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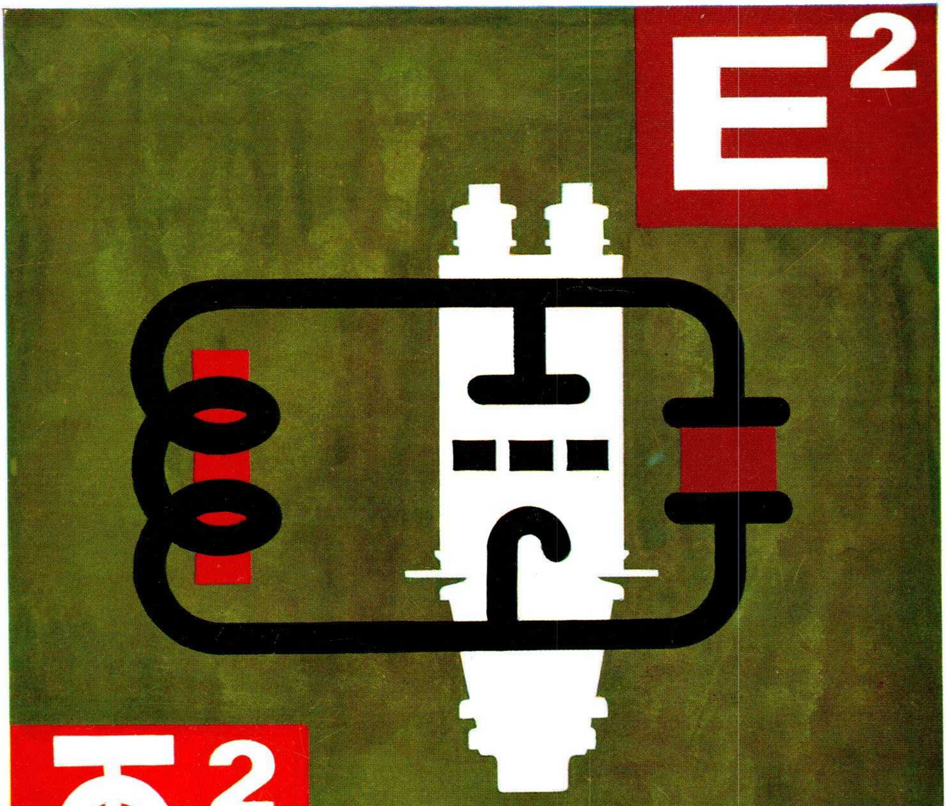
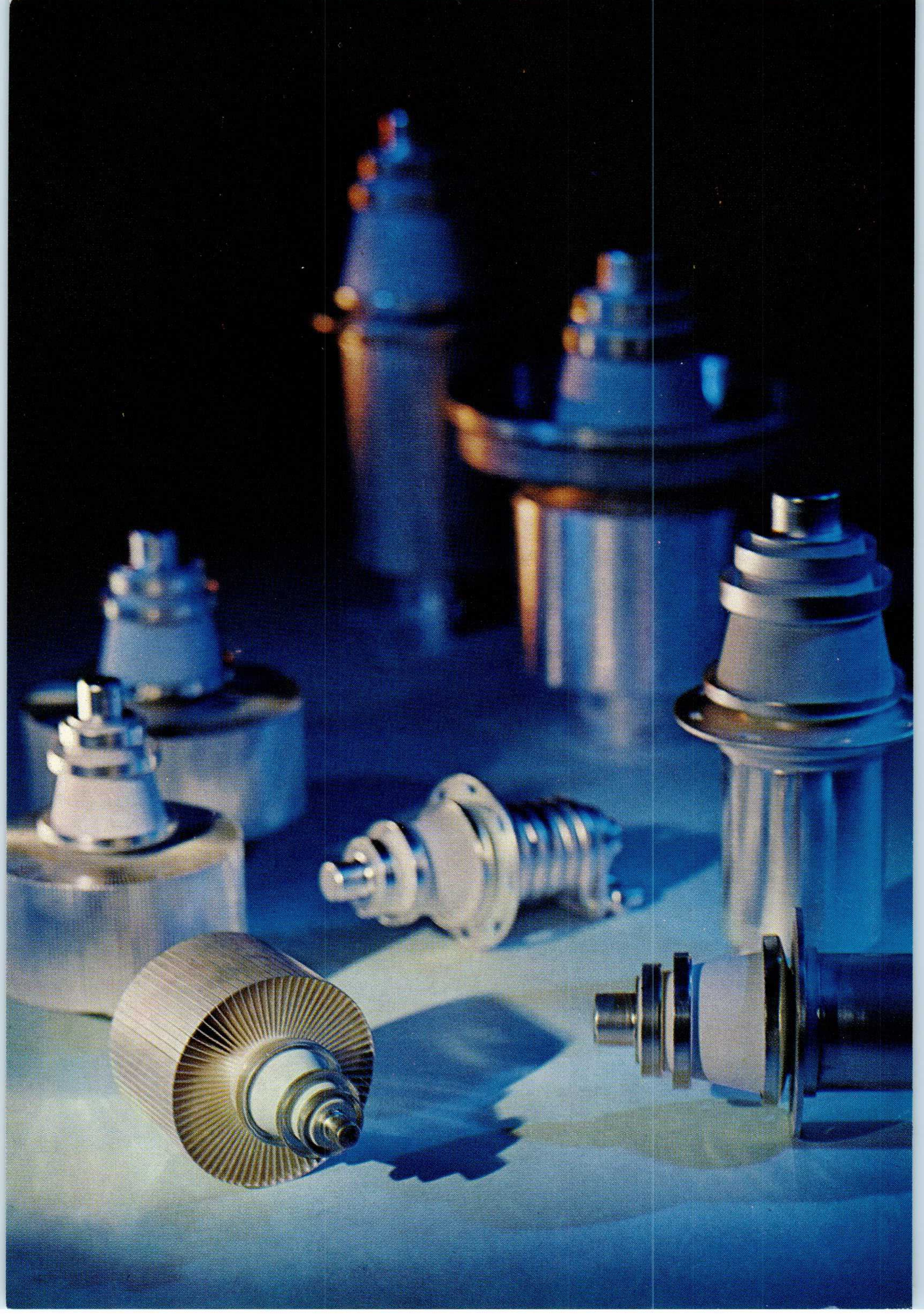
APPLICATION BOOK

ELECTRONIC COMPONENTS AND MATERIALS DIVISION

TUBES FOR R. F. HEATING



Tubes for R. F. Heating



Tubes for R.F. Heating

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Edited by **H. Stanley**

PUBLICATIONS DEPARTMENT
ELECTRONIC COMPONENTS AND MATERIALS DIVISION

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Acknowledgement

Although this book merely touches some of the problems met in the design of generators for r.f. heating, it brings together the work of many people active in that field.

Among the problems it deals with, a number have provided stimulus for further development. Solutions that were technically sound as well as economically acceptable were, in many cases, found only through close cooperation between the user of the equipment, the designer of the r.f. generator, and the manufacturer of the r.f. tube. Members of sales, development, manufacturing, quality control, and applications staffs have all participated in this cooperation. To these, and to the staff of the development laboratory in Eindhoven, without whose active and enthusiastic cooperation practical realization of much of the applications work described in this book would not have been possible, the author expresses his appreciation and heartfelt thanks. Especial thanks are also due to his colleagues P.G. Giles, D. E. Nightingale, and C.W. Touch who have contributed some of the chapters.

Introduction

Until very recently man had at his disposal only two sources of heat, the sun's rays and fire. Both radiate the same kind of energy in the infra-red band of the electromagnetic spectrum. Primarily it is the irradiation of matter by energy in this band that produces the molecular phenomenon known as heat.

But infra-red energy penetrates only a short distance into most substances. To reach the interior, heat has to be carried by one or both of two inherently slow and inefficient processes, conduction and convection. Though heating by these means was adequate for the needs of earlier centuries, it long ago became obvious that efficiency would be far greater if heat could somehow be generated within a substance, and greater still if it could be concentrated in the region where it was most needed.

The discovery of longer waves in the electromagnetic spectrum, and mastery of the technology needed to generate them, opened the way to making this possible. Such waves penetrate far deeper than infra-red, and if they are properly directed they can be made to excite the same sort of molecular phenomena. Essentially two distinct ways of doing so have been applied, each being most effective in a different part of the spectrum. In terms of frequency (rather than wavelength), these are:

50 Hz - 30 MHz	Eddy current heating of metals in a magnetic field.
1 MHz - 3 GHz	Dielectric heating of non-metals in an electric field.

Frequency plays a part not only in how the energy is applied, but also in how it is generated:

50 Hz - 300 Hz	Single- and multi-phase transformers working from a.c. mains.
300 Hz - 50 kHz	Rotary converters, now giving way to thyristor converters.
50 kHz - 100 kHz	Thyristor converters, electron tubes.

100 kHz - 200 MHz	Electron tubes of more or less conventional construction.
200 MHz - 1 GHz	Electron tubes of special construction.
1 GHz - 2.5 GHz	Magnetrons and, to a lesser extent, klystrons.

In this book we shall concern ourselves only with those frequencies at which tubes of conventional construction — mainly triodes — are used.

What we shall talk about most is the design and construction of practical circuits for use in conjunction with such tubes. Treatment of the subject from this angle may already be long overdue. Experience has repeatedly shown that circuit design — and at the frequencies with which we shall be concerned, that term can be taken to include layout and construction as well — is every bit as important as theoretical calculations of the operating conditions of the tube. Without proper regard to both, an r.f. generator cannot operate as it should. All too often, attempts to blame failure on either the tube or the circuit are meaningless. Both are inextricably linked and must work together as a unit.

This is not to say that all of the burden must fall on the circuit designer; the tube manufacturer must bear his part too. With that in mind, a series of tubes has been developed which can be integrated into the circuit to the fullest possible extent. Two fully worked-out examples of r.f. heating generators discussed in the last chapter of this book show how the use of such tubes, in conjunction with sound circuit design criteria, can ease the way to consistently satisfactory results.

1 Tubes for R.F. Heating

Traditionally, high power r.f. amplifier or oscillator tubes have been classed as transmitting tubes, regardless of whether they are used in radio transmitters or in r.f. heating equipment. This somewhat arbitrary classification tends to give the impression that any tube of the required power capability is suitable for either application. However, this assumption is not completely valid as the tube operating conditions in the two applications are very different.

In radio transmitting fixed station equipment, power tubes operate under almost ideal conditions. Such installations are virtually free from mechanical shock and vibration, are almost always supplied from a regulated power source and, most important, always operate with a matched constant load. Such operating conditions are rarely found in industrial r.f. heating applications. As these installations are normally associated with production lines and other electro-mechanical equipment, power regulation is often less than ideal and the equipment is often subjected to various degrees of vibration. Also, because of the nature of the work being done, the load is rarely matched correctly and can be expected to vary considerably. From this it can be seen that power tubes intended for industrial applications require different design criteria from those intended for transmitting service.

1.1 Design Requirements

The design criterion used for industrial heating tubes differs in several ways from that used for transmitting tubes. As mentioned in the previous paragraph, two new properties required are better shock and vibration resistance, and higher overload capability resulting from the necessity to operate under unmatched conditions. However, certain properties normally found in transmitting tubes are not required for industrial r.f. generators. Among these are wide bandwidth capabilities, low distortion and high power gain.

It may be asked why a high power gain is not necessarily required in industrial tubes. The reason for this is somewhat complex and is discussed in the following paragraphs.

Given a graph of constant current curves for a power triode, a load line may be drawn which represents the operating conditions of the tube, given certain operating requirements. From the load line and other data, the grid bias, anode voltage, grid impedance, anode load impedance and output power may be determined. If, on this set of curves, a new set of load lines is drawn for different values of load impedance, while anode voltage, feedback factor and grid impedance remain constant, a plot connecting the ends of the resultant load lines will represent the line of constant grid impedance. Similarly the grid impedance can be made to vary while the other factors are kept constant, thus giving a plot representing the line of constant anode load impedance. Finally, both load impedance and grid impedance can be made to vary in such a manner that the output power remains constant; thus a line of constant output power may be plotted.

From this it can be seen that if the lines of constant grid impedance and constant output power could be made to coincide, the output power would remain constant, regardless of changes in load impedance. While such an ideal situation is impossible, it has been found that if a low power gain is combined with a grid structure that has low primary and secondary emission, the output power remains quite constant over a significant range of load impedance variation. It is true that this requires a relatively high drive power, but this is considered of secondary importance when compared to the stable performance of the tube. In a typical example, the output power varies by about 4% for variations in load impedance between 0.6 and 1.4 times the assumed value.

The required drive power for our tubes is of the order of 2% of the available output power.

1.2 Tube Types

To meet the need for industrial applications, a range of tubes has earlier been designed specifically for such applications. This range includes the TBL(W)(H)7/8000, TBL(W)(H)6/17, TBL(W)12/38 and TBL7/9000, all of which have been well received by equipment manufacturers and equipment users. The most important single feature of these tubes is the "*K*" grid, so named because of the material used in its construction (see 1.4).

In their development, a maximum operating frequency of 30 MHz was accepted because, at the time, virtually all industrial equipment operated below this frequency.

However, there has recently been a growing requirement for power tubes to operate at higher frequencies for applications in dielectric heating of low-loss materials. Also, even higher power levels are becoming necessary for induction heating equipment. In view of these requirements and the success of the earlier range of industrial tubes, a series of power tubes with ceramic envelope has now been introduced, which features the "*K*" grid and offers many other advantages.

Bearing in mind the requirements for high frequency dielectric heating and high power induction heating, the following combinations of output power and limiting frequency have been selected:-

Power (kW)	Frequency (MHz)	Type No.
2.5	160	YD1240
4.0	160	YD1150 series
7.5	150	YD1160 series
16	120	YD1170 series
32	80	YD1180 series
62	30	YD1190 series
122	30	YD1202 series
247	30	YD1212 series
480	30	YD1342 series

The stated output power figures represent the anode output power. The power delivered to the load will be less than this figure and will depend on the tank circuit efficiency and drive power required to sustain oscillation.

The combinations of maximum frequency and output power may seem to be arbitrary. However, the choice of these combinations is based on two sound reasons. First, most dielectric heating applications use frequencies in the 25 MHz to 30 MHz range and those requiring higher operating frequencies generally require power levels below 10 kW. Second, the higher the power capability of the tube the more difficult it becomes to achieve high frequency performance. This is because the necessarily larger electrode structures introduce higher capacitive and inductive values and increase the transit time of the electrons. Therefore, the combinations chosen offer the best compromise between tube cost and frequency capability.

During the development of this new series of ceramic envelope industrial triodes, the following points served as guides:

1. A d.c. to r.f. efficiency of at least 75 %.
2. Grid current constant over tube life.
3. Relatively low operating anode voltage.
4. Rugged electrode construction.
5. Suitability for high frequency operation.
6. Good safety margins on high voltage and other parameters.
7. No additional electrode seal cooling at frequencies lower than 4 MHz. (On the higher power tubes a, low velocity air flow may be required to prevent ionisation of the air adjacent to the terminals, which could result from the higher operating voltages).
8. Low water consumption for water cooled types.

1.3 Envelope Construction

The electrodes are mounted on heavy ring-shaped terminals arranged coaxially on conical high purity alumina ceramic spacers. This method of construction, which is standard for the entire series of ceramic envelope tubes, guarantees maximum mechanical strength and ensures low inductance for high frequency operation. The use of low-loss ceramic also results in minimum capacitive current heating in the electrode seals.

Ceramic insulation has previously been used mainly for communication tubes primarily for economic reasons. However, for industrial tubes this material offers several advantages. Two of these advantages mentioned above are high mechanical strength and low capacitive loss. In addition, ceramic has a much higher melting point than glass, which means that the tubes can be processed at a higher temperature for more efficient degassing, thus providing a better vacuum. Also, ceramic can be machined to much closer tolerances than glass, which allows very accurate alignment of the electrodes. Finally, ceramic can withstand temporary overheating better than glass, a feature which is very important to the user of industrial equipment.

1.4 The "K" Grid

This grid is ideally suited to industrial applications because of three properties:

1. It has a high work function, even when covered with thorium evaporated from the cathode.
2. It has a rough surface, ensuring good heat radiation.
3. It has high thermal conductivity.

A "*K*" grid can withstand a continuous load of 25 W/cm^2 (161 W/in^2) and even under this severe condition, complicated by the presence of deposited thorium, the thermionic emission is only about 1 μA/cm^2 (6.4 μA/in^2). This value does not change significantly over the normal tube life.

From this it can be seen that the "*K*" grid fulfils one of the most important requirements for constant power operation, that of low grid emission. In addition, the excellent thermal conductivity and heat radiation properties of the "*K*" grid assure good heat dissipation, another requirement of operation under unmatched load conditions.

As the load impedance varies from the matched value, anode and grid dissipation also vary. With a load impedance lower than the nominal value, grid dissipation falls while anode dissipation increases. The opposite occurs when the load impedance is higher than nominal, i.e. the anode dissipation falls but the grid dissipation rises, reaching a maximum at a value of load impedance just short of the unloaded value. The "*K*" grid, therefore, is well suited to such operating conditions, having as its primary features excellent heat conduction and radiation properties, as well as low emission.

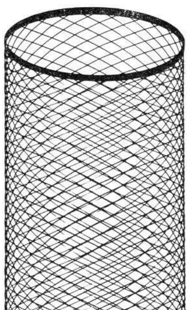


Fig. 1.1. The "K" grid.

1.5 Electrode Construction

Two different internal constructions are used; one for tubes rated up to 7.5 kW, and another for tubes of 15 kW and higher.

In the lower power tubes, the grid and cathode constructions are similar to those used in the TBL(W)(H)7/8000 and TBL(W)(H)6/4000. The cathode consists of two helically wound wires in parallel and has proved its strength, reliability and long life. With the use of coaxial metal-ceramic seals, the cathode construction becomes even more rugged.

While this helical cathode provides an excellent solution for the lower power tubes, its use becomes impracticable at high power levels. Since a desirable feature of these tubes is low anode voltage, there is necessarily a requirement for high cathode current for a given power level. This would require the use of a very long helical or multistrand cathode which would not be sufficiently rugged and which would tend to deform with age. For these reasons, a special mesh type cathode was designed for the tubes that operate at power levels of 15 kW and higher.

Among the features of the mesh cathode are high emission, mechanically rugged construction, small dimensions, low inductance (particularly important at higher frequencies), a desirable ratio between filament voltage and current, and high slope. Switching tests performed on mesh cathodes show that deformation during life is negligible.

Fig. 1.2. Mesh cathode construction.

1.6 Anode Voltage Supply

An industrial tube operating with a relatively low anode supply voltage has the following advantages over the more conventional high voltage low current type of operation:

1. Seal losses due to internal capacitive circuits are lower.
2. Ionization and corona effects are minimised.

Additionally, a circuit designed around such a tube and using solid state rectifier stacks offers substantial cost savings. The reasons for this are fully discussed in Ch. 2 under the heading "Anode Voltage Supply".

1.7 Additional Considerations

The small size of the mesh cathode has led to tubes whose overall dimensions are quite small relative to power capability. The advantage of small size and weight may not be immediately apparent, especially to the equipment manufacturer, because size and weight are normally not problems in industrial installations.

However, the size and weight of power tubes can be very important to the equipment user, from the point of view of shipping, storing, and handling spare tubes. Also, in equipment which may be portable or transportable, size and weight can be very important.

This low size to power ratio places higher than usual demands on anode and grid structures with regard to heat dissipation. "*K*" material easily meets the requirements of grid dissipation. With regard to anode dissipation, all tubes, both air and water cooled types, have been designed with high efficiency anode cooling systems to ensure adequate anode cooling under even the most adverse operating conditions. In the water cooled types operating at 30 kW and higher, the anode cooling system consists of helical water flow grooves machined in a thick-walled anode structure. This system is extremely efficient; a water consumption rate of 0.34 litres (0.16 gallons) per minute per kilowatt of anode dissipation is the normal requirement.

Throughout this chapter, a great deal of emphasis has been placed on rugged construction and ability to withstand temporary overloading. As an indication of how this has been accomplished with this new series of tubes, laboratory tests performed on the YD1212, the 240 kW tube, showed that an output power as high as 450 kW could be reached for a short period of time without damage to the tube.

1.8 Summary

To summarise, the new series of ceramic envelope industrial triodes offers the following features:

1. Efficiency of at least 75 %.
2. Very rugged construction.
3. Small size.
4. Highly efficient water cooling system.
5. Good high frequency characteristics.
6. High safety margins on all parameters.
7. Constant grid current and drive through life.
8. No additional cooling required for low frequency operation.
9. Relatively low anode voltage.
10. No cathode or grid deformation through life.

A quick selection guide is given in Table 1.1. In addition, full technical information is available on request.

SELECTION GUIDE — INDUSTRIAL CERAMIC TRIODES

oscillator output power (continuous) (kW)	type no.	cooling	typical operation					
			class C industrial oscillator					
			frequency	filament		anode		
			(MHz)	(V)	(A)	(kV)	(A)	output power (kW)
2.22	YD1240	forced air	160	6.3	33	4.5	0.7	2.4
3.55	YD1150	forced air	160	6.3	33	5.0	1.0	3.9
3.55	YD1151	water (separate jacket)	160	6.3	33	5.0	1.0	3.9
3.55	YD1152	water (helix)	160	6.3	33	5.0	1.0	3.9
7.15	YD1160	forced air	150	6.3	66	5.0	2.0	7.5
7.15	YD1161	forced air	150	6.3	66	5.0	2.0	7.5
7.15	YD1162	water (separate jacket)	150	6.3	66	5.0	2.0	7.5
15.4	YD1170	forced air	120	5.8	130	6.0	3.4	16.2
15.4	YD1171	water (separate jacket)	120	5.8	130	6.0	3.4	16.2
15.4	YD1172	water (helix)	120	5.8	130	6.0	3.4	16.2
30.0	YD1180	forced air	80	7.0	175	7.5	5.4	32.4
30.0	YD1182	water (integral jacket)	80	7.0	175	7.5	5.4	32.4
60.0	YD1190	forced air	30	8.4	235	8.0	10.0	62.3
60.0	YD1192	water (integral jacket)	30	8.4	235	8.0	10.0	62.3
120	YD1202	water (integral jacket)	30	12.3	255	10.0	16.0	122.5
120	YD1203	vapour	30	12.3	255	10.0	16.0	122.5
240	YD1212	water (integral jacket)	30	12.6	380	14.0	23.5	247.5
240	YD1213	vapour	30	12.6	380	14.0	23.5	247.5
480	YD1342	water (integral jacket)	30	14.0	555	16.0	42.0	489.0
480	YD1343	vapour	30	14.0	555	16.0	42.0	489.0

Full details on these tubes are given in the Electron Tubes series of our Data Handbook System

2. H. T. Supply

2.1 Mains Voltage Fluctuations

The H.T. supplies for r.f. generators are, normally, suitably rectified single-phase or three-phase mains supplies. Before discussing the methods of rectification we should give some consideration to the mains voltage fluctuations that can be expected.

A common factor, for instance, in plastics moulding factories is the fairly heavy electrical consumption of compressors to run hydraulic presses. The mains feeder to the factory is usually well loaded, if not actually overloaded, with the net result that the mains voltage to equipments in the factory is low when some of the large compressors are running. Such compressors are usually hydrostatically controlled. The switching of compressors is random but at certain times they can either be all on or all off. A local sub-station can usually improve the position at, for example, full load but then the off-load mains voltage rises considerably. The r.f. generators have to withstand the total mains fluctuations aggravated by such prevalent local conditions. The electrical design of the generator tube stake into account such voltage fluctuations although it is not usually possible to guarantee a long life if the mains fluctuations are greater than a level that can be reasonably incorporated in the design. It follows that precautions may be necessary in the equipment to guard the tube against *wide* mains fluctuations.

Finally, the mains voltage fluctuations may have a considerable influence on the design of automated or timed processes using r.f. heating.

2.2 Rectification

The various methods that can be used for rectifying either a single phase or three phase mains supply are given in Table 2.1 on page 14, together with the relevant data for transformer and rectifier ratings.

2.3 Self Rectification

In addition to the rectifier systems shown in Table 2.1 the possibility exists of using an industrial tube with raw a.c. applied to the anode and allowing the tube to do its own rectifying. Such a practice is, however, not recommended for the following reasons. The oscillator will operate at its maximum efficiency at the crest of the positive-going sinewave where the h.t. potential equals the permitted maximum tube voltage rating. Over the remainder of the positive half cycle, both before and after the crest, the oscillator efficiency will fall as the applied anode voltage falls. In the positive half cycle the total tube operating time is $2/\pi$. During a large part of this time the tube is operating inefficiently so that the *power* delivered by the tube is considerably less than $2/\pi$. Since the tube is not operating over the negative half cycle, the net result is that the power output is reduced by a factor $> 1/\pi$. In practice, this factor approximates to 2/9. Finally, of course, it is impracticable to use self rectification with three-phase mains supplies.

2.4 Power Supply Ripple

The arguments of the previous section also apply, to an extent governed by the amplitude of the ripple from the h.t. supply, to the generator tube. Thus with a very "rough" d.c. line it is necessary to ensure that the tube ratings are not exceeded on the crest of the ripple. This rough d.c. is the result of single-phase full-wave rectification without filtering. Three-phase rectification generally gives a sufficiently low ripple.

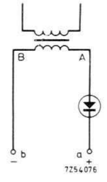
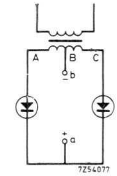
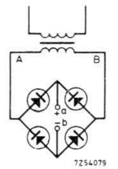
Excessive ripple voltages can be an additional source of undesirable radiation from the generator by modulating the r.f. energy at the ripple frequency. The radiation level of the equipment is determined by the maximum radiation obtained at the crest of the ripple modulation.

In some processes, because of the speed of the process, e.g. seam welding, modulation ripples may occur on the seam. In these applications the ripple must necessarily be kept very low.

2.5 Filament Supplies

All modern power tubes have thoriated tungsten filaments of spiral, rod, or mesh construction. To ensure adequate emission and the maintenance of the original electrical characteristics over a long period of use, certain precautions must be observed.

Table 2.1 Idealized rectifier circuit performance (part 1a)

<table>
<tr><th></th><th>single-phase
half-wave</th><th>two-phase
half-wave</th><th>single-phase
full-wave</th></tr>
<tr><td>type of rectifier circuit</td><td>
</td><td>
</td><td>
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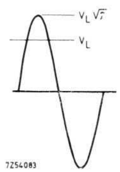
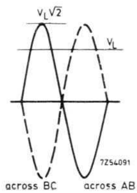
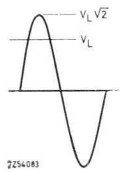
<tr><td>secondary input voltage per phase</td><td>

across AB</td><td>
</td><td>

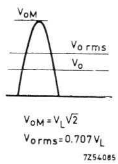
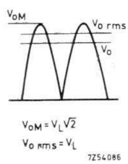
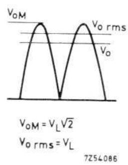
arcoss AB</td></tr>
<tr><td>output voltage across a-b</td><td>
</td><td>
</td><td>
</td></tr>
<tr><td>output voltage pulses per cycle (N)</td><td>1</td><td>2</td><td>2</td></tr>
</table>

Table 2.1 Idealized rectifier circuit performance (part 1b)

		single-phase half-wave	two-phase half-wave	single-phase full-wave
output voltage				
d.c. output voltage V_o		$0.45\ V_L$	$0.90\ V_L$	$0.90\ V_L$
r.m.s. output voltage $V_{o\ \mathrm{rms}}$		$1.57\ V_o$	$1.11\ V_o$	$1.11\ V_o$
peak output voltage V_{oM}		$3.14\ V_o$	$1.57\ V_o$	$1.57\ V_o$
output current				
average current per rectifier leg I_{FAV}		I_o	$0.5\ I_o$	$0.5\ I_o$
$I_{F\ \mathrm{rms}}$ per rectifier leg	R	$1.57 I_o$	$0.785 I_o$	$0.785 I_o$
	L	$0.707 I_o$	$0.707 I_o$	$0.707 I_o$
I_{oM} per rectifier leg	R	$3.14 I_o$	$1.57 I_o$	$1.57 I_o$
	L	I_o	I_o	I_o
transformer rating				
r.m.s. voltage per secondary phase		$2.22\ V_o$	$1.11\ V_o$ (to center tap)	$1.11\ V_o$ (total)
r.m.s. current per secondary phase	R	$1.57\ I_o$	$0.785\ I_o$	$1.11\ I_o$
	L	$0.707\ I_o$	$0.707\ I_o$	I_o
secondary volt-amperes VA_s	R	$3.48\ V_o I_o$	$1.74\ V_o I_o$	$1.23\ V_o I_o$
	L	$1.57\ V_o I_o$	$1.57\ V_o I_o$	$1.11\ V_o I_o$
primary current (transf. ratio 1 : 1)	R	$1.21\ I_o$	$1.11\ I_o$	$1.11\ I_o$
	L	$0.5\ I_o$	I_o	I_o
primary volt-amperes VA_p	R	$2.69\ V_o I_o$	$12.3\ V_o I_o$	$1.23\ V_o I_o$
	L	$1.11\ V_o I_o$	$1.11\ V_o I_o$	$1.11\ V_o I_o$
fundamental ripple frequency f_r		f	$2f$	$2f$
percentage ripple $100\ V_{fr\ \mathrm{rms}}/V_o$		111	47.2	47.2
crest reverse voltage across diode		$3.14\ V_o$	$3.14\ V_o$	$1.57\ V_o$
		$1.41\ V_L$	$2.82\ V_L$	$1.41\ V_L$

R = resistive load L = inductive load f = supply frequency

In the calculation of the above circuit performances, the rectifier forward voltage drop and the transformer impedance have been ignored. Figures for primary transformer volt-ampere ratings neglect magnetizing current.

Table 2.1 Idealized rectifier circuit performance (part 2a)

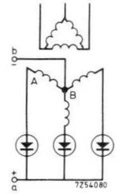
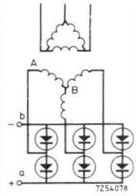
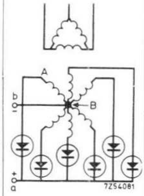
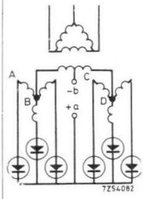
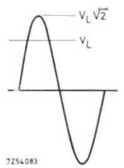
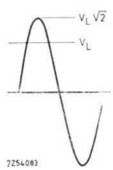
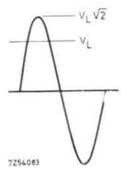
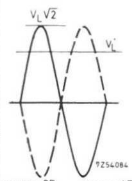
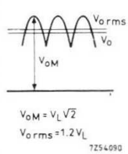
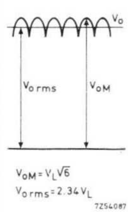
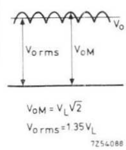
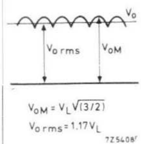
	three-phase half-wave	three-phase full-wave	six-phase half-wave	double-star with inter-phase transf.
type of rectifier circuit				
secondary input voltage per phase	 across AB	 across AB	 across AB	
output voltage across *a-b*				
output voltage pulses per cycle (N)	3	6	6	6

Table 2.1 Idealized rectifier circuit performance (part 2b)

		three-phase half-wave	three-phase full-wave	six-phase half-wave	double-star with interphase transf.
output voltage					
d.c. output voltage V_o		$1.17\ V_L$	$2.34\ V_L$	$1.35\ V_L$	$1.17\ V_L$
r.m.s. output voltage $V_{o\ \mathrm{rms}}$		$1.02\ V_o$	V_o	V_o	V_o
peak output voltage V_{oM}		$1.21\ V_o$	$1.05\ V_o$	$1.05\ V_o$	$1.05\ V_o$
output current					
average current per rectifier leg I_{FAV}		$0.333\ I_o$	$0.333\ I_o$	$0.167\ I_o$	$0.167\ I_o$
$I_{F\ \mathrm{rms}}$ per rectifier leg	R	$0.588\ I_o$	$0.577\ I_o$	$0.408\ I_o$	$0.293\ I_o$
	L	$0.577\ I_o$	$0.577\ I_o$	$0.408\ I_o$	$0.289\ I_o$
I_{oM} per rectifier leg	R	$1.21\ I_o$	$1.05\ I_o$	$1.05\ I_o$	$0.605\ I_o$
	L	I_o	I_o	I_o	I_o
transformer rating					
r.m.s. voltage per secondary phase (V_L)		$0.885\ V_o$	$0.428\ V_o$	$0.741\ V_o$	$0.855\ V_o$
r.m.s. current per secondary phase	R	$0.588\ I_o$	$0.816\ I_o$	$0.408\ I_o$	$0.293\ I_o$
	L	$0.577\ I_o$	$0.816\ I_o$	$0.408\ I_o$	$0.289\ I_o$
secondary volt-amperes VA_s	R	$1.50\ V_oI_o$	$1.05\ V_oI_o$	$1.81\ V_oI_o$	$1.50\ V_oI_o$
	L	$1.48\ V_oI_o$	$1.05\ V_oI_o$	$1.81\ V_oI_o$	$1.48\ V_oI_o$
primary current per transformer leg (transf. ratio 1 : 1)	R	$0.484\ I_o$	$0.816\ I_o$	$0.577\ I_o$	$0.414\ I_o$
	L	$0.471\ I_o$	$0.816\ I_o$	$0.577\ I_o$	$0.408\ I_o$
primary volt-amperes VA_p	R	$1.24\ V_oI_o$	$1.05\ V_oI_o$	$1.28\ V_oI_o$	$1.05\ V_oI_o$
	L	$1.21\ V_oI_o$	$1.05\ V_oI_o$	$1.28\ V_oI_o$	$1.05\ V_oI_o$
primary input current (transf. ratio 1 : 1)	R	$0.830\ I_o$	$1.41\ I_o$	$0.817\ I_o$	$0.707\ I_o$
	L	$0.817\ I_o$	$1.41\ I_o$	$0.817\ I_o$	$0.707\ I_o$
fundamental ripple frequency f_r		$3f$	$6f$	$6f$	$6f$
percentage ripple $100\ V_{fr\ \mathrm{rms}}/V_o$		17.7	4.0	4.0	4.0
crest reverse voltage across diode		$2.09\ V_o$	$1.05\ V_o$	$2.09\ V_o$	$2.42\ V_o$
		$2.45\ V_L$	$2.45\ V_L$	$2.83\ V_L$	$2.83\ V_L$

R = resistive load $\quad$ L = inductive load $\quad$ f = supply frequency

In the calculation of the above circuit performances, the rectifier forward voltage drop and the transformer impedance have been ignored. Figures for primary transformer volt-ampere ratings neglect magnetizing current.

1. Although filaments of mesh construction are more rugged than the more conventional types, the tube should not be exposed to undue shock or vibration before installation or during operation.
2. Current surges should not exceed the stated maximum values (usually 2.5 times the operating current).
3. The nominal voltage measured at the filament terminals of the tube should ideally be within $\pm 1\%$ of the published value. The permitted excursions of $+5\%$ and -10% about this value should only be of a temporary nature.

The filament transformer usually operates at mains frequency. By using a suitably designed transformer some of the above requirements can be met.

A number of primary taps will permit the accurate setting of the filament voltage on installation at the final site. Adequate provision should also be made, when specifying the nominal secondary voltage, for the voltage drop in the leads between filament transformer and tube terminals.

To limit switching surges, step switching in the primary can be used or the transformer can be designed with a built in leakage reactance that will limit the maximum current to a predetermined value. If the transformer can be adequately cooled either by natural convection or by a small blower, the resistance of the secondary windings may also assist in this respect.

2.6 Anode Voltage Supply

Solid state rectifier stacks simplify the design of high-voltage high-current power supplies*. They eliminate the need for a high-voltage insulated heater transformer to run the heaters of gasfilled rectifier tubes and the thermal delay circuits that are needed to prevent h.t. being switched on before the rectifier tubes are warmed up.

The avalanche characteristic of silicon rectifier diodes eliminates the additional capacitors or resistor ladder networks that were required to ensure equal voltage distribution sharing across each diode in a stack, since it absorbs transient energy within the rectifier circuit without damage.

The BYX25 and BYX39 series of controlled avalanched rectifiers have been designed for use in high-voltage high-current power supplies.

* See our publication "High-voltage rectifier stacks" (ordering code 9399 254 02301)

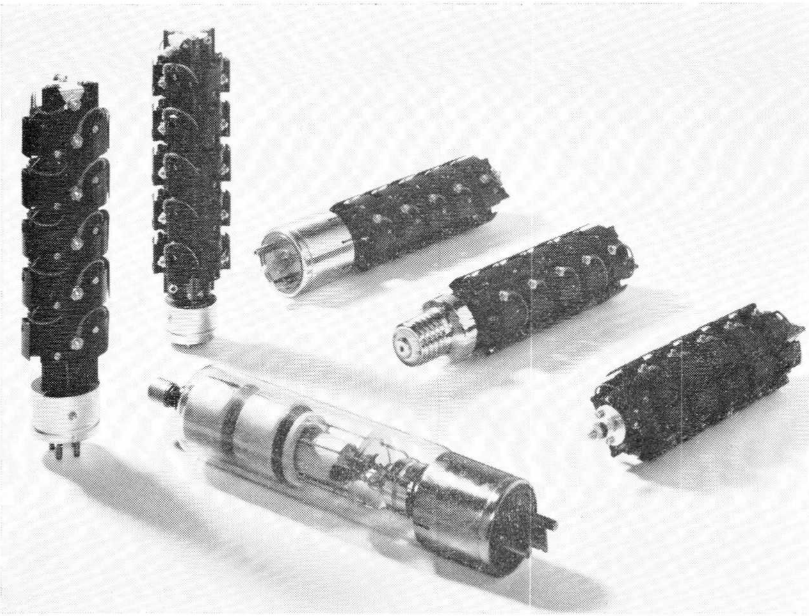


Fig. 2.1. Solid-state high-voltage rectifier stacks, compared with a gas-filled rectifier diode. Two basic stack mounting arrangements are available — stud mounting and tube-base mounting (for direct replacement of tubes); there are four types of tube-base mounting in our range, as shown.

Each diode is capable of a maximum mean forward current of 20 A with a maximum crest working reverse voltage of 1 kV. The diodes are stacked in series to give the required crest reverse voltage plus a suitable safety factor.

It follows that it is preferable, in the design of a generator tube for a given power, to keep the working voltage as low as possible and the current high in order to reduce the number of rectifier diodes in the stack.

Further advantages in lower voltage operation of high-power generators can be seen in the manufacture of the h.t. transformer. "Insulation costs money", and it is cheaper to make a transformer of a given kVA rating at a lower voltage than at a higher voltage. The same principle applies to the design of the oscillatory circuit where r.f. potentials are likely to be at least double those of the h.t. transformer, and where the voltage ratings of items such as tank capacitors, blocking capacitors or decoupling capacitors are one of the main cost features.

Since the essential requirement of dielectric heaters is the production of a high-voltage across the load coupling plates, it may be argued that a higher generator tube operating voltage is to be preferred. However, the problem of transferring r.f. energy from a tube to the work is a *power* transfer problem and the potential applied to the load is dependent on the load coupling circuit design. A simple approach to this problem is given in Chapters 6 and 8.

The arguments for a low-voltage high-current industrial tube, so far as power supply requirements are concerned, may be summarized as follows:

(a) H.T. transformer design is simpler and cheaper.
(b) Tank capacitors, blocking capacitors and decoupling capacitors are smaller and cheaper.
(c) Shorter stacks of a high current rating can be used.

For certain applications, however, gasfilled rectifiers are preferred. Fig. 2.1 shows examples of solid-state rectifiers and a rectifier tube.

3. Cooling Systems and Temperature Measurements

To ensure that the r.f. generator tubes operate satisfactorily and have a long life, some form of cooling is essential. The parts of the tube that require cooling are the electrodes (mainly the anode) and the seals.

The tubes in the new industrial ceramic range are available with any desired method of anode cooling, such as forced air cooling, water cooling or integral water cooling. Vapour cooling may also be used on certain of the high power tubes. In addition to the anode cooling requirement, and depending on the circuit configuration, it may be necessary to cool the grid and filament cathode seals by forced air. At induction heating frequencies and frequencies up to 4 MHz, forced air cooling of the seals is not required provided that the recommended terminal connectors (or connectors of equivalent thermal mass and contact area) are used.

3.1 Anode Cooling Systems

Air, water, or vapour may be used for cooling the anodes of the tubes described in this application book, according to type. The factors which may determine whether air or water cooling is to be used are discussed in the following paragraphs.

3.1.1 Air Cooling

If water is not required for any other purpose, consideration may be given to forced air cooling of the anode. One advantage of air cooling is that the warm exhaust air may be used to prevent condensation inside the cabinet, especially with certain types of dielectric heating load.

One disadvantage may be the noise of the exhaust air. If air is used the supply must be properly filtered, otherwise blockage of the anode radiator may occur or foreign matter, deposited in parts of the circuit, may give rise to short circuits or to inefficient operation. It is also essential that the air filtering system is correctly maintained and the cost of this must be considered. The air blower circuit should be interlocked with the main switching system to prevent the tube being operated without cooling. The interlock switch should be mounted in the exhaust stream.

Where the ambient temperature is high, or for operation at high altitudes, care should be taken to see that the air flow is adequate in terms of mass per unit time.

The amount of air required at sea level is given approximately by the formula:

$$Q = \frac{50\ P}{T},$$

where Q = air flow in m^3/min.
P = anode dissipation in kW.
T = temperature rise in degC.

3.1.2 WATER COOLING

Water cooling is quiet and may well be preferred if water is already supplied to another part of the equipment. The water may be drawn either from the mains or from a recirculating system. Apart from saving water the recirculating system helps to ensure a high standard of purity. Some of the requirements for satisfactory cooling water are that it should not be corrosive or deposit scale, should not contain insoluble material which might cause blockages, and should not be a good electrolyte. Its mineral content and electrical conductivity should therefore be periodically checked especially when it is not drawn from a circulating system (Fig. 3.1). A non-corrosive water should be low in chlorides, oxygen and carbon dioxide. Scale formation can be avoided by keeping down the amount of silica and bicarbonates, especially calcium bicarbonate. No exact figures can be given for impurities as they are interdependent.

The cooling water system must be interlocked with the electrical supplies. As an added safeguard, the electrical supplies should automatically switch off if the water outlet temperature rises too high.

The amount of cooling water required is given by the formula

$$Q = \frac{14.3\ P}{T},$$

where Q = water flow in litres/min.
P = anode dissipation in kW.
T = temperature rise in degC.

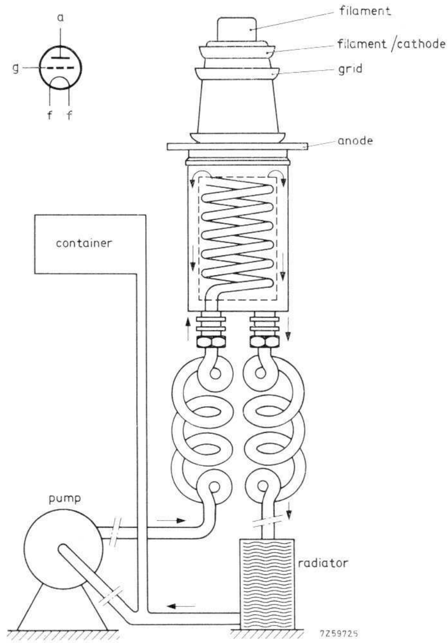


Fig. 3.1. A typical water cooling system.

The above requirements indicate that from a long-term view a recirculating system filled with distilled water in its primary will be the best. The additional initial cost will be well offset by reduced maintenance costs if the system is truly closed and the risk of contamination of the primary water removed.

The new tubes with their integrated water cooler will permit a substantial saving in the original installation because their cooling requirements of 0.34 litres per minute per kW anode dissipation are really low in comparison to the more conventional water jacket types.

The secondary cooling can then be effected through a conventional air or water heat exchanger. The purity requirements for the secondary water are, of course, far less stringent since such a water-water heat exchanger can be designed to suit the local water supply.

3.1.3 VAPOUR COOLING

Providing the installation of a boiler does not seriously inhibit good electronic circuit design, vapour cooling can be used. However, it offers noticeable economic advantages over other indirect water systems only at anode dissipations above 50 kW and with the tube operated near its maximum published anode dissipation. If the condenser pipes inside the boiler are of adequate diameter, enough cooling water may be fed directly through the cooler without a secondary heat exchanger. To maintain the advantage of the higher temperature gradient between condensing steam and condenser water, the flow of the condenser water will, in most cases, be much the same as that needed for conventional water-cooled tubes.

Finally, the factors governing the choice of anode cooling are summarised below.

Forced Air Cooled

Advantages

Air is everywhere freely available.

Anode can be 'live' to r.f. and d.c. with no serious problems.

No water joints to leak.

No condensation problems – the warm air can be ducted around in the oscillator.

Can also provide cooling for seals if designed correctly.

Disadvantages

Air filters are required on air intake and need to be cleaned regularly.

Cooling system may generate noise.

With several generators in a confined space, ambient temperatures may increase unless large and expensive hot air exhaust ducts are fitted in the work space.

Water Cooled

Advantages

No noise.

Air filters not required.

No increase in the ambient temperature of the surrounding work space.

Less overall electrical consumption than the air cooled system as only a low-power pump is used (closed circuit systems).

Disadvantages

Insulated hose coil required if the anode is operated at high r.f. and/or high d.c. potentials.

Condensations problems due to atmospheric conditions.

Mineral content of cooling water may produce 'fur' in cooler unless demineralised water or a recirculating system is used.

Water Cooled
(integral heli-cooled)

The heli-cooled tube does not require a separate water jacket and is fitted with a more efficient cooler.

Vapour Cooled

This system offers no economic advantages over indirect water systems unless anode dissipations in excess of 50 kW are required and the tube is working near its operating maximum.

3.2 Temperature Measurement

It is essential, when designing new equipment, that the temperatures of the electrode seals and coolants are carefully measured. A number of paints and waxed papers are available for this purpose. Care should be taken with paints as some of them are lossy at radio frequencies.

When thermometers are used for coolant temperature measurement it is advisable to use the spirit variety where r.f. fields are present. If only mercury thermometers are available they must be electrically screened.

4. Environmental Design Considerations

4.1 Interference Problems

Unless special precautions are taken, radiation of r.f. energy at fundamental or harmonic frequencies, through direct propagation or leakages from the generator, is inevitable. Because such radiation could cause interference, certain proposals have been internationally agreed, and are indeed binding by law in some countries. These proposals set limits to the frequency and the amount of radiation caused by industrial equipment.

The following extracts from these proposals are given for the guidance of v.h.f. generator designers.

Four harmonically related frequency bands are here of interest, within the limits of which, free radiation from industrial equipment is acceptable. These four bands are as follows:

$$\begin{aligned} 13.65 \text{ MHz} &\pm 0.05\% \\ 27.12 \text{ MHz} &\pm 0.6\ \ \% \\ 40.68 \text{ MHz} &\pm 0.05\% \\ 461.04 \text{ MHz} &\pm 0.2\ \ \% \end{aligned}$$

Other limitations restrict the radiation of harmonics, if the generator operates in one of these bands, to 225 μV/m at 100 metres distance.

For all other frequencies harmonic radiation is restricted to 45 μV/m at 100 metres distance.

Further safety margins have in some countries been imposed on the above specifications to guard against component tolerances, component ageing, incidental fault conditions, etc.

If such limitations are applied they become one of the prime design considerations, often taking precedence over the equipment performance and are never satisfied cheaply.

4.2 Radiation Suppression

R.F. generators, even when working inside the recommended frequency bands, must suppress the incidental harmonic frequencies that result from any generation or amplification of r.f. power.

The selection of operating conditions (choice of angles of current flow in the tube) suggested by harmonic analysis is, although helpful in some instances, largely of theoretical interest. The quality factor Q of unavoidable component resonances can magnify the amplitude of a harmonic out of proportion to the calculated value. Even push-pull operation, where the elimination of all even harmonics is theoretically possible, may not always provide a satisfactory solution. Industrial loads can, by their reactive variations, cause an unbalancing of a carefully adjusted push-pull circuit that is sufficient to make some of the even harmonics re-appear to an objectionable degree.

Some suppression, particularly of the higher order harmonic frequencies, can be undertaken at the source, i.e. the tube. Tubes with an effective high output capacitance are here of help since a relatively large anode capacitance will represent a low impedance path to the higher order harmonics at source. This then influences the choice of oscillatory circuits in favour of those where the anode-to-grid capacitance C_{a-g} can be used for this purpose, that is, those circuits where the tank circuit is connected between grid and anode.

The tank circuit capacitance can also assist in this respect, but only if it can be tied so tightly to the internal tube capacitance that the intervening connector reactances are much smaller at the harmonic frequencies than that of the C_{a-g} of the tube or of the external capacitance. Generally such a construction can only be realised with coaxially connected tubes as found in our new series of industrial ceramic envelope triodes.

If component resonances cause certain harmonic frequencies to appear at an undue magnitude, a search for resonant effects, preferably with varying load reactances, should be made. It can then often be quite a simple matter to reduce the magnification at the offending frequency by a change in component layout or the substitution of components with others of the same basic value but different outline and different stray reactances. These objectionable component resonances could, for example, be caused by the self resonance of tank circuit or blocking capacitors, connectors between tube and circuit, or filtering chokes. The latter, even if wound to give a correct quarter-wave voltage pattern at the fundamental frequency, will, of course, have several self resonances at higher frequencies, any one of which could lie close enough to a harmonic to cause its magnification. If such a magnification is severe, power may be dissipated, creating hot bands on these chokes. A change of length-to-diameter ratio, choice of different wire gauge or insulation or perhaps the use of alternative supporting formers may offer the correct solution.

Any residual harmonic propagation, after exploration of all the above possibilities, can only be suppressed by filtering. If necessary, double screening around the generator proper must be used with capacitive and inductive filter networks in all supply leads, the only r.f. "exit" being through a Faraday screen at the load coupling circuit position (Fig. 4.1).

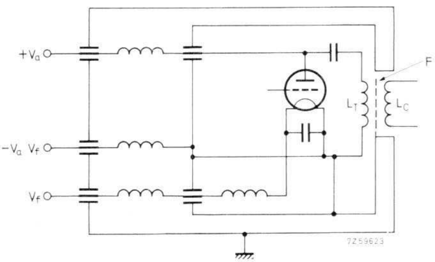


Fig. 4.1. Example of double-screened generator box with π filtering networks consisting of feedthrough capacitors and chokes. A Faraday screen, F, is inserted between the tank circuit coil L_T and the load circuit coupling coil L_C.

A few remarks can be made on the construction of these Faraday screens. The finer the mesh, the better the suppression of higher order frequencies. As a general rule the dimensions of one slot should not exceed a tenth of the wavelength of the highest frequency that is to be suppressed and neither should one slot, or a combination of slots, resonate near this frequency (Fig. 4.2).

If no other way can be found of suppressing one or more of the remaining harmonics, absorption circuits may be used. These are series resonators inserted in the basic generator circuit or the feedline to the load position at high voltage points of the harmonic frequencies. Since their intensity may, however, be so low that exact location is difficult, the final position of the absorption circuits may have to be established by trial and error. Such a change of position may in itself affect the resonance of the absorption circuit and necessitate further tuning for optimum results (see Figs. 4.3 and 4.4).

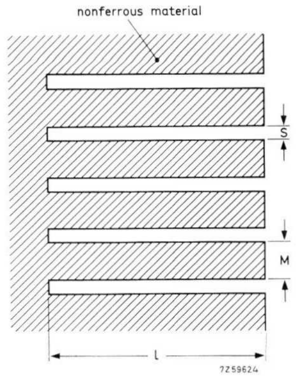


Fig. 4.2. Sketch of Faraday screen. The cross hatched area is the metallic panel out of which the screen has been cut. The mark-space ratio M/S will determine the degree of higher order harmonic suppression, and the path length along the edges of the whole screen any possible resonant slot properties. l should be a significant multiple of $M + S$.

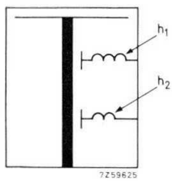


Fig. 4.3. Capacitively loaded line or Kolster circuit with harmonic absorption circuits for the frequencies h_1 and h_2. Accurate resonance is obtained by trimming either the inductive or capacitive components of these circuits.

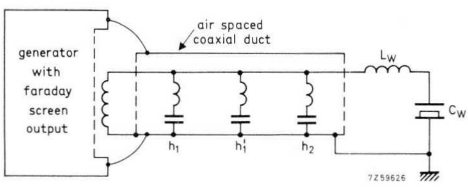


Fig. 4.4. System for removing residual harmonics between Faraday screen of generator and load position. The three traps shown in the coaxial link are in this example for two frequencies; h_1 and h_1' are located at two separate high voltage points of the same harmonic and h_2 at the voltage maximum of another harmonic.

If the processing requirements are such that no adequate frequency control or harmonic suppression can be achieved or the processing frequency and its harmonics fall outside the allocated bands, total screening is the one remaining solution. The machine is then housed in a screened room consisting of a double-walled wire cage with good r.f. filters in the mains power supply and good earthing of the outer cage (Fig. 4.5). The effectiveness of bonding on windows and doors must also be checked.

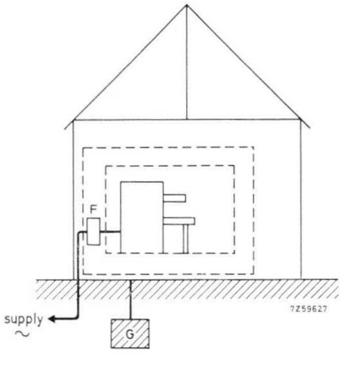


Fig. 4.5. Sketch of double-screened work room with filter (F) in mains supply and adequate earthing (G). Metallic sections of the building structure should not be relied upon to serve as screening or earthing points because their resistance and reactance at high frequencies can be appreciable.

4.3 Frequency Stability of Oscillators

Oscillators in the v.h.f. bands can, of course, be designed and built to satisfy the necessary stability requirements. However, from an applications and economic point of view these oscillators are almost completely restricted to the widest of the four bands, namely around 27.12 MHz.

Industrial oscillators are subjected to loads that appear resistive, at best, only during a small part of the total processing cycle. During the remainder of the cycle, reactances will be reflected to the tank circuit, changing its frequency in proportion to their magnitude. This is because the frequency of a tuned circuit always adjusts itself to such a value that the reactive values of circuit capacitances and inductances are balanced.

It is, therefore, an obvious solution to make these reactances as small as possible in comparison with those appearing in parallel with them. This will minimize the tank circuit frequency shift.

However, small reactances signify high circulating currents that may cause uneconomically high power losses unless suitable tank circuit constructions are used.

The formula

$$Q_L = \frac{1}{\frac{2\,\Delta f \cdot \tan\delta}{f}}\,\eta_C,$$

where

Q_L = the loaded Q of the tank circuit

$$\frac{\Delta f}{f} = \frac{\text{frequency deviation}}{\text{operating frequency}}\ (\%)$$

$\tan\delta$ = the loss factor of the load material

η_C = the circuit efficiency (%)

gives an indication of the Q_L figures required.

Assuming for example that

$$\frac{\Delta f}{f} = 0.5\,\%$$

$\tan\delta = 0.04$

and

$\eta_C = 80\ \%.$

one obtains

$$Q_L = \frac{80}{2 \times 0.5 \times 0.04} = 2000,$$

a rather large figure that might not easily be realized with simple tank circuit constructions.

The problem becomes, however, less serious when it is seen that the above formula assumes a coupling factor that is constant and near unity throughout the processing cycle between tank circuit and load circuit, as will occur only with the simplest form of directly-coupled load circuits.

With other coupling systems this high coupling factor is, because of the tuning characteristic of the load circuit, only maintained over a small part of the total processing time. The coupling factor integrated over the total processing time will thus be much less, and loaded Q figures between 100 and 500 are usually sufficient.

The choice lies therefore between two circuits; one with a high coupling factor over a large part of the processing time and requiring tank circuits of extremely high Q_L, or one with tank-load circuit combinations with a reduced integrated coupling factor and a proportionally higher peak power demand during the "in tune" time. Both solutions demand an economic sacrifice.

A third method, more sophisticated and possibly more expensive, uses automatic means to cancel reactive load variations (Fig. 4.6). The oscillator frequency is monitored and deviations due to reactive load changes are translated into a servo signal that will compensate these reactive load changes by electromechanical means. The time taken for this compensation is dependent on the response time of the mechanical parts of the servo system, thus limiting the use of this control method to applications where correction times are measured in tenths of seconds.

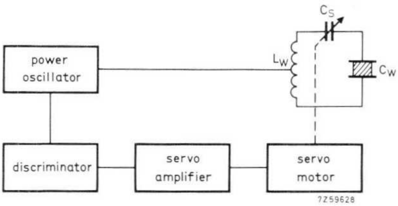


Fig. 4.6. Block diagram of load tuning with servo system. Series tuning is shown, but parallel tuning could also be used.

If the servo response is adequate with respect to the changes of the load material characteristics the load will appear near resistive during the major part of the processing time, and a better processing time-power transfer function will also be obtained. Load tuning servo systems are also suitable where the r.f. power source is a driven and possibly crystal controlled amplifier.

4.4 Safety Circuits

To prevent prolonged and costly production hold-ups through damage to tubes, components, handling equipment and possibly the work material, safety circuits are usually fitted. These circuits render the radio frequency generator and its supply source inoperative when faulty operation occurs. Three groups of safety measures need to be considered.

1. Normal operation safety devices and operator protection.
2. Circuit fault protectors.
3. Load position fault protectors.

The functions of the devices required under the above three headings will, of course, overlap in certain cases, and one safety circuit may satisfy all three requirements.

4.4.1 NORMAL OPERATION AND OPERATOR PROTECTION

Protection devices in this group include air- or water-to-electricity interlocks as well as mechanical interlocks that prevent access to high tension or radio frequency sections of the generator.

In this group fail-safe protection is the primary requirement and speed of action somewhat less important. The action of ordinary relays, micro-switches and circuit breakers will usually be adequate.

Operator protection can be obtained with interlocking microswitches that prevent the generator being switched on so long as access doors or panels are open or protection cowlings removed from the load position. To increase the safety factor it is common practice to duplicate these circuits to make operation impossible unless both circuits are in the "operate" condition. At the load position, for example, this may include two push-button switches, so spaced that the operator *must* use both hands to bring the machine into operation.

The equipment protection circuits must prevent supply or r.f. energy being dissipated in any part of the generator unless all cooling installations are operating. Air and water flow switches, preferably mounted at the respective outlets, are normally used. They can be connected in series with temperature sensors to ensure that normal operating temperatures cannot be exceeded. Delay timers, in a secondary safety circuit, can prevent the application of high tension before gasfilled tubes or the filament of the generator tube itself have reached their correct operating temperature (Fig. 4.7).

Filament surge currents can normally be limited by step switching or self limiting transformers with a built-in leakage reactance. The limiting characteristics of these transformers should be such that the specified filament current surges cannot be exceeded. Very-high-power filament transformers can be fed through primary voltage regulators, keeping the filament voltage within, say, 1 % of the nominal value. The cost of such an installation might be more than offset by the longer life of individual tubes.

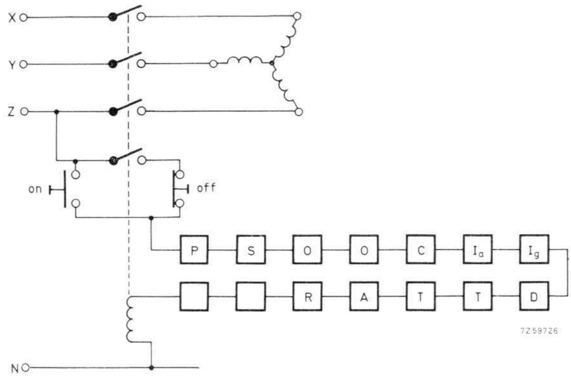


Fig. 4.7. General safety circuit with interlocks for: P access panels, S safety cover, O operators push-buttons, C cooling plant, I_a anode current, I_g grid current, D delay switches, T, process timers, A anode temperature sensing, R arc protector.

Anode and grid current relays can be set to trip at current values that prevent over-dissipation in either of these electrodes. If used for this alone, the relay response time is in no way critical. An additional precaution against serious anode over-dissipation is a thermal fuse attached to the anode by a low melting point metal link that will sever when undue temperature rises occur, releasing a switch in series with the general safety circuit (Fig. 4.8).

4.4.2 CIRCUIT FAULT PROTECTORS

For use with an ideally designed oscillator that remains on the fundamental frequency even under extremely heavy overloads – up to the point

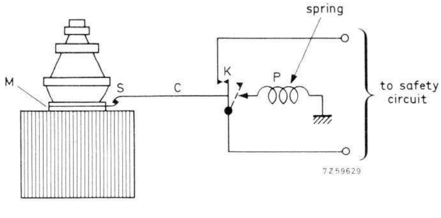


Fig. 4.8. Fuse link to tube anode. A metal band M is fixed around the neck of the tube and has on one side a soft solder joint, S. An insulated cord C holds contact K closed against spring tension of P. In case of overheating S melts and releases cord, causing K to go open-circuit.

where oscillations cease and the total input power is dissipated in the tube anode – the safety cut-outs described above would be adequate. However, circumstances may necessitate a very fast removal of the supply energy if the tube or circuit components are not to suffer permanent damage. The main dangers are the extremely high r.f. potentials due to parasitic oscillations that can cause internal arcs in the tube or external arcs on the circuit components. Arcing at the fundamental frequency, initiated perhaps by contamination of insulating surfaces, may cause component ringing. Because such arcs can be equivalent to an effective short-circuit across the fundamental frequency controlling circuit, they produce similar very high potentials. Under such conditions the supply energy should be removed within less than one cycle of the supply frequency. The system used will depend primarily on the type of power supply used. Where no smoothing is incorporated, grid controlled rectifiers may be used for fast switching of the supply at the secondary of the h.t. transformer (Fig. 4.9). Alternatively, thyristors can be connected in the h.t. transformer primary.

In a system with smoothing capacitors, the stored energy must also be removed from the generator tube and circuit in a similar short time. These fast switching systems resemble an ordinary relay that cuts out at a predetermined current. They all make use of a sensing element that will give a signal proportionate either to the rate of rise or, more frequently, the absolute value of the anode current of the generator tube. The energy stored in smoothing circuits can be absorbed with a "crowbar" circuit (Fig. 4.10).

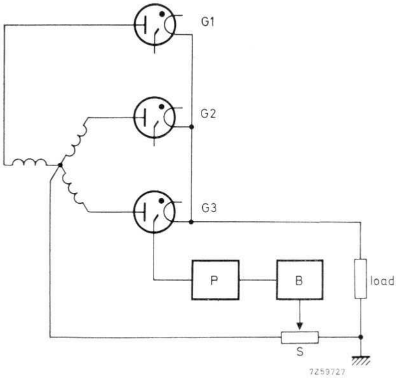


Fig. 4.9. Schematic of quick switching safety circuit for power supply without smoothing. G_1, G_2 and G_3 are grid controlled rectifiers that are, under normal operation, kept in a conducting state by the pulse generator P, which is locked to the mains frequency. The firing pulses delivered by P can, however, be stopped by the action of the blocking network B which derives its command signal from the series resistor S. The network can be designed to respond either to an absolute current value or to a predetermined rate of rise of current through the load, e.g. the generator tube. A similar system could be used to control thyristors in the primary of the h.t. transformer.

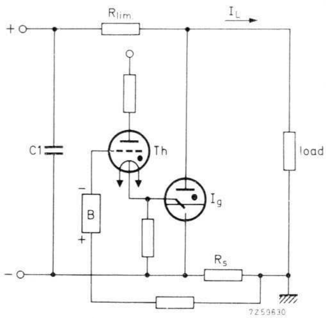


Fig. 4.10. Schematic of "crow bar" circuit. Such a circuit can be used in those cases where a rapid removal of stored energy is necessary, the source in this example being C_1. In the event of increased load currents I_L, a larger potential drop will occur across the sampling resistor R_S, counteracting the bias B that holds off thyratron Th. Above a predetermined threshold, thyratron Th will fire, causing, in turn, the ignitron Ig to conduct and bypass the available energy. At the same time the change in anode current of the thyratron can be used to activate a relay that will drop out the mains circuit breakers in the normal manner.

The circuit described can remove the high tension supply from the generator tube in a few tens of microseconds, but it is good practice to test their operation to ensure complete safety. This can be done by a very simple "test wire" method. The oscillator tube is removed from its position and replaced by a copper wire connected between the anode and cathode terminals. The length of the test wire should be about 3 cm per kV of supply voltage. It should not fuse when the full supply voltage is applied. If subsequent tests and further adjustments of the safety circuit are necessary, a new piece of wire for each test should by used. At the power levels where such safety measures are commonly used, the test wire thickness is 0.23 mm for the YD 1202 (120 kW) and 0.25 mm for the YD 1217 (240 kW).

To avoid excessive switching-on surges two mains contactors can be used (Fig. 4.11). The first (K_1) connects the mains supply through series resistors to the h.t. transformer. The second contactor (K_2), activated by a separate contact on the first, short-circuits the series resistors, so that after the elapsed switching time the full mains voltage appears on the primary terminals of the h.t. transformer. The natural switching time of most commercial circuit breakers is adequate for this purpose.

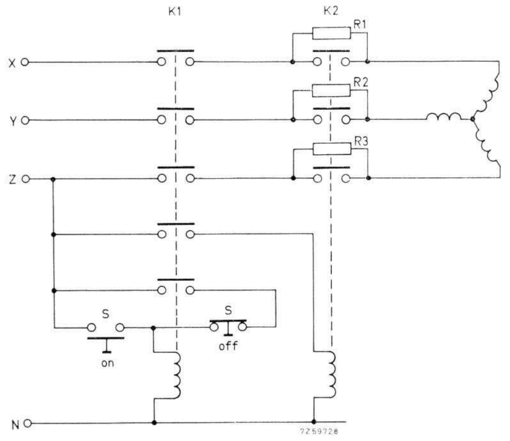


Fig. 4.11. Diagram of three-phase step switch. S is the initiating switch and $R_{1,2,3}$ the series resistors that are short-circuited by K_2 when K_1 has completed its function.

4.4.3 LOAD POSITION FAULT PROTECTORS

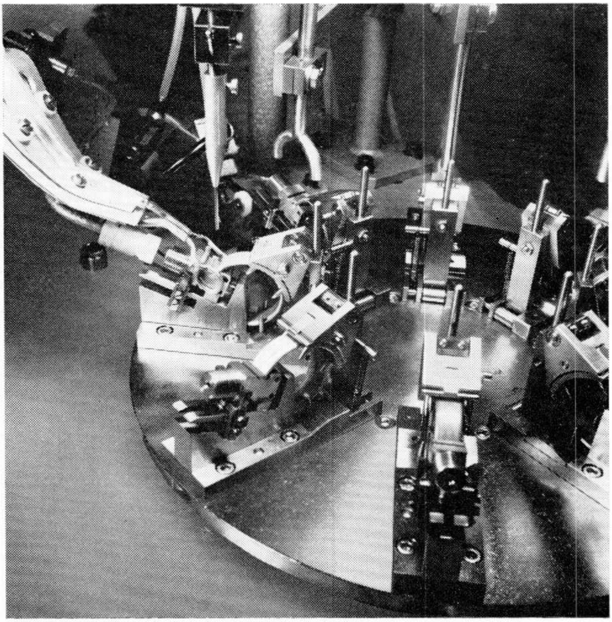
Load position fault protectors may be necessary in both induction and dielectric heating to protect load coils, pressing dies, and jigging and handling equipment, since their destruction, through arcing, can lead to costly production hold-ups.

When arcing is caused through accidental misalignment of the load position tools, contamination, or work material faults it will be necessary to cut the r.f. power rapidly. Cut-off times from 50 milliseconds down to only several tens of microseconds may be required. Arc anticipators are sometimes used that sense a rapid rise in anode current and cause the cut-out to operate before the arc causes damage.

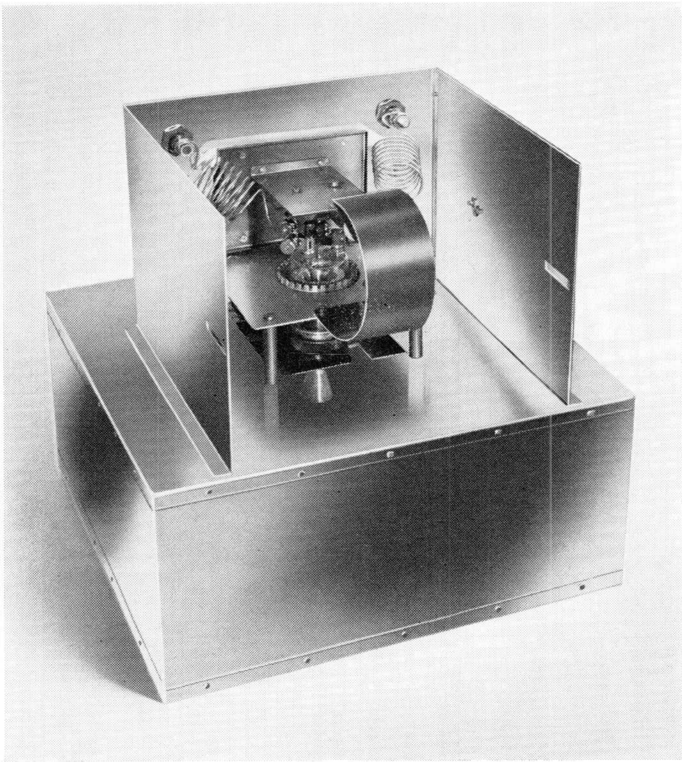
A simpler circuit uses, as the fault detecting element, the resistive path that appears between the electrodes as a result of an arc-induced puncture of the work material. A d.c. source of high internal resistance (several MΩ) is placed across the welding electrodes. In normal operation this potential is enough to block, through thyristors or small thyratrons, a grid switching bias supply in series with the generator tube grid resistor. Under fault conditions the electrode resistance will decrease, firing the triggering element which, in turn, will make the grid switching bias operative and thereby stop the generation of r.f. energy.

Load position sensors may be mechanical-electronic, optical-electronic, thermo-electronic or plain electronic and be made to activate, with their fault signals, the rapid cut-out circuits or grid switching circuits mentioned in the previous paragraphs.

Grid switching circuits are especially useful where a rapid or automatic restart of the generator is required. These circuits are fully described in Ch. 5.11.



A 6 kW r.f. generator for soldering manometer parts, using tube type TBL7/8000 (by courtesy of Himmelwerk A.G., Tübingen, Western Germany).



Laboratory model of a 85 MHz, 5 kW Kolster circuit, using the TBW7/8000, suitable for dielectric heating.

5. Circuit Techniques

Radio frequency power sources for industrial heating are, with few exceptions, self oscillating generators fitted with thermionic triodes operating under class *C* conditions. If they are to work at the efficiencies theoretically obtainable from the tube characteristic curves and be free from excessive harmonic and parasitic oscillations, care must be taken in circuit dimensioning and layout so that the required r.f. potentials are presented to the tube electrodes in the correct phase relationship.

The class *C* calculations are adequately covered in standard text books. This chapter and chapter 6 consider the practical circuit design and construction so that the final circuit, or perhaps even the prototype, will meet the design requirements.

The equipment designer should, when contemplating the component layout, include a three-dimensional concept in his development rather than rely on the two-dimensional theoretical circuit diagram. However, one complements rather than excludes the other, and in practice it will be found that the more efficient circuit is not only the more elegantly realised construction but that this construction can be functionally related to a logically drafted circuit diagram. Extending this design philosophy, or rather retracing the above sequence, it can be said that as r.f. potentials appear in their correct locations and no parasitic circuits are evident from a diagram drawn with a maximum economy of lines, then efficient operation, free from parasitic symptoms, can be expected of the circuit construction. It follows that the close integration of tube and circuit components must be one of the first design aims and the exclusion of stray reactances as well as the grouping of the components should be evident from the basic circuit diagram.

It will be seen that this is valid for both low (0.5 MHz) and high (30 MHz) frequency equipment, and that the ceramic coaxial construction of the new industrial triodes is eminently suited to these techniques.

5.1 Basic Oscillatory Circuits

5.1.1 Single Tube Circuits

For a given power level and operating frequency the selection of a tube type is determined by its electrical characteristics and mechanical construction. However, a wide choice of oscillatory circuits are available in which the tube can function. Up to about 10 MHz all the well known tank circuit and feedback arrangements are acceptable and the preference for one or the other will be influenced by constructional convenience. This choice will, with increasing frequency and power level, be narrowed down in favour of those circuits with common grid or anode terminations and, therefore, primarily capacitive feedback systems. The reasons for this are that the essential low inductance connections are more easily made to these two electrodes and that capacitive feedback systems are, because of their relatively smaller physical dimensions, less likely to cause undesirable phase shift or constitute parasitic resonators.

In this context the term "high frequency" is relative and depends on the power level and construction of the tube. A tentative demarcation line can be drawn for any given loaded Q along which the product of power (in kW) and frequency (in MHz) is constant throughout a range of tubes. For the series of ceramic coaxial tubes under discussion this product would be about 500 for a loaded Q of 20, 200 for a loaded Q of 40 or 100 for a loaded Q of 80 (see Fig. 5.1). These lines can then help in

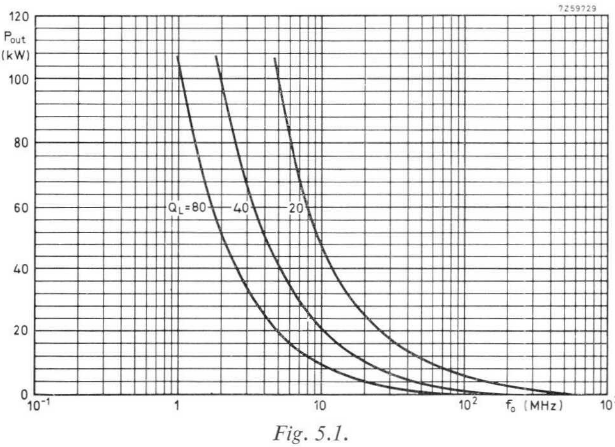


Fig. 5.1.

further determining the physical construction of the oscillator circuit. Figures on the low side of these demarcation lines signify that ordinary *LC* circuits can be satisfactory. Figures on the high side make the use of lumped circuits, cavities or lines, with the tube capacitance truly integrated with the circuit capacitance essential,.

Most types of oscillators circuit can thus be used for induction heating whereas dielectric heating applications will require those that are more suited to high-frequency operation. In both categories, however, preference should be given to the circuit containing only one *LC* combination at or near the operating frequency. This is particularly important in dielectric heating where changes in load reactance can be large and if reflected to the tank circuit, may cause its frequency to be shifted sufficiently far away from that of the "drive" circuit to prevent satisfactory oscillation.

Table 5.1 shows a selection of oscillatory circuits most widely used in radio frequency heating. Some of these are merely derivations from one of the basic circuits, modified to suit requirements.

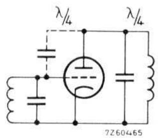
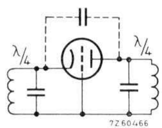
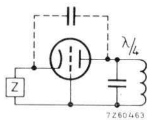
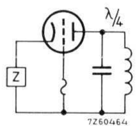
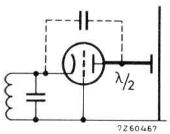
The suggested frequency ranges and fields of application are only intended to be used as a guide. The frequency controlling *LC* assembly in the anode is common to all the circuits. Their classification into three groups is dictated by the usually preferred construction that places the common electrode at nominal earth potential.

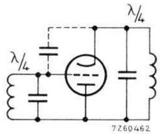
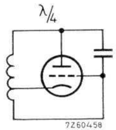
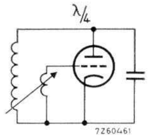
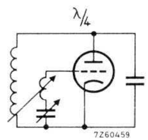
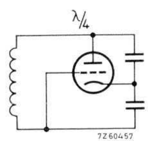
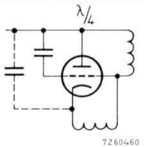
The circuits shown in Table 5.1 are restricted to the tube and the reactances necessary for oscillation. Supply feedlines, blocking capacitors and filtering elements are discussed under their appropriate headings in this chapter. Their choice and method of construction can, however, contribute greatly to the final performance of the generator.

The r.f. potentials developed by the various elements of the oscillatory circuit must be held at a predetermined level with respect to their surroundings which are generally assumed to be at r.f. earth potential.

If correctly adjusted, any oscillator may be regarded as a self-contained "floating" entity as far as r.f. potentials are concerned, and any one point in this circuit may theoretically be chosen as the earth reference point, i.e. the point that can deliberately be made to be the same r.f. potential as its structural surroundings. Unless a specific application demands otherwise, the preferred earthing point is the lead from the electrode that is common to both the tank and the drive or feedback circuit of the oscillator. Two reasons are given for this choice of earthing point.

Table 5.1 Selection of oscillatory circuits most widely used in r.f. heating

Oscillatory Circuit	Frequency of Drive Circuit in Relation to Tank Circuit	Feedback	Preferred Earthing Point	Preferred Construction	Frequency Range and Application
 (1) Common Cathode	Higher	Capacitive (C_{ga})	Cathode	Coil + cap. Lumped circuits Lines Cavities	All up to 50 MHz
 (2a) Common Grid	Lower	Capacitive (C_{fa})	Grid	Lines Cavities	Dielectric 20-600 MHz
 (2b) Common Grid	$\lambda/4$ chokes in filament circuit	Capacitive (C_{fa})	Grid	Coaxial lines Cavities	Dielectric 20-600 MHz
 (2c) Common Grid	Lower $Z = \lambda/4$ or $n \times \lambda/4$ resonator, where n is odd number	Inductive "S" loops	Grid	Coaxial lines Cavities	Dielectric 100-600 MHz
 (2d) Common Grid	Lower	Capacitive probes (C_{af})	Grid	Coaxial lines	Dielectric 10-600 MHz

	(3) Common Anode	Higher	Capacitive (C_{ag})	Anode	Coil + cap. Lumped circuits Lines Cavities	All up to 50 MHz
	(4a) Hartley		Inductive pot. divider	Cathode	Coil + cap	Induction heating up to 10 MHz
	(4b) Modified Hartley	A-periodic	Inductive through separate feedback coil	Cathode	Coil + cap. Lumped circuits	Induction dielectric up to 50 MHz
	(4c) Modified Hartley	Higher	Inductive through separate feedback coil with series cap. for phase correction	Cathode	Coil + cap. Lumped circuits	Induction dielectric 10-50 MHz
	(5a) Colpitts		Capacitive potential divider	Cathode, grid	Coil + cap. Lumped circuits	Induction dielectric up to 10 MHz
	(5b) Modified Colpitts	$\lambda/4$ chokes in filament circuit	Capacitive division in tube augmented by C_{af} probes	Grid anode	Lumped circuits Lines Cavities	Dielectric 10-200 MHz

1. It is probable that the common electrode and its associated components will have the greater share of stray reactance with respect to the surroundings and may, if not bonded to these, cause a substantial part of the generated r.f. power to be transferred into the surrounding structure. These same stray reactance paths could, with their relatively low values at higher frequencies, promote parasitic oscillations.
2. The circuit components representing the path to the common electrode have to carry both the output and drive (feedback) signals, producing a certain amount of mutual cancellation. This mutual cancellation, if allowed to become excessive, may affect the grid signal in phase and amplitude to the point where the oscillator efficiency is impaired and, in the extreme case, oscillations at the operating frequency cease. In most circuit layouts this common reactance can be suitably reduced if it is made to generous dimensions and used as the earth reference point.

Since the effect of the common electrode lead reactance is of such importance, the fullest use can be made of the large-area coaxial tube electrode terminations that are a feature of the new industrial triodes. This is equally important for low and high frequencies, except that in the latter case the use of coaxial line and cavity circuits makes this advantage more obvious.

5.1.2 Multiple Tube Circuits

Although single tube operation is preferred, both from a design as well as operational point of view, there are instances where, despite the complications, multiple tube operation is chosen. This may take the form of parallel, push-pull or even parallel-push-pull operation to increase the available load power. Other more specific requirements may determine the choice of one or the other of these systems.

Parallel Operation

Parallel operation of smaller tubes may, for example, be preferred by the designer if more power is required than a single tube can deliver, at a frequency outside the operating range of a single large one.

For some other applications, notably induction heating, load coupling problems may be simpler if the generator itself has already a very low impedance. In such cases, two smaller tubes in parallel may provide more suitable matching than one single tube of double power rating.

The oscillatory circuits for parallel operation will in no way differ from those shown for single-tube operation in Table 5.1, but it will be necessary to prevent tube-to-tube resonance from acting as a parasitic circuit.

This resonance will, with most types of tube, be in the region of 50 MHz to 300 MHz and will have a very high Q, already determined by the low-impedance low-loss connections to the circuit.

In exceptional cases, where the internal tube capacitances are low and the tube shape allows a very close mounting of both tubes, the inter-tube resonancy may be so high that it will lie well inside the transit time region of the tube. Oscillations would then be unlikely but the excitation of this circuit by high-order harmonics would, of course, still be possible. To render the inherent inter-tube resonance harmless, a separate resistance or heavily damped inductance between one, and only one, of the electrode pairs will usually be sufficient. This can be a heavily damped choke of a few turns connected between each anode and the tank circuit, or a similar arrangement between each grid and the remainder of the grid circuit (Fig. 5.2). In the latter case, nickel alloy tapes with a very low d.c. resistance may suffice.

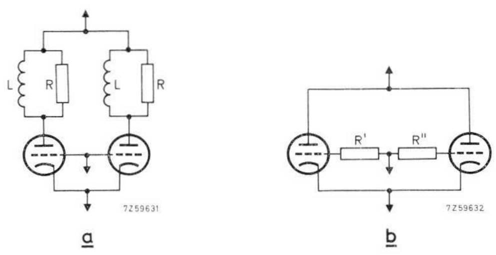


Fig. 5.2. (a) anode and (b) grid damping circuits for the suppression of inter-tube resonance in parallel operation. The resistors R′ and R″ need often be no more than strips of resistive material such as ferrometal alloys etc.

Because of the danger of parasitic oscillation the use of parallel tubes in conventional LC circuits should be limited to frequencies up to, say, 10 MHz. Parallel working is *not* suitable for lumped circuit or line operation.

Push-Pull Operation

Push-pull operation also is often used for higher power levels and/or higher operating frequencies. For certain applications other properties of the push-pull system may be of interest to the designer. The push-pull circuit can be a convenient high impedance r.f. power source and ease transformation problems, notably where anode supply voltages are low but load impedances are relatively high. Since the anode-to-anode voltage swing, at about double the value of single tube operation, appears across the tank circuit, only a quarter of the circuit capacitance will be needed to obtain the same loaded Q. The r.f. operating voltage of these capacitors will, however, have to be twice that of those used for the equivalent single-ended circuit and careful costing is necessary. Much will, of course, depend on the type of capacitor used, airspaced plates – ceramic – or vacuum, but it is by no means a foregone conclusion that to quarter the capacitance at double the working voltage will be the more economic solution.

When radiation problems are a fundamental design consideration, the use of push-pull circuits can help with the elimination of the even harmonics at the source. An effective cancellation of these frequencies will, however, be achieved only if the circuit is well balanced physically and electrically. Great care is needed to ensure that an otherwise well functioning circuit is not thrown off balance by loading changes, since most industrial loads are single ended. A Faraday screen between tank circuit and load coil can often help to reduce the unbalancing effects of capacitive load variations.

Circuits for push-pull oscillation are derived from their corresponding single-ended equivalents and a few of the more common ones are shown in Fig. 5.3.

It can be seen from the increased complexity of the diagrams shown in Fig. 5.3 that the stray reactances have multiplied and consequently have thus increased the possibilities of parasitic oscillation. These will usually be generated through interaction between the tubes and will be at similar frequencies to those found with parallel operation. The chief cause of such oscillations is the position of the tank circuit capacitor which cannot easily be integrated with the tube capacitances. So far as parasitic oscillations are concerned, it is in series with both tube capacitances and will, particularly with large values, merely be an insignificant series reactance at high frequencies. The use of coaxial tubes at low operating frequencies also offers a structural advantage as the ratio of lead inductance between the tank capacitor and the tube-to-tank circuit inductance can be very much reduced.

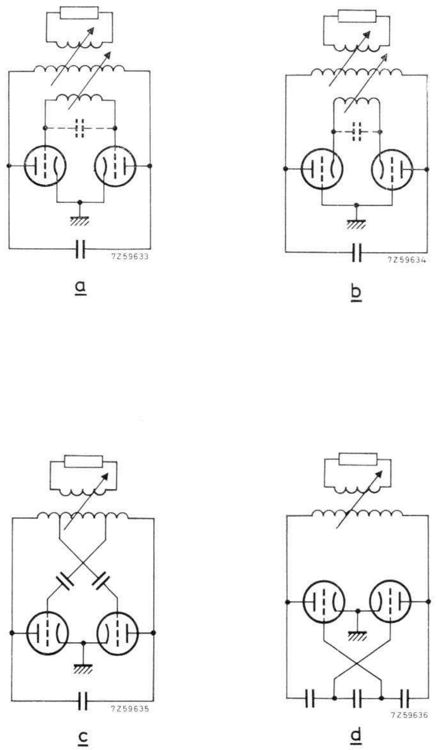
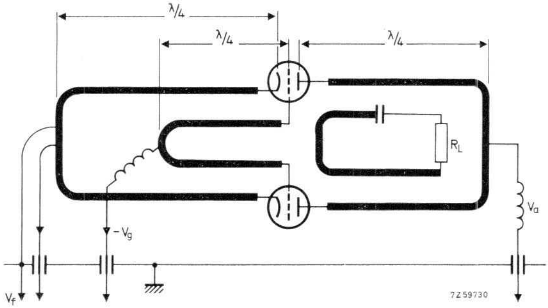


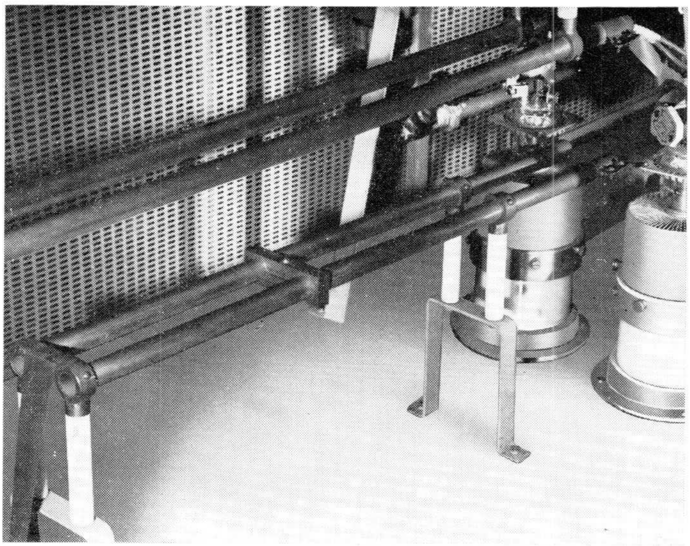
Fig. 5.3. (a) Common cathode, aperiodic or TATG;
(b) Common grid, aperiodic or TATG;
(c) Common cathode Hartley derived;
(d) Common cathode Colpitts derived.

It follows that layout becomes more difficult with increasing tank circuit C/L ratio and increasing frequency. Good results can, however, be obtained at very high frequencies (50 MHz-150 MHz) if lecher lines are used in the tank as well as grid-cathode circuits (Fig. 5.4).

This type of line is ideally suited for push-pull operation at very



(*a*)



(*b*)

Fig. 5.4. A lecher line oscillator shown in its schematic form (a) and a practical circuit (b).

high frequencies and can easily be made self decoupling by feeding supply lines through the line members. If the line length is too great for a given equipment size, the line members may be folded.

To maintain the balance of a push-pull circuit, feedlines for the anode, grid or filaments can be conveniently brought out at the zero voltage point that is present on any inductor used in a push-pull circuit. Only one of these, however, will normally be returned to the nominal r.f. earth potential, either directly or via a low inductance filtering capacitor, preferably of the feedthrough type. The remainder are connected through a $\lambda/4$ choke, which should not appear critical, to their respective feedline filtering capacitors. Such measures will permit the oscillator to "find its own balance" with respect to the r.f. earth potential, and safer push-pull operation can be obtained.

During the development stage separate grid resistors, grid meters and cathode current meters should be used to give an indication of the push-pull operation of tubes and circuit.

5.2 Tank Circuits

Having determined the power level and frequency to suit a particular application and decided the type of oscillatory circuit to be used, consideration should next to be given to the tank circuit, i.e. the combination of inductance and capacitance that governs the frequency of oscillation and from which the generated radio frequency energy is ultimately drawn by the load. Its design in terms of circuit constants (C/L) ratio is based on the high frequency power available from the tube and the ratio of this to the amount of stored energy as dictated by the final application. This ratio is known as loaded Q (Q_L) and can be directly calculated from the tube operating conditions and the tube characteristic curves given in the published data:

$$Q_L = \frac{v_a^2}{X \cdot P_{out}}, \tag{1}$$

where

Q_L = loaded Q (no dimension)
v_a = r.m.s. value of r.f. anode swing (V)
X = reactance of tank circuit inductor or capacitor at operating frequency (Ω)
P_{out} = r.f. power available from the tube anode (W)

To maintain stable oscillation the loaded Q of a power generator should never be allowed to fall below 15, and any possible overloaded conditions should be included in this limit. This means that a well designed industrial generator should have a loaded Q of 25 to 50. Higher values, up to several 100, may be needed where a high degree of frequency stability under conditions of large load reactance changes is required. High loaded Q figures, however, imply low tuning component reactances and hence high circulating currents, and unless the inherent circuit resistance R_s is kept appropriately low, large power losses will occur in the tank circuit. R_s in turn is, for a given Q_L, dependent on the type of circuit and its construction.

5.2.1 Power Losses

The amount of power lost in the tank circuit is given by

$$P_C = I_C{}^2 \cdot R_s \tag{2}$$

where

P_C = r.f. power lost in the resistive circuit elements (W)
I_C = circulating currents (A) due to applied r.f. potential and reactance of resonator elements; $I_C = V_a/X$
R_s = resistance of circuit assembly at operating frequency (Ω)

A further circuit characteristic, the unloaded Q, Q_u, can then be given and expressed in terms of circuit constants by:

$$Q_u = X_L/R_s \tag{3}$$

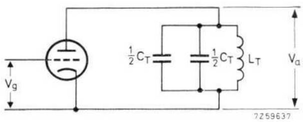
The relation of Q_u to Q_L will define the circuit efficiency η_c

$$\eta_c = \frac{Q_u - Q_L}{Q_u} \times 100. \tag{4}$$

Equations (1) to (4) show that the design and construction of a tank circuit with minimum circuit losses should be a truly integrated combination of its physical components rather than a mere joining together. Long interconnection leads, if not part of the required circuit inductance, sharp discontinuities at right angles to the flow of current, and the use of capacitors with inductive terminals should be avoided. At higher frequencies surface treatment of these components can be beneficial.

5.2.2 Effect of Stray Reactances

Whilst the above considerations are valid for all frequencies, their significance increases with increasing frequency and power. In the ideal case, the assembled circuit will bear a close resemblance to its diagram, in as much as an attempt will have been made to concentrate the largest possible share of the design value of the circuit capacitance and circuit inductance in their respective locations and that each contain as little reactance of the opposing vector as possible. Again, with increasing frequency and power, the inclusion of the inherent generator tube reactances becomes more important. An increase in frequency also causes the tube capacitive reactances to decrease and its inductive reactances to increase, thus impairing the overall performance. The use of appropriate low impedance accessories, developed for use with the new series of tubes, will help to solve such problems. Where an alternative to the standard accessories is required the use of finger strip contacts will provide the all important low impedance r.f. paths.



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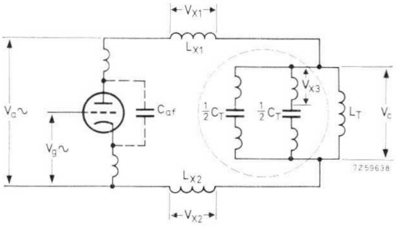


b

Fig. 5.5. Tube and anode tuned circuit as commonly drawn, disregarding stray reactances. A composite tank circuit capacitor has been assumed.

Fig. 5.6. The same circuit with some of the more obvious stray reactances inserted.

The r.f. voltage swing v_c shown in Fig. 5.6 across the tank circuit components that should be equal to the anode swing v_a is reduced by $v_{x1,2}$, the voltage loss along the lead inductance $L_{x1,2}$. A further significant degradation of Q_L can be caused by the voltage drop v_{x2} along the connections within the tank capacitor assembly if these are allowed to become too great. In addition to this, the tube and circuit stray reactances shown could easily form a number of parasitic resonant circuits. Since there is an r.f. potential decay, along the inductive section of any tuned circuit, this must be taken into account when placing additional items like tubes or capacitors in this circuit. These voltage sensitive components should be grouped together as close as possible to the voltage maximum of the tuned circuit. Their physical size, however, will often prevent their ideal positioning and a finite r.f. potential difference between various items placed along the circulating current path must be accepted.

Results more closely in agreement with the calculations for the circuit constants will be obtained if the complex (ringed in Fig. 5.6) of the tank capacitor assembly and its inherent lead inductances is transferred from its original position to the left hand side of the diagram, close to the generator tube. This is shown in Fig. 5.7, where a separation and clear grouping of the important capacitive and inductive circuit components has been achieved.

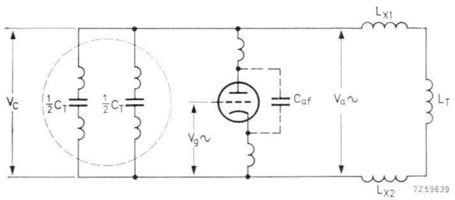


Fig. 5.7. Capacitive and inductive components, including strays, are so grouped that only the minimum of the opposing vector is effective in each composite reactance. $L_{X1,2}$ *are now part of the main tank inductor.*

From the conventional circuit diagram that does not show the stray reactances, a simpler drawing can be made that nevertheless indicates the all-important component grouping and indicates the preferred layout (Fig. 5.8).

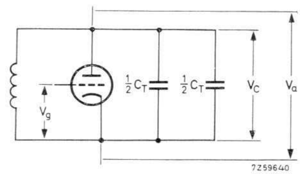


Fig. 5.8. Preferred diagram, indicating the logical component layout in a generator circuit.

The component furthest away from the tank inductor will therefore be at the highest voltage, and any other component, spaced at a distance from this point, will be at a lower voltage.

It is, however, convenient to base the first calculation of the tank circuit constants on the r.f. voltage swing obtainable at the tube since this is a figure readily available from the tube data. Next, adjust the nominal values of any additional capacitors "above" or "below" the tube in proportion to their distance from the tube if these distances are an appreciable fraction of the total path length of the tank circuit inductance. Remember that tank circuit capacitors themselves have finite dimensions and should be oriented with their major dimension at right angles to the current flow in the tuned circuit, whether composed of manufactured units such as ceramic, or constructed as plate capacitors from sheet metal (Fig. 5.9).

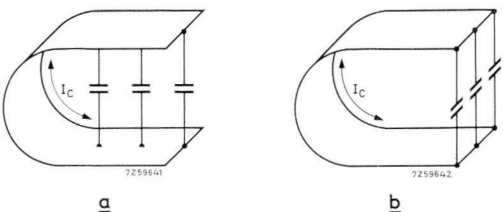
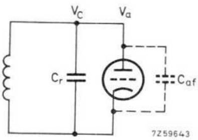


Fig. 5.9. Example of incorrect (a) and correct (b) positioning of individual capacitive components on a U-shaped strip inductor. The incorrect disposition of the capacitive components will result in a higher frequency and degraded Q_L as well as uneven current sharing through the individual capacitors.

A parasitic circuit, formed between the tube and the remaining tank circuit capacitance through the inductance of the connectors linking these items, is inevitable. The likelihood of parasitic oscillations in this circuit is, however, greatly reduced if the tube is on the "low" side of the tank capacitor. In this position the tube internal capacitance (which would represent the tank capacitor of the parasitic resonator) is at a lower r.f. voltage at the operating frequency. Because the fundamental r.f. amplitude must be regarded as the supply voltage for the parasitic circuit, the operational Q of this circuit will be lower than with a reversed order of tube and components. Such a difference of position may make only a difference of a few hundred volts but the parasitic circuit Q will, of course, change with the square of this voltage difference. The two types of circuit configuration given in Fig. 5.10 show that the circuit in (*b*) is less likely to generate parasitics.



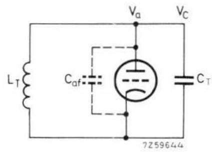


Fig. 5.10. The layouts have the following in common: $C_{af} < C_T$, $X_{C_{af}}$ *high and* X_{C_T} *low at the parasitic frequency but:*

in (a)	*in (b)*
$v_a > v_c$	$v_a < v_c$

therefore under (a) the generator tube could respond to the parasitic circuit more readily.

5.2.3 Tank Inductors

Probably because the tank circuit inductor is the one circuit component that is required in as many different shapes as there are applications, ranging from multi-turn coils to cavities, equipment makers prefer to manufacture this item to their own design. Adequate methods of calculation for inductors of a variety of shapes can be found in most standard text books and their dimensioning will usually present no problems. It should however be noted that the difference between the calculated and realised value becomes greater with decreasing inductive values and the large dimensions normally associated with high power levels. Since tank circuits for industrial generators usually have a high C/L ratio this applies equally at the lower frequencies, and adjustment by trial and error is needed in a prototype construction. Since inductor formulae assume that there is no extraneous coupling, a further change of inductive value will be caused by the coupling of the coil with its surroundings. In industrial generators where, due to the high Q_L values, the magnetic field is appreciable and close screening is required, these changes may be large.

It is also assumed that the designer will make allowance for the inherent surface resistance R_s when dimensioning the inductor. However, the total tank circuit inductance will extend beyond the structure of the inductor into the connections of the other components. These should, therefore, have the same r.f. current rating and be joined to the tank inductor with large area mating surfaces. Surface discontinuities at right angle to the flow of the circulating current should be avoided. When the method of construction permits, soldered, brazed or seam welded joints with a large meniscus should be used in preference to overlap and screw joints or spot welding.

The choice of materials is limited, in practice, to copper, aluminium or brass. Copper and brass can easily be silver plated, but the effectiveness of a commercial plating depth of about 6–8 μm will be evident only at higher frequencies (150 MHz–200 MHz) where skin effects cause a higher current concentration near the surface. Note that only the commercial grades of aluminium (99.5 %) should be used as all Dural grades are excessively lossy at radio frequencies. Examples of tank inductors can be seen in Fig. 5.11.

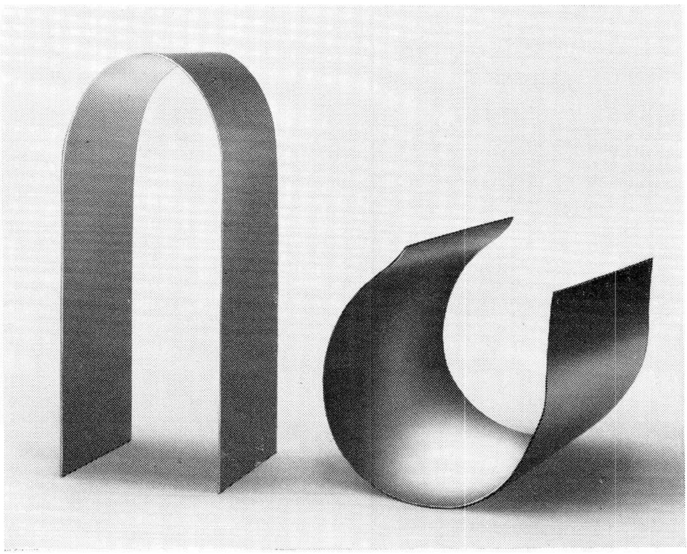


Fig. 5.11. Examples of tank inductors of a high C/L ratio in the 27 MHz range. At this frequency and with a copper construction, the inductor loop widths must be approximately 2 cm per kilowatt of the tube output to avoid undue circuit losses.

5.2.4 Tank Capacitors

From the theoretical circuit diagram as well as the potential distribution and current paths in a tuned circuit it appears that the ideal capacitive component should, to be truly unipotential, have no physical dimensions. Though this is unrealistic, it should nevertheless be the aim to concentrate as much electrical capacitance into as little effective plate area as sound engineering practice will permit. This will ensure that the currents due to the energy exchange between inductor and capacitor in the tuned circuit will not cause excessive potential changes over these areas, and decrease the effectiveness of the capacitor as a store of energy.

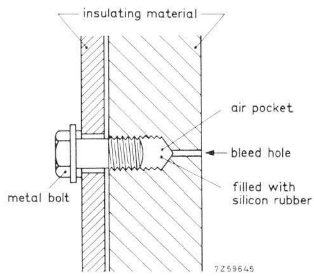
If a single or plate capacitor serves the purpose it should be connected with its major dimension at approximately right angles to the current flow through the circuit inductor. In the same way an assembly of several capacitors mounted along lines of equal r.f. potential, as indicated in Fig. 5.9 (*b*), is to be preferred. A more even sharing of the circulating current will result, avoiding unnecessary electrical and thermal stresses between the individual units. This is very important since the current carrying limitations, imposed on a given type of capacitor, are largely governed by the electrical and thermal conductivity of its connectors. Capacitors with large connectors or several connectors per side should therefore be used.

Though apparently contradicting some of the preceding statements, it may, in certain cases, be better to use a large number of small capacitors in parallel. In this way a wider current sharing can be achieved, the inductive components inherent in each capacitor are paralleled, and their resultant is therefore, very small. The paralleled capacitors also provide a larger total thermal contact area. The frequency of parasitic inter-component resonances is, for the same reason, increased to a value that may lie well inside the transit time region of the tube, thus reducing the possibilities of parasitic oscillation.

Capacitors of very high kVA rating are often manufactured as water or oil cooled assemblies and can thus, for a given volume, be operated at much higher current densities.

When constructing multiple capacitor assemblies, e.g. banks of ceramic capacitors, all metallic parts must be electrically tied to a definite d.c. potential. The polarity and level of this potential are of no consequence but are chosen so that undue stresses in the insulating materials are avoided. Failure to observe this precaution could produce high charges on the "floating" structural elements that might give rise to corona effects and arcing.

Air pockets in insulating materials under high electrical stress can ionize and cause decomposition of the surrounding material. Where practicable, the free air space should be filled with one of the silicone rubbers used for electrical purposes. Further precautions against ionization are shown in Fig. 5.12.



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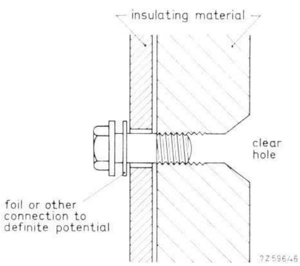


b

Fig. 5.12. To avoid ionization damage the brass bolt shown should be tied to a definite potential and so dimensioned that it will fill the threaded hole completely, or alternatively the hole drilled right through the insulating material. Where requirements for mechanical strength are not so high bolts manufactured from insulating material can be used.

A further possible cause of trouble with multiple tank capacitor assemblies is their stray capacitances to those parts of the surrounding structure that are associated with the feedback circuit. This could, if substantial, provide feedback paths at higher frequencies and cause parasitic oscillations.

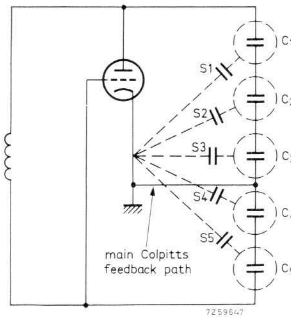


Fig. 5.13. Example of Colpitts oscillator with multiple capacitor assembly in the tank circuit. Each sub-assembly is assumed to have a relatively large stray capacitance (S_1, S_2, S_3, S_4, S_5,) to ground. Separate high frequency circuits are thus created and since their stray capacitances extend to the cathode potential they will respond in the original Colpitts mode. A badly distorted waveform or violent parasitic oscillations will result. Because these parasitic oscillations follow the fundamental oscillator configuration their suppression through antiparasitic devices will be very difficult.

5.2.5 Coil and Capacitor Assemblies

The L/C type of tank circuit is the most conventional one and can be satisfactory, depending on the chosen loaded Q and power level, up to frequencies of about 50 MHz. When the circuit capacitor is constructed from sheet metal, in preference to ready-made components, the capacitance can be calculated from the standard formula for plate capacitors:

$$C = \frac{A_c \times 1.1}{4\pi \times d} \text{ (pF)},$$

where

A_c = the effective plate area (cm^2)
d = the distance between the plates (cm).

The complementary inductor will in all practical cases be a single layer coil. Its inductance can therefore be determined by the following formula, where it is assumed that its length will be greater than a third of its diameter and the spacing between turns equal to the diameter of the conductor:

$$L = \frac{n^2 D^2 \pi^2}{l + 0.45 D} \quad (\mu\text{H}),$$

where

n = number of turns
D = coil diameter (cm)
l = coil length (cm).

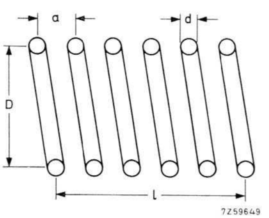
Where the spacing between turns is not equal to the conductor diameter, a correction factor must be introduced so that

$$L_{\text{eff}} = L + \Delta L \quad (\mu\text{H})$$

where

$$\Delta L = \mu \times K \times D.$$

K is a factor dependent on the ratio of the coil pitch to the conductor diameter. Values for K are given in Fig. 5.14.



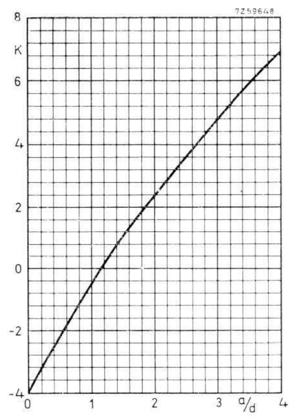


Fig. 5.14 (a). The derivation of the factor K from the coil dimensions.

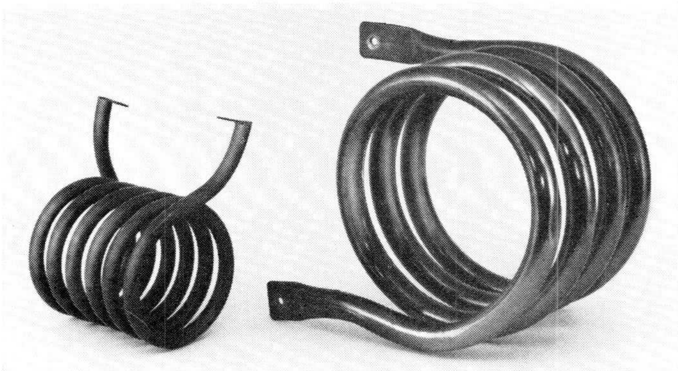


Fig. 5.14. (b). Typical self-supporting coils used in coil and capacitor tank circuits.

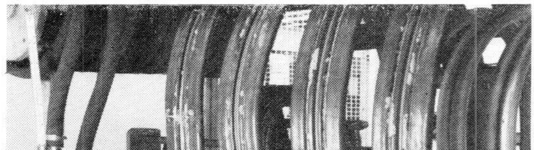


Fig. 5.15. A general view of a coil and capacitor tank circuit. This tank circuit is used with a YD1212 in a 240 kW induction heater developed by our application engineers.

5.2.6 Lumped Circuits

With increasing frequency and Q_L, the reactive values of C and L decrease to the point where the circuit inductance is provided by no more than perhaps a half or three quarter turn of the coil, and an appreciable amount of the total circuit inductance will be in-built in the structure of the tank circuit capacitor. To avoid undue losses in such a circuit due to the increased circulating currents, a smoother mechanical and electrical transition between the two components is necessary. The extreme case, where both are so closely merged that they appear as one component, is known as a lumped circuit. The calculation of the capacitive section will still be possible with the help of the plate capacitor formula (5.2.5) but that of the inductor is somewhat different.

$$L = 0.002 \times l\,(2.303 \log_{10} \frac{4l}{b+c} - \Theta) \quad (\mu\text{H})$$

where for a loop in the shape of a

circle	$\Theta = 2.45$
square	$\Theta = 2.85$
triangle	$\Theta = 3.197$

and l, b and c in cm as shown in Fig. 5.16 (a).

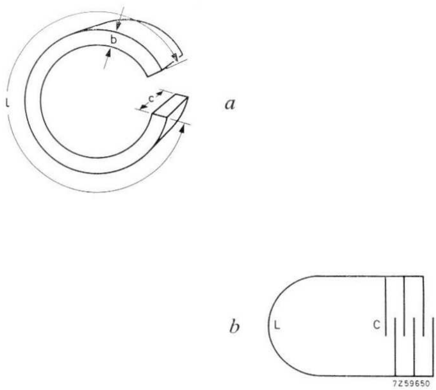


Fig. 5.16. Schematic diagram of (a) lumped circuit inductor and (b) of a lumped circuit.

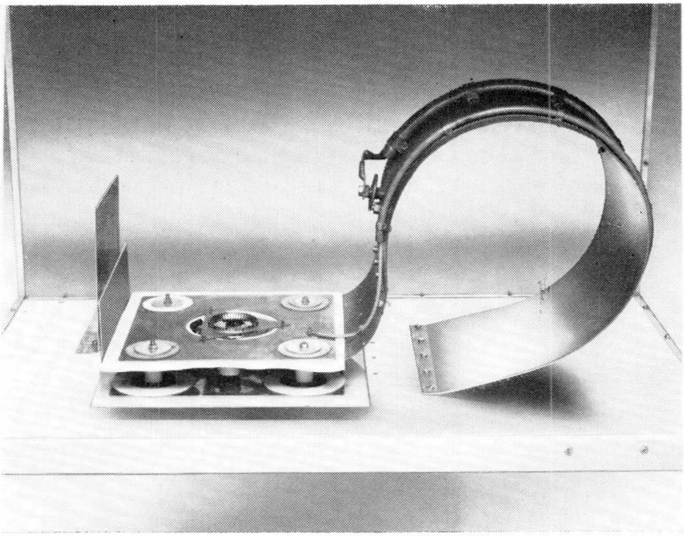


Fig. 5.17. An example of a well integrated coil and capacitor assembly forming a 27 MHz tuned circuit. The faces of the tank capacitor are used to accommodate the respective anode and grid blocking capacitors and the tubes running along the inductor section carry the feedlines to the grid and filament voltage. This is an excellent example of the self decoupling techniques discussed in Ch. 5. The tube and the feedback components are not shown.

5.2.7 Laminated Circuits

If the C/L ratio increases yet further both components can be manufactured integrally. An example of such a construction is the laminated circuit (Figs. 5.18 and 5.19), where each section of a plate capacitor is extended into the inductive loop of the circuit. The paralleling of a number of such plates then results in a LC combination, where, with an increasing number of plates, and therefore rising Q_L, the surface necessary for the consequently increased circulating current is automatically provided. For any given plate size the frequency changes are small when further plates are added to a minimum number (usually 10 to 15, see also the graphs given in Figs. 5.20 and 5.21). Good circuit efficiencies can be obtained and compact assemblies can be made for anode supply potentials up to 5 kV.

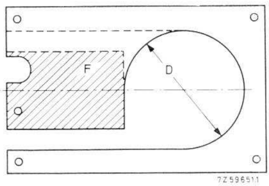


Fig. 5.18.

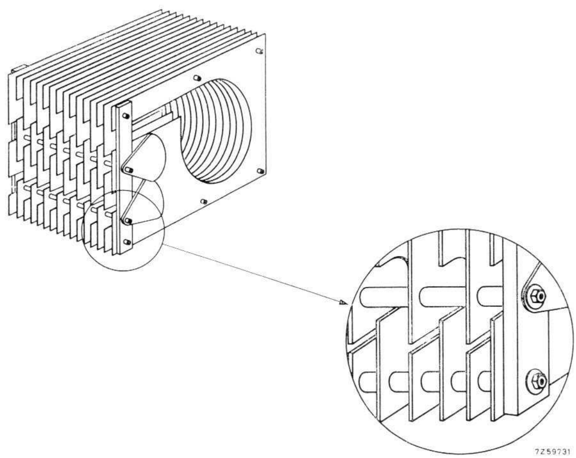


Fig. 5.19. Sectioned view of a laminated circuit.

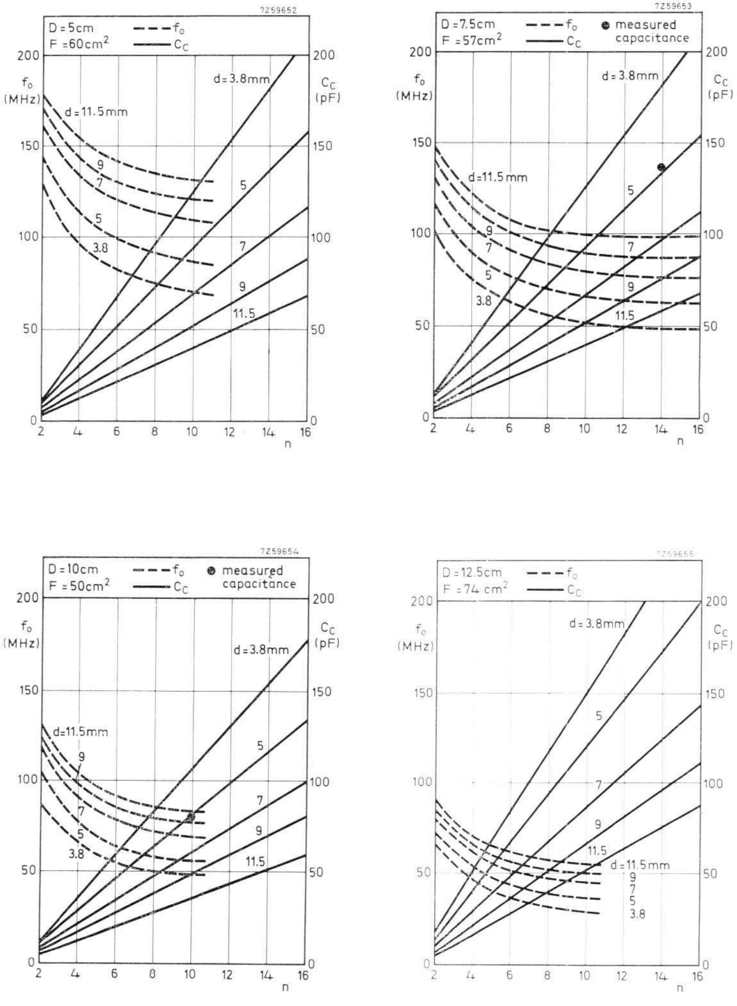


Fig. 5.20. Resonant frequency (f_0) and circuit capacitance (C_c) as a function of the number of plates (n) with plate spacing (d) as a parameter for several selected values of the cut-out diameter (D) and the capacitive area (F) of one resonator plate.

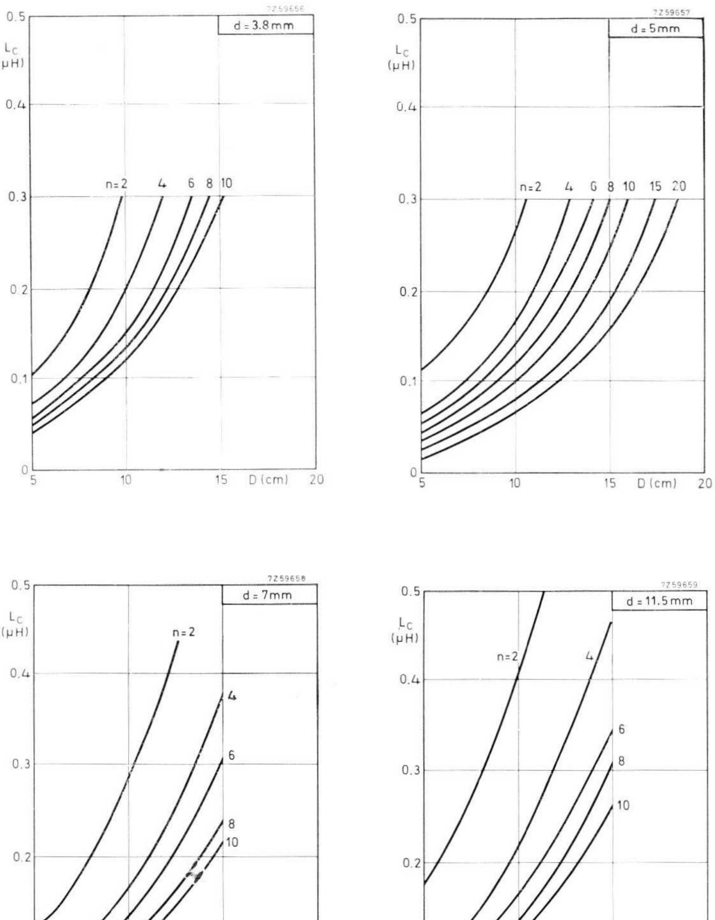


Fig. 5.21. (a). Circuit inductance (L_c) as a function of the cut-out diameter (D) with the number of plates (n) as parameter for several selected values of the plate spacing (d).

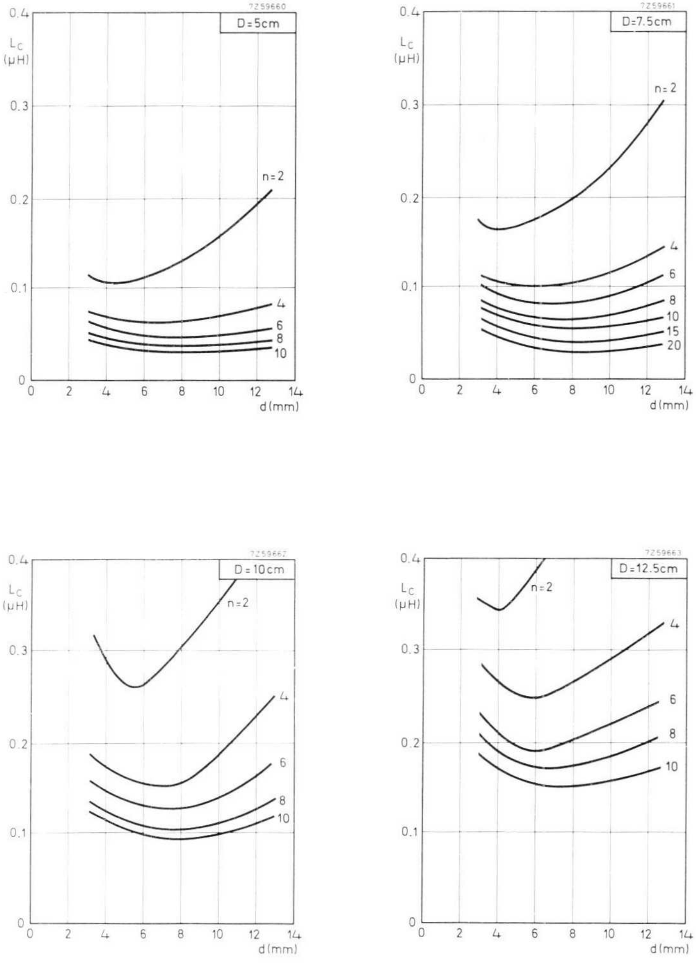


Fig. 5.21. (b). Circuit inductance (L_c) as a function of the plate spacing (d) with the number of plates (n) as parameter for several selected values of the cut-out diameter (D).

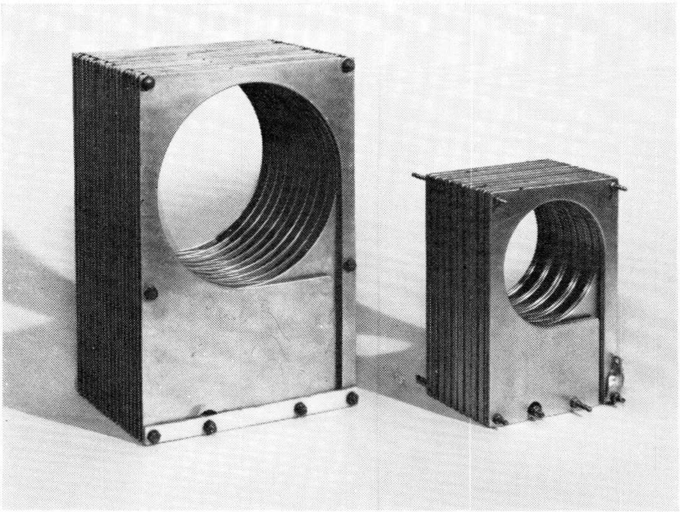


Fig. 5.22. A comparison of the laminated circuits used for 75 MHz (120 pF) and 20 MHz (560 pF).

5.2.8 Toroidal Cavity

For still lower reactive values of the frequency controlling components, the circuit inductance will become so small that a shortening of the inductive path is structurally no longer possible and its desired value will have to be realised by placing several such paths in parallel. To achieve this paralleling in a practicable form, a number of inductive loops can be arranged in a circle so that their open (capacitive) ends face the axis passing through the centre of the circle (Fig. 5.23).

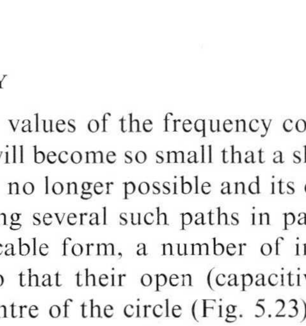


Fig. 5.23. Development of toroidal cavity by the rotation of a lumped circuit around the axis A.

Since the greater part of the circuit losses in a tuned circuit are caused by the resistivity of the inductive path, more efficient circuit performance can be obtained if the ratio of this path length to the enclosed area of magnetic field is kept to a minimum. The ideal inductor should have a circular cross-section as, of all the possible geometric shapes, it is the circle that encloses the largest possible area for a given circumference.

A toroidal cavity designed for minimum circuit losses should ideally be of circular cross-section. Since the manufacture of such a shape is rather complicated and costly, some circuit efficiency is sacrificed and the cross-section is usually made square with perhaps rounded corners. The final shape will then resemble that of a flat cylinder with the capacitive section of the circuit near its axis (Figs. 5.24 and 5.25).

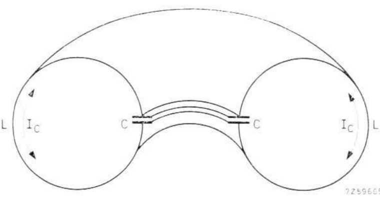


Fig. 5.24. Section of toroidal cavity with ideal inductive path length.

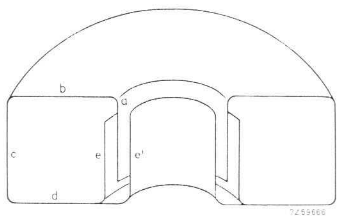


Fig. 5.25. Section through toroidal cavity tank circuit. If the area enclosed by the sides a, b, c, d and e/e′ approximates to a square, the value of the inductive component can be given by

$$L = 0.0004 \times l\,(\mu H)$$

where

$$l = a + b + c + d + e\,(cm).$$

The formula given in Fig. 5.25 remains reasonably accurate, even if more capacitive sleeves than shown in the figure are used so long as their own cross-sectional area, including the airgaps, remains small in comparison to the area enclosed by the inductive path.

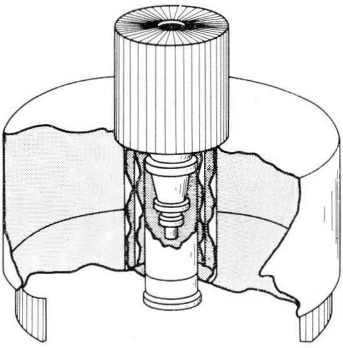


Fig. 5.26. A toroidal cavity showing one of the new ceramic triodes mounted with its cooling fins at the top of the cavity.

5.2.9 The Kolster Circuit

The Kolster circuit is, in its construction, a compromise between the capacitively loaded toroidal cavity and a similarly terminated co-axial line. This construction has a circulating current path of approximately square cross-section, except that the loading capacitance now takes the form of a plate that is parallel to one of the end faces of the circuit "box" (Fig. 5.27).

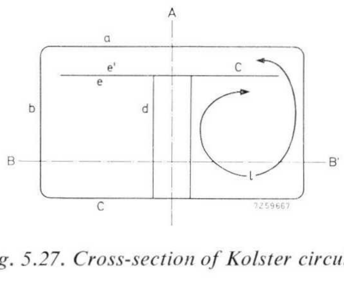


Fig. 5.27. Cross-section of Kolster circuit.

The component value for the inductive section can, with sufficient accuracy, be calculated from the formula given under Fig. 5.25 for toroidal cavities.

When the shape of the Kolster circuit along the section B-B' is square rather than circular, the calculation of b and d is based on the mean of the ratio of the inscribed and circumscribed circle of this square shape.

Because the Kolster circuit has usually only one capacitive plate, limits are imposed on the maximum loading that can be realized. For this reason Kolster circuits are normally used in the frequency range of 20 MHz to 200 MHz where a relatively low capacitance will still give an adequate loaded Q.

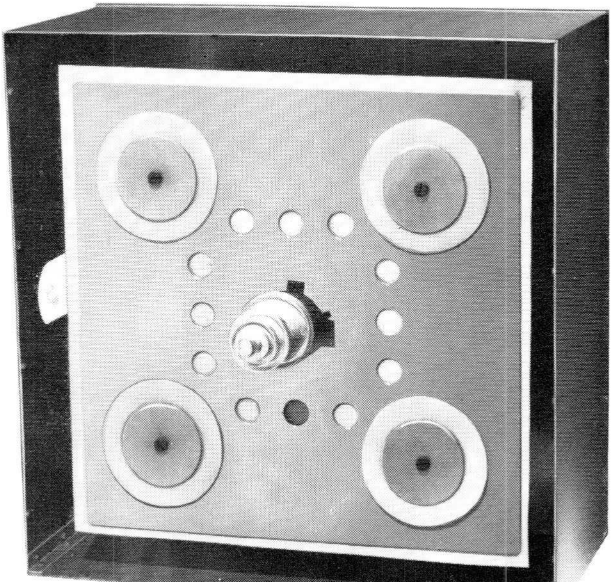


Fig. 5.28. Kolster circuit designed for operation at 85 MHz.

5.2.10 LINE CIRCUITS

Transmission lines, either open (lecher line) or co-axial, can be used as tank circuit resonators in a generator in the same way and for the same frequency range as cavity or Kolster circuits. When shortened by an appreciable amount of end-loading capacitance, there is little difference in performance or structural disposition between a Kolster circuit, a co-axial line, a lumped circuit or a lecher line. In many instances, however, lines offer a wider choice of impedance, the use of operational multiples of $\lambda/4$, the solution of transformation problems as well as the physical arrangements of tube mounting, and the application of self decoupling feed systems.

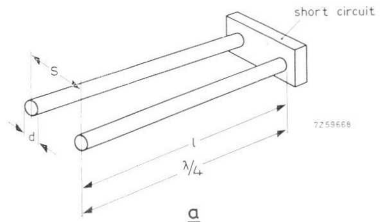


Fig. 5.29. (a). Lecher line with $\lambda/4$ characteristic.

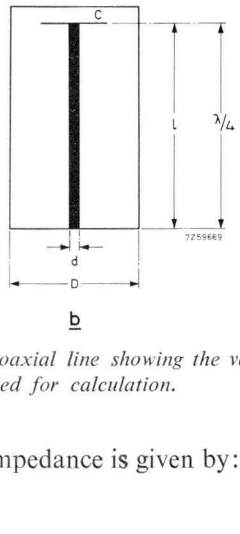


Fig. 5.29. (b). Coaxial line showing the various dimensions required for calculation.

For the lecher line (parallel wire) the natural impedance is given by:

$$Z_o = 276 \log_{10} s/d$$

where s is the spacing from the centre of one line member to the centre of the other and d the diameter of the line members (Fig. 5.29*a*).

For the co-axial line (Fig. 5.29*b*) this value is given by

$$Z_o = 138 \log_{10} D/d$$

where

D = inner diameter of the screen

d = outer diameter of the inner conductor.

The physical length of either system, when capacitively end-loaded, is

$$l = \lambda\, \Theta/2\pi$$

where

l is the physical length of line
and Θ is related to the circuit constants by

$$\tan \Theta = \frac{\lambda}{2\pi\, vCZ_o},$$

where

λ = wavelength
v = velocity of light
C = the end-loading capacitance.

If two quarter-wave systems are combined to a half-wave structure a transformer may be constructed where the voltage ratio of the capacitively loaded line ends is given by

$$T = \frac{C_G \cos\, \Theta_L}{C_L \cos\, \Theta_G},$$

(see Fig. 5.30).

A practical example of the use of such a line is given in Chapter 8 'Applications'.

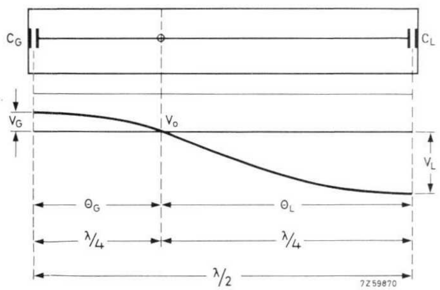


Fig. 5.30. Voltage distribution along a half-wave line with capacitively loaded ends.

5.3 Drive and Feedback Circuits

An oscillator can, in principle, be compared to an amplifier, as part of the generated output is fed back to the input of the tube. Thus, the input of the oscillator will, in many cases, closely resemble, both in terms of circuit and layout, the input of an amplifier. In other instances the comparison is not so obvious, as both feedback and driving are carried out by the same components. Some of these may have yet a third function and constitute part of the main tank circuit.

All systems must, however, present a suitable impedance across the input electrodes of the generator tube that will permit the application of a radio frequency potential of the appropriate magnitude and phase relationship.

An alternative grouping of oscillatory configurations can therefore be made when these drive and feedback circuits are considered from a structural as well as from electrical viewpoints.

5.3.1 Tuned Input-Tuned Output

Tuned-input tuned-output circuits are those most closely resembling an amplifier and may, as can be seen in Table 5.1, be separated into the three further categories of:

a. Common cathode.
b. Common grid.
c. Common anode.

The necessary feedback may, for all three types, be obtained by supplementing the appropriate inter-electrode capacitances or by the use of inductive link coupling.

The tuned input circuits themselves will structurally resemble the main tank circuits. However, their power handling ability will need to be only that fraction of the tank circuit that corresponds to the required drive power. Their C/L ratio should be considerably lower than the main tank C/L ratio to prevent the drive circuit from taking over the frequency control of the oscillator. The ideal ratio of the two respective C/L values would be about

$$\frac{C/L \text{ tank}}{C/L \text{ drive}} = 10.$$

With increasing frequency the tube internal capacitances contribute an ever increasing share to this ratio and values attainable in practice will decrease and may, at frequencies around 100 MHz, be no larger than 2. A deliberate reduction of this ratio may have to be accepted when the suppression of the higher-order harmonics becomes important and use is to be made of the low-impedance characteristic of a heavily capacitively loaded grid circuit. The designer must then decide how far the risk of oscillatory instability is acceptable. The drive circuit with the higher C/L ratio will cause more rapid changes in the feedback conditions when reactive changes in the load are reflected to the tank circuit. Normally such a circuit characteristic is undesirable, since these changes, usually downward, will produce a deterioration of the tube operating conditions with a consequent reduction of load power.

There are, however, a few exceptional instances where these changes and the resultant frequency shift can be used to advantage. If fault conditions develop in the processed materials and the load circuit or the handling equipment cause a sufficiently large reactive change in the tank circuit to de-tune it with respect to the grid circuit, the feedback will be reduced. The r.f. voltages in all parts of the generator circuit will be decreased and further damage to material or equipment prevented.

5.3.2 R.F. Potential Divider

All other oscillator circuits which have no well defined structural separation of drive and feedback circuits are considered under this heading. In some instances, the drive and feedback components will be part of the main tank circuit. Both the Hartley and Colpitts circuits as well as their variants can be placed in this group. The first derives the necessary drive voltage from a tapping on the main tank circuit inductor to the cathode (the common tube electrode). The second is the classic Colpitts circuit, where the equivalent voltage division is effected by a capacitive divider in the main tank capacitor, the lead from the dividing point again being connected to the tube cathode. In both cases substantial r.f. currents flow in these leads, and the circuits are only practical where the inductance of this lead is very small in comparison to that of the main tank inductor. Their use is therefore limited to oscillator circuits of low frequency.

The provision of adjustable drive for the tube often complicates the construction of this critical tapping point. The unavoidable additional mechanical arrangements also add further stray reactances (mainly inductive) at a position where they can be least tolerated.

For higher frequencies (10 MHz – 200 MHz) a modified version of the Colpitts circuit is often successfully used. The capacitive division is made with the help of the tube's internal capacitances, augmented, if necessary, by an external variable C_{a-f}. The only lead inductances affecting the feedback in such a circuit are those of the tube electrodes and the external C_{a-f} assembly.

Low value electrode inductances are an inherent design feature of the new ceramic triodes, and the large coaxial electrode terminations allow the electrically effective attachment of low inductance external capacitors such as the external C_{a-f} mentioned above. Excessive phase shifts between the tube electrodes and external components can thus be avoided and a better oscillator performance obtained. It is possible to realise the constructions shown in Fig. 5.31 where the additional C_{a-f} is made up in the form of capacitive probes mounted on the cathode ring and facing the anode cooler. The use of three probes in parallel further reduces the component inductance and helps to maintain optimum feedback phasing at high frequencies up to 160 MHz. An improvement in performance of tubes of higher power rating than the one shown in Fig. 5.31 can be obtained at lower frequencies using a similar feedback system.

Since such an arrangement effectively separates the feedback path over most of its inductive path length from the main tank circuit (as opposed to the construction based on the more usual Colpitts circuit) the incidence of phase and r.f. voltage changes along this path is further reduced.

Aperiodic feedback circuits are commonly understood to be of the transformer type where the tuned primary of the transformer is represented by the main tank circuit. The secondary, an untuned coil of a few turns, is connected across the input electrodes of the tube in such a way that the necessary 180° phase shift is generated. The amount of feedback required may then be set by adjusting the magnetic linkage between the primary and the secondary of this transformer. Since, however, even the untuned secondary will have its own resonance through the association with other stray reactances, the use of such oscillators is recommended only for frequencies up to about 5 MHz. At higher frequencies the resultant self resonance of the drive coil plus stray reactances would be too close to that of the tank circuit, and incorrect oscillations (probably in a *T-A-T-G* mode) could result.

Both the r.f. potential divider and aperiodic feedback circuits have the advantage of being relatively insensitive to reactive changes reflected from the load and will maintain the originally determined feedback conditions over a large operational frequency shift.

All types of feedback circuit are made up from either inductive or capacitive components. The most stable performance is obtained if care is taken to exclude the opposing reactive components as much as structurally feasible and to restrict all r.f. current paths, notably those in common with other sections of the circuit, to a minimum length.

Notes on the quarter wave chokes, used with some of the feedback systems shown in Table 5.1, are given in Ch. 5.7 as their method of construction is identical to that of chokes more widely used in supply and bias feedline filtering networks.

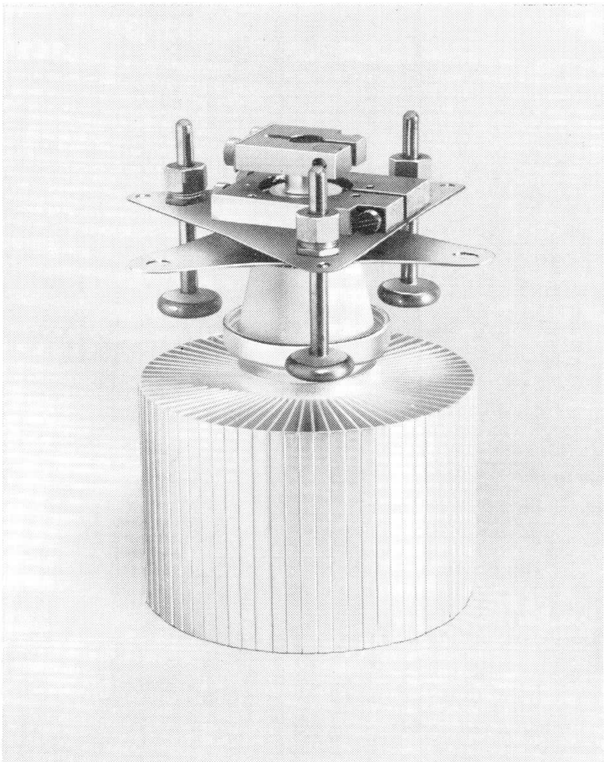


Fig. 5.31. Grid and filament connectors and feedback probes fitted to YD1150 when used as a common grid oscillator. The necessary feedback capacitance C_{af} is formed by the probes looking at the anode cooler (or mounting flange if water cooled tubes are used).

5.4 Cathode Circuits

The layout of the cathode circuits will largely depend on the type of oscillator configuration used, that is, common cathode, common grid or common anode. In each case, however, the cathode circuit must fulfil the dual function of providing a suitable radio frequency path as well as a means of supplying the necessary power to the filament at the mains supply frequency. As most industrial tubes use a thoriated tungsten filamentary cathode, both these requirements are made more exacting by the need to make this filament as unipotential as possible with respect to radio frequencies. This is important, no only to ensure the correct radio frequency operation of the tube, but to ensure an acceptable tube life.

With any type of filament, however, additional filament decoupling with a suitable capacitor across the filament terminals, is a further essential precaution to ensure this uniformity and prevent r.f. heating of the filament. Even when the frequency of operation is so low, that the reactance of filament and leads becomes negligible, such measures will help to minimise the effects of any incidental harmonic or parasitic potentials.

Capacitors used for this purpose should be of such a value and outline and be mounted in such a position that the resultant resonance, due to filament inductance, connector inductance and capacitive value, lies *below* the operating frequency. Although this will normally only be possible down to frequencies of about 1 MHz, some protection is still provided for any incidental higher frequencies. If large capacitive values, with their inherently larger stray inductive values, are necessary to bring the filament resonance to the desired frequency, these should be parallelled with other capacitors of, perhaps, lower capacitance, but with good v.h.f. characteristics.

Failure to observe these precautions may lead to absorption effects at the fundamental or one of its harmonic frequencies and the resultant circulating currents could destroy the filament. A wrongly dimensioned circuit could also act as an effective cathode grid impedance and cause, if an unsuspected stray feedback path is present, parasitic frequencies to be generated. It can, therefore, in certain circumstances be expedient to decrease the resonant frequency of the filament and its decoupling capacitor to the desired value by mounting the latter a few centimetres further away from the tube terminals, rather than increasing its capacitance. This decreases the C/L ratio of the circuit and reduces the possibility of magnetic coupling to other circuits in the vicinity. A final check on the resonance of the filament decoupling circuit should be made with the help of a grid dip meter. This will also give a rough indication of the location, direction and strength of the magnetic field of the circuit.

Fig. 5.32. Grid-cathode circuit and feedback assembly. Three ceramic capacitors (total 140 pF) are grouped around the tube header and connected between filament (cathode) and grid plates. The three adjustable rods holding the feedback ring can be seen protruding from the collets on the filament (cathode) plate.

5.5 Blocking Capacitors

Blocking capacitors, whether in anode, grid, or cathode circuits, are used to hold off supply voltages from other parts of the circuit but are, at the same time, required to form no major obstruction to the flow of r.f. currents. They will, therefore, have the appropriate supply voltage rating and be capable of carrying the radio frequency currents in their part of the oscillator circuit. Their VA rating with respect to the operating frequency need not however be large since their reactance, in series with the main components, should be so small that the consequent r.f. voltage drop is negligible in comparison to the tuned circuit r.f. voltage.

5.5.1 Location

The preferred location of any blocking capacitor is between a tube electrode and its corresponding circuit section, rather than in a part of the circuit carrying substantial reactive currents.

From Figure 5.33 it is evident that example (*a*) would be electrically as well as economically the preferred solution. This ideal positioning of the blocking capacitor will, however, generally be possible only at relatively low frequencies, say up to 30 MHz. With increasing frequency, where the tube capacity forms an increasing proportion of the total tank circuit capacitance, some, and eventually all, of the reactive circuit currents will have to flow through the blocking capacitor.

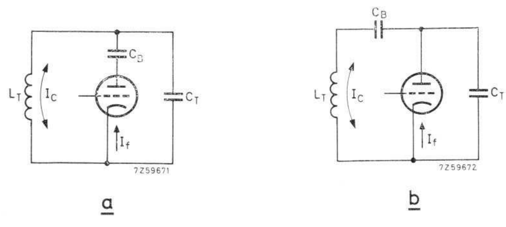


Fig. 5.33. Two possible locations for an anode blocking capacitor are shown. In (a) the blocking capacitor C_B is required to carry only the fundamental current through the tube, holding the d.c. supply potential off the remainder of the tank circuit. In (b) C_B must be dimensioned to carry the total circulating current and would probably have to be of a higher capacitance to prevent an undue r.f. potential drop. C_T would have to withstand the d.c. supply voltage in addition to the r.f. peak voltage.

At these higher frequencies the inherent component inductance of the blocking capacitor can constitute a significant proportion of the total circuit inductance, and most types of ready manufactured components may prove to be unsatisfactory in this respect. Individually constructed items, that have been designed to fit into the circuit layout, will give a better performance.

5.5.2 Choice of Capacitor

For both low and high frequency circuits the major dimension of the blocking capacitors should be at right angles to the flow of the r.f. currents. This will minimize the inductive effects of the physical structure of the blocking capacitor. When ready made components are used, the paralleling of a number of capacitors with lower individual capacitance and current rating can often offer a better solution than the use of a single capacitor. The advantages of such a construction are the same as those outlined for the tank circuit capacitors; the individual component resonances are much higher and likely to be at frequencies well into the transit time region of the tube. Their integrated resonance will, because of the small but inevitable manufacturing tolerances, be of a much lower Q than an equivalent single capacitor would produce. The current densities through the individual capacitors can also be substantially reduced.

For lower frequencies, banks of ceramic capacitors or compact mica disc stacks are suitable. For higher frequencies blocking capacitors can be made in the form of sandwiches (Figs. 5.34 and 5.35), where the two conducting plates are manufactured from aluminium or copper sheet, spaced by a sheet of high grade, low loss insulating material such as Teflon or polypropylene. If one of the metallic plates is well integrated with the appropriate tube electrode connector a most effective sandwich can be formed. If this connector is also of the large area coaxial type the residual component inductance will be very low indeed.

The danger of ionization or arcing, in any accidental air pockets between the metallic surfaces and the dielectric, is negligible if the capacitive reactance is low enough to limit the r.f. voltage drop across the capacitor to a few hundreds of volts.

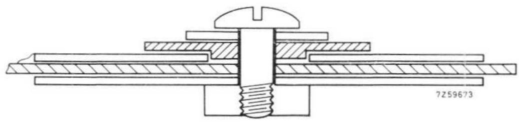
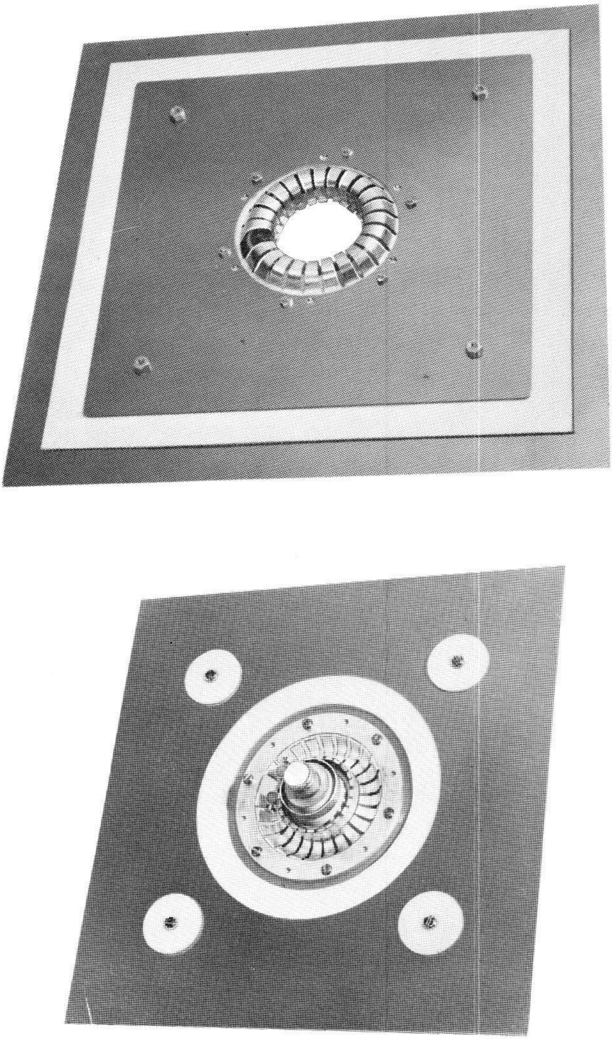


Fig. 5.34. Sandwich capacitor. Shaded areas are Teflon or other high grade insulating materials. Bolts are used in preference to screws because they gave a smoother surface in the area of maximum voltage stress.



a

b

Fig. 5.35. The front (a) and reverse side (b) of a sandwich type grid blocking capacitor used in an 27 MHz Kolster circuit.

5.6 Filtering Inductors

Any two points in a radio frequency circuit that are of different r.f. potential, but are intended to have the same d.c. potential, may be connected by a suitable inductive reactance. Ideally this reactance will represent an infinite impedance at the operating frequency and therefore cause no fundamental power losses, and minimum resistance to the supply currents. Any parallel tuned circuit with a quarter wave characteristic or, in exceptional cases, an odd multiple thereof, could serve this requirement. It must also have a low C/L ratio and be broadly tuned so that it will be effective over a wider operational frequency band without showing unduly sharp resonances near the fundamental frequency or any of its harmonics. Sharp resonances in such a component would render it more critical with respect to the operating frequency and possibly induce high circulating currents that might, in the extreme case, lead to its destruction through overheating. The r.f. power thus dissipated would be lost to the useful load power of the generator. Any additional sharp resonances at higher frequencies could give rise to parasitic ringing or the accentuation of certain harmonic frequencies.

5.6.1 Quarter-Wave Chokes

A suitable parallel tuned circuit with the lowest possible C/L ratio is then an inductor that will resonate with its inter-turn and stray capacitances at the required frequency. Such a component is known as a quarter wave choke since its inductive component is predominant and the r.f. voltage distribution parallel to its axis will follow a quarter wave pattern. If a further broadening of its tuning characteristic is desired the C/L ratio can be reduced by increasing the spacing of the turns or increasing its length to diameter ratio.

In many cases, the broadening of the response curve may be achieved by resistive damping; i.e. the wire used for winding the choke can be a resistive material, or a resistor can be connected in parallel with the choke windings. Such measures are, however, only recommended as a last resort, since, whilst suppressing unwanted resonances, they will dissipate a certain amount of fundamental r.f. power.

The formulae for the calculation of quarter wave chokes that have been published are essentially theoretical. In practice however, a very small change in inter-turn or stray capacitances changes the calculated value of the resonance appreciably. These capacitances are, to a large extent, determined by the electrical properties of the wire insulation (if any) and the material and shape of the supporting coil former. The completed choke will thus inevitably require a certain amount of trimming by trial and error.

A quick and simple choke construction may be carried out at follows. Since the length of wire required will be less than the physical half-wave length of the operating frequency, a piece of wire of adequate gauge to carry the supply current may be cut to this length and wound on a prepared former. The resultant quarter wave resonance will thus be found to be somewhat lower than required. Next, shorten the length of the choke, in discrete steps, by removing part or complete turns. At the beginning and throughout each step of this process the resulting resonance can be conveniently checked with a grid dip meter until the required frequency is obtained. The lengths of wire unwound from the coil former when reducing the turns, should be cut at each step to avoid erroneous results, and the coil should rest on an insulating support of low dielectric constant when making absorption checks. It is important to remember that the inductances of both end connections of the choke wire, used for connecting the choke into circuit, form part of the total inductance of the choke.

Any inductor of this type will show several resonance points, usually in a harmonic relationship, and the one lowest in frequency is the required quarter wave resonance. A well designed oscillator will operate correctly if these quarter-wave chokes are within $\pm 20\%$ of the operating frequency.

If malfunctioning of an oscillator fitted with quarter wave chokes wound to this tolerance is observed, the fault is likely to be in some other part of the oscillator circuit, rather than in the chokes themselves.

The correct functioning of these chokes can finally be verified during operation, by passing a neon probe along their surface and parallel to their axis. A change of ionization, corresponding to a quarter wave voltage distribution, should be visible in the neon probe.

5.6.2 Self Decoupling Systems

Filtering chokes do not perform any of the basic functions in an oscillator circuit. Their inherent stray reactances may, however, be responsible

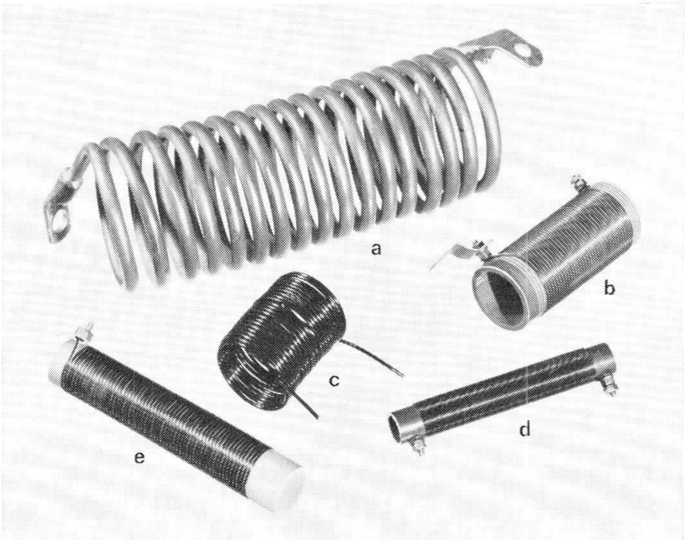
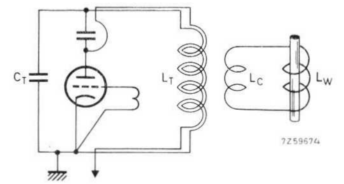


Fig. 5.36. Various types of filtering chokes that are normally used in their λ/4 mode. (a) A 27 MHz filament choke for 140 A filament current of self supporting construction. (b) A 27 MHz choke on a threaded Paxolin former. (c) A 40 MHz self supporting enamelled wire choke. (d) A broadband 27 MHz choke with a high length to diameter ratio. (e) A 40 MHz choke wound on a Teflon former.

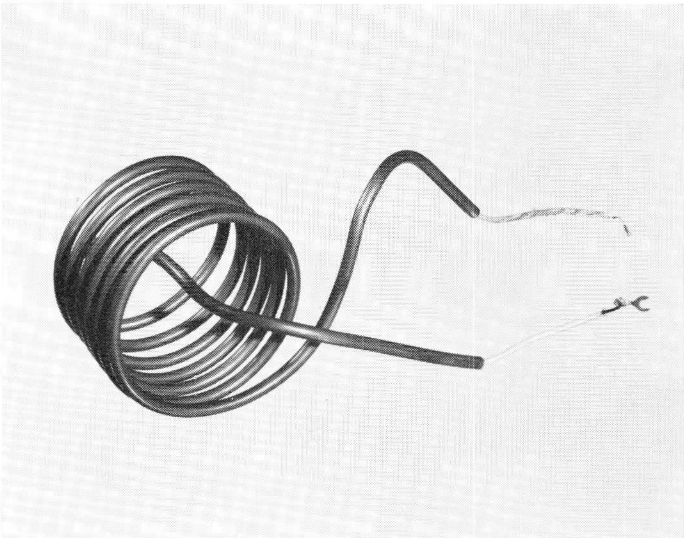
for faulty operation of the oscillator. It is, therefore, good practice to avoid their use wherever possible.

In many instances these chokes can be omitted and their function assumed by the natural r.f. voltage gradient along the circuit elements, giving better oscillator performance, more effective decoupling, and a more elegant circuit layout.

Using this method, supply or bias leads can be dressed to follow the contours of coils, cavities, lines or flat sheet inductors or be fed through metallic pipes or channels. The leads must, however, be terminated at a point that has, as a function of its position in the oscillatory circuit, a minimum r.f. voltage with respect to the nominal r.f. earth reference. Often the same r.f. path can be shared by several leads; for example, filament and bias cables. Figure 5.37 shows the essential details for such a layout.



a



b

Fig. 5.37. (a). Total self decoupling of anode supply line in induction heating generator with separate load circuit. At the point of emergence from the oscillatory circuit the h.t.line is free of radio frequency and no further decoupling elements are necessary. (b) Inductor designed as feedback for a low-frequency induction heater using the self decoupling method. The Fibreglass covered grid lead is fed through the centre of the inductor thus avoiding the use of separate grid chokes.

To avoid resonance effects between feedlines and the corresponding r.f. conductor, the feedlines should be screened cables, or fed through copper pipes dressed along the r.f. conductor. If the latter is of suitable dimension and shape, perhaps itself a pipe or other hollow conductor, the supply can be fed through its hollow centre. In addition, decoupling capacitors may be required at the point of entrance or emergence (or both) of the supply lines. This precaution becomes particularly important if the resonance of supply line – r.f. conductor is close to the operating frequency.

In the applications described in Chapter 8, extensive use has been made of such decoupling methods, simplifying the circuits considerably as well as materially contributing to their troublefree performance.

Since, in Fig. 5.38, the work coil must be interchangeable at the points x, x', the h.t. lead cannot be made to follow the remainder of the tank inductor path represented by L_w. Self-decoupling is therefore used up to the point x, whence the residual decoupling is effected by a choke. The dimensioning and the operational influence of this choke is however much less critical since the r.f. swing across it is small and the choke itself is well removed from the anode, where its stray reactances could otherwise exert disturbing influences.

A laboratory example of the use of "self decoupling" for the filament and grid bias leads on a 27 MHz, 25 kW Colpitts oscillator with grounded anode is shown in Fig. 5.39. Both cables follow the contour of the main circuit inductor and emerge from their tubular screening below the level of the earth plate. At this point they are completely free of r.f. energy although the r.f. swing between cathode and anode is some 7 kV in this oscillator.

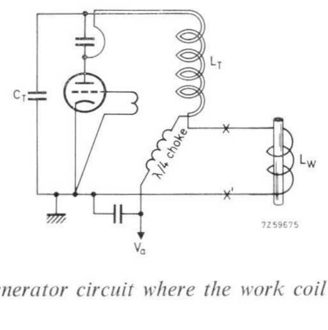


Fig. 5.38. Schematic of generator circuit where the work coil is a part of the tank coil.

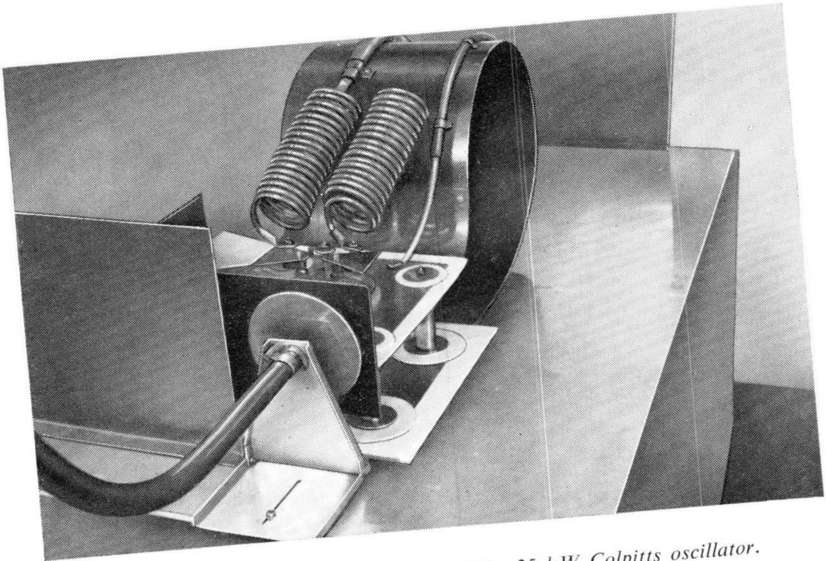


Fig. 5.39. Self decoupling on a 27 MHz, 25 kW Colpitts oscillator.

5.7 Supply Line Filtering

5.7.1 Capacitive Filtering

All supply and metering leads associated with an r.f. generator must be kept free from radio frequency energy to prevent damage to other circuits, the loss of fundamental power, and the generation of spurious feedback paths. To achieve this, filtering circuits are normally incorporated. Filter circuits, in their simplest form, need be no more than a low reactance path at radio frequencies across the leads in question or a similar path to the nominal earth potential of the generator, i.e. the chassis or cabinet. Both the inductive and capacitive series reactance of the component used must be as low as practicable.

Though many types of capacitor can serve this purpose, the structural and electrical advantages that can be gained by the use of feedthrough capacitors cannot be over-stressed. Their large connecting flanges of extremely low inductance make an effective earth connection when mounted on metal panels and partitions. Their central bolt will further reduce lead inductance as it is unnecessary for separate wiring to be fed through channels or cut-outs in partitioning.

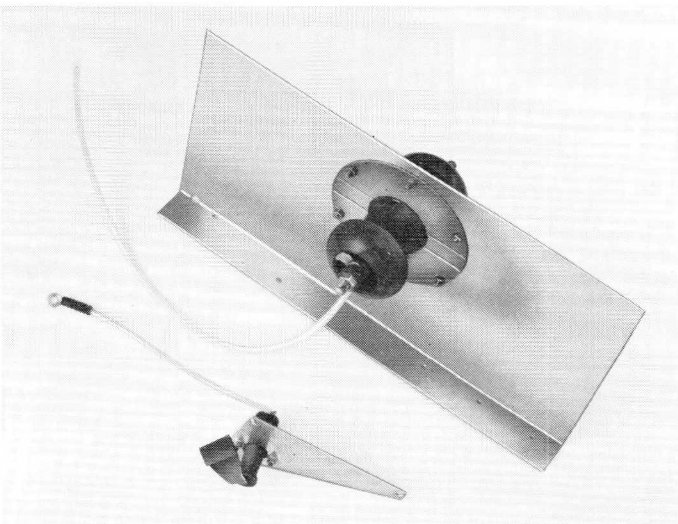


Fig. 5.40. Two types of feedthrough capacitor mounted on panels.

The screening effect of feedthrough capacitors can be so thorough that all the r.f. currents in a particular location will flow on the 'live' side of the partition only, and the reactive load may be shared unevenly between the 'inside' and 'outside' halves of the capacitor. In this case a capacitor of higher nominal current rating, or higher capacitance to reduce the reactive r.f. potential drop, may then be required.

At lower frequencies where larger capacitances than those obtainable in feedthrough types are required, the paralleling of ordinary capacitors with feedthrough ones will help with the prevention of higher frequency parasitic oscillations or the suppression of higher order harmonics.

5.7.2 Composite Filters

Although in many cases the provision of capacitive filtering alone may be adequate, it is good practice to use a combination of both inductive and capacitive filtering elements in the form of inverted L or π filters. Where the protection of sensitive auxiliary equipment or the suppression of low level radiation is important, several such filters may be used at critical points along the supply lines. The effectiveness of each will,

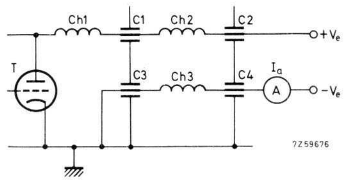


Fig. 5.41. Example of double filtering network in the h.t. supply of a generator. $Ch_{1,2,3}$ *are* $\lambda/4$ *chokes or other suitable inductors.* $C_{1,2,3,4}$ *are feedthrough capacitors mounted on partitioning panels. The negative supply line is also filtered and connected near the tube cathode.*

however, depend entirely on the true relation of its 'cold' side to the nominal earth potential points. Figure 5.41 shows an example of this filtering technique applied between the r.f. generator and its power supply.

5.8 Parasitic Suppression Circuits

In the few cases where the parasitic oscillations cannot be eliminated by changes in layout or the use of other components, parasitic suppressor circuits will have to be employed as a last resort. Their main function is to provide a high impedance path in series with, or a low impedance one in parallel to, other circuit elements at the parasitic frequency. In some instances the insertion of a purely resistive element near one of the tube electrodes may be sufficient. If the resultant fundamental power losses are unacceptable, this resistor may be shunted by a suitable inductor, which must have a low reactance at the operating frequency but much higher inductance at the parasitic frequency. In other cases, insertion of a low impedance path, relative to the higher parasitic frequencies, in the form of a capacitor between two of the tube electrodes, may be adequate. It follows, however, that such methods can be used only where the fundamental and parasitic frequencies are sufficiently far apart, so that excessive fundamental frequency changes, changes of drive conditions and resistive losses are avoided.

5.8.1 Resistive and Capacitive Stoppers

For the suppression of more violent parasitic oscillations, or those that occur in the neighbourhood of the fundamental, heavily damped tuned circuits are commonly used. Depending on their position in the circuit these can be either series or parallel tuned (Fig. 5.42). The tuned circuits should have a C/L ratio as low as practicable and be suitably damped to give a broadband effect.

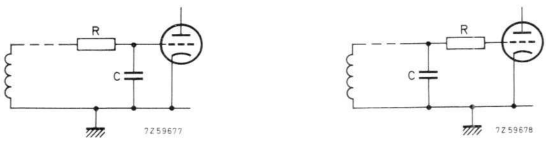


Fig. 5.42. The diagrams show simple RC suppressors used when fundamental and parasitic frequencies are well separated. The resistive component need often be no more than a nickel alloy strip with a d.c. resistive value of a fraction of an ohm and therefore consume only little fundamental power.

5.8.2 Tuned Circuits

The necessary damping can be achieved by paralleling the antiparasitic circuit inductor with a resistor, or by making the inductor from resistive material. For very high frequencies this damping effect can sometimes be provided by suitable choice or shaping of the inductor surface. A matt surface rather than a polished surface, or a threaded rod instead of a straight rod, may provide sufficient surface resistance.

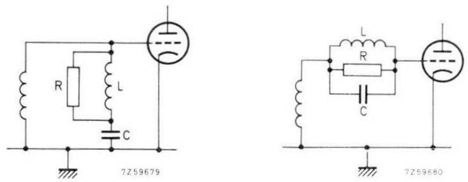


Fig. 5.43. Series and parallel tuned suppressor circuits.

Either of the two suppressor circuits shown in Fig. 5.43 can be equally effective, whether mounted in the grid or the anode circuit of the tube. In the first case fundamental power losses will, however, be less and the circuits may be assembled from components with a lower resistive and reactive rating. The C/L combination, as low a ratio as possible, should resonate at the parasitic frequency.

5.8.3 Current Obstacles

A further method of parasitic suppression, suitable for cavity or line type circuits, is the insertion of r.f. current obstacles at right angles to the probable flow of the parasitic currents. The advantage of this method is that it can be used up to frequencies of several 100 MHz, when there is little difference in the frequencies of the fundamental and the parasitic mode, without noticeably reducing fundamental power.

Fig. 5.44 shows a toroid cavity (or line) in which the fundamental current flows in a radial pattern as indicated by the arrows. The parasitic mode has a circumferential current path (– · – · –) that can be broken by radially disposed vanes without interfering with the radial mode currents. Narrow slots, parallel to the fundamental current flow, can have the same effect, but their use is not always practicable since they may cause radiation as well as a weakening of the structure.

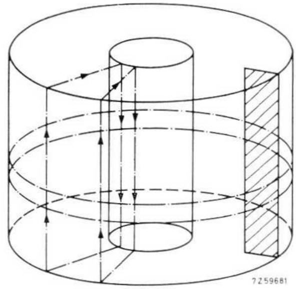


Fig. 5.44. Toroid cavity.

The same suppression system may, of course, also be used to eliminate the intercomponent resonances existing between the units of a composite tank circuit capacitor or between the latter and the generator tube (Fig. 5.45).

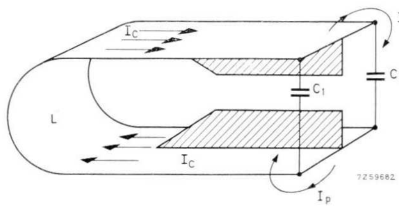


Fig. 5.45. Antiparasitic obstacles in a conventional LC circuit. L represents a half turn strip inductor and C_1 and C_2 two capacitors forming the total circuit capacitance or one tank circuit capacitance plus tube capacitance. I_c is the path of the fundamental circulating current and I_p that of the probable parasitic current. Since the two current paths are at right angles to each other, vanes may be mounted, as shown in the sketch, parallel to the fundamental current path thus forming an effective obstacle against the generation of parasitic oscillations.

5.9 Earthing Points

Earthing points are, as already defined in Ch. 5.1.1 on basic oscillator circuits, those points on r.f. generators that are deliberately made to be of the same r.f. voltage as the surrounding structure. Reasons have been given in Ch. 5.1 why the common electrode is usually preferred for this function.

The choice and construction of the earth point can critically influence the performance of a generator. The word 'point' is, strictly speaking, a misnomer because any practical construction will have a finite dimension that, however small, will still produce an r.f. potential gradient when r.f. currents flow through it. Operationally satisfactory earthing connections can, however, be made if the necessary structural components are of the lowest possible inductance and parallel capacitive paths from other parts of the circuit are kept to a minimum.

At lower frequencies, for example, where the common electrode is often the cathode, all other connections with the same nominal r.f. voltage, can be brought back to this point. A plate or sub-chassis, no larger than necessary to accommodate the essential components, of low inductance, good conductivity and well spaced from the main chassis or cabinet structure, can be a suitable connecting point (Fig. 5.46). It will also be of negligible dimensions with respect to the wavelength of higher order or parasitic oscillations. Even under such near ideal conditions, the components placed on this sub-chassis should preferably be positioned so that the current paths, associated with their various circuit functions, will not intersect.

This sub-chassis can then be bonded to the main structure or enclosure with a single low inductance strap, all other supports being made from low-loss insulating material. This arrangement reduces the risk of returning stray r.f. currents from any part of the main enclosure to the feedback circuit.

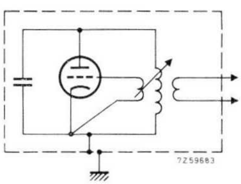


Fig. 5.46. Schematic diagram of oscillator where all connections at nominal earth potential are returned to a small sub-chassis.

At higher frequencies, where cavity or line type circuits are used, the earthing point problem is somewhat simpler since most of these circuits are self-screening. Provided that such a circuit is correctly designed and adjusted, its operation will not be affected by making the earthing connections to any point on its outer skin.

Loads with large reactive changes during the processing cycle are common in industrial heating. At the higher frequencies the reactive changes may cause standing waves on the feedline between the generator and load position. The effective earthing of the load (operating position) is therefore important to prevent possible injury to the operator or malfunctioning of the circuit. These risks will be reduced if the bonding connection between the generator and load position is of very low inductance compared to the remaining structure. Earthing points on load cables or at the work position are treated more fully in the next chapter under the heading "Load earthing points".

5.10 Grid Biasing Circuits

5.10.1 Resistors

Industrial generators, the majority of which operate under self oscillating class *C* conditions, require a negative grid bias to hold the tube at the desired operating point on its characteristic. A simple bias source can be provided by the insertion of a resistor between the grid circuit and the cathode of the tube. Since this resistor is outside the radio frequency functions of the oscillator circuit, any component, of a suitable ohmic value and power rating, can be used. For higher powers, substantial savings can be achieved by a careful choice of this component and the method of disposal of its dissipated energy.

Air blown resistors of large surface area, often in the shape of resistance mats, are frequently used. If a cooling water supply is available, a worthwhile space saving can be made by using water- or oil-and-water-cooled resistors. A very economic solution is, for example, the use of commercially produced immersion heater elements. These are made in a wide range of operating voltages and dissipations and are readily available. The correct ohmic value and dissipation capability can be obtained by suitable series parallel arrangements of the individual units.

The electrical characteristics of all our new ceramic triodes are such that the off load grid current and, therefore, grid dissipation are compatible with the recommended full load operating conditions and the grid dissipation limits. However, certain applications may require that a limit is set on the maximum grid current. One method of limiting the maximum grid current is to construct the grid resistor, partly or wholly, of incandescent lamps. In most lamps the resistance of the filament increases with increasing temperature and limits the flow of current. By suitable choice of lamp ratings a reduction of the maximum grid current of about 20 % can be achieved. For optimum regulation the lamps should show the largest possible brightness variation over the range of oscillator loading changes. The lamp rating should be chosen so that only about 80 % of it is used at the highest dissipation likely to occur.

5.10.2 Regulator Tubes

For applications where an accurately controllable grid current is required, a tube may be used as grid resistor (Fig. 5.47). Such a regulator tube will have a purely d.c. function and be connected with its cathode toward the grid circuit and its anode to the cathode of the oscillator tube. Its internal resistance, equivalent to the grid resistor value, can then be set by the application of an appropriate d.c. voltage between grid and cathode.

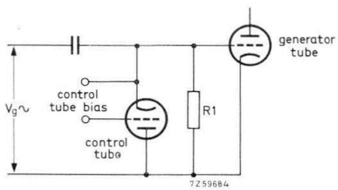
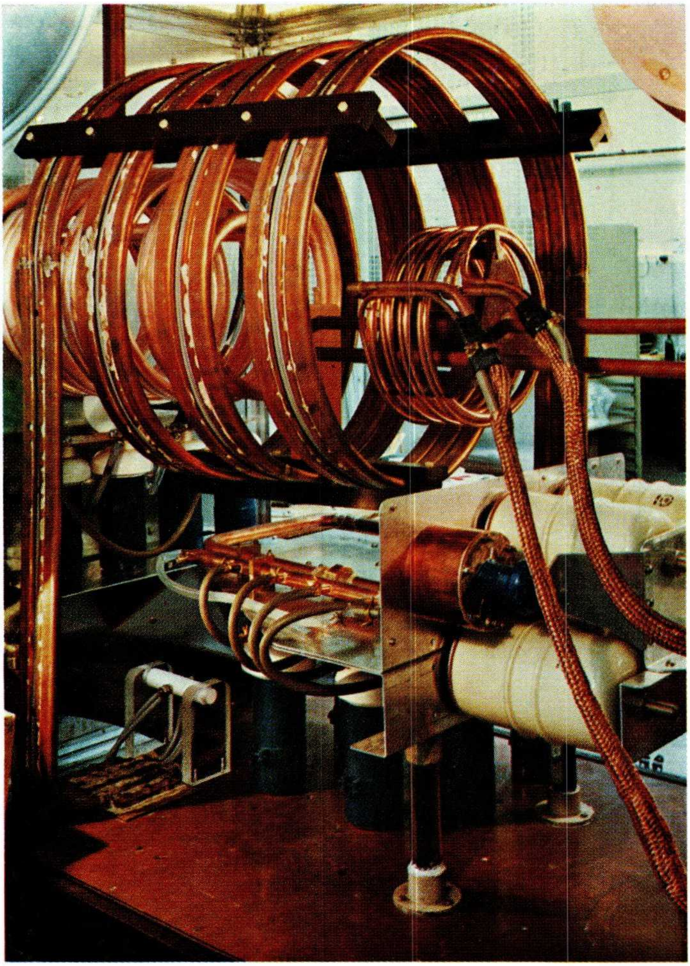


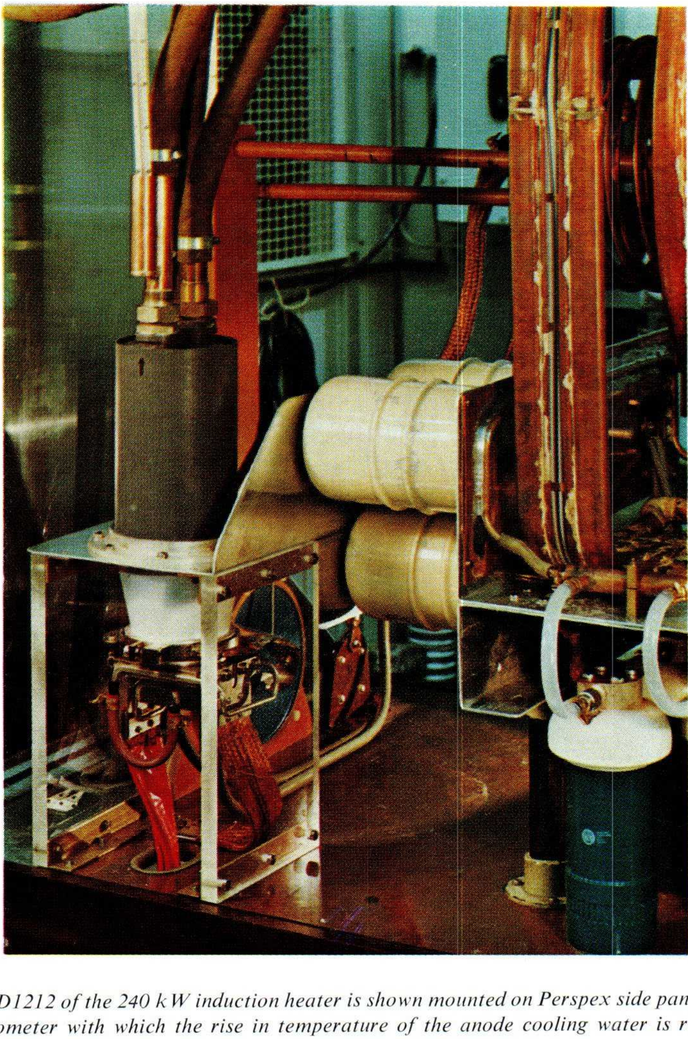
Fig. 5.47. Circuit of tube controlled grid resistor. R_1 is in parallel with the control tube and should have a value several times that required for normal operation. Its purpose is to provide a grid-cathode path for the oscillator tube in case of control circuit failure. The r.f. section of the circuit is not shown.

5.11 Grid Switching Circuits

Switching operations, required for process timing or safety reasons, are normally carried out by circuit breakers on the mains side of the anode power supply. They can also be carried out applying a negative bias to the grid of the generator tube in excess of the operational grid swing.



General view of the oscillator circuit of a 240 kW, 300 kHz laboratory model induction heater (see also Ch. 8.1). The tank coil, and the grid coil with its flexible braid connections are clearly shown.



The YD1212 of the 240 kW induction heater is shown mounted on Perspex side panels. The thermometer with which the rise in temperature of the anode cooling water is read can be seen attached to one of the water feed pipes.

A grid switching action is many times faster than the conventional circuit breaker. The switching takes place within a few cycles of the operating frequency and will, therefore, allow more accurate process timing as well as preventing damage to costly load coils or handling equipment in the event of an r.f. flashover at the load position. With grid switching, the mains circuit breakers are used only for the initial switching on or re-setting operations and the wear on these components is greatly reduced.

To make grid switching an economic and reliable proposition it is essential that the grid characteristic of the generator tube permits the use of a cheap biasing circuit. The non-emissive *K* grids are ideal for this purpose. The complete absence of grid emission permits the use of relatively low current supplies for these biasing purposes, as, of course, no grid emission currents will have to be counteracted.

A circuit suitable for the grid switching of an oscillator is given in Fig. 5.48, and an oscilloscope trace showing the decaying r.f. amplitude obtained when switching a 400 kHz oscillator is given in Fig. 5.49.

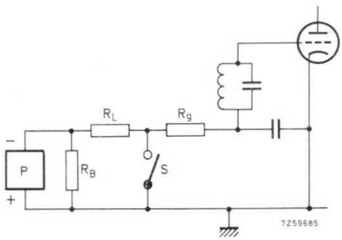


Fig. 5.48. Basic grid switching circuit.

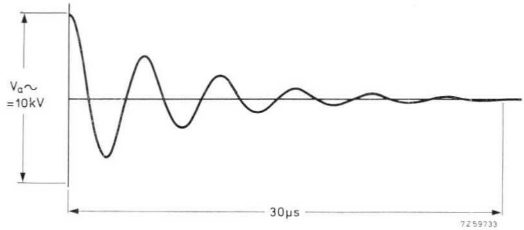


Fig. 5.49. Oscillogram of decaying r.f. anode swing of a 400 kHz oscillator after application of switching bias. The fundamental amplitude has decayed to an insignificant value within 30 µs.

During normal operation switch S, which could for example be a thyratron, is closed and the tube grid current will flow through R_g. In this condition power supply P will have to feed the limiting resistor R_L and the ballast resistor R_B. When S is opened, the full voltage supplied by P will appear at the grid of the tube since the grid is non-emissive. With a biasing potential slightly in excess of the peak drive swing, the grid and anode currents of the generator tube will cease and the power supply P will have to feed only its ballast resistor R_B. The value of $R_g + R_L$ should not exceed that of $R_{g\ max}$ given in the published tube data.

When using grid switching systems the regulation characteristics of the anode power supply must be taken into account and the rating of d.c. and r.f. components must be compatible with the no-load voltage. Power supplies with good regulation are therefore more suitable for this purpose.

5.12 Arc Suppression Circuit

To prevent damage to the work material, or the welding electrodes if an arc is struck, the incorporation of an arc suppression circuit is worthwhile.

A suitable circuit is given in Fig. 5.50. The voltage developed across R_1 is applied to the working electrodes C_w through a $\lambda/4$ filtering choke. It also holds the thyratron Th in the non-conducting state. When an arc pierces the work material, an additional resistance R_p appears in parallel with R_1, reducing the voltage and causing the thyratron to conduct. The consequent voltage changes on R_3 can then be used to activate a grid switching bias supply that will render the generator tube inoperative.

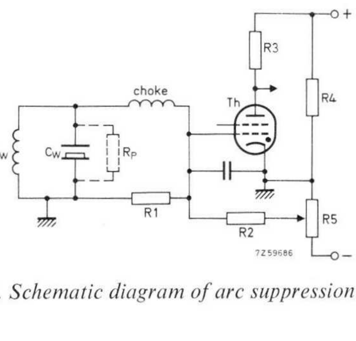


Fig. 5.50. Schematic diagram of arc suppression circuit.

6. Load Coupling Circuits

6.1 Requirements

Load coupling circuits, whether used for inductive or dielectric heating, serve three purposes.

1. To make the r.f. energy generated by the oscillator available at the position of the work material.
2. To provide a match between the electrical properties of the load material and the impedance of the generator tube.
3. To provide a physical match between the generator components and mechanical handling arrangements for the load material.

6.1.1 Power Formulae

The power requirements for both groups of applications may in the first instance (disregarding incidental electric and thermal losses) be calculated with the help of the standard calorimetric formula, where the power re-required for a certain process is given by

$$P_w = \frac{(T_2 - T_1) \times m \times S \times 4.2}{t}, \tag{5}$$

where

P_w = required power (W)
T_1 = temperature of work material at start of process (°C)
T_2 = temperature of work material to be attained at completion of process (°C)
m = mass of work material in grammes (g)
S = specific heat of work material (cal/g.degC)
t = processing time in seconds (s).

It will be found that the value calculated will be much closer to the actual requirements in the case of dielectric heating. In these applications the thermal losses incurred in terms of heating time, radiation due to elevated temperatures and ultimate required temperatures are much smaller than for induction heating. For dielectric heating the derived load power figure may have to be increased by about 10%, but an increase of some 30% may be necessary for induction heating. Load circuit coupling losses must, of course, in both cases be added to this figure.

Having derived the power requirements, it remains to relate these to the work material properties and the appropriate magnetic (induction heating) or electric (dielectric heating) fields. The formulae for both applications then give an immediate appreciation of the processing requirements in electrical terms, when changes of work material properties due to temperature changes are ignored in the first instance.

Thus it can be deduced that

Induction Heating		*Dielectric Heating*	
$P_w = H_o^2 \cdot \mu \cdot f \cdot K$	(6A)	$P_w = E^2 \cdot \pi \cdot \varepsilon' \cdot f \cdot \tan\delta$	(6B)

where

P_w = power absorbed by a unit volume of work material

and

H_o = peak magnetising force
E = peak r.f. field strength
μ = permeability of work material
ε' = capacitance per unit volume
f = operating frequency
$\tan\delta$ = loss factor of work material.
K = a factor depending on the material used and containing the resistivity ϱ and penetration depth δ.

When the material properties are given and the operating frequency has been chosen, the chief operating factors influencing the circuit design are the magnetising force H_o (induction) and the field strength E (dielectric).

6.1.2 Circuit Layout

Load coupling circuits, like any other circuit, are assembled from inductive and capacitive elements. The circuit reactances (real and stray) will produce a number of resonances that may, by coupling into the tank circuit and feedback system, cause fault conditions in the oscillator. These faults can show in the form of accentuated harmonics, parasitic ringing or downright parasitic oscillations.

Fundamental and stray C/L ratios of load coupling circuits should, therefore, be kept as low in comparison with the tank circuit C/L ratio as the loaded Q and application will allow. The effect of stray capacitive coupling paths between the loading circuit and the remainder of the generator can be cancelled, if necessary, by the insertion of Faraday screens. To avoid stray coupling paths, particularly in the feedback system of the generator, the load circuit earthing point will have to be carefully chosen and preferably made near the load position and referred to the external structure of the generator. For greater distances, some form of transmission line can be used and, in such cases, the earth return should be to the line-outer. Lines of higher characteristic impedance ($Z_o > 50\ \Omega$) are more suitable for this purpose as their inherent Q is lower and is less likely to cause oscillator moding through resonant line effects under fault conditions.

6.2 Inductive Loads

To relate the general induction heating formula to the electrical constants of the load circuit, H_o must be expressed in ampere turns and the formula elements converted to definite work-piece dimensions and material properties. The necessary ampere-turns to dissipate a given power are:

$$n\,I = \sqrt{\frac{P_w\,\delta_w\,l_w \times 10^9}{\pi\,\varrho_w\,d_w}}, \qquad (7),$$

where

P_w = required work power (kW)
δ_w = penetration depth (cm)
l_w = length of work-piece (cm)
ϱ_w = resistivity ($\mu\Omega$·cm)
d_w = diameter of work-piece (cm).

The load can thus be seen as the secondary of a transformer. As the primary will have to be of a different diameter (generally larger) a coupling factor is introduced. This can, with any degree of accuracy, only be given for regular shapes when work coil and load are completely meshed and will, for all practical purposes, be

$$k = \frac{d^3}{D^3},$$

where
d = diameter of work-piece
D = diameter of load coil.

Optimum coupling, irrespective of the size of the work-piece, can, moreover, be achieved if the length to diameter ratio of both load and load coupling coil can be kept at a constant ratio

$$\frac{l_{wc}}{D} = \frac{l_w}{d} = 1.5$$

with the additional dimensions of
l_{wc} = length of work coil
l_w = length of work-piece.

Having established the necessary ampere-turns, or the current through the load coil, it will be useful to find the resultant inductive value so that coupling methods to the tank circuit may be considered. Results of acceptable accuracy can be obtained with the help of the following formula for cylindrical coils and a length-to-diameter ratio of greater than one.

$$L = 0.013 \times n^2 \times \frac{A_{e\,LC} - A_{cL}}{l} \; (\mu\mathrm{H}),$$

where
n = the number of turns on the load coil
$A_{e\,LC}$ = cross-sectional area of load coil (cm^2)
A_{cL} = cross-sectional area of load (cm^2)
l = length of load coil (cm)

Formula (7) contains two further factors that influence the practical load coil design and the choice of operating frequency
1) volume resistivity ϱ
2) penetration depth δ

Both of these factors have a pronounced effect on the efficiency of the load coupling arrangement.

Fig. 6.1 gives *d* as a function of frequency for some of the more frequently used materials.

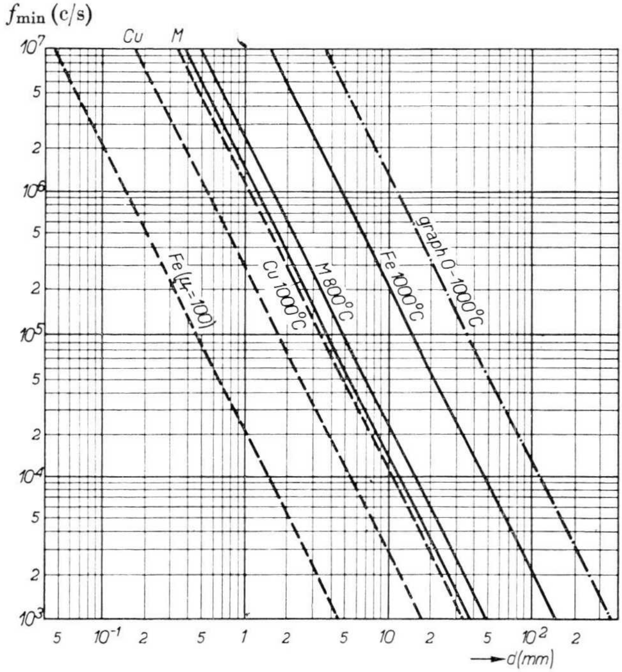


Fig. 6.1. The lowest frequency f_{min} at which the efficiency is not unnecessarily low, as a function of the diameter d of the work-piece, for graphite (graph), iron (Fe), copper (Cu) and brass (M), at room temperature (dotted lines) and at 800-1000 °C (fully drawn lines). For cold iron it has been taken that $\mu_r = 100$.

The load coil efficiency can be expressed by

$$\eta_{LC} = \frac{1}{1 + \frac{D^2}{d^2} \cdot \left(1 + 6.25\frac{\delta_2}{d^2}\right) \sqrt{\frac{\varrho_1}{\mu_r \varrho_2}}}$$

where

D = diameter of load coil (cm)
d = diameter of work-piece (cm)
δ_2 = penetration depth in work-piece (cm)
ϱ_1 = volume resistivity of work material ($\mu\Omega\cdot$cm)
ϱ_2 = volume resistivity of load coil material ($\mu\Omega$.cm)
μ_r = relative permeability

and δ is related to ϱ and μ_r by

$$\delta = 5.03 \sqrt{\frac{\varrho}{\mu_r . f}} \quad \text{(cm)}$$

Several practical conclusions can be drawn from the above formulae.

a. The generator efficiency will be higher if the work coil-to-work-piece diameter ratio is made as small as processing conditions will allow.
b. To keep δ_2/d^2 small, the processing frequency should be chosen so high that the penetration depth will not exceed one eighth of the work-piece diameter. A minimum frequency, necessary for efficient operation, can therefore be established and is given by

$$f_{\min} = 16 \times 10^2 \frac{\varrho_w}{\mu_r d^2} \quad \text{(Hz)},$$

where

ϱ_w = resistivity of work-piece ($\mu\Omega$.cm)
d = work-piece diameter in cm.

c. The power requirements given in equation (5) and (7) determine if the material passes through the Curie point during the process. This results in a change of μ_r and must be considered when calculating the load coupling. Similarly, when loosely packed scrap metal is melted, its change of volume must be taken into account, since this influences the effective value of μ.

Fig. 6.2 gives the efficiency η as a function of D/d at different values. Frequencies above f_{min} give no noticeable increase of efficiency. Higher frequencies can be used where they offer the possibility of simplified generator construction.

d. The factor $\sqrt{\varrho_1/\mu_r\varrho_2}$ suggests that higher efficiency will be obtained if the resistivity of the coupling coil material is as low as possible. For this reason alone it is necessary to use artificially cooled coupling coils.

With the operating conditions and dimensions of the load coil established, it remains to couple it to the generator circuit. Depending on the requirements, several methods are possible.

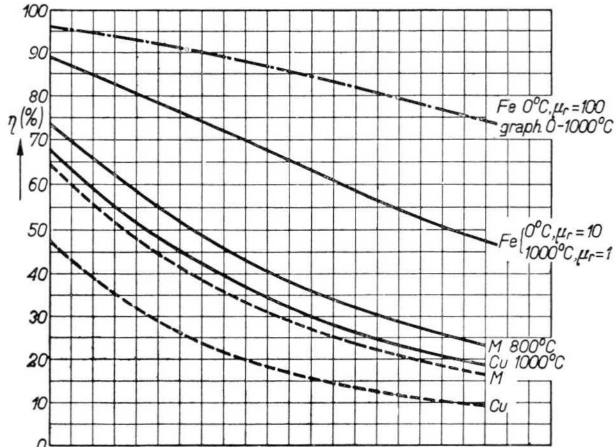


Fig. 6.2. Efficiency η of the work coil as a function of D/d (D = diameter of the work coil, d = diameter of the work-piece) for various materials, at 0 °C (dotted lines) and at 800-1000 °C (fully drawn lines). For cold iron a curve has been plotted for μ_r = 10. The frequency is f_{min}.

6.2.1 Direct Coupling

In some applications the tube operating conditions, in conjunction with the loaded Q (C/L/ratio) of the tank circuit, can be made so that the resulting circulating current and the total number of turns on the tank circuit coil give the necessary ampere-turns. This enables the tank circuit coil to act directly as a load coil (Fig. 6.3). Although such a system can be very efficient, since r.f. losses in intermediate coupling elements are avoided, it is less flexible and more suited to single-purpose equipment where the work position is not too far away from the generator.

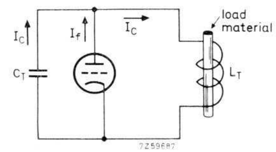


Fig. 6.3. Schematic of directly coupled induction heater where the whole of the tank circuit inductor forms the load coil. The current through this coil $I_c = I_f \times Q_L$ where I_f is the r.m.s. value of the fundamental tube current.

In a modification to the above direct coupling system, only part of the total tank circuit inductor is used to provide the necessary ampere-turn product (Fig. 6.4). This permits a greater flexibility as work coils of different shape can readily be interchanged. Variations of the inductance from coil to coil should, however, not be so great as to affect the C/L ratio and operating frequency of the generator adversely.

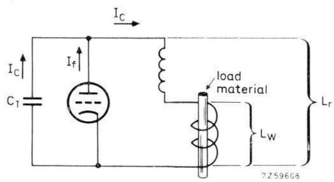


Fig. 6.4. Directly-coupled induction heater with a work coil that constitutes only part of the main tank circuit inductor.

Parallel connection of the load coil (Fig. 6.5) may be used as a form of direct coupling when, for instance, other considerations require a tank circuit of extremely high loaded Q. The current through the load coil is then a part of the total tank circuit current. The division of the current between tank circuit and the load coil is inversely proportional to their respective reactive values.

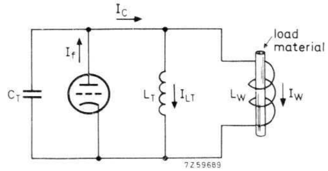


Fig. 6.5. Diagram of parallel connected load coil. L_w *will generally be of a greater inductance than* L_T *and may be made interchangeable. The useful load coil current* $I_w = I_C - I_T$ *will depend on the ratio of the inductive values of* L_T *and* L_w.

6.2.2 Untuned Transformer Coupling

Where the work position is remote from the r.f. power source, transformers are frequently used to maintain matching between the tube and its tank circuit secondary (Figs 6.6 and 6.7). In these cases, the coupling from work to work coil is one transformer, and a further transformation to the tank circuit is also required. This second transformer can also be used for power control in the work-piece by varying its coupling to the main tank inductor.

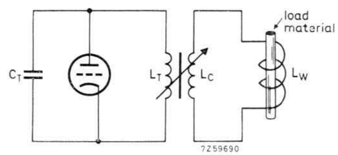


Fig. 6.6. Transformer-coupled work coil. Optimum results will be obtained if the load coupling inductor L_C *is as near as possible to the effective inductance value of* L_w, *i.e. when the work-piece is inserted.*

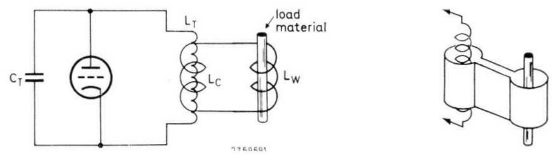


Fig. 6.7. Schematic diagram and sketch of single turn "wrap around" load coupling coil.

The required flux linkage between L_T and L_C will be determined by the ampere-turns of both these inductors, bearing in mind that the ampere-turns of L_T are given by both the tube operating condition and the C/L ratio or loaded Q of the tank circuit.

To achieve an efficient power transfer with such untuned transformers it will, in most cases, be necessary to keep the coupling factor large. This means that a close mesh between L_T and L_C, in terms of their respective diameters and mutual immersion, must be achieved. Although the coupling factor can, in some cases, be improved by the use of ferrite core materials, the closeness of the mesh is limited by the voltage breakdown characteristics of the material. Depending on the nature of the load material the inductors L_w and L_C may range from $\frac{3}{4}$ turn loops to multi-turn coils.

6.2.3 Tuned Transformers

To provide even greater flexibility of application at the output terminals of a generator, as well as avoiding some of the voltage breakdown problems mentioned above, a capacitive tuning element can be introduced (Fig. 6.8). This will, by bringing the load coupling circuit resonance close to the operating frequency, increase the flux linkage, or conversely, allow greater physical distances between the tank and load coupling coil.

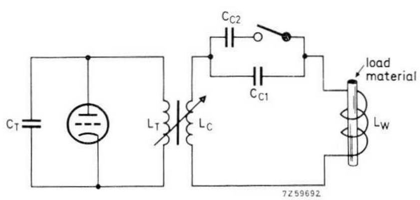


Fig. 6.8. Tuned load coupling transformer.

An increase in the operational safety and the flexibility of the application is achieved if $L_C + L_w$ are tuned close to the operating frequency by the capacitors C_{C1}, C_{C2}. A step change of this tuning capacitance can also bring the load circuit back to resonance when work coils of different inductive values are used. Tapping points on L_C can also be used for this purpose. For optimum performance L_C should have the same inductive value as L_w with the load inserted; it may, of course, be of different physical dimensions, determined by the size and shape of the tank circuit coil L_T.

6.2.4 Current Concentrators

When work-pieces are of irregular cross-section, or have a diameter so small that a normally shaped coupling coil would give inefficient coupling, further transformers are inserted. These normally consist of single-turn windings whose inside surface is shaped to either fit the work-piece, or to produce the desired localised heating patterns. Because of the increased magnetic flux that is achieved they are commonly termed 'current concentrators' (Fig. 6.9). These auxiliary transformers may be used with any of the previously described coupling systems.

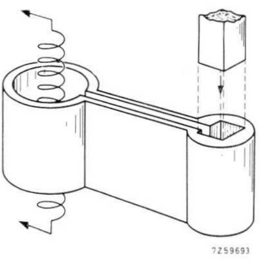


Fig. 6.9. Current concentrator.

6.2.5 Work Coils and Heating Patterns

An induction heating generator will be only as good as the "work coil", i.e. the inductor that forms the ultimate link to the work-piece. Special work coils have to be designed for the more intricately shaped work-pieces or to achieve specific heat patterns (Figs 6.10, 6.11 and 6.12).

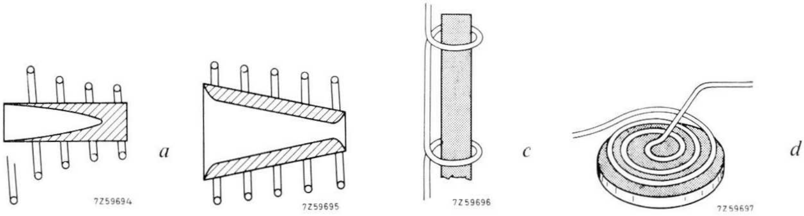


Fig. 6.10. (a). Straight cylindrical work-piece with defined temperature gradient along its axis.
(b). Conical work-piece with work coil to give even heating. The cone of the work coil is of a lesser angle than that of the work-piece since there is less mass and also because of the reduced diameter less coupling on the r.f. side. The divergence of the two cones will also depend on the wall thickness of the work-piece.
(c). Work coil designed to heat the end sections only.
(d). "Pancake" coil for the even heating of flat discs or similar objects. A radial temperature gradient may be introduced by making the spiral slightly dome-shaped or conical.

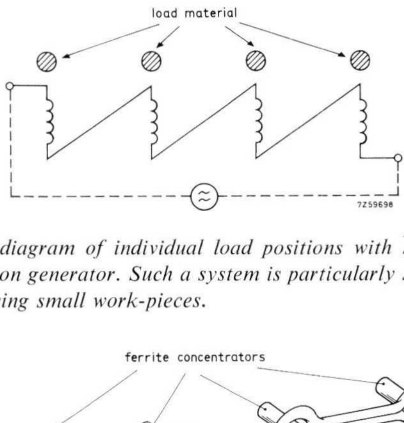


Fig. 6.11. Schematic diagram of individual load positions with series connected work coils fed from a common generator. Such a system is particularly suitable for continuous production lines involving small work-pieces.

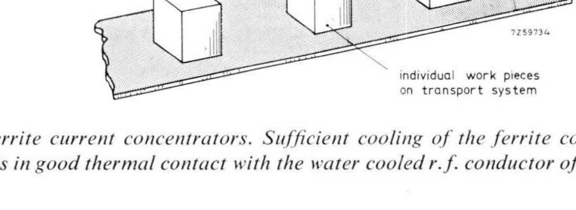


Fig. 6.12. Ferrite current concentrators. Sufficient cooling of the ferrite concentrator is ensured if it is in good thermal contact with the water cooled r.f. conductor of the load coil.

6.2.6 Load Earthing Points on Induction Heaters

Fig. 6.13 shows the preferred earthing points on an induction heater with a separate tuned load coupling circuit.

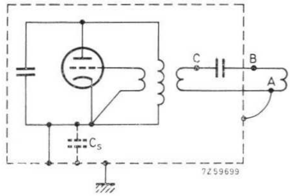
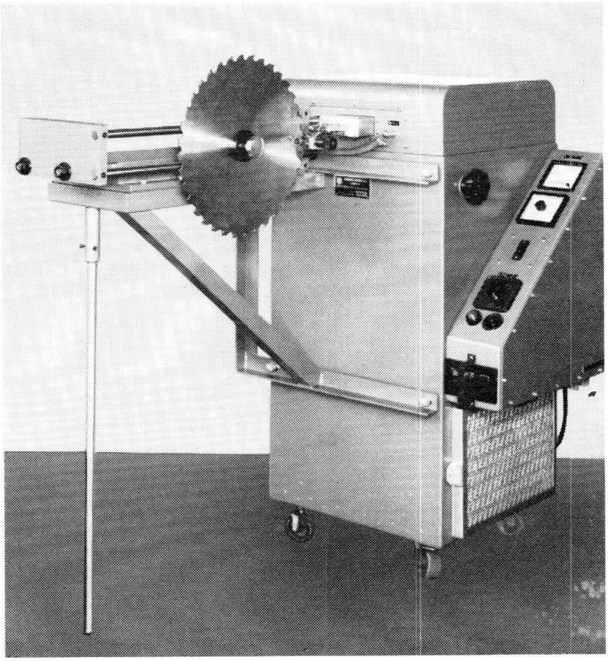


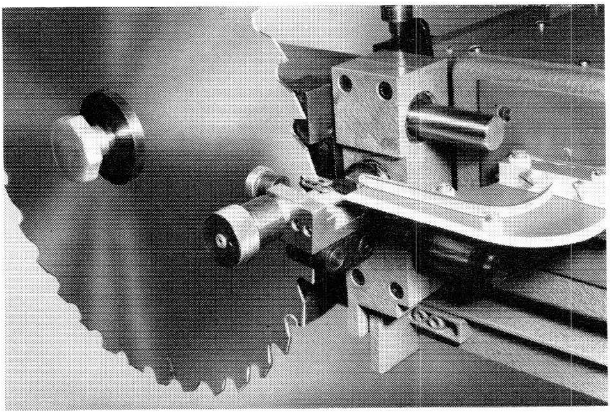
Fig. 6.13. Example of correct location of earthing points in an induction heater.

Cathode (filament) and feedback return are connected at a single point on a sub-chassis which itself is connected to the main structure by a single connector of low inductance. To avoid the creation of capacitive parasitic feedback paths, the stray capacitance C_S between sub-chassis and main structure is kept to a minimum. The load circuit earthing point can then be made independently on any part of the main structure from either of the points A, B or C.

6.2.7 Power Control

Where the power transferred to the load must be varied during processing or, for example, work-pieces of different shapes are to be processed in succession at the same load position, various methods of r.f. power control can be used. The main distinction is made between those involving some mechanical movement to produce the desired electrical effect and pure electrical systems. The latter group of systems is, of course, much quicker in action and therefore more suited to automated process control.





A 2 kW generator for soldering hardened material on to saw blades, using tube type TB5/2500 (by courtesy of Himmelwerk A.G., Tübingen, Western Germany). The inductor arrangement is shown enlarged.

Electro-mechanical Systems	*Electrical Systems*
1. Mechanical adjustment of h.t. supply voltage by variable transformer or series reactor in the primary of the supply transformer.	1. Phase cutting with thyristors in the primary of the h.t. supply transformer
2. Change of magnetic coupling between tank and load circuit by physical movement of one or other, or both, inductors.	2. Phase cutting with grid controlled rectifiers in the secondary of the h.t. supply transformer.
3. Similar change of magnetic couling through relative movement of other transformer type coupling elements.	3. Grid pulsing of the r.f. oscillator tube through thyristors, thyratrons or pulse amplifier tubes.

The resultant power output can, at these low frequencies, be monitored with reasonable accuracy and economy with the aid of rectifier type volt and ampere meters, the latter being coupled to one lead of the load coil through a current transformer. Other suitable instruments are the electrostatic voltmeter, hot wire and thermocouple current indicators.

6.3 Dielectric Loads

The second of the two analogous heating formulae given in Ch. 6.1.1 has, as its chief factor, the voltage of the applied high frequency field that must appear across a dielectric material if a certain power is to be dissipated in the material. For practical use this expression may be rewritten as follows:

$$V_{\text{r.m.s.}} = V \cdot P_w \cdot X_C \cdot \tan \delta^{-1}$$

where

$V_{\text{r.m.s.}}$ = the r.m.s. value of the r.f. voltage applied across the faces of the work-piece (V)

P_w = The power to be dissipated in the work-piece as calculated from the standard calorimetric formula (W)

X_C = The reactive value of the capacitance of the work-piece calculated from its physical dimensions and electrical properties with the help of the standard plate capacitor formula (Ω)

$\tan \delta$ = The loss factor of the work material at the operating frequency.

It should be noted that most materials have a loss factor that rises with temperature and may eventually go through a peak.

Loss factors for likely dielectric heating materials will range between the first and third decimal of tan δ, and their dielectric constants between 1 and 5. The voltages required across the work-piece will range from several hundred to a few thousand volts. A limit to these voltages is, however, set by the electric strength of the work material beyond which internal breakdown will occur. If more power is to be dissipated in the work an increase of processing frequency often becomes necessary. The required voltages occur at the tube anode or the tank circuit capacitor and the load coupling circuit is required to make these available at the load position. A secondary requirement of the load coupling circuit is to provide a match between tube and load material. Depending on the application, a step up or reduction of the generated voltage may be required. The step up can be achieved with a transformer system whilst the reduction of generated potential need often be no more than a capacitive potential divider, where the lower section is formed by the load material capacitance itself.

The main divergence of requirements is between those circuits that are used for bulk material heating such as pre-heating or process drying, and others that are used for plastic welding. In the first group, an air-gap or other low-loss dielectric gap is used, on at least one side, between the load position electrode and the material, to ensure even heating and to allow the free passage of the material on a conveyor belt or other loading mechanism. These dielectric gaps will then act as series capacitors, causing a potential division whose value will be determined by the respective dielectric constants of the gap and load material. This means therefore that they require a much higher load position voltage than the processing conditions of the material alone would suggest.

In the second group one finds "contact processes", i.e. those applications where a direct contact between work electrode and material is required, as in plastic welding. Since the voltages needed for this type of work are only up to about 1000 volts, capacitive potential divider systems, between the r.f. power source and the work position, may be used for the necessary reduction.

A few of the more common load circuits of both systems are outlined in the following sub-sections.

6.3.1 Directly-Coupled Systems

The three examples shown in Figs. 6.14, 6.15 and 6.16 are all directly-coupled systems and will, provided that connecting leads are of low inductance, show a good processing time-power transfer characteristic, since the work material capacitance, and the series capacitance, are in parallel with the tank circuit capacitor. They are thus an integral part of the frequency controlling components and will, because of dielectric changes in the load material during the process, cause a shift of the operating frequency that may under certain conditions be unacceptable. Up to a point, the resulting frequency shift can, however, be limited by making

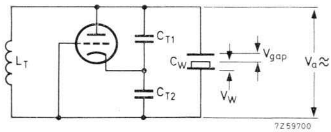


Fig. 6.14. Example of directly coupled dielectric heater where V_a is sufficiently large to equal $V_w + V_{gap}$. The oscillator circuit shown is a Colpitts type, but other systems could be used.

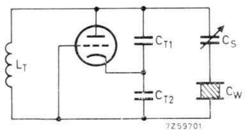


Fig. 6.15. The same directly coupled circuit as shown in Fig. 6.14 but adapted for plastic welding by using a separate and possibly variable capacitor C_s in series with the welding electrodes.

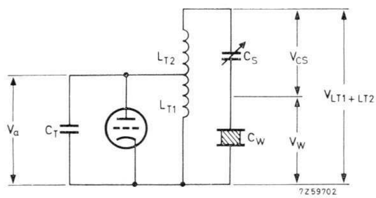
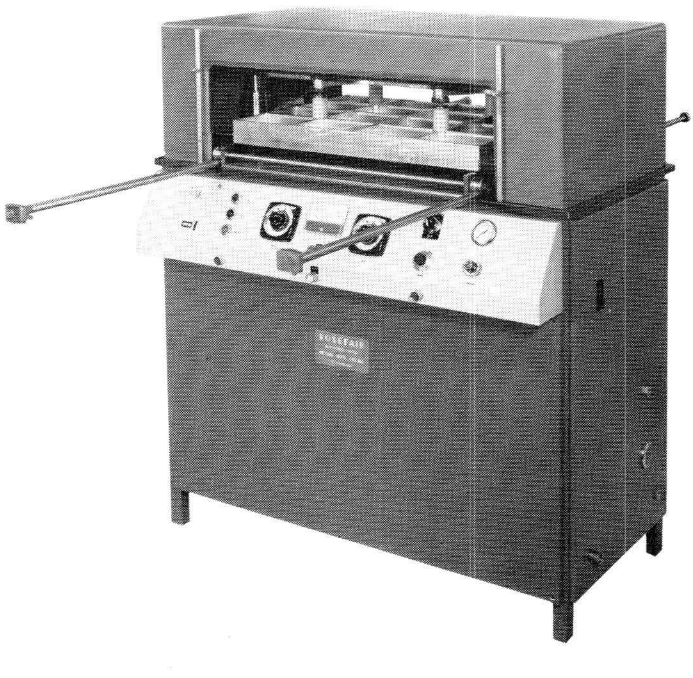


Fig. 6.16. Schematic of auto-transformer circuit that can be used where $V_{C_S} + V_w$ must be greater than V_a. The transformer section $L_{T1} + L_{T2}$ could, for very high frequencies, also take the form of a line.

the tank circuit capacitor large in comparison to that of the loading system or by using a series capacitance that is small in comparison to that of the work material. A compromise between the available r.f. potential and the acceptable frequency shift may have to be found. Directly-coupled systems are also likely to radiate all the harmonic frequencies generated by the tube because no frequency filtering takes place between the tube and the load. This situation is aggravated in welding machines where screening of the working electrodes is virtually impossible and the welding electrodes act as effective aerials, especially at higher frequencies.



A 7 kW plastic welder using tube type YD1160 (by courtesy of Rosefair Electronics, Watford, England).

6.3.2 Indirectly-Coupled Systems

Indirectly-coupled systems (Figs 6.17 to 6.20), using either capacitive probes or transformer coupling, afford a greater flexibility in several ways. A wider choice of the frequency controlling tank circuits is possible, larger distances between generator and load may be bridged, r.f. voltages may readily be stepped up or down, and filtering networks may be introduced to eliminate unwanted frequencies. By making the coupling factor between generator and load relatively small, the fundamental frequency shift may be reduced, albeit at a cost of the relationship between power transfer and processing time. Unless automatic retuning is used, indirectly-coupled systems will have a "tuning" characteristic in their power transfer. This characteristic makes the use of generators with a greater peak power capability necessary if the work is to be accomplished in a time comparable to that achieved with a directly-coupled load. (Faraday screens may have to be inserted between inductive coupling elements to minimize harmonic power transfer.) The effect of the "tuning characteristic" is sometimes used to influence the process sequence (see Fig. 6.18).

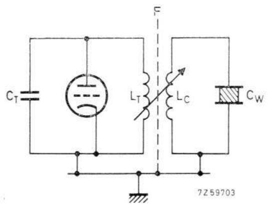


Fig. 6.17. Example of simple transformer-coupled load. For maximum power transfer the load circuit L_C/C_W should be tuned near the operating frequency of L_T/C_T at the peak of the processing cycle. The power transfer may also be varied by adjusting the coupling between L_T and L_C. F is the Faraday screen, preventing capacitive coupling between oscillator and load circuit.

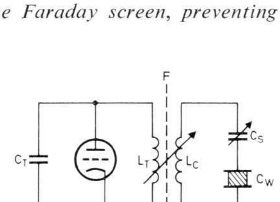


Fig. 6.18. Transformer-coupled load circuit containing a series capacitor C_S that may be used for power control purposes or the retuning of the load when the work material characteristics change during the process.

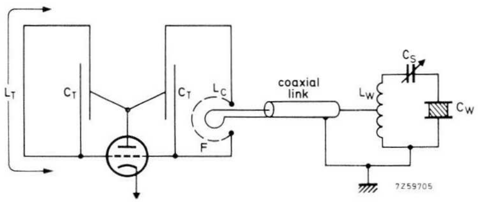


Fig. 6.19. Coaxial link coupling of cavity type generator to work coupling circuit. The frequency stability of such circuits can be very good and the load C_W can be re-matched by the capacitor C_S thus keeping the load circuit L_W—C_S—C_W in tune over the greater part of the processing cycle. Since mismatches will occur, the coaxial link should preferably be of the airspaced type to avoid irreparable voltage breakdowns. F is again the Faraday screen.

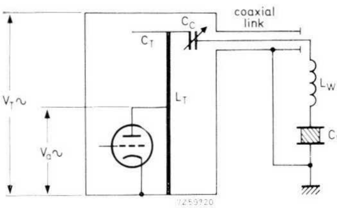


Fig. 6.20. Schematic of coaxial line generator with coaxial output and capacitive probe coupling. The tank circuit C_T L_T is formed by a capacitively loaded $\lambda/4$ line which itself is a step-up auto-transformer. C_C is a variable capacitive probe. L_W and C_W are the inductive and capacitive components of the series-tuned load circuit.

6.3.3 Integrated Coupling Systems

The generator-load coupling systems described in the previous sub-sections have all, whether coil-capacitor, cavity or line tank circuits are used, a fundamental quarter wave operational mode. The shortcomings that have been noted for these circuits can be largely overcome if integrated generator-load coupling circuits are used. As opposed to direct systems, generator and load circuit are still two separate entities, but joined in such a way that they form one resonant circuit and therefore have unity coupling, resulting in a much improved power transfer-processing time relationship.

The most practical form for this type of circuit appears to be a half-wave line that is capacitively shortened at one end by the tube and generator tank capacitance, and at the other end by the loading capacitance (Fig. 6.21). A transformation from generator tube to load can thus be arranged by suitably apportioning the end loading capacitances. As it is only their ratio, and not their absolute value, that determines the transformation ratio, an appreciable loading capacitance may be placed in parallel at both ends to improve the overall frequency stability.

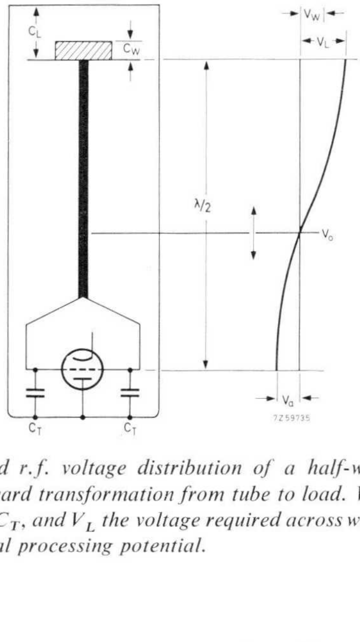


Fig. 6.21. Schematic and r.f. voltage distribution of a half-wave line generator load circuit, assuming an upward transformation from tube to load. V_a is the r.f. swing across tube and tank capacitor C_T, and V_L the voltage required across work material plus air gap. V_W is the actual material processing potential.

Yet a further advantage is the improved oscillatory stability (freedom from parasitic oscillations) that is due largely to the "floating" zero voltage point on the half-wave line. A change in load capacitance will cause a relatively small change of operating frequency, transformation ratio and displacement of the zero voltage point, but will still allow the load to appear as a pure resistance to the tube. (A fuller description of this system and calculating methods for the circuit constants can be found in Ch. 8 'Applications'.)

Such an integrated system, with its excellent oscillatory stability and work efficiency, can also be used for plastic welding (Fig. 6.22). The layout of the generator remains substantially unchanged except that the material in the load position is replaced by a capacitive probe that feeds the r.f. energy through a coaxial duct to the welding press.

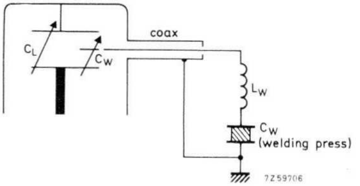


Fig. 6.22. Diagram of half-wave line load section used for plastic welding. The voltage across C_L, and C_L itself can be made large. C_W represented by the capacitive coupling probe, can be relatively small thus causing little change of the effective line and loading capacitance. Consequently, the resulting changes of operating frequency are quite small. As in the case of the dielectric heater, the load will always look purely resistive to the tube.

6.3.4 Load Position Capacitors

The majority of load position electrodes are in the form of single flat capacitor plates with the load material inserted between them (Fig. 6.23). The number of possible configurations of practical value is certainly much smaller than, for example, the diversity of load coil shapes used in induction heating. A modification to the flat plate system is usually necessary only for larger bulk loads where higher powers are used or an even voltage distribution in the material is difficult to achieve.

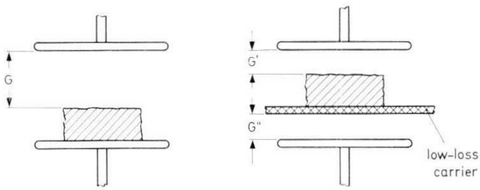


Fig. 6.23. Sketch of simple flat plate loading capacitor (a) with air gap on one side of load material and (b) with air gap on both sides, the load material being supported by a loss-free carrier, perhaps a moving belt. To ensure even heating the load capacitor plates should always extend well beyond the dimensions of the load material. Such a construction will be adequate for most types of plastic pre-heaters and process dryers. If the thickness of the low-loss carrier is negligible, G in (a) equals G′ + G″ in (b).

When a more even heat distribution over a large area is required, balanced applicator electrodes may be used (Fig. 6.24). The load material rests on a third plate at ground potential. The load coupling circuit construction is simplified and the loading tray or belt can be made from conducting materials.

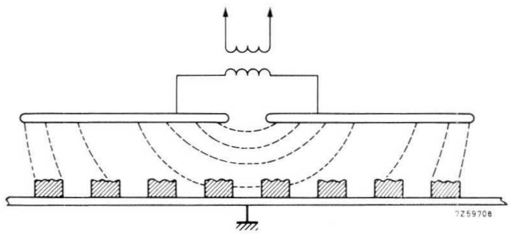


Fig. 6.24. Balanced applicator electrodes.

Another large load area system giving good results is the distributed or stray field capacitor method (Fig. 6.25). Each electrode is made up from a number of rods lying parallel to each other in the same plane. All the odd numbered rods are connected to one side of the generator output and all even numbered rods to the other. Interconnections, within the groups of rods, may have to be duplicated to keep the inductance within the electrodes low.

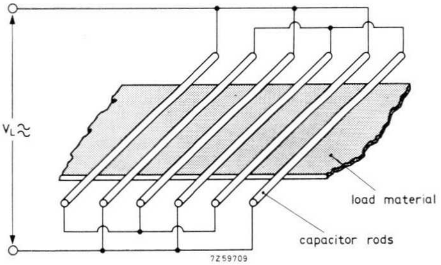


Fig. 6.25. Stray field capacitor system with large area load underneath.

Similar applicators may be used for edge gluing of wood since, with the stray field method, the two dielectrics, wood and glue, can be made to appear as parallel capacitors. With glues of a sufficiently high loss factor, setting will be much faster, as the thermal losses in the wood are very low.

For wood shaping, contact electrodes of suitable outline are used so that only the essential pressure need be applied to the work. To avoid voltage breakdowns and overcome uneven heating due to variations in the timber, a low-loss barrier material of high dielectric strength and high compression strength can be inserted and will, for mass production runs, probably be an integral part of one of the electrodes.

Composite dielectric loads containing dielectrics of different characteristics are shown in Fig. 6.26. In Fig. 6.26(*a*) the two materials are in series and, because tan δ of the glue is larger than that of the wood, a smaller voltage drop V_G will appear across the glue with the result that both materials will heat at approximately the same rate. As only the glue need be heated to make a satisfactory joint, the power dissipated in the wood, due to V_U, is largely wasted. The arrangement shown in (*b*) places both materials under the same r.f. voltage so that the one with the higher loss factor, in this case the glue, will heat much faster.

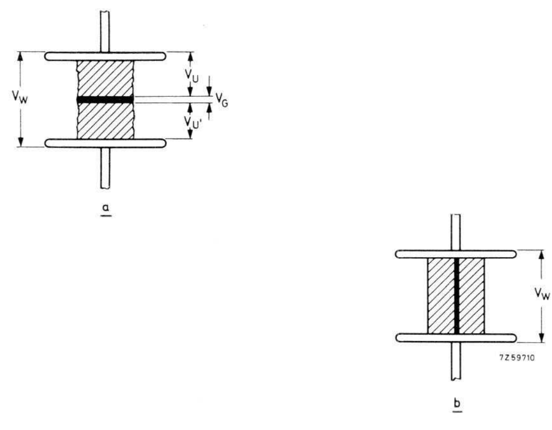


Fig. 6.26. Composite dielectric loads.

Wood shaping electrodes with a barrier dielectric are illustrated in Fig. 6.27. E_1 is the pressure electrode, E_2 the supporting electrode, w the material to be processed and B the barrier dielectric of high mechanical strength. By choosing a barrier material of suitable dielectric constant in relation to that of the work-piece, the total r.f. voltage $E_1 - E_2$ may be kept somewhat lower, thus easing some of the load circuit problems.

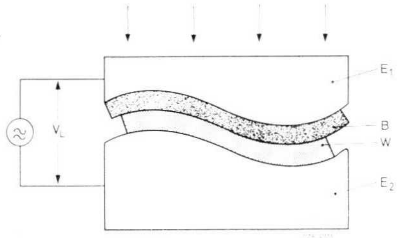


Fig. 6.27. Wood shaping electrodes.

6.3.5 Welding Electrodes

Welding electrodes, though appearing in a great variety of shapes, are always contact electrodes. As well as applying the r.f. potential they exert pressure on the work material to obtain a satisfactory weld. The contact electrodes can also be used to cool the outer surface of the work material. A well engineered press system, exerting even pressure, is at least as important in a welding installation as a good r.f. power source (see Fig. 6.28). Barrier materials, in the form of low-loss sheets, are inserted between the work material and the opposing plane electrode. This gives a better r.f. field distribution, minimises the risk of flashes and prevents heat conduction away from the welding material.

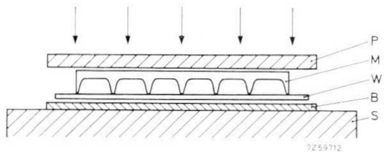
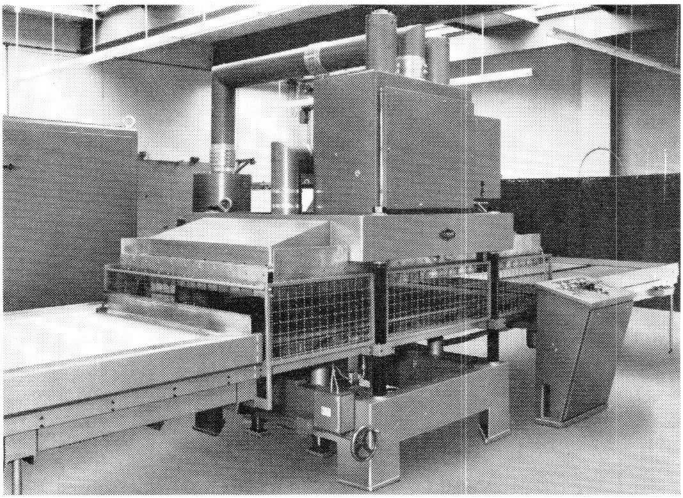


Fig. 6.28. Sectional view of press head.

In Fig. 6.28, P is the pressure electrode, M the welding electrode with the appropriate pattern, W the work material consisting of two or more layers of plastic sheet, B the barrier material such as Fibreglass, polypropylene, or Pertinax sheet, and S the supporting electrode. The polarity of P and S is not of great importance; either could be earthed or even the sequence of M, W and B reversed. Sections of the welding electrode can be fitted with cutting edges, so that welding and trimming can be achieved in one operation.



An HF welding press (10 ton pressure) fed from a 40 kW, 27.12 MHz generator using 2 tubes TBL12/25 (by courtesy of Herfurth GmbH, Hamburg, Western Germany).

6.3.6 Earthing Points on Dielectric Heaters

To avoid unnecessary mismatches and the appearance of radio frequency energy on nominally earthed parts of the installation, certain precautions in the choice of feeder connection and earthing points are necessary. Earth returns of the feedline or load position are, as shown in Fig. 6.29, for preference returned to their own independent connecting point and only from there returned to the generator casing.

For greater distances some form of transmission line will probably be used and, in this case, the return will be to the outer of the line. Lines of higher characteristic impedance ($Z_o > 50\ \Omega$) are more suitable for this since their inherent Q will be lower and less likely to cause oscillator moding through resonant line effects.

In *a* and *c* the load circuit is independent of the tank circuit, the only linkage being through magnetic coupling. A Faraday screen may be inserted, if necessary. In *b* and *d* the load circuit is returned to the inside of the cavity and therefore shares part of its return path with the tank circuit. Since this return point of the coupling loop cannot be maintained at the equivalent of the external earth potential under all conditions, r.f currents may flow between these points and cause an r.f. voltage to appear on the outside of the cavity that could easily upset the correct operation of the oscillator.

Screening partitions between conventional coil-capacitor type tank circuits as shown in *e* and *f* can produce the same effects.

Capacitive load coupling probes into similar tank circuits are simpler in construction but can be successfully used only at relatively high frequencies and high source voltages if their capacitance and dimensions are to be kept small in comparison to the remainder of the circuit. To avoid secondary resonance effects, a coaxial connection between the capacitive loading probe and the load position is essential.

Examples of practical load circuits of both separate and integral type construction are given in Chapter 8 'Applications'.

6.3.7 Power Control of Dielectric Heaters

Control over the power dissipated in the load in dielectric heaters can be achieved by similar methods to those described for induction heaters. The use of variable capacitors in coupling or loading circuits offers a further method of control. At the higher frequencies used for dielectric

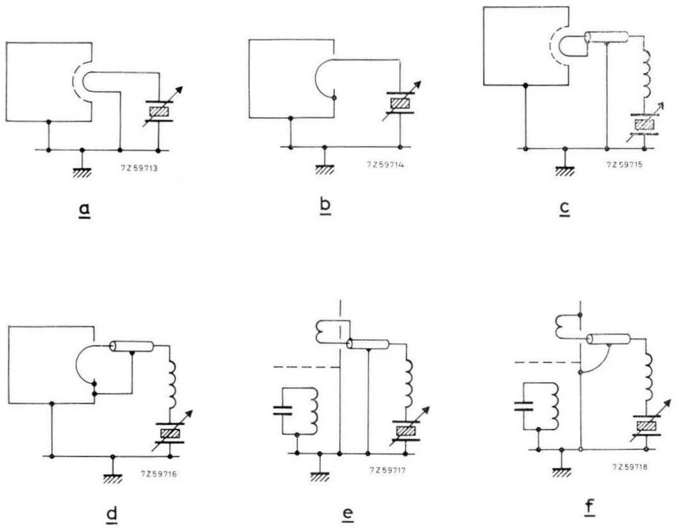


Fig. 6.29. Correct (a, c, e) and incorrect (b, d, f) load circuit and earth return connections on cavity type generators with inductive coupling.

heating, the physical sizes and construction of the capacitors make the use of electromechanical control methods possible.

Servo systems can be used that adjust the load coupling or the load tuning capacitors in accordance with a predetermined reference signal such as the tube anode current or the frequency shift of the oscillator. The limitations on the use of such control systems are set by the operating time of their mechanical components. They can, therefore, be used successfully only for processes of relatively long duration where the necessary periods of adjustment or re-setting are measured in tenths of seconds.

Within economically acceptable limits instantaneous load power indications can, at these higher frequencies, be given only through reference to the d.c. input conditions of the tube. An anode current meter may, for example, be calibrated in kW load power, the calibration points being established with a set of dummy load measurements (see chapter 7.5.1).

7. Test Instruments and Methods

From the large number of electronic measuring and testing aids available many can be found that will in some ways be useful during the design, construction and final measurements of an industrial generator. Others will have to be adapted to the specific problems of the industrial heating field; and the methods of their use may vary with the application.

The requirements for checking the correct operation of all the oscillatory functions are, however, simple and can be met with a few well chosen items.

Before discussing the instruments proper, we must consider power levels required at various stages of testing. Initially it will be helpful if the oscillator under test is operated from an infinitely variable supply at reduced anode voltage. It will then be possible to detect the onset of the fault conditions and perhaps "hold" these for some time for closer investigation at an input level that is safe for the tubes and components. The majority of power supplies for industrial use are unsmoothed, and the use of variable-input transformers is therefore preferred. Other methods of regulation, such as phase cutting, where the supply waveform is spiky at low input levels, and peak-to-mean ratios are high, are useful only in special cases.

Where fault conditions do not occur until a relatively high supply voltage has been reached, the use of phase cutting supply systems or the omission of a phase from a multiphase supply can be used to simulate high-voltage operation with a reasonable safety margin. Parasitic circuits, if present, will respond to the peak voltage but the parasitic power will be much less. A steady and well-defined display of the investigated high frequency can be obtained if the time base of the oscilloscope is locked to a sub-multiple of the supply waveform. To examine high frequency waveforms within the supply waveform envelope, a time base expansion facility is necessary.

The identification of active parasitic circuits is often simplified at low voltage operation if the fundamental oscillations are deliberately suppressed, allowing the parasitic circuits to become predominant. Suppression of the fundamental oscillation will depend on the type of generator and the following methods are offered as a guide:

a. removal of the feedback coil
b. change of value of the feedback coil
c. insertion of screen between feedback and tank coils
d. r.f. grid/filament short
e. damping of tank circuit using a resistive overload.

Ensure that any parasitic voltages or currents do not exceed the tube or component ratings.

7.1 Grid Dip Meters

Grid dip meters are variable oscillators, usually constructed with the tank circuit inductor outside the case housing the remainder of the components. The case and coil are shaped to enable them to be placed in the proximity of components or circuits requiring measurement. The coupling is usually inductive. The component or circuit under investigation will then absorb a part of the radio frequency energy generated by the grid dip oscillator. This absorption will be maximum when the grid dip oscillator is at resonance with the circuit under test and part of the generated energy is dissipated by its resistive component.

The damping thus imposed on the grid dip oscillator circuit can then be monitored with a moving coil meter, registering the change of grid current in the oscillator tube. At the same time the corresponding frequency is read from a dial coupled to the built-in variable capacitor.

Grid dip meters are essentially 'yes-or-no' indicators, since their calibration will change slightly with their coupling into the test object. However, after some practice in their use, other rough quantitative information on the investigated circuit can be deduced. The amount of absorption indicated can serve as a guide to the circuit quality, and, by substitution, component values may be determined.

Despite its inherent crudeness such an instrument fulfils many of the functions of a Q meter, and the results obtained are often more meaningful as, for such checks, the circuit under test does not have to be disturbed. The majority of commercially available grid dip meters can also be used as tuned-circuit diode wavemeters when the anode supply to the oscillator tube is switched off.

A good instrument covers all frequencies from 100 kHz to 1000 MHz with two or three probes and a range of plug-in coils for each probe.

Higher sensitivity can be obtained if the changes in grid current are registered through a d.c. bridge or a simple differential amplifier. The more sophisticated grid dip oscillator is designed so that the natural changes of grid current, caused by the varying C/L ratio within any range, are small and smooth; no 'self dips' i.e. absorption symptoms due to enclosure or component resonances of the instrument itself should occur at the higher frequencies.

When investigating a circuit or group of components for resonances, some judgement must be exercised to distinguish between harmless ones and those that present potential parasitic danger sources.

Every conductor, component or group of components will have at least one and probably several inherent resonances but it is their relative electrical position in the oscillatory circuit, rather than their mere presence, that determines their effect on the operation of the oscillator. The more dangerous points are those that have a higher r.f. voltage during normal operation. Probably the most dangerous resonances are those that are coupled, either capacitively or inductively, to both the input and the output of the oscillator tube. These reflected resonances may often

Fig. 7.1. A grid dip meter showing the plug-in coils and their stowage position (by courtesy of Edison Electronics Division, McGraw-Edison Company, Grenier Field, Manchester, N.H. 03103).

be discovered with the help of a grid dip meter by searching other parts of the circuit, remote from their origin.
A physical re-positioning of one of the offending circuits or the replacement of a component with one of equal value but different stray reactances will always prove a more acceptable solution than the insertion of parasitic suppressor circuits that may introduce yet further stray resonances.

7.2 Neon Probes

Neon (or other inert gas) lamps of almost any rating may be used for the detection of radio frequency energy and, as their luminosity changes with the intensity of the applied field, rough comparisons of field strength and distribution can be made (Fig. 7.2).

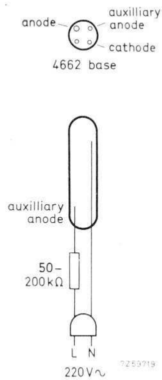


Fig. 7.2. Diagram of probe.

If relatively low energy levels are to be detected the ionization threshold of several tens of volts of the indicator lamp must be overcome. For this purpose, ‚mains priming' is used, that is, the ionization is initiated by the application of a 50 Hz mains voltage to the terminals of the lamp. A 50 kΩ-200 kΩ resistor in series with one of the leads will limit the current through the lamp. Priming keeps the level of the ionization low so that indications of weak r.f. fields are not swamped. With suitable indicators, energy levels as low as 5 mW/cm^2 can be detected.

When using our neon indicator 4662, shown in Fig. 7.3, further useful information can be deduced from the correct interpretation of the ionization phenomena. From several kHz to about 5 MHz the ionization produces an orange glow similar to the ‚priming' glow but varying in intensity. Above 5 MHz and well into the microwave region a pale blue glow is superimposed. If during the investigation of a low frequency oscillator, say 500 kHz, a dinstinct pale blue glow is present, or appears under certain operating conditions of the generator, the presence of parasitic oscillations or exceedingly strong harmonics of a higher order is indicated. The location of their greatest field strength can then be found and methods for their suppression considered.

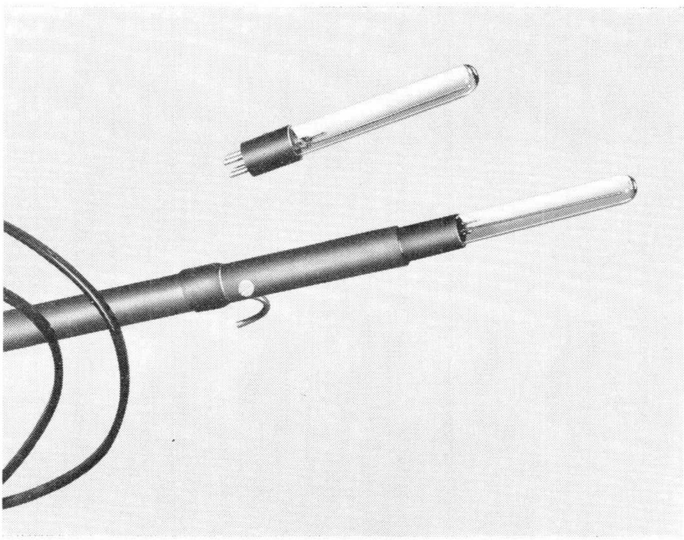


Fig. 7.3. Neon tube shown separately and connected to its wander probe.

Conversely the incidence of a weak orange glow, if produced by a high frequency generator, can indicate that this equipment is not operating at the correct frequency or that squegging is taking place.

Neon probes may also be used for checking the fundamental r.f. voltage distribution along inductors, capacitors and filtering chokes as well as the effectiveness of screening.

7.5 kW, 27 MHz generator used for profile drying of paper (by courtesy of Intertherm, London, England).

7.3 Waveform Monitors

Fault conditions on r.f. power generators may often be precipitated by excessive distortion of the fundamental waveform or parasitic oscillations. As the latter may be present in a latent form or only under certain operating conditions, when otherwise no signs of malfunctioning are evident, a visual display of the waveform may help in a quick diagnosis.

If a meaningful picture of the waveform is to be obtained the monitoring point should be at one of the generator tube electrodes. The anode is the preferred monitoring point since it is here that the r.f. energy – fundamental, harmonic or parasitic – is generated. Other points in the circuit are less suitable for this purpose because of intervening component reactances. The tank and load circuits in particular tend to mask any irregularities in the waveform because of their filtering action.

Oscilloscopes with a substantially flat response up to 100 MHz are recommended but a simple display unit with the deflection plates directly accessible and without the added complications of amplifiers can also be used. The r.f. swing at the chosen measuring point is usually several times the value required for full screen deflection, and potential dividers have to be used.

Carefully constructed potential dividers reduce the risk of introducing irregularities into the response of the display system. The common earthing point of generator and display unit is important and, since the display unit is usually external to the generator, it should be on its outer screening but close to the oscillator tube position. A suitable divider may be constructed from appropriately dimensioned capacitors with low-inductance terminals (the effect of terminal inductances can be compensated by parallel resistors).

In extreme cases, where physical connection to the generator anode is not practicable, the top end of the divider can take the form of a capacitive aerial. Some care is then necessary in spacing this aerial from the r.f. voltage source, since too great a distance will act as a low capacity-high-pass filter which will discriminate in favour of the higher frequencies and present the waveform much worse than it really is.

The diagnosis of incipient or existing fault conditions will depend on the correct interpretation of the displayed trace. Discrimination must be made between those irregularities, that may although undesirable, merely indicate the presence of low-level harmonics or component ringing and others that may lead to disastrous fault conditions.

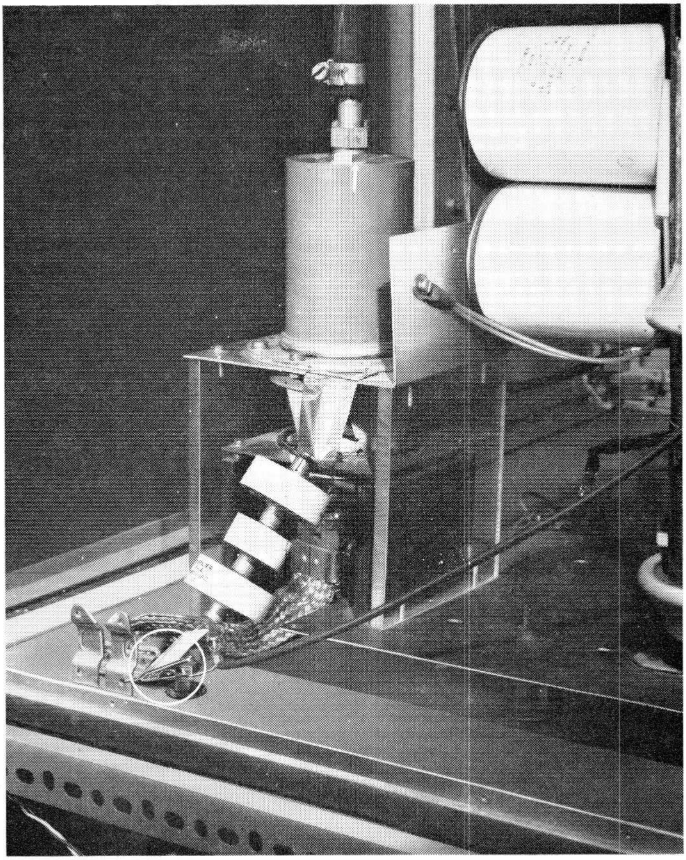


Fig. 7.4. Capacitive divider in position. The monitoring leads are shown inside the white circle in the lower left hand corner.

Oscillation will usually remain unimpaired if the waveform distortion consists of no more than perhaps a 10 % ripple, attributable to harmonics or component ringing. However, the amplitude ratio of this ripple with respect to the fundamental must remain substantially constant and the ripple waveform itself must be reasonably smooth under changing operating conditions.

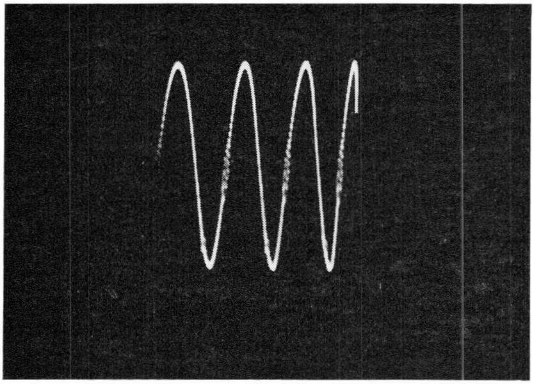


Fig. 7.5. Oscillogram of 400 kHz waveform. The slight ripple on the waveform is due to component ringing. Since its amplitude is less than 5 % of that of the fundamental it can be ignored.

Changes that can be deliberately introduced during a test include variations of drive, loading, and the introduction of a varying reactive component into the load circuit. Harmonic distortions of the display can then be distinguished from others by their relative position to the fundamental. If their position remains constant on a changing fundamental, caused by a deliberate change of fundamental frequency, their origin is of a harmonic nature and therefore less critical. A change of relative position between fundamental and ripple on the other hand indicates component ringing or the presence of parasitic oscillations. In the latter case it will then often be found that this ripple may change its amplitude and waveform, the rate of change, spikiness and amplitude being a good indication of the severity of the parasitic danger.

A further symptom of parasitic oscillation is the presence of high frequency "bubbles" on an otherwise satisfactory fundamental waveform. These "bubbles" are formed by the high-frequency envelope of a few cycles of the parasitic oscillation, triggered when the fundamental r.f. swing on grid and anode passes through critical points (Fig. 7.6). These points are conducive to the generation of such parasitic oscillations, and occur more frequently on the positive slope. It should be noted that the danger of parasitic oscillation is usually greater outside the design operating conditions, i.e. under extremely heavy overloads.

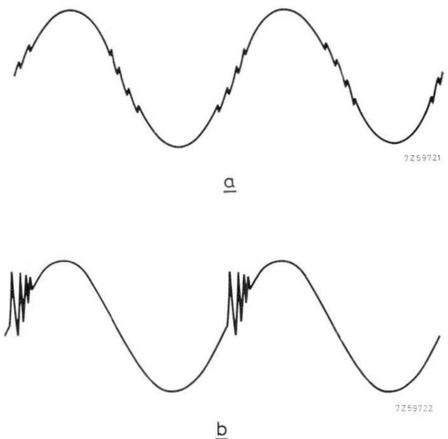


Fig. 7.6. Tracings of parasitic high frequency envelopes on a fundamental waveform.

The likelihood of damage to the tube or circuit components is greater if these oscillations appear near the positive peak of the fundamental anode waveform as shown in Fig. 7.6(*b*). This is because their amplitude, and this may assume a multiple of that of the fundamental, if added to the fundamental, brings the total amplitude well above the tube and component ratings.

Even if such high frequency bursts are harmonically related to the fundamental, their elimination is necessary since they indicate the presence of a parasitic feedback path and may, under varying operating conditions, grow to dangerous proportions.

In all cases, the prevention of parasitic oscillations by repositioning or substituting components is the recommended method rather than fitting parasitic suppressors to reduce their effect.

7.4 Panoramic Receivers

A panoramic receiver, used in conjunction with a communications receiver or signal generator, can be an invaluable help during the final testing stages of radio frequency heaters, particularly those that are intended to operate within a specific frequency band whilst being subjected to reactive changes in the load (Fig. 7.7). Frequency drift during operation, sidebands, moding and some types of parasitic oscillation can be readily observed.

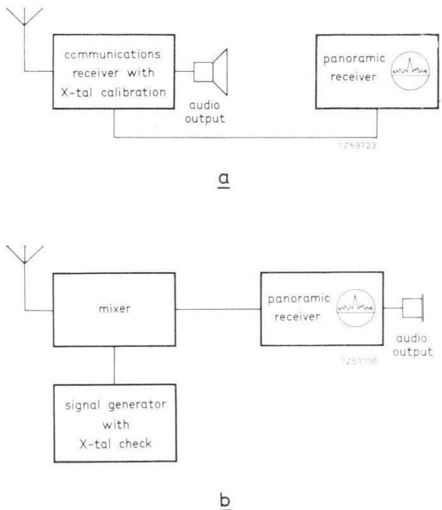


Fig. 7.7. Schematic diagram of panoramic receiver installation. Either (a) or (b) will form a testing set-up, suitable for the frequency checks outlined in the text.

The audio output of the communications receiver can also indicate any identical modulation or squegging that may be present.

Instruments, such as are available as companion sets for the moderately priced communication receivers, are adequate so long as their display scale covers a deviation of about $\pm$ 500 kHz from the centre frequency. Facilities for varying the sweep width will be an advantage but need not extend below $\pm$ 50 kHz. The resolution, adequate for industrial purposes, need not be higher than 5 kHz.

7.5 Power Measurements

An industrial r.f. heating equipment may be divided into two parts, the generator and the loading arrangement. The latter may take the form of, say, an inductive hardening equipment or perhaps a press for plastic sheet welding. In any case, a series of power and efficiency measurements is desirable (at least during the development and production of the equipment) and this section gives some information of the best methods to use.

7.5.1 Methods of Power Measurement

In practice it may be necessary to measure power dissipation in the range of 2 kW-250 kW at frequencies from a few hundred kHz to about 200 MHz. Various proprietary measuring devices are available but most of these have some limitation which may make then unsuitable for the work in hand. It has been found that the most suitable form of load is a water cooled resistor, calculation of power dissipation being by a measurement of water flow and temperature rise.

The type of resistor and its method of connection to the generator will depend upon the frequency concerned. At the lower frequencies, normally used for induction heating, a simple method of measurement is to couple a steel can to the generator by means of a work coil and to water cool the can. This method has been used to measure load powers exceeding 230 kW. At the higher frequencies a 50 Ω carbon resistor can be capacitively coupled to the generator by a coaxial cable, the resistor being water cooled.

In dielectric heaters other methods may be used for measuring power at the work position. A loss-free container can be inserted between the electrodes. This is filled with a known quantity of water and heated by the r.f. field for a given time and the temperature rise of the water measured. Alternatively, for higher powers, a closed container can be used and a measured flow of water passed through it. The water flow and temperature rise will give the load power.

The load power can be calculated from the formula

$$P_L = 0.070\, Q \times \Delta T \quad \text{(kW)},$$

where P_L is the load power in kW, Q is the water flow in litres per minute and ΔT is the temperature rise in degrees centigrade.

For most practical purposes a temperature rise of about 20 degC is reasonable. With thermometers calibrated to 0.1 degree this will allow accurate measurements, but the water flow must be not so low that hot spots, which may damage the resistor or cause steam formation, can appear.

Load power may also be measured by raising the water to boiling point and converting it to steam. The latent heat of evaporation of water may be taken as 2260 J/g and so the load power is given by

$$P_L = \frac{m \times 2.25}{t},$$

where P_L is the load power (in kW), m is the mass (in grammes) of water converted to steam and t is the time (in seconds). Care should be taken to see that any condensed steam does not fall back into the boiling water otherwise the results will be inaccurate.

One practical measurement method is to fill a sponge with water and to heat it until steam is generated. It is then quickly weighed and replaced in the oscillator. After a convenient time, say 60 seconds, it is reweighed and the amount of water converted to steam calculated.

7.5.2 Efficiency Analysis

To ensure that the maximum r.f. power is available for the load the tube anode efficiency must first be checked. If this is low then the output power will be lower than necessary. A thermometer should be positioned in the water outlet pipe close to the tube and the temperature taken with the filament switched on and the water flow adjusted to a convenient value. The required water flow is given in the published tube data. The generator should then be operated under the expected running conditions and the temperature rise checked. At this point it should be remembered that the water is being heated not only by the anode dissipation but by power lost in the grid and filament. The grid power can be calculated in the usual way, where

$$P_g = 0.9 \times I_{g\ \mathrm{mean}} \times V_{g\ \mathrm{pk}}.$$

This figure should be subtracted from that previously obtained and this will give the anode dissipation P_a. Anode efficiency η_a is given by:

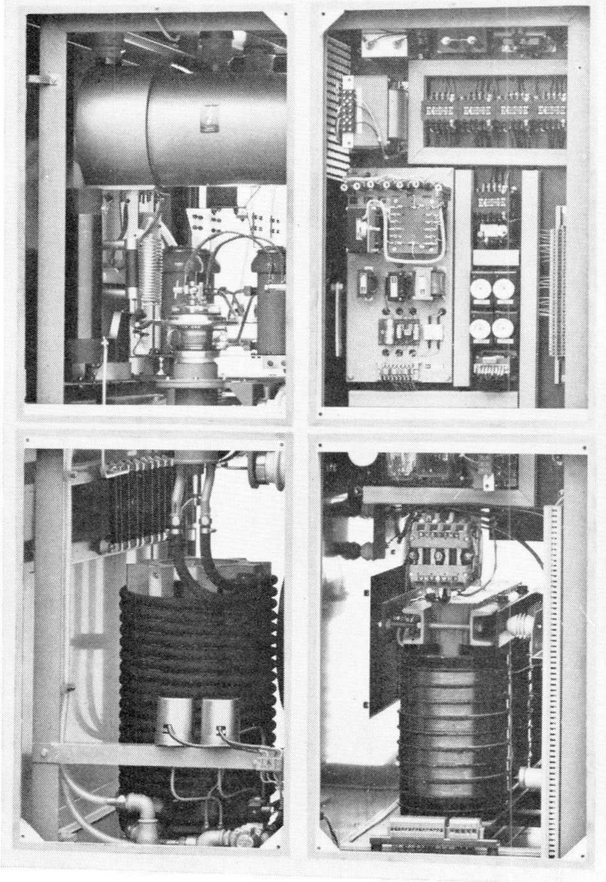
$$\eta_a = \frac{P_{\mathrm{in}} - P_a}{P_{\mathrm{in}}} \times 100 \quad (\%),$$

where $P_{\text{in}} = V_a \times I_a$.

Figures for anode efficiency are given in the published tube data.

Similar techniques can be used for measuring circuit losses of various kinds, and thus a complete generator can be analysed to check that the losses are reasonable.

One precaution should be mentioned. Where a thermometer must be placed in an r.f. field, a spirit-filled type should be used. If only mercury-filled varieties are available, they should be electrically screened.



7 kW generator (frequency 200-800 kHz) using tube type TBW6/14 (by courtesy of Fritz Hüttinger Elektronik GmbH, Freiburg, Western Germany).

8. Applications

In this chapter, details are given of the design of an r.f. induction heater and a dielectric heater. Both designs are based on the information given in Ch. 5 and Ch. 6 and both use one of the new range of ceramic envelope industrial triodes.

It is intended that this chapter will be supplemented by up-to-date application notes as they become available, thus keeping the r.f. heating engineer in touch with the latest developments.

Certain application notes are already available and will be forwarded on request.

8.1 Induction Heating Generator

8.1.1 Design Considerations

Experience of high power industrial heating generators has shown that instability in the generator can often be attributed to parasitic resonances. The presence of parasitic oscillations has tended to be accepted both in normal operation and in fault conditions and the use of anti-parasitic devices which are palliatives has become widespread. By careful circuit design and construction such parasitic oscillations can be avoided under all operating conditions including severe overloads.

Instability in generators can usually be attributed to one or more of the following:

1. resonance of the h.t. feed choke with its associated capacitances
2. blocking capacitors inadequate, i.e. they cause large voltage drops if their self inductance is too high or their effective capacitance is too low
3. resonance of the grid feed choke and blocking capacitor and by-pass capacitor
4. resonance of filament capacitors and chokes
5. inductive interconnections between components, including the tube
6. chassis currents giving rise to unwanted coupling.

The LC filtering arrangements for the supply lines to the tube can cause parasitic circuits to be set up by virtue of their own self resonances. These are caused by their physical size and disposition relative to other components. Damping circuits and parasitic stoppers can also contribute to instability as they are further sources of unwanted circuit resonances. If the normal LC arrangement can be dispensed with, these problems can be avoided and component costs reduced.

Although the frequency of operation is only of the order of hundreds of kHz, to reduce unwanted coupling, undesirable phase shift effects and r.f. voltage drop, it is necessary that intercomponent connections be kept as short and of as low inductance as possible. The result of this careful design is the elimination, at source, of any potential parasitic resonances due to lead inductance.

Chassis currents can cause undesirable feedback effects and should be avoided as far as possible for the following reasons:

1. unwanted coupling due to more than one current following a common path
2. unwanted coupling to components on or near the chassis or screening
3. power loss in the chassis.

These effects have been avoided in the present design by using only one common earth return to the oscillator.

The use, and the disadvantage, of the usual filter system in the anode d.c. supply has been avoided by passing the supply cable down the inside of the copper tube forming the anode inductor.

An oscillator (Fig. 8.1), designed on these principles, was constructed to show the performance of the ceramic envelope industrial triode tubes type YD1172, YD1182 and YD1192 in such a generator.

8.1.2 Tank Circuit Calculations

To calculate the tank circuit components the following figures for normal operation are required:

	YD1172	YD1182	YD1192	
V_a	6.0	7.5	8.0	(kV)
V_a (pk)	5.4	6.8	7.1	(kV)
V_a (r.m.s.)	3.82	4.8	5.03	(kV)
P_{out}	16.25	32.4	63.0	(kW)

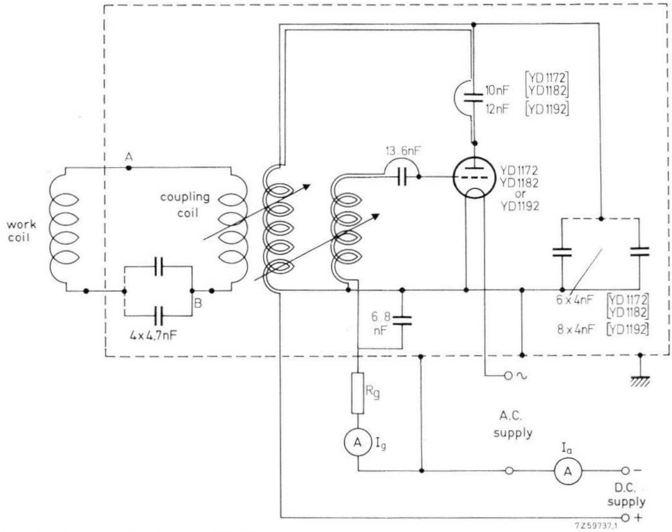


Fig. 8.1. Circuit of induction heater oscillator.

To minimise frequency drift and to ensure reasonable stability without excessive circuit loss a loaded circuit Q of approximately 50 is desirable.

If X_C is the tank circuit capacitance reactance, then:

$$X_C = \frac{V_{a(\mathrm{r.m.s.})}{}^2}{Q_L \times P_{\mathrm{out}}},$$

so for each tube:

	YD1172	YD1181	YD1192	
$X_c =$	$\frac{3.82^2 \times 10^6}{50 \times 16.2 \times 10^3}$	$\frac{4.8^2 \times 10^6}{50 \times 32.4 \times 10^3}$	$\frac{5.03^2 \times 10^6}{50 \times 63 \times 10^3}$	(Ω)
$=$	18.0	14.2	8.03	(Ω)

Therefore, C at 400 kHz

$=$	22.2	28.2	50	(nF)

To operate the three tubes without changing the tank circuit inductance, and accepting a lower Q_L for the highest power tube (YD1192), the tank coil inductance was fixed at 6.5 μH. The capacitance values were then chosen as follows:

	YD1172	YD1182	YD1192	
	24	24	32	(nF)
giving Q_L values of:				
	58	46	34	

The capacitors had to be selected from a standard range and, to keep the volume low and interconnections as short as possible and to meet the ratings required, capacitors type T.C.C. HLC4150 of 4 nF each were chosen to build up the required tank capacitance. It was felt that the resulting small change of frequency from the design frequency of 400 kHz was not important and that the circuit losses when the YD1192 was used would be reduced because of the lower relative circulating current.

8.1.3 Construction

The oscillator was assembled inside an aluminium cabinet 1.0 m high $\times$1.4 m long$\times$1.1 m deep (Fig. 8.2). The filament transformer and grid

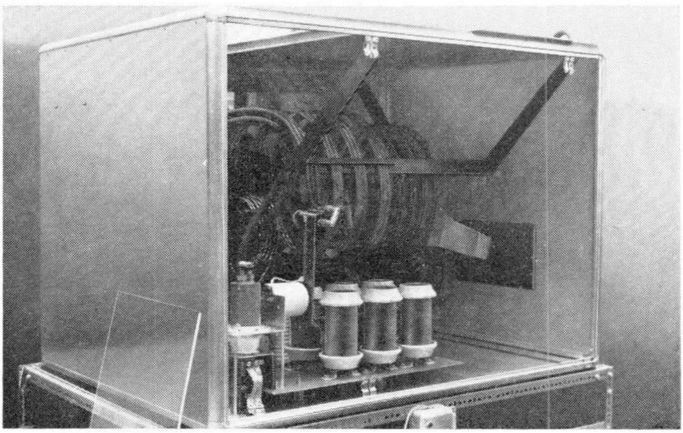


Fig. 8.2. General view of the equipment. The work coil and load are not visible as they are mounted outside on the right-hand panel.

resistor were mounted underneath. The work coil was also mounted outside the cabinet as it would be in practice (Fig. 8.3). The framework and panels were designed to fit together with well butting joints tightly bolted to minimise losses due to any induced currents.

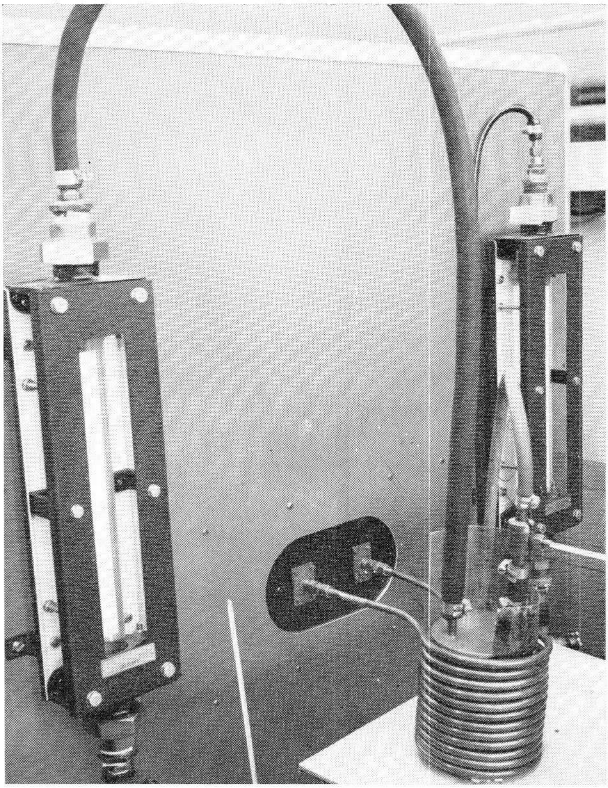


Fig. 8.3. View showing work coil and water cooled load can with mica insulation to prevent shorting of the coil. The flow meters and thermometers used in some of the measurements are also shown.

The tank coil was formed from three 13 mm diameter copper tubes brazed together side by side at 150 mm intervals (Fig. 8.4). The coil internal diameter was 480 mm and the total length was 355 mm made up by evenly spacing 3.75 turns of the three tube assembly. The coil was secured in position with Paxolin clamps.

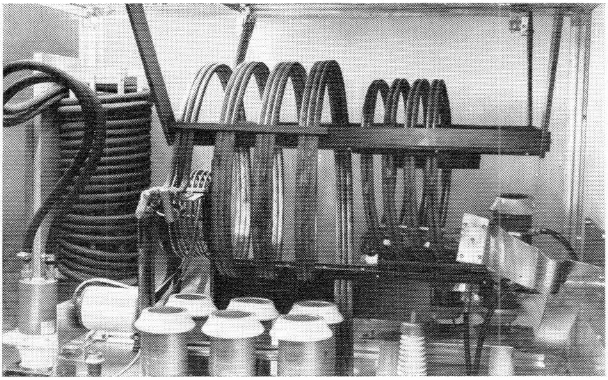


Fig. 8.4. A detailed internal view showing typical positions of grid, tank and coupling coils.

The earthy end of the tank coil was clamped to a copper plate 305 mm $\times$690 mm $\times$2 mm thick carrying the tank capacitors. The other end was clamped to a low inductance copper strap connecting the live ends of the tank capacitors and one end of the anode blocking capacitor together. The maximum tank circulating current (425 A) requires that the coil be cooled. Water for this purpose was passed along one outer tube and back through the other, both input and output being at the earthy end of the coil. The centre tube carries the d.c. supply cable to the anode. The cable, fed in at the earthy end of the coil, emerged at the live end and was dressed so as to lie across the blocking capacitor and thence to the tube anode.

The copper strap connecting the live ends of the tank capacitors to the tank coil was also cooled by the tank coil cooling water by means of a pipe soldered to the strap. This assists the cooling of the capacitors.

The tank capacitors were mounted closely together on the copper plate connected to the earthy end of the tank coil. The plate was supported 50 mm above the chassis by insulators and bonded to it at the tube end (Figs. 8.2 and 8.6).

27.12 MHz generator supplying 10 kW at the electrodes, using tube type YD1173 (by courtesy of Körting Radio Werke GmbH, Grassau, Western Germany).

Considerations of tube-to-circuit connection suggested that the tube be mounted anode upward. This gives short anode and cathode connections and has the advantage of keeping the anode, with its high potential, clear of the cabinet (Figs. 8.5 and Fig. 8.6). Special measures must then be taken, however, to protect the equipment against water spilled during tube replacement.

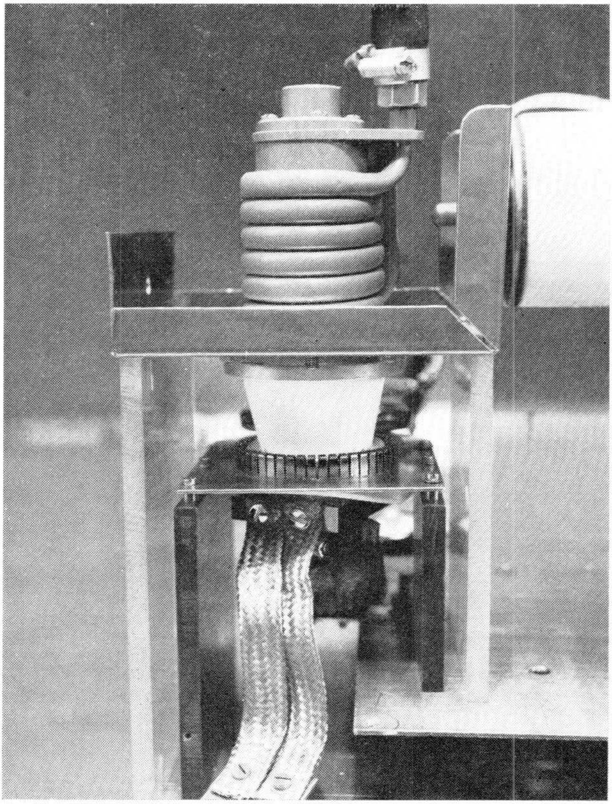


Fig. 8.5. The YD1172 mounted in the cabinet.

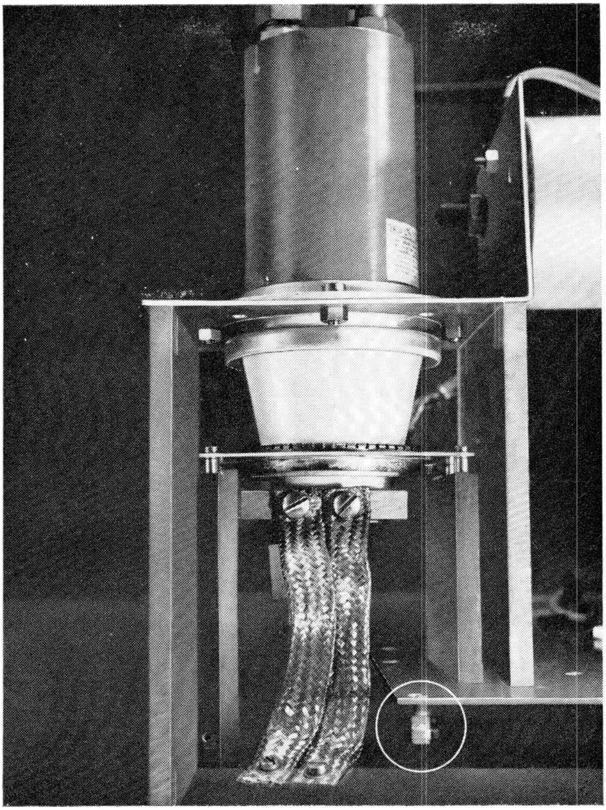


Fig. 8.6. The YD1182 mounted in the cabinet. The earthing bolt connecting the tank capacitor plate to the chassis is visible at the bottom centre.

The grid coil, consisting of 7 turns of 190 mm internal diameter, was constructed from a single tube of 6 mm diameter copper with the connection from the grid terminal to the grid resistor running down the centre of the tube. The coil, 90 mm long, was wound on a simple insulating framework mounted so that it could be moved relative to the tank coil to vary the drive. The ends of the coil terminate in copper braid to provide the flexibility required to make drive adjustments. The requirement for a separate grid choke was thereby eliminated. The dimensions and coupling position were determined by experiment to provide adequate adjustment of drive for all three tube types.

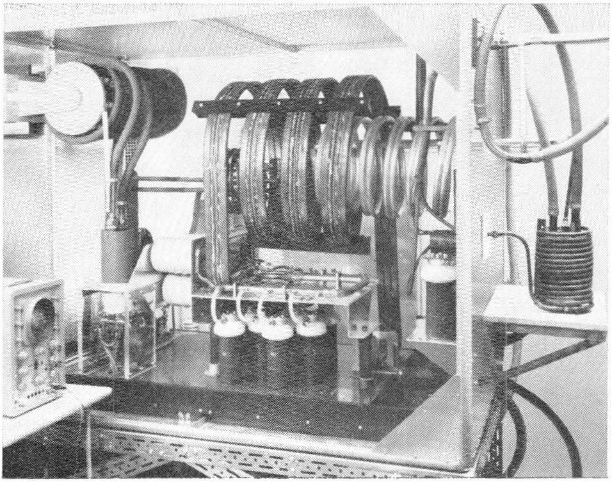


Fig. 8.7. View of a 240 kW induction heater, constructed on similar principles as the 60 kW equipment described in the text.

A single blocking capacitor of 10 nF, T.C.C. type 1052, was used for the YD1172 and YD1182. Because of the higher anode potential and increased r.f. current of the YD1192 a bank of three higher voltage capacitors was used for this tube, mounted vertically one above the other.

The coupling coil (Fig. 8.4) was manufactured from two 13 mm copper tubes spot brazed together at 150 mm intervals to form a double tube. It consisted of 4 turns of 380 mm internal diameter spaced to give a total length of 200 mm. Cooling water was passed along one tube and back through the other. The coil was mounted on a simple Paxolin slide attached to the tank coil to enable the coupling between the tank circuit and load to be adjusted.

A bank of 4 nF capacitors, similar to those used in the tank circuit, and variable in number, was provided to complete the series connection between the coupling coil and the work coil. By varying the number of capacitors and the relative position of the coupling and tank coils an optimum power match could be achieved.

The work coil was made from a single, water cooled, 10 mm copper tube. It consisted of 12 turns closely spaced with an internal diameter of 120 mm (Fig. 8.3).

It was required that the test load should simulate a typical eddy current heated work piece and that the load power should be measured at the same time. To this end a closed can made from 1.5 mm steel sheet was used as the load, provision being made to pass cooling water through the can and to measure both the rate of flow and the temperature rise. The can was tightly coupled to the work coil and the power dissipated in the can by induced eddy currents was transferred to the cooling water. The can measured 150 mm high $\times$ 115 mm diameter (Fig. 8.3).

8.1.4 Measurements

The water flow was arranged in three separate circuits enabling the sources of power loss to be separately determined, namely: anode loss, load power, and the tank coil, coupling coil and work coil losses taken together as a single unit. Thus anode dissipation, load power and circuit losses can all be calculated independently (allowance being made for drive power and filament heating in the temperature rise of the anode cooling water).

The water flow rates were as follows:

	YD1172	YD1182	YD1192	
Anode	7.15	7.15	14.3	(l/min)
Load	7.15	14.3	19.0 approx.	(l/min)
Coils	2.4	7.15	3.6	(l/min)

For most tests the water flow was chosen to simplify the calculation of power dissipation. (The coil water flow for the YD1182 could be halved.)

In each water circuit a single thermometer was placed in the outlet side and was used to measure initial and final temperatures, the former being checked before and after heating. Test results are shown in Table 8.1.

Losses unaccounted for (P_{loss}) include cabinet and radiation losses and these are given by:

$$P_{loss} = P_{out} - P_{load} - P_{cct} - P_{drive}.$$

Therefore

	YD1172	YD1182		YD1192	
P_{loss}	0.245	1.21	1.64	1.04	(kW)

These figures as a percentage of P_{out} are:

	1.5	3.8	5.0	1.7	(%)

Note: Since the figure for P_{loss} is obtained by subtraction of relatively large measured or calculated powers from the input, small errors in these measurements produce greater errors in the figure obtained for P_{loss}.

Table 8.1. Results obtained when the operating conditions were set for the three tubes from the published data.

	YD1172	YD1182		YD1192	
V_a	6.0	7.0	7.5	8.0	kV
I_a	3.4	5.72	5.4	10.0	A
$I_{g\ (loaded)}$	0.91	1.46	1.41	2.3	A
$I_{g\ (off\ load)}$	1.36	2.4	2.23	3.5	A
R_g	500	375	450	300	Ω
$I_{g\ (max.\ publ.)}$	1.5	2.4	2.4	3.5	A
$P_{(load)}$	13.5	25.5	25.8	53.5	kW
P_a	4.15	8.45	7.85	18.4	kW
$P_{(cct)}$	1.83	3.5	3.8	5.0	kW
$P_{(out)}$	16.25	31.6	32.65	61.6	kW
η_a	79.5	79.0	80.5	77.0	%
$\eta_{(load)}$	66.0	63.0	63.5	67.0	%
$\eta_{(trans)}$	85.0	83.0	82.0	87.0	%
$*P_{(grid\ resistor)}$	0.414	0.80	0.894	1.32	kW
$*P_g$	0.303	0.592	0.52	0.74	kW
$*P_{(drive)}$	0.717	1.392	1.414	2.06	kW
$T_{(anode\ seal)}$	149	149	149	149	°C
$T_{(grid\ seal)}$	170	149	149	170	°C
$T_{(cathode\ seal)}$	191	185	185	185	°C
$T_{(filament\ seal)}$	204	170	170	195	°C

* Calculated.
Seal temperature measurements were made without any air cooling.
Final approximate temperature.

Table 8.2. The following results were obtained with the 240 kW induction heater shown in Fig. 8.7,* using the YD1202 and the YD1212.

YD1202

V_a (kV)	I_a (A)	P_{in} (kW)	I_g (A)	P_g (kW)	P_a (kW)	$P_{(out)}$ (kW)	η_a (%)	P_{ld} (kW)	η_{ld} (%)	$P_{(\text{tank cct.})}$ (kW)	$P_{(\text{load cct.})}$ (kW)	η_{cct} (%)	R_g (Ω)	$P_{(\text{drive})}$ (kW)
10.0	15.8	158	3.5	1.8	33	125	79.2	111	70.6	3.2	4.5	92.5	200	4.25
12.0	13.0	156	2.7	1.2	31	125	80.0	110	70.6	5.0	4.5	91.0	300	3.38

YD1212

14.0	23.5	329	4.5	1.4	88	241	73.2	226	68.6	5.7	10.7	95.5	160	4.65

*) A full description of this heater is available on request.

Switched overload and off-load switching tests on the oscillator were satisfactory; no parasitic oscillations were evident. The oscillator was overloaded to the point where I_g was reduced to 0.1 A and 0.9 A for the YD1172 and YD1182 respectively. Power supply limitations prevented an overload test on the YD1192.

With the YD1192 in circuit the grid coil became warm. For a final design a larger diameter tubing (8 mm) would alleviate this problem.

Tests were carried out with pieces of mild steel block and tube and the time required to heat such samples to red heat was consistent with the expected load power, having regard to the mass of the individual test pieces.

With the YD1182 in circuit the effect of earthing the load circuit was examined. Each of the points *A*, *B* and *C* in Fig. 8.1 was earthed in turn to the cabinet by a low inductance copper strap, but no change in operation could be detected.

8.1.5 Summary

An improved form of high power induction heater, operating at 400 kHz, has been designed without chokes and anti-parasitic devices. The r.f. circuit elements are used to feed the d.c. potentials to the generator tube terminals. High load efficiencies and complete freedom from parasitic oscillations have been obtained even at severe overloads.

The oscillator has operated satisfactorily at output powers up to 60 kW with three of the new range of ceramic envelope industrial triodes, namely the YD1172, YD1182 and YD1192.

Table 8.2 shows the results that were obtained from a similarly constructed induction heater rated at 240 kW, using the YD1202 and the YD1212. (See Fig. 8.7).

8.2 Dielectric Heater Using Half-Wave Line at 30 MHz

A general purpose dielectric heater operating at 30 MHz and suitable for the pre-heating of plastics or process drying is now described. The circuit is so arranged that the nominal power level would be 15 kW when fitted with the YD1172 or 30 kW with the YD1182.

For a given frequency the required heating losses in a dielectric load material are proportional to the square of the applied r.f. voltage. It was therefore decided to make the half-wave line transformation from tube to load at a ratio compatible with the highest possible voltage at the load position without giving undue rise to corona or ionization discharges. Allowing for variations in normal ambient conditions such as air temperature, humidity and contamination, r.f. peak voltages of about 12 to 14 kV were considered practicable.

8.2.1 Design Considerations

The operating conditions of the two tubes on which the further circuit design is based are as follows:

	YD1172	YD1182	
V_a	6	7	(kV)
I_a	3.4	5.72	(A)
$-V_g$	460	564	(V)
$+V_g$	340	436	(V)
I_g (on load)	0.92	1.5	(A)
I_g (off load)	1.35	2.2	(A)
R_{g-f}	500	375	(Ω)
P_a	4.2	8.6	(kW)
P_{out}	16.2	31.44	(kW)
P_{load}	13.5	27	(kW)
η_a	79.4	78.5	(%)

The resultant instantaneous values for the peak anode swing are

	YD1172	YD1182	
$v_{a(pk)}$	5.4	6.4	(kV)

and therefore, if the tank circuit is to be connected between grid and anode,

$v_{g-a(pk)}$	6.2	7.4	(kV)

With a generator designed for both tube types, and hence for two different sets of operating conditions, the tank circuit capacitor must be so dimensioned that the resultant Q_L values will sustain stable oscillations under industrial conditions with the larger tube (lower anode impedance) and cause no excessive circuit losses with the smaller one (higher anode impedance), that is, Q_L should lie between 25 and 50.

From the above tube data a transformation ratio of 2 : 1 can be seen to be satisfactory and since this, together with the load position capacitance, will determine the generator tank capacitance, an assessment must be made to establish whether the resultant loaded Q will satisfy the above conditions.

If the noted load powers are to be dissipated in a dielectric load, a specific load volume must be considered. The design example load was formed by a number of cylindrical pellets of wood-loaded urea formaldehyde, each having a diameter of 5 cm, a height of 3.5 cm and weighing 75 g.

The material has, at room temperatures, a dielectric constant (ε'/ε) of 5.1, a loss factor (tan δ) of 0.05, and a specific heat (S) of 0.4.

Assuming, in the first instance, no changes of these figures during the process, the weight of load (m) which could be raised in temperature by 100 degC with 13.5 kW (YD1172) in 15 seconds, can be derived from the equation

$$P_{\text{load}} = \frac{(T_2 - T_1) \times m \times S \times 4.2}{t}.$$

Therefore,

$$m = \frac{P_{\text{load}} \times t}{(T_1 - T_2) \times S \times 4.2} = \frac{13\,500 \times 15}{100 \times 0.4 \times 4.2} = 1210 \text{ g}.$$

As soon as heating starts, the dielectric constant and, to a much larger extent, the loss factor will increase (the latter to about double). The processing conditions can therefore only apply during the power peak of the processing cycle. It is this figure that is the generator rating.

Further calculations were based on the changed loss factor and, assuming that there will also be some thermal losses, showed that 1000 g (or 14 test pellets) represent a reasonable test load for the YD1172. For the YD1182 this would be doubled. The capacitive area occupied by these pellets will then be 275 cm^2 and the resultant load capacitance given by

$$C_w = 35 \text{ pF}$$

and

$$X_C = 150\ \Omega.$$

For the increased loss factor the potential difference V across the load will be given by

$$V = \sqrt{\left(\frac{P_{\text{load}} \times X_c}{\tan \delta}\right)} = \sqrt{(13\,500 \times 150 \times 10)} \qquad \text{(V)}$$

$$= 4.5 \text{ kV}_{\text{r.m.s.}} \text{ or } 6.3 \text{ kV}_{\text{pk}}$$

By capacitive division the necessary air gap above the load material will then be 7.5 mm.

Although load plates made to the above dimensions would be adequate, other practical considerations suggested that their area should be much larger than the calculated values.

1. If the load plates are no larger than the dimensions of the load material, the latter will be subject to the capacitive edge effects of the load plates, and uneven heating will result.
2. Larger area load plates will allow the accommodation of a wider range of load shapes in a general purpose dielectric heater.
3. The total resultant load position capacitance can thus be made considerably larger and thereby used to swamp some of the capacitive changes that take place in the load material during processing. An improved frequency stability of the generator will result.

For these reasons a load plate size of 50×50 cm $= 2500$ cm^2 was considered adequate for a large range of possible pre-heating and process drying applications. The opposing capacitive plate, for reasons similar to those above, must be adjustable. A mean plate spacing of 5 cm was assumed in the calculations of the load position capacitance, making this capacitance 44 pF. This is sufficiently low to permit the addition of a set of differential capacitive plates that will partly compensate for any capacitive changes caused by different settings of the dielectric gap.

The final load position capacitance was accordingly made

$C_L = 50$ pF.

When calculating the generator end-capacitance, allowance was made for the fact that capacitors, comparable in their dimensions to the inductive path length of the circuit in which they are mounted, will show a significant voltage decay along their surface in the direction of the r.f. current flow, with the consequence that their effective contribution to the circuit capacitance is smaller than their calculated value would suggest.

This effect will increase with increasing capacitance, since, for a given dielectric and frequency, both the physical size as well as the r.f. currents causing the potential decay along the surface will increase, and the inherent inductive component will constitute a larger proportion of the total circuit inductance.

If, then, a line transformer is to be constructed with a transformation ratio larger than one, the ratio of the end-loading capacitances must be larger than the theoretical calculation suggests, so that instead of

$$T = \frac{C_G \cos \theta_L}{C_L \cos \theta_G} \quad \text{(see Ch. 5.2.10)}$$

the empirical relation

$$T' = \sqrt{\frac{C_G}{C_L}}$$

more accurately satisfies practical requirements. In the above expressions T and T' are the transformation ratio, C_G and C_L the end-loading capacitances and $\cos \theta_1$ and $\cos \theta_2$ the corresponding lengths of the shortened line measured in radians.

With the stated requirement that

$$T' = 2 : 1$$

the end-loading capacitances need to have a ratio of

$$C_G/C_L = 4.$$

Therefore

$$C_G = 200 \text{ pF}.$$

This value includes the C_{a-g} (24 pF) of the YD1172 and will be only slightly larger (208 pF) for the YD1182 for which the C_{a-g} is 32 pF.

The resultant Q_L values for the two operating conditions will then be

		YD1172	YD1182
Q_L	=	47	38

Both are acceptable for industrial operation. (These figures ignore the distributed line capacitance which is very small compared with the loading capacitance.)

8.2.2 Line Dimensions

All further line dimensions are mainly determined by the electrical requirements as well as physical size of the end-loading capacitances, including the fact that their major dimensions, in order to exert the maximum capacitive effect on the line, should be at right angles to the line axis. As the load plate is 50×50 cm and one of the end-faces of the line-outer conductor forms the opposing capacitive area, the line-outer was given a cross-sectional area of 70×70 cm, allowing for an adequate voltage gap to the edge of the load plate.

The dimensions of the line-inner conductor are then such that the conductor is subject to an amount of undesirable losses, but these are still acceptable for most applications. One of the aims of the present design is the construction of a circuit with minimum losses. The natural line impedance was 77 Ω and was chosen as a compromise between line length and losses. For this the ratio of line-outer diameter (D) to line-inner diameter (d) is

$$D/d = 3.6.$$

This results in a line-inner cross-section of 19.5×19.5 cm. The line length was then calculated by treating the generator section of the line and load section of the line as separate quarter-wave resonators and the following values were found:

Length of generator section	$l_G =$	40 cm
Length of load section	$l_L =$	130 cm
Half-wave length	$l_{\lambda/2} =$	170 cm.

Some modification of these figures was necessary to account for the physical size of the end-loading capacitances. The load-position capacitive plate is equivalent in its inductive value to about 20 cm of the calculated line length, reducing l_L to 110 cm. To an even larger extent this argument applies to the generator end capacitance of 200 pF, because this value must be accommodated within the given line cross-section whilst maintaining an industrially safe dielectric spacing of 3.5 cm. The available capacitive area at the generator end-face of the line is thus only 63×63 cm, resulting in a capacitance of 97 pF. The remainder was made up by extending the capacitive area, with the same dielectric spacing, parallel to all four sides of the line outer, by 25 cm and adding the C_{a-g} of the smaller tube. The capacitance thus obtained would be 118 pF, but is reduced to the necessary 103 pF through the loss of area caused by a ventilation grille.

The inductive contribution to the line length of this box-shaped capacitive extension, together with a 10 cm taper between the latter and the 77 Ω line-inner, is approximately equivalent to 60% of the previously calculated generator line length of 40 cm so that the total overall line length is:

Load line section:	130 cm
Equivalent inductive length of load plate:	—20 cm
Calculated generator line section:	40 cm
Taper:	10 cm
Capacitive box depth:	25 cm
Equivalent inductive length of taper and capacitive box:	—25 cm
Total length of $\lambda/2$ line-inner:	160 cm

An undesirable result of the abrupt dimensional changes from the 77 Ω line-inner to the capacitive box, as well as the relatively great length of the box in comparison with the originally calculated generator section line length, is the creation of a secondary resonance. However, since its origin is known, measures to render this resonance ineffective so far as the oscillator functions of the tube are concerned, are fairly simple.

8.2.3 Oscillator Circuit

For the oscillatory functions, only the generator section of the half-wave line need be considered. With the anode connected to the line-outer and the grid to the line-inner, as shown in Fig. 8.8, the effective tank circuit appears between grid and anode, and the basic configuration for a common grid oscillator is established.

The necessary 'drive circuit' reactance between grid and cathode is furnished by a tuned circuit of relatively high C/L ratio assisting in the suppression of higher frequency modes or harmonics. The resonance frequency of this circuit is, moreover, set about 40% below that of the tank circuit to make the feedback conditions less dependent on the frequency changes that may occur in the tank circuit, especially with varying settings of the dielectric gap at the load position. For the nominal tank circuit frequency of 30 MHz this drive circuit was tuned to 18 MHz.

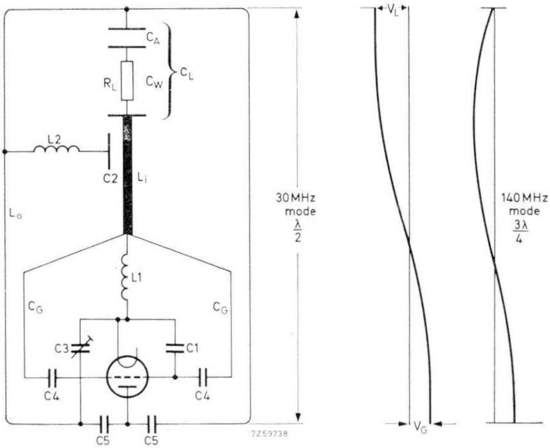


Fig. 8.8. Basic circuit of a half-wave line dielectric heater, not to scale. The components are as follows: R_L, resistance of load material; C_W, capacitance of load material; C_L, load position capacitor, 50 pF; C_G, generator tank circuit capacitor, 200 pF; C_1, drive circuit loading capacitor, 140 pF; C_3, feedback capacitor, 2 to 20 pF; C_4, grid blocking capacitor, 2.5 mF (p.t.f.e. sandwich); C_5, anode blocking capacitor, 1.5 nF (p.t.f.e. sandwich); L_i, 77 Ω line section; L_0, line-outer case; L_1, drive circuit inductor. The components L_2 and C_2 are for absorbing a secondary mode at 140 MHz and are described in the text.

Although this setting is in no way critical, care was taken to avoid any of the sub-harmonics of the oscillator as these might increase the harmonic content in the anode waveform.

The feedback system for a common grid-tuned anode-tuned cathode circuit is provided by the addition of external cathode-anode capacitance, the amount being governed by the feedback ratio of the tube. The appropriate values of capacitance for the two tubes operated under the conditions given in Ch. 8.2.1, are then,

	YD1172	YD1182	
C_{feedback}	5.7	9.4	(pF)

Because of the r.f. voltage losses that are to be expected in the feedback structure, these values will in practice be somewhat higher, and are adjusted with the help of a variable capacitive component. Practical useful limits of this capacitive variation lie between -20% and $+50\%$ of the design value. This precaution, if necessary, also caters for operating conditions other than those specified.

In addition to the basic half-wave circuit, Fig. 8.8 also shows the approximate r.f. potential distribution at the fundamental frequency as well as the high frequency mode generated by the tank circuit capacitive box. The resonance of this latter mode was found to be at 140 MHz and was rendered harmless by the insertion of an absorption circuit at one of its voltage maxima.

8.2.4 Circuit Layout

The practical layout of the half-wave dielectric heater follows essentially the outlines suggested by the basic circuit of Fig. 8.8 and is shown in more detail in Fig. 8.9 where blocking capacitors, feed lines etc. are included.

The line-outer, which is also the main supporting structure, forms a prismatic box of $70 \times 70 \times 195$ cm; the extra height contains the motorized load matching mechanism. The structure is shown in Fig. 8.10.

The tube anode blocking capacitor, a p.t.f.e. sandwich with 1.5 nF capacitance, is mounted at the bottom end-face of the line-outer. The tube, being supported at this level by its mounting flange, protrudes with the bulk of its anode structure and cooling system below the line box. Since the outer surface of the anode is thus outside the r.f. field of the line, the water hose coil need be dimensioned to cope only with the d.c. supply voltage drop. A further capacitive plate of the same dimensions as the blocking capacitor is electrically connected with, and placed parallel to, the blocking capacitor so that the dielectric spacing between anode and grid level will be 3.5 cm.

The line-inner, consisting of the tank circuit capacitor box and the 77 Ω line section and loading plate, is supported by the bottom face of the tank capacitor box, at the level of the grid ring of the tube, by four polypropylene pillars (Fig. 8.11). Its lateral positioning is determined by two 10 cm diameter polypropylene tubes between two opposing vertical panels of the line-outer, 25 cm from the top end of the line. These supporting tubes serve at the same time as air intake ducts for the fan that is mounted inside the line-inner and used to provide the necessary forced air cooling for the tube header.

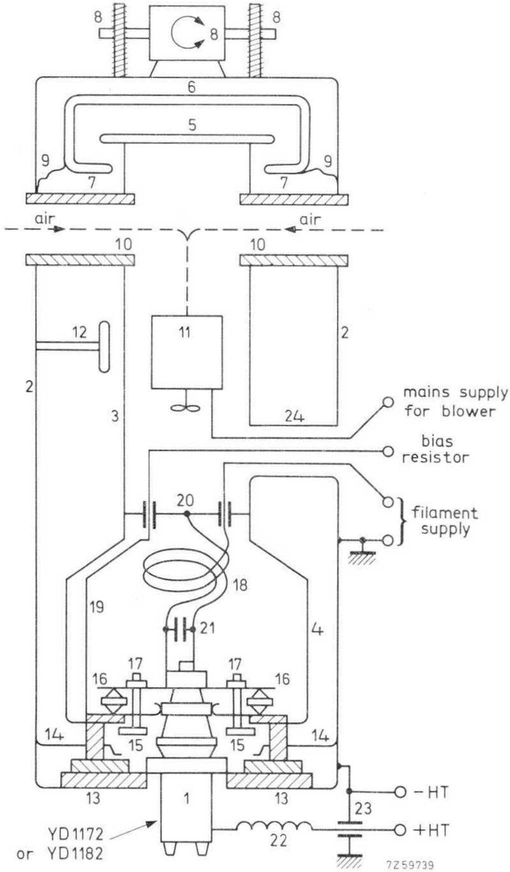


Fig. 8.9. Sectional sketch of half-wave dielectric heater (not to scale). Components are as follows: (1) tube (2) line-outer (3) 77 Ω line-inner (4) generator tank capacitor box (5) lower load capacitive plate (6) upper load capacitive plate (7) differential capacitor plates (8) motor and gears for load setting (9) copper foil straps (10) air ducts (11) coaxial blower (12) 140 MHz mode suppressor circuit (13) anode blocking capacitor (14) grid blocking capacitor (15) feedback capacitor (16) drive circuit loading capacitors (17) collets for feedback adjustment (18) drive circuit inductor and filament supply (19) grid bias lead (20) bias and filament decoupling (21) filament decoupling capacitor (22) h.t. filtering choke (23) h.t. filtering capacitor (24) feed-through tube.

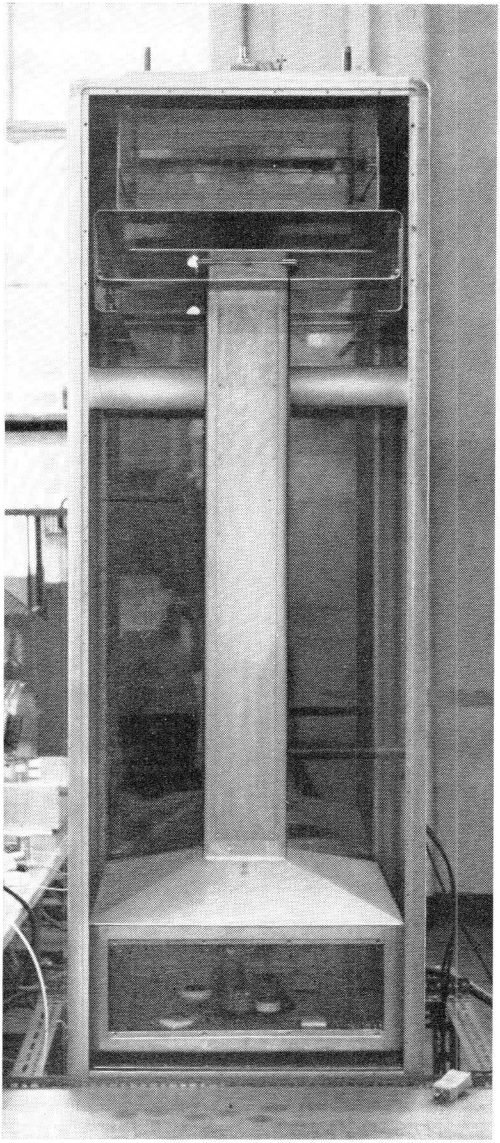


Fig. 8.10. Overall view of 30 MHz half-wave line dielectric heater, with the front panels of the line-outer and the tank circuit capacitor box removed.

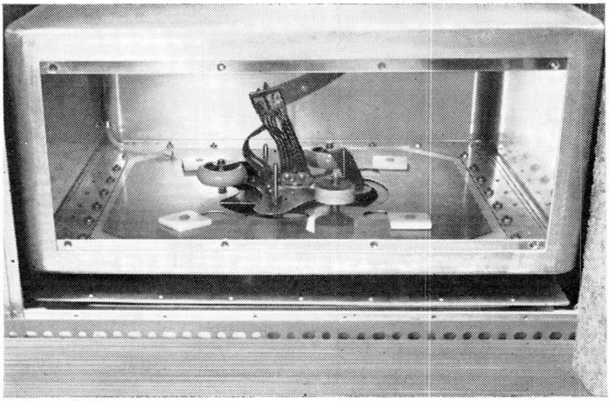


Fig. 8.11. Tank circuit capacitor box, containing the grid-filament circuit. The large p.t.f.e. sandwich capacitor at the bottom of the box is the grid blocking capacitor. The grid bias lead is connected at the rear left-hand corner and is dressed along the inside of the box for decoupling purposes.

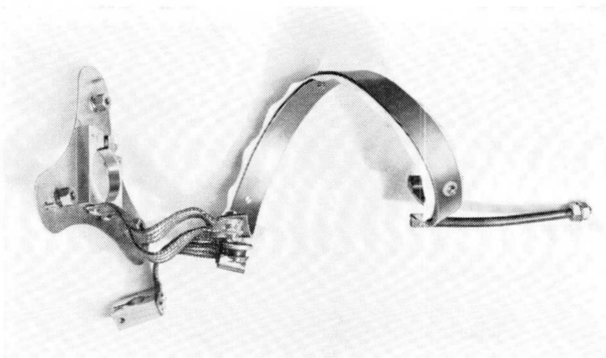


Fig. 8.12. Grid-cathode circuit inductor with filament connectors.

The inside of the tank circuit capacitor box contains the grid-filament circuit and is connected through the grid blocking capacitor and a finger strip connector to the grid of the tube. Fig. 8.12 shows a $1\frac{1}{2}$ turn coil of 17 cm diameter which serves the dual purpose of drive circuit inductor and filament supply leads. It is manufactured from two 55 cm copper bars, of cross section 5×25 mm, wound in parallel and interleaved with 0.5 mm thick p.t.f.e. strip. At the lower end it is joined with short pieces of flexible copper braid to the respective filament connectors (Fig. 8.13) and decoupled with a 2 nF capacitor. One of the upper limbs of this coil is fixed to the centre of the top of the tank circuit capacitor box and the other with a feedthrough capacitor to the inside of the lower part of the 77 Ω line-inner section. The lower filament connector (r.f. cathode) is fitted with a triangular plate, similar to that of the grid connector, but off-set by 60°, and carries on its upper surface three collets holding the C_{a-f} feedback capacitor. Between the underside of this plate and the grid blocking p.t.f.e. sandwich capacitor, there are three ceramic capacitors serving as grid-cathode circuit loading capacitances with a total value of 140 pF. Into the collets on the cathode connector are fitted three 6.5 mm rods which protrude downwards through gaps in the grid blocking capacitor and carry a ring-shaped electrode facing the anode blocking capacitor, thereby forming the adjustable feedback capacitor.

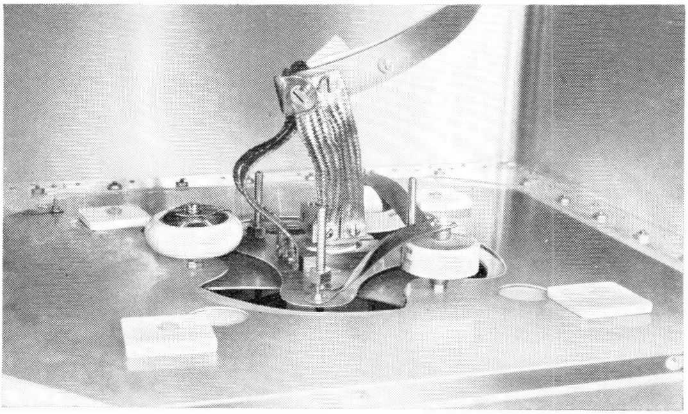


Fig. 8.13. Grid-cathode circuit and feedback assembly. Three ceramic capacitors (total 140 pF) are grouped round the tube header and connected between filament (cathode) and grid plates. The three adjustable rods, holding the feedback ring, can be seen protruding from the collets on the filament (cathode) plate.

Near the zero voltage point on the 77 Ω line-inner, a brass tube with a 3 cm diameter is fitted between the line-inner and the line-outer. This can be seen in Fig. 8.14 and it fulfils several functions: firstly, it serves as a d.c. earth connection to the line-inner; secondly, it carries one limb of the tube filament current; finally, it screens from r.f. the other filament limb as well as the grid bias lead and the mains supply leads to the blower motor (mounted inside the line-inner). At first sight, it might appear that

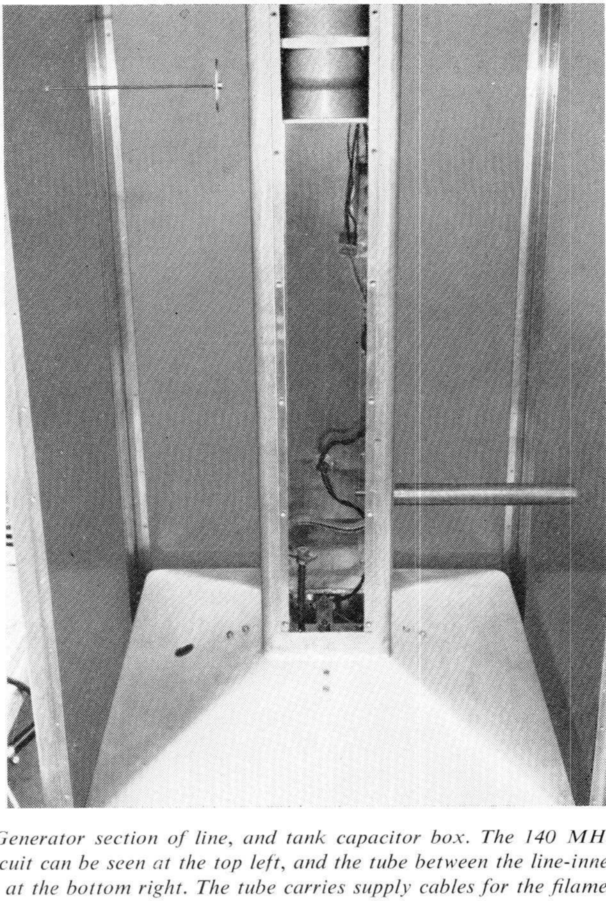


Fig. 8.14. Generator section of line, and tank capacitor box. The 140 MHz mode absorption circuit can be seen at the top left, and the tube between the line-inner and outer can be seen at the bottom right. The tube carries supply cables for the filament, the grid bias, and mains for the coaxial air blower. Part of the blower is visible at the top, inside the line-inner.

this brass tube would tie the line-inner to the line-outer so far as their respective r.f. functions are concerned, and thereby split the half-wave line into two quarter-wave circuits, losing the advantage of the floating zero voltage point. This is not so, however, since the diameter of this tube is very much smaller than that of the line-inner and will therefore represent a relatively high impedance in parallel with the much lower one of the 77 Ω line-inner. Its resultant effect on the position of the zero voltage point, the operational frequency and the transformation ratio is negligible.

The load position (Fig. 8.15) differs little from that found in a conventional dielectric heater. It consists of a horizontal 50 × 50 cm loading plate at the top end of the line-inner together with the opposing capacitive plate connected with copper foil straps to the line-outer. Two smaller

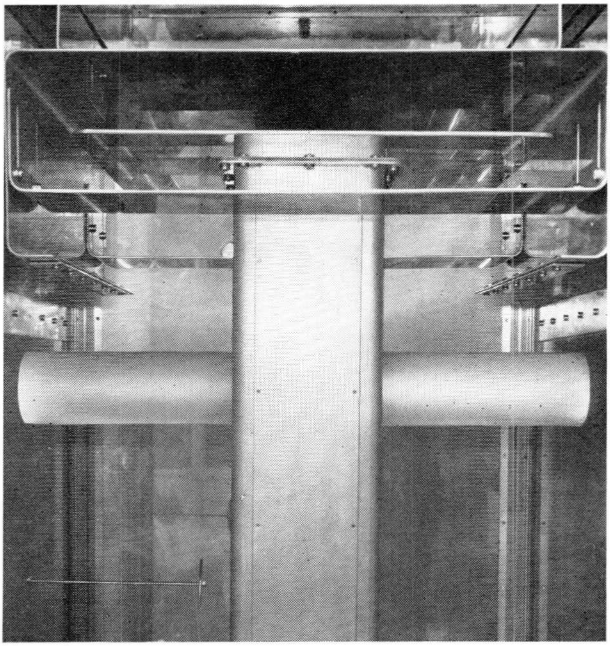


Fig. 8.15. Load section of the half-wave line with the load position capacitor. The differential capacitive plates can be seen to be preset, adjustable in relation to the main variable plate. The horizontal polypropylene tubes can also be seen.

capacitive plates connected to this opposing plate but facing the underside of the load plate act as a differential capacitor, minimising the frequency shift when the dielectric gap is adjusted with the aid of the motor and gear train located above.

The acceptor circuit, suppressing the 140 MHz mode referred to earlier in the text and shown in Fig. 8.8, consists of a 22 cm $\times$ 4 mm threaded rod fixed at one end to one side of the line-outer and carrying on its other end a capacitive disc. This disc has a diameter of 6 cm for the YD1172 and 8 cm for the YD1182, and faces the line-inner at a distance of 4 cm. Its vertical position is 110 cm from the bottom of the line. This position was determined by the location of the most accessible voltage maximum of the 140 MHz mode.

8.2.5 Performance

With the differential capacitor plates fitted at the load position, the dielectric gap set to 5 cm and no load inserted, the operational frequency (f_o) was found to be 31.4 MHz. Changes in this frequency with varying dielectric gap, both with and without differential plates, are shown in Fig. 8.16, where it can be seen that the stabilisation of frequency with the help of the differential plates is considerable. They were therefore used in all the measurements noted.

Power measurements, made with the conditions given in Ch. 8.2.1, were, in the first instance, made with the help of a 50 Ω matched and water-cooled load. The energy transfer from the half-wave line to the load cable was effected by the insertion of a capacitive probe between

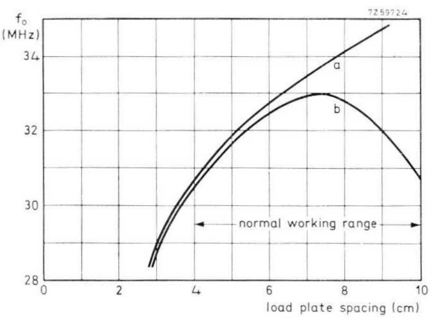


Fig. 8.16. Variation of operating frequency with spacing of the load capacitor (a) without compensation (b) with compensation by differential plates.

the lower and the upper load plates. The results obtained are shown in Table 8.3.

The overloaded and unloaded conditions shown in the above columns were measured without further adjustments to the drive circuit and show satisfactory tube and circuit performance. The 'normal load' results were used to calculate the approximate circuit losses (P_c) as follows:

	YD1172	YD1182	
P_{out} =	16.30	31.25	(kW)
$-P_{drive}$ =	$-$ 0.78	$-$ 1.16	(kW)
$-P_{load}$ =	$-$14.50	$-$28.50	(kW)
P_c	1.02	1.59	(kW)

These give the corresponding circuit transfer efficiency figures (η_c) below.

(YD1172) $\eta_c = 94\%$ (YD1182) $\eta_c = 97\%$

For practical operation, the design example material of woodloaded urea formaldehyde pellets was used. Press moulding technology requires that the temperature of this type of material should rise by a minimum of 100 degC in less than 45 s of preheating. Table 8.4 gives sets of figures for the start and finish of the process.

To check both the operational stability of the oscillator and the functioning of the 140 MHz-mode suppression, switching tests were carried out under off-load, full-load and overload conditions. There were no parasitic or moding symptoms. For the given dimensions and mounting position the suppressor circuit has an operational bandwidth of about 20 MHz at the 140 MHz mode. Since this mode also has a voltage minimum at the load position and is therefore little changed by capacitive changes at this point, these figures show that the 140 MHz suppression is adequate.

Throughout these tests the r.f. potential distribution along the half-wave line was monitored with a neon probe and conformed under all conditions to the half-wave pattern.

Because the generator circuit and load circuit are integral and self screening, no appreciable radiation was expected. To confirm this, a detector probe fitted with a 4662 neon indicator was passed over the outer skin of the line box and the supply leads. Such a device will normally register radiation intensities of the order of 5 mW/cm^2 and as no ionisation could be observed, the intensity of any residual radiation is below this level.

Table 8.3. Performance of YD1172 and YD1182 in half-wave dielectric heater into matched load.

	YD1172			YD1182			
	normal load	over-load	off load	normal load	over-load	off load	
V_a	6	6	6	7	7	7	kV
I_a	3.4	4.7	0.54	5.7	7	0.8	A
I_g	0.95	0.67	1.36	1.32	1.1	1.85	A
R_g	500	500	500	375	375	375	Ω
p_a	4.1	8.8	—	8.75	15.1	—	kW
η_a	80	69	—	78	69	—	%
$P_{(load)}$	14.5	17	—	28.5	30	—	kW
$\eta_{(load)}$	71	60	—	71	61	—	%

Table 8.4. Performance of prototype dielectric heater with practical test load.

	YD1172		YD1182		
Weight of material Processing time	1 15		2 15		kg s
	start	finish	start	finish	
V_a	6	5.8	6	5.6	kV
I_a	2.8	3.6	5.2	6.1	A
I_g	1.1	0.91	1.28	1.1	A
Temperature	22	124	22	125	°C
f_o	31.0	30.7	31.6	31.0	MHz
Maximum deviation of f_o during process	−300		−600		kHz

Although the off-load currents noted in Table 8.3 may seem large for a 'minimum loss' design the following circumstances must be taken into account. The circuit losses of the half-wave line generator can, in a first approximation, be assumed to be double that of a corresponding quarter-wave line generator in the off-load condition. This is because the coupling of the load circuit to the quarter-wave line generator is very much reduced by the action of removing the load, eliminating most of the losses in the secondary circuit. Under full load conditions, however, and assuming a correct match of the separate load circuit into a quarter-wave generator, both systems will dissipate the same amount of circuit losses. The higher off-load current apparent with a half-wave circuit does not therefore reduce the transfer efficiency under full-load conditions.

8.2.6 Summary

The advantages of the half-wave line over more conventional circuits can be summarised as follow :

1. The construction of a half-wave generator-load circuit is extremely simple as it differs very little in its physical layout from the theoretical circuit diagram.
2. High r.f. potentials may be obtained by line transformation without the introduction of additional reactances that normally cause parasitic oscillations.
3. The load will appear resistive because unity coupling is maintained throughout the processing cycle, and a favourable 'load power/processing time' relationship is maintained.

The ceramic coaxial construction of the new industrial triodes, YD1172 and YD1182, facilitates their electrical integration into this form of half-wave line generator.

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