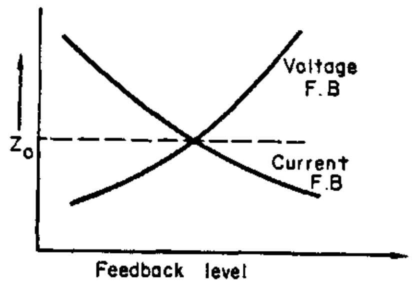


Fig. 13.13 Change in input impedance with increasing feedback level.

operating frequency band. Compensation can best be obtained with feedback, because feedback also improves linearity. Feedback also changes the impedance of transistors, thereby providing mismatch, with consequent degradation of linearity. It so happens that with increasing feedback level, voltage feedback increases the output impedance, but current feedback reduces it. By applying

voltage and current feedback in parallel, it is possible to obtain a substantially constant output impedance over a wide range of frequencies, even though the feedback level changes (Fig. 13.13).

Wideband circuit arrangement

A basic circuit arrangement for a class B wideband transistor amplifier is shown in Fig. 13.14. It will be seen that this incorporates a bias supply arranged to compensate for changes in bias point with rising temperature, together with both voltage and current feedback. This arrangement is quite satisfactory as a linear amplifier from 2 MHz to 30 MHz, with a suitable choice of component values. If more gain is required by two such stages in cascade, it is preferable to

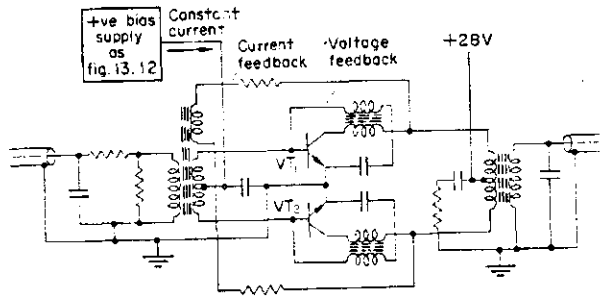


Fig. 13.14 Single-stage wideband class B r.f. amplifier.

avoid the use of a feeder by using transformer matching directly into the second-stage input.

With the very low impedance of transistors, the impedance of quite short leads can be troublesome and transformers must be designed with the lowest possible leakage inductance. Even so, as the power level is raised it becomes increasingly difficult to couple the two sides of the class B stage sufficiently tightly together. For higher powers it may be necessary to revert to partial tuning to overcome the side-to-side coupling problem, and effectively reduce the bandwidth. Experimentally it has been found that an improvement in linearity in terms of i.p.'s can be obtained with amplifiers slightly mistuned. This is due to cancellation, and is not consistent at all signal levels. It is not a suitable method to use for a production equipment.

REFERENCE

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

Appendix I

A Graphical Method of Harmonic Analysis

If often happens that a quick harmonic analysis is required in circumstances where speed is more important than a high order of accuracy. The following method is a ready means of determining the amplitude and phase relationships in a complex waveform, provided that the levels of the fifth and higher orders are relatively low. The accuracy obtainable depends upon the plotting accuracy and the relative levels of harmonics and fundamental, with higher harmonic levels giving a greater accuracy.

Before proceeding with the description however, it should be mentioned that the method is not new [1]. It was devised by J. Harrison many years ago and has since appeared in several textbooks, one of which is Castle's *Manual of Practical Mathematics*, published as long ago as 1920. However, a recent sampling (admittedly a small one) taken among young engineers and students seems to indicate that it has fallen into disuse—indeed, it was quite unknown to them.

DESCRIPTION OF THE METHOD

Perhaps the simplest way of describing the process of waveform analysis by this method is by practical illustration. Figure AI.1 is a plot of a complex waveform, the main constituents being fundamental, second harmonic and third

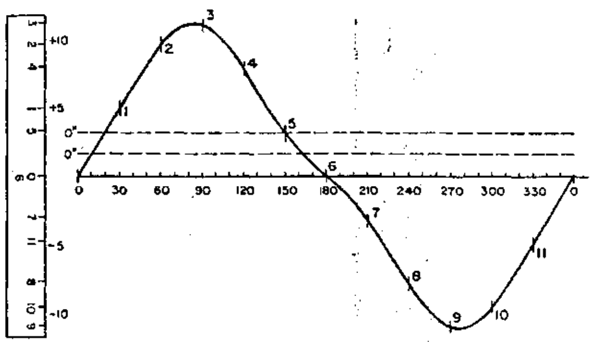


FIG. AI.1 A complex waveform with second and third harmonic content. The points identified numerically every 30° on the waveform are projected to determine the graduations on the paper strip.

harmonic. Points 0–11 are marked on the waveform at 30° intervals: 0 at 0°, 1 at 30°, 2 at 60°, etc. On a separate strip of paper, also shown in Fig. AI.1, the amplitude of points 0–11 are marked with respect to zero level. In practice, the paper strip is moved along in 30° steps, to obtain the best accuracy.

This marked strip is used to plot curves A, B and C in shown Fig. AI.2 and described later, of the constituent components of the waveform, by algebraic subtraction and addition of the amplitudes of the various points in accordance with Table AI.1.

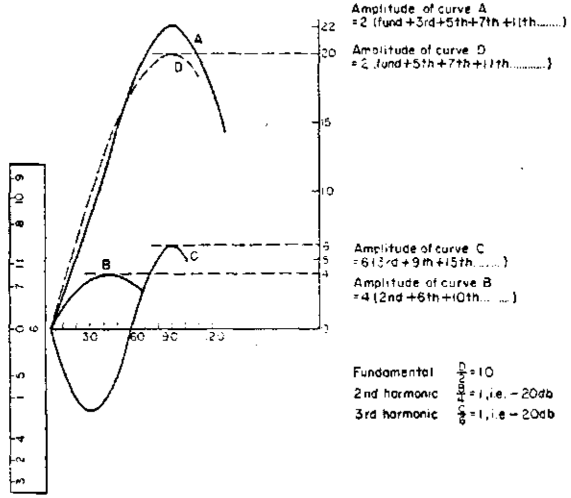


FIG. AI.2 The constituent components of Fig. AI.1 waveform plotted from the information given in Table AI.1. *Note:* For clarity subdivisions are not shown, but 10 on the linear amplitude scale represents twenty small divisions.

In Table AI.1 mathematical expressions representing component parts of a Fourier expansion are given for the three curves. As can be seen from these expressions, half the amplitude of curve A gives the peak amplitude of the fundamental, third harmonic, ignoring the fifth, seventh, etc. One-quarter of the amplitude of curve B gives the peak amplitude of the second harmonic, ignoring

TABLE AI.1

θ	A 2 (fund. + 3rd + 5th)	B 4 (2nd + 6th + 10th)	C 6 (3rd + 9th + 15th)
0°	0–6	(0–3) + (6–9)	(0–2) + (4–6) + (8–10)
30°	1–7	(1–4) + (7–10)	(1–3) + (5–7) + (9–11)
60°	2–8	(2–5) + (8–11)	(2–4) + (6–8) + (10–0)
90°	3–9	(3–6) + (9–0)	(3–5) + (7–9) + (11–1)
120°	4–10	(4–7) + (10–1)	(4–6) + (8–10) + (0–2)

the sixth, tenth, etc. One-sixth of the amplitude of curve C gives the peak amplitude of the third harmonic, ignoring the ninth, fifteenth, etc.

The peak amplitude of the fundamental is obtained by subtracting one-third of the amplitude of curve C from curve A, at every 10° point, giving curve D, which is twice the fundamental amplitude.

The relative phase of the three components of the original waveform is obvious from Fig. AI.2.

Because the peak amplitudes of curves A and C are coincident at 90° in the example, both relative phase and amplitude could be obtained without drawing curve D, but in the general case it is necessary to plot curve D for an accurate assessment.

PLOTTING THE CURVES

To plot the curves, the strip is reversed and points marked at 30° intervals, as shown in Fig. AI.3. The method of plotting the points for curves A, B and C is shown in Fig. AI.3 (*A*), (*B*) and (*C*), respectively. Points S.T.U. and V are points for curve A, with points L, M and N for curve B, and points G and H for curve C, all at 30° intervals.

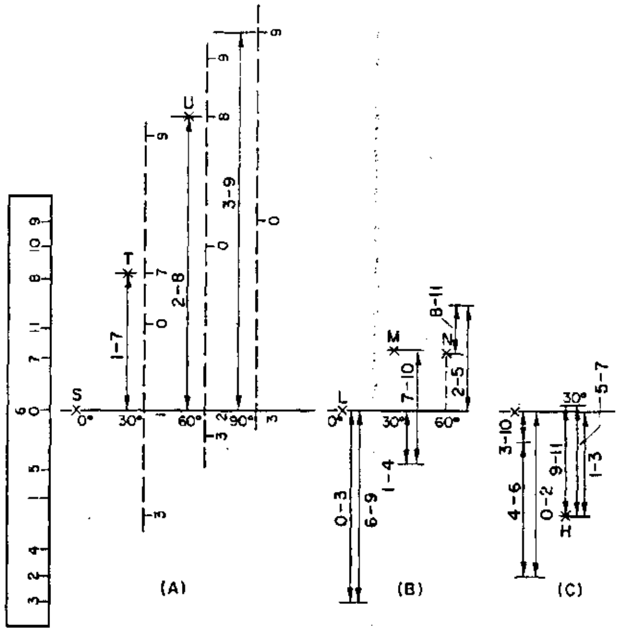


Fig. AI.3 The method of marking points at 30° intervals for curves A, B and C, in accordance with Table AI.1 and using the paper strip inverted.

Figure AI.3 is used to give an explanation of the method of using the inverted strip. The actual curves should be plotted as shown in Fig. AI.2, in order that the relative phase of harmonics and fundamentals can be determined.

From these points a rough approximation of curve A could be drawn, but they are quite inadequate to draw B and C.

Additional points can be obtained at 30° intervals, starting at 10° and 20°, by using two more paper strips. On these strips, points 0–11 are marked, as before, but with reference to new zero lines, 0′ and 0″ (see Fig. AI.1). Points on the 0′ strip are marked 0 at 10°, 1 at 40°, 2 at 70°, etc., giving points for curves A, B and C when inverted and transferred to Fig. AI.2, at 10°, 40°, etc.

Similarly, the other strip 0″, will give points at 20°, 50°, 80°, etc., when marked with reference to level 0″. It is seen that 0′ is the baseline drawn at the level of the complex waveform at 10°, and 0″ is the level at 20°.

ACCURACY

For a second harmonic level of 20 dB and the scale used (see Fig. AI.2, note), the limits of measurement are about half of one small division in eight small divisions (curve B), giving an accuracy of approximately 0·5 dB. Also, the limit of measurement of about one small division corresponds to a harmonic level of –40 dB (one-tenth of curve B or C). At this harmonic level half of one small division represents a tolerance of ±3 dB.

Table AI.2 gives the amplitude of curve D at every 10°, showing that in the example the plot of the fundamental does not depart from a sine wave by more than 2%.

Table AI.2

θ	Level from Fig. AI.2	Rationalized level	$\sin\theta$ level
0°	0	0	0
10°	3·5	0·175	0·1736
20°	6·7	0·335	0·342
30°	10·0	0·5	0·5
40°	12·6	0·63	0·6428
50°	15·0	0·75	0·766
60°	17·0	0·85	0·866
70°	18·5	0·925	0·9397
80°	19·5	0·975	0·9848
90°	20·0	1·0	1·0

It should be pointed out that these limits of accuracy assume that the plot of the original waveform is in itself correct within fine limits. In practical cases where the method of obtaining the original waveform does not enable the plot to be very accurate, the result will not be within such fine limits. For example, when the waveform is obtained from an oscilloscope, photographically or by tracing, the initial error is of the order of 5%. Thus the results obtained will be less accurate by this amount, giving total tolerances of approximately ±1 dB for harmonics 20 dB down and ±5 dB for harmonics 40 dB down.

APPLICATIONS

It is obvious that there are many applications for this method of harmonic analysis, particularly in making early assessments, where quick answers are of more immediate importance than extreme accuracy.

One example is the determination of the harmonic content to be expected from valves and transistors under various operating conditions, based on the published characteristics.

Another example is the measurement of the harmonic content in a coaxial feeder at v.h.f., obtained from a plot of the feeder voltage along a slotted line. In an application of this nature several precautions are necessary, but with a relatively high harmonic content and low v.s.w.r., the results obtained can be substantially correct.

REFERENCE

[1] Stokes, V. O. 'A graphical method of harmonic analysis'. *Wireless World*, 121–123 (March 1966).

Appendix II

The Self-Inductance of Single Straight Conductors of Circular Cross-Section

Although the size of conductors used for interconnections in high-power r.f. circuits is mainly dependent upon current-carrying capacity, in some instances it may be advantageous to select a size of conductor to provide some inductance in a connection which is part of an r.f. circuit. Such a connection may permit reduction in the size of an inductor or other variable element. In some cases the

size of conductor is chosen to reduce the lead inductance in a connection to a component. The extent of reduction possible with typical conductor sizes can be seen from Fig. AII.1. These curves are also useful in estimating the likely total inductance of connections, in order to determine the range of variable element necessary to cover a specified frequency range.

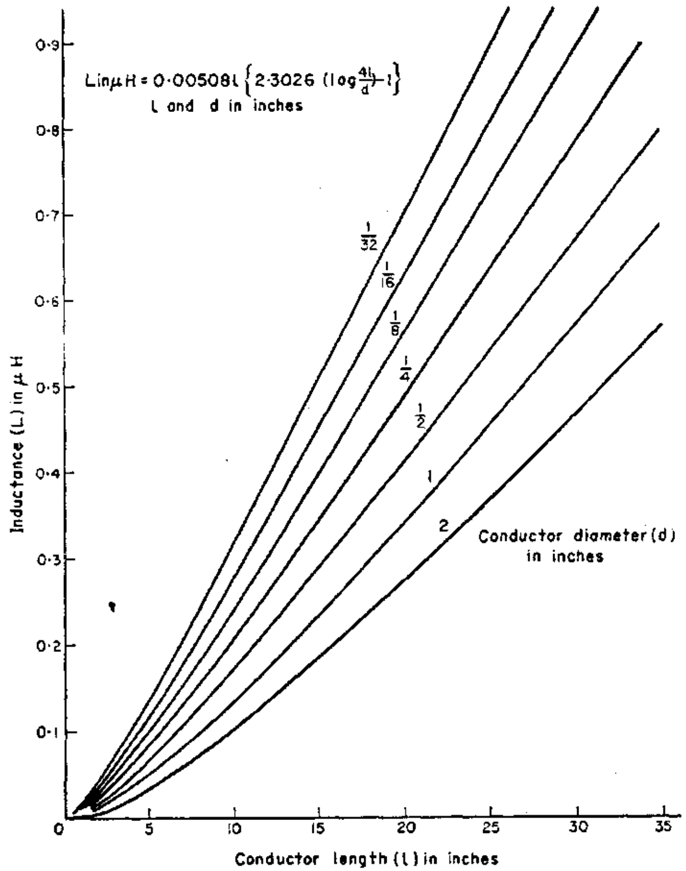


Fig. AII.1 Self-inductance of single straight conductors of circular cross-section.

Appendix III

Self- and Mutual Inductance of Turns of Large Diameter

For the large-diameter conductors necessary to carry the r.f. current and the low values of inductance required in the high-frequency ranges, general formulae for the inductance of multi-turn solenoids are insufficiently accurate. At the same time, these inductors are quite expensive items to manufacture, so any design should be substantially 'right first time'. However, inductors can be designed

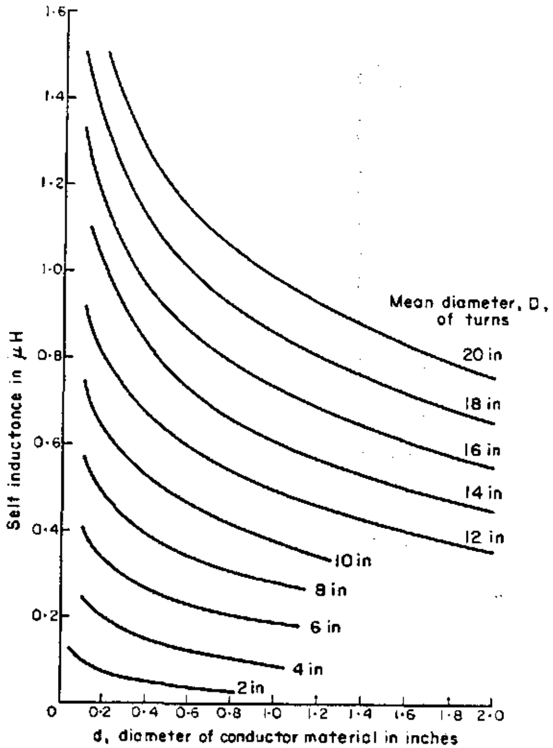


FIG. AIII.1 Self-inductance of single circular turns of circular cross-section material.

with sufficient accuracy based on the inductance value of each turn and the mutual inductances between the various turns. It is for this reason that Figs AIII.1 and AIII.2 have been reproduced, showing the self- and mutual inductance, respectively, of single turns with dimensions typical of those used for inductors in high-power applications.

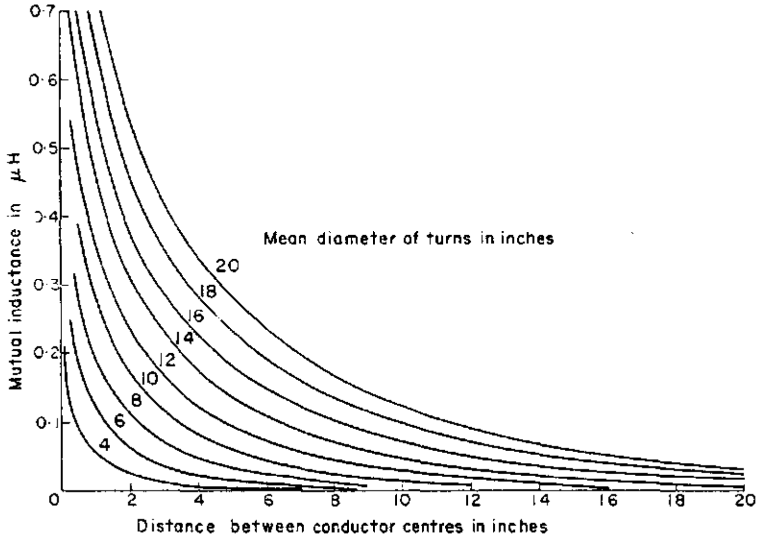


FIG. AIII.2 Mutual inductance of equal coaxial turns at various spacings.

The method of calculation based on Figs AIII.1 and AIII.2 is given below.

(1) With the material diameter, d, and the mean turn diameter, D, known, read off the self-inductance, L, of each turn from Fig. AIII.1.

(2) Assume a spacing between turns of some value, probably determined by voltage clearance, and read off Fig. AIII.2 the mutual inductance between adjacent turns M_1, alternate turns M_2, every third turn M_3, etc.

(3) If N is the total number of turns, then the total inductance is given by;

$$N \cdot L + (N-1)(2M_1) + (N-2)(2M_2) + (N-3)(2M_3) + \cdots +$$

As an example, consider a four-turn inductor, 10 in. mean diameter of 1·0 in. diameter material and 1·8 in. between the centres of adjacent turns. From Figs AIII.1 and AIII.2, $L = 0{\cdot}375\ \mu H$; M_1 (at 1·8 in.) $= 0{\cdot}19\ \mu H$; M_2 (at 3·6 in.) $= 0{\cdot}095\ \mu H$; M_3 (at 5·4 in.) $= 0{\cdot}055\ \mu H$; and total inductance $L = 4(0{\cdot}375) + 3(0{\cdot}38) + 2(0{\cdot}19) + 1(0{\cdot}11) = 3{\cdot}13\ \mu H$.

One of the advantages of the inductance being presented in this form is that it provides a ready means of determining the effect of either a change in turn diameter or a change in material diameter. An application of this type can be useful in controlling the angular position of inductor tapping points.

Appendix IV

Voltage Flashover

The voltage which can appear between two electrodes before flashover occurs is dependent on both the spacing between them and the diameter of the individual electrodes. In high-power applications the most usual concern is with parallel circular conductors. For this type of conductor Fig. AIV.1 shows the voltage at which flashover will occur between electrodes of various diameters and spacing at 0°C at sea-level, with correcting factors for higher temperatures and altitudes.

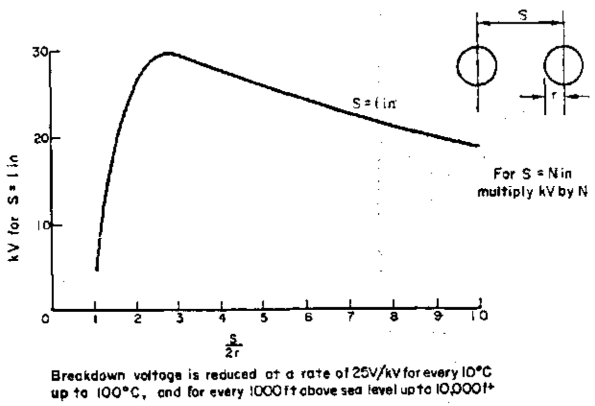


FIG. AIV.1 Flashover voltage between parallel circular conductors at 0°C at sea-level.

It is interesting to consider the effect of selecting ratios of $S/2r$ on either side of that giving the maximum voltage at 2·5. If a ratio of say, 2, is selected, as the voltage approaches breakdown surface ionization is likely to occur. This effectively reduces the ratio by a small amount, but the characteristic is very steep, so that any ionization is invariably followed by breakdown. On the other hand, if ionization occurs when $S/2r$ is greater than 2·5, the effective reduction of the ratio increases the breakdown voltage, and instead of breakdown it is more likely that the ionization will cease.

For the voltage breakdown between a circular conductor and a flat plate, reference should be made to Fig. AIV.2. First, consider an image of the circular conductor at a distance S equal to twice the distance between the centre of the conductor and the surface of the flat plate. Next, determine the breakdown

voltage which would occur between the conductor and its image from Fig. AIV.1. Then the effective breakdown voltage between the circular conductor and the plate is half the voltage which would produce flashover between the conductor and its image.

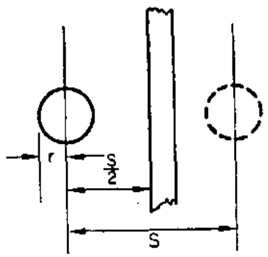
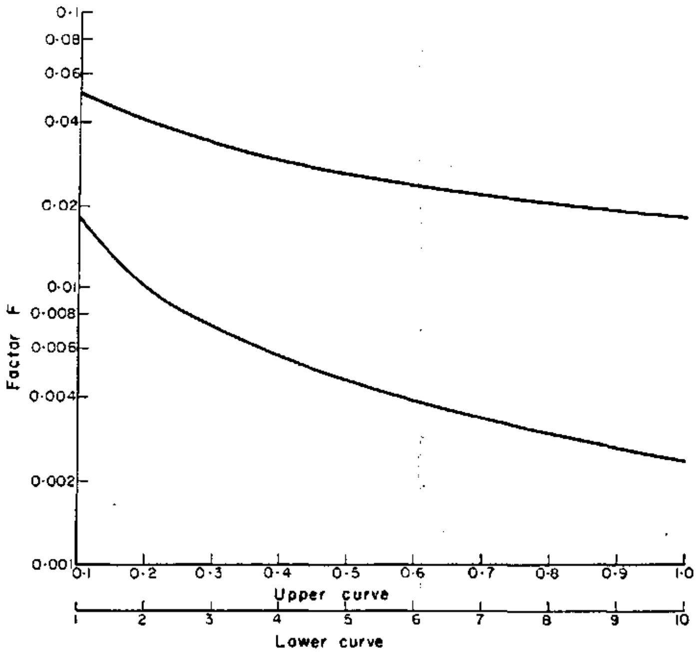


FIG. AIV.2 Flashover voltage between a circular conductor, or a flat plate with a radiused edge, and a flat plate.

Appendix V

Inductance of Single-Layer Solenoids

The curves given in Fig. AV.1 provide a ready means of calculating the inductance of single-layer inductors of constant diameter and pitch. The method is particularly applicable to solenoids of many turns of up to 2 in. mean diameter, so they will normally be wound with wire not exceeding 0·125 in. diameter. Consequently they are appropriate for low-power inductors and chokes.



Ratio of length/diameter (l/d)
Inductance in microhenrys = 'F'. d. N^2 where d = coil diameter in inches, N = total turns, 'F' from l/d ratio curve

FIG. AV.1 Inductance of single-layer solenoids.

For inductors of a small number of turns, wound with gauges greater than 0·125 in. diameter on large mean diameters, reference should be made to Appendix III.

Appendix VI

S.I. Units

The Eleventh Conférence Générale des Poids et Mesures in 1960 adopted the rationalized and coherent system of units linked to the Giorgi units, and this has been accepted by the Electrotechnical Commission (I.E.C.) and the International Organization for Standardization (I.S.O.). British industry, through the Industry Standards Committee of the B.S.I., has opted to go straight to the use of S.I. units when changing to a metric system for the United Kingdom in the 1970s.

The important feature of the S.I. system is that it is a coherent system of units which covers the majority of quantities needed in all technologies and is based on six fundamental units. A coherent system is one where the product or quotient of any two unit quantities in the system is the unit of the resultant quantity; a simple example of this is that unit area is the result of unit length multiplied by unit width.

BASIC UNITS

Quantity	Name	Symbol
Length	metre	m
Mass	kilogramme	kg
Time	second	s
Current	ampere	A
Temperature	kelvin	K
Luminous intensity	candela	cd

Note: 1K = 1°C, and 0K = −273°C.

Most radio engineers are familiar with the basic units, but some of the derived units may be difficult to assimilate. One of these is the new unit of force, the newton (N) which is expressed as kilogramme metre/s² ($N = kg\ m/s^2$). It is a coherent unit, in that the force applied to unit mass produces unit acceleration and is independent of gravitational acceleration (g). Some rethinking will also be necessary regarding the joule (J), which becomes the sole unit for work, energy and quantity of heat. It is expressed in newton metres (N m).

Derived Units

Quantity	Name	Symbol
Force	newton	$N = kg\ m/s^2$
Work, energy, heat	joule	$J = N\ m$
Power	watt	$W = J/s$
Charge	coulomb	$C = A\ s$
Potential	volt	$V = W/A$
Capacitance	farad	$F = A\ s/V$
Resistance	ohm	$\Omega = V/A$
Frequency	hertz	$Hz = s^{-1}$
Magnetic flux	weber	$Wb = V\ s$
Flux density	tesla	$T = Wb/m^2$
Inductance	henry	$H = V\ s/A$
Luminous flux	lumen	$lm = cd\ sr$
Illumination	lux	$lx = lm/m^2$

Definitions

A *newton* is that force which gives a mass of one kilogramme an acceleration of one metre per second squared.

A *joule* is the work done when the point of application of a force of one newton is displaced through one metre in the direction of the force.

A *watt* is one joule per second.

A *coulomb* is the quantity of electricity transported in one second by a current of one ampere.

A *volt* is the p.d. between two points of a conducting wire carrying a constant current of one ampere when the power dissipated between these points is one watt.

A *farad* is the capacitance between the plates of a capacitor when a quantity of one coulomb of electricity produces a p.d. between the plates of one volt.

An *ohm* is the resistance between two points of a conductor when a constant p.d. of one volt produces a current of one ampere.

A *hertz* is the frequency of a periodic phenomenon of which the periodic time is one second.

A *weber* is the flux which, linking a circuit of one turn, produces in it an e.m.f. of one volt as it is uniformly reduced in one second (1 tesla $= 10^4$ gauss).

A *henry* is the inductance of a closed circuit which produces an e.m.f. of one volt when the current varies uniformly at one ampere per second.

Density is kilogramme per metre cubed (kg/m^3).

Pressure is newton per square metre (N/m^2).

Further information on S.I. units is given in the following publications.

The Use of S.I. Units, PD5868. British Standards Institution, London.
ANDERTON, P. and BIGG, P. H. *Changing to the Metric System*, 3rd edn H.M.S.O., London (1969).

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