SPECIAL HIGH-FREQUENCY ENGINEERING NUMBER

THE BROWN BOVERI REVIEW

10 kW Air-cooled Transmitter Triode, also available for Short Waves

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INTRODUCTION.

THE first special number of this journal to be dedicated entirely to the high-frequency field was issued in December, 1941, thus concluding the volume commemorating the fiftieth anniversary of the foundation of the firm in 1891. It is therefore a matter for some satisfaction that it should again be possible, just under three years afterwards, to present a further enlarged number devoted exclusively to the same topic.

Whereas at the time of the appearance of the previous high-frequency number the department itself was still in the development stage, a remarkable all-round consolidation has taken place in the interim, and there is an imposing number of orders from at home and abroad to testify to the success of the Company's efforts in the field to date.

The reasons for the new line of manufactures having been taken up were clearly set out in the first special number. In point of fact, it is highly desirable that Swiss industries should base their activities as exclusively as possible on their own designs since in no other way can an absolutely free hand be ensured in the export trade. Every manufacturing licence acquired from a foreign patent-holder necessarily entails an immediate and far-reaching restriction of potential exports. Moreover, there can no longer be any doubt that the chief economic mission of Swiss industry consists not only in the setting up of the necessary financial means to permit of the importation of essential commodities unobtainable in the country itself, but to provide adequate employment for the far from sparse population. This does not imply, however, that the interests of the intellectual worker should be neglected in favour of those of the manual worker. On the contrary, the work of technical development must be raised to the same status as that of production proper. From the present special number it will be seen that the Company has already met with a certain amount of success on the road to creating designs of its own in the new field on which it has embarked.

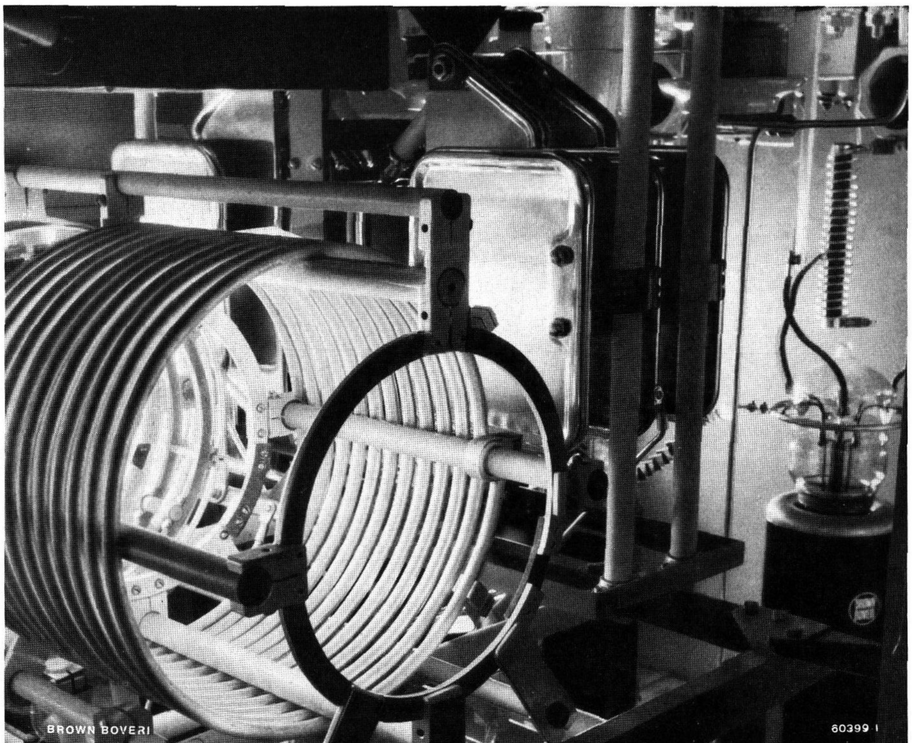
In the course of the last three years both the high-frequency research and manufacturing facilities have been greatly extended. The disturbed market conditions in particular rendered it more and more imperative to take up the manufacture of certain vital components. For instance, workshops have been equipped for the production of transmitter and receiver valves, although for the time being only high-power and special types are being turned out, chiefly for the ultra-short-wave and micro-wave fields. For the same reason the cultivation of synthetic crystals for filters and the manufacture of high-frequency powdered iron cores and the like have been taken up. This new range of manufactures is considered not only to meet a long-felt want in Switzerland, but is expected to form a not unimportant export line upon the return of more or less normal conditions.

Special recognition is due to Prof. P. Scherrer of the Swiss Federal Institute of Technology and to Dr. H. Staeger of the Department for Industrial Research at the same Institute for their contributions to the present special number. They bear striking testimony to the cordial relations existing between Swiss scientific research and industrial circles.

Space could unfortunately not be found in the present issue for articles on carrier-current telephony and remote supervisory control over transmission lines, which, however, will appear in one of the next numbers of this journal.

Perusal of the following pages will reveal that the Company's activities in this field already cover a wide range, while readers delving more deeply into certain of the articles presented cannot but notice the enthusiasm of the engineers and designers for their work and the spirit of personal sacrifice with which they tackle the problems they have been set. This circumstance, coupled with the fact that the high-frequency research, design, manufacturing, and sales departments are now fused into a homogeneous whole and can be counted upon to back up those of longer standing in the solution of the post-war problems, augurs well for the future.

(MS 554) *Th. Boveri / E. Klingelfuss. (E.G.W.)*



High-frequency output coil for a 10 kW broadcasting transmitter.

High-frequency Communications Engineering:

SPECIAL TRANSMITTERS FOR WIRELESS BROADCASTING, TELEPHONY, AND TELEGRAPHY.

Decimal Index 621.396.61

The recently developed Brown Boveri 10 kW transmitters are designed for short and medium wave-lengths. Being built up from self-contained panels they are readily transportable and can be simply and quickly erected on site. Features of the transmitters are ease of control, rapid change of wave-length, clear layout, ready accessibility to all components, and small space requirements. They are switched in, out, and over to the different methods of operation automatically. Each wave in the wave-band can be directly adjusted. Moreover, the transmitters can be supplied with full-automatic wave-change if required.

Fundamental Design of Brown Boveri Medium and Short-wave Transmitters.

APART from small transmitters Brown Boveri constructs medium and high-power units for short and medium waves. In the following notes special features of the design of these transmitters are touched upon. Several outfits with an output of 10 kW have already been supplied. They are suitable both as independent plants and as preliminary stage for large transmitters, but are mostly employed as independent transmitters, e. g. for regional broadcasting stations or in the case of short waves for trans-continental or trans-oceanic telephony and telegraphy.

These 10 kW transmitters, which have been developed during the last few years, are so designed that they can be completed down to the very last detail in the works and tested under service conditions. A noteworthy feature is the sub-division of the equipment into enclosed panels, which keeps the time required for erection as well as the necessary technical staff and the corresponding costs down to a minimum. Semi-automatic operation can be provided which considerably facilitates operation and supervision. The daily frequent change of wave-length involved in the case of short-wave transmitters for telephony and telegraphy can be simply and reliably effected in a very short time due to the special design features incorporated. The resulting curtailment of service interruptions is an important factor from the point of view of economy. When necessary, automatic wave change can be provided for. The transmitters have a high efficiency due to the anode modulation in the output stage.

The *10 kW medium-wave transmitters* type SO 26/10k are constructed for a wave-range of 160—600 m. In the case of the *10 kW short-wave transmitters* type SO 25/10k the normal wave-range is 13—90 m, but can be extended down to about 10 m in exceptional cases. The wave-ranges are smoothly adjustable in both cases, i. e. any desired wave-length can be obtained with the control knobs. The short-wave transmitter is suitable for telephony and unmodulated and modulated continuous-wave telegraphy. With the latter method of operation it has a maximum output of 20 kW.

In external appearance there is no essential difference between the medium and short-wave transmitters of the above-mentioned types, as, moreover, will be clear from Figs. 1 and 2. The necessary power equipment with the high-voltage mutator needs three panels and the transmitter proper with the high-frequency preliminary and output stages and the low-frequency (i. e. modulation) preliminary and output stages four panels, to which must be added the panel with the 8 kW modulation transformer. The latter, however, can be located anywhere in the transmitter room. Due to this arrangement such 10 kW transmitters set no installation problems even in existing rooms. The normal layout is with the power and modulation transformer panels in one row and the transmitter panels proper in a second row in front of them, but separated by a gangway. A noteworthy feature is the small amount of space required by the whole transmitter. Control and supervision of the entire equipment can be conveniently effected from a common control desk (see Fig. 1).

Electrical Design.

Class C anode modulation has been adopted in the 10 kW output stage, this ensuring the highest efficiency. The preliminary stages can be operated alone as 1 kW transmitter by simply changing over their connections. The mode of operation and relative power conditions remain the same at this power stage, i. e. with unmodulated continuous wave telegraphy the maximum output is of the order of 2 kW. On the high-frequency side the driver unit for 10 kW operation of the transmitter acts as anode modulated 1 kW output stage, while on the low-frequency side there is a separate amplification channel for operation with this power.

The high frequency is generated by the oscillator on the h. f. preliminary stage panel in the left foreground and progressively amplified from left to right until the h. f. output stage is reached in the centre of the row of panels. Conversely, the low frequency is led from the control desk into the l. f. preliminary stage panel in the right foreground and from there is amplified to practically the same power level as the high frequency from right to left. Both powers

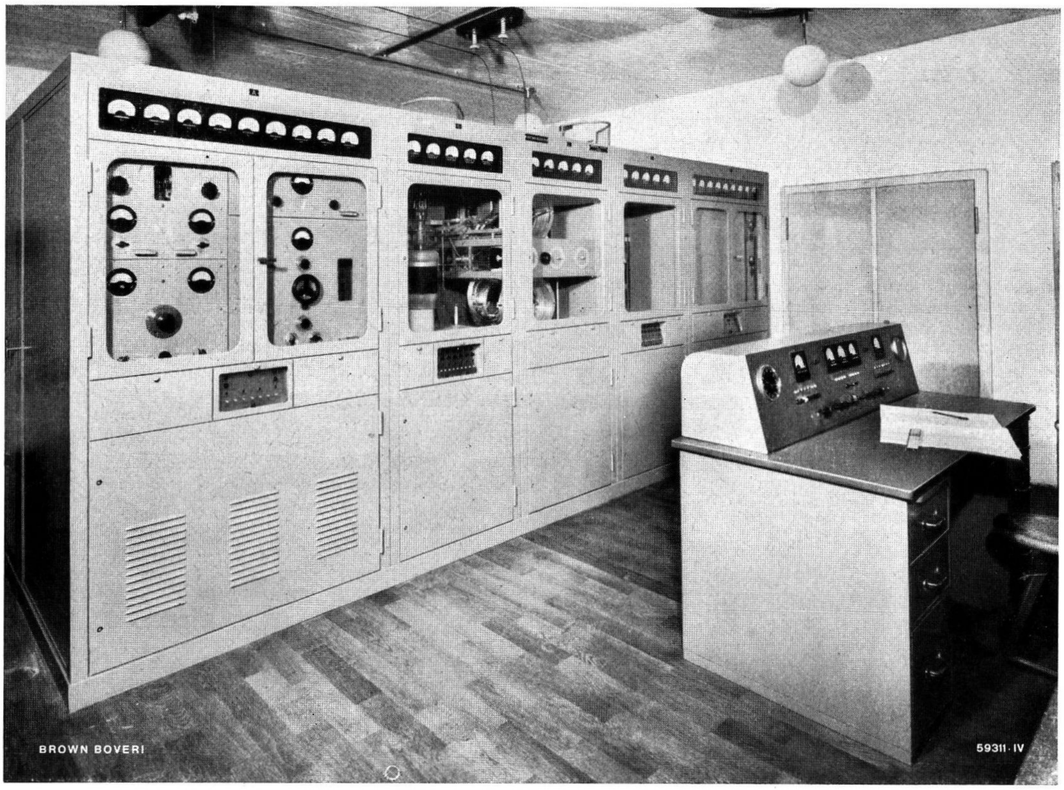


Fig. 1. — 10 kW medium-wave transmitter type SO 26/10 k, with control desk, ready assembled.

Space required excluding desk: Length 4800 mm, depth including gangway 3200 mm, height of panels 2120 mm.

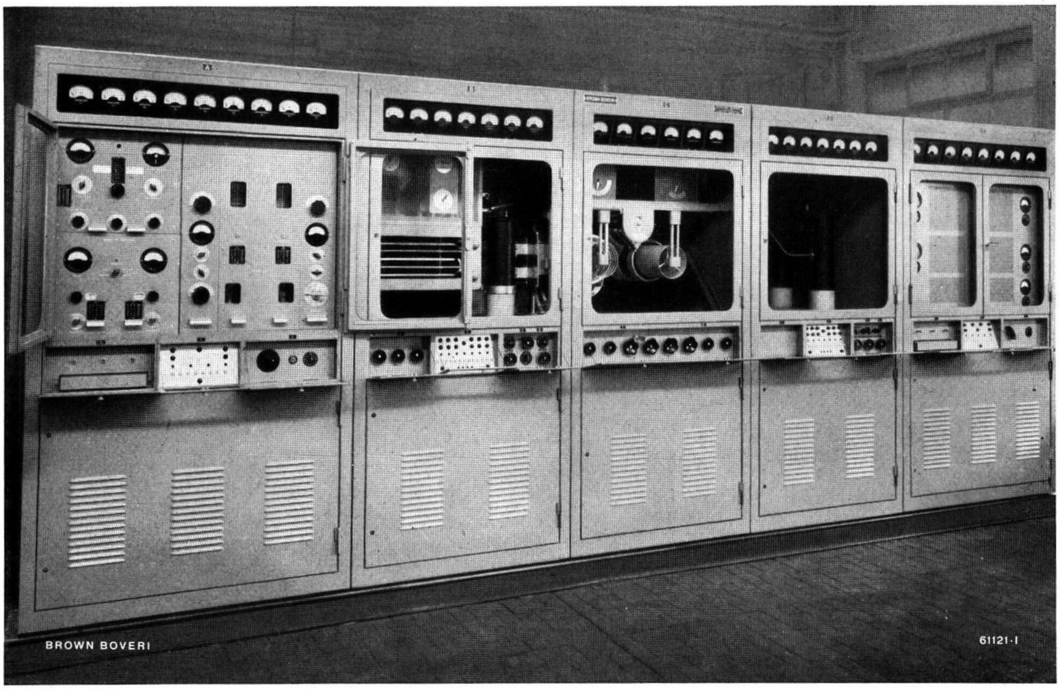


Fig. 2. — 10 kW short-wave transmitter type SO 25/10 k after acceptance in factory.

Space required excluding desk: Length 5400 mm, depth including gangway 3400 mm, height of panels 2120 mm.

The above engravings give a good idea of our semi-automatic 10 kW transmitters which can be readily transported and rapidly erected. The individual panels contain from left to right: The h. f. preliminary stages, the h. f. output stage (valve and coil panel), the valves of the modulation output stage, and the modulation preliminary stages. The short-wave transmitter has a similar control desk to the medium-wave transmitter. In the lower engraving the doors are open to show the control knobs of the transmitter.

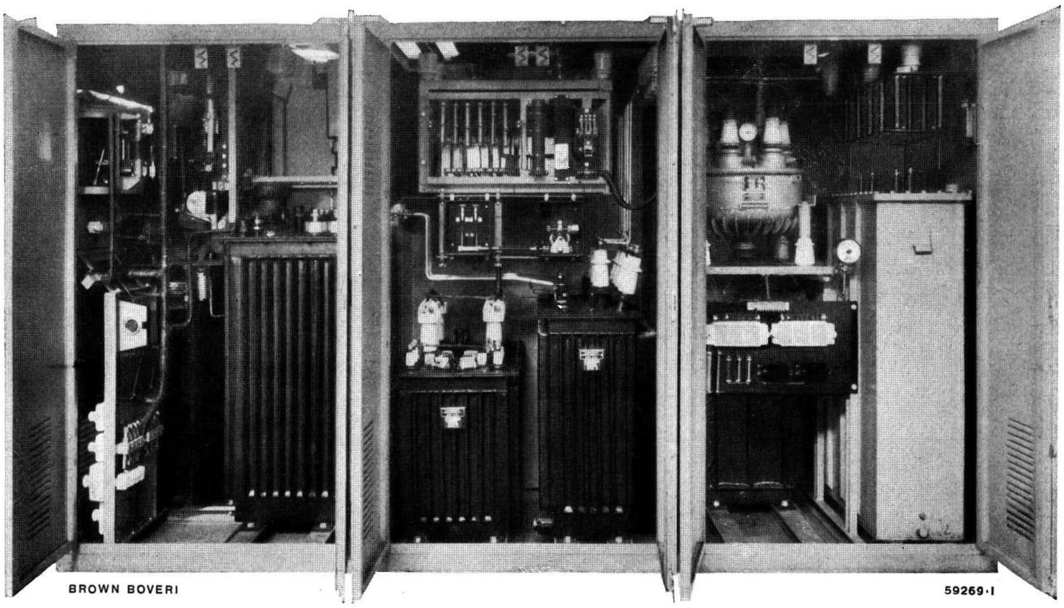


Fig. 3. — Transmitter power plant with associated high-voltage rectifier.

On left: Distribution and control equipment with automatic transformer-type voltage regulators.
In centre: Transformer and mutator grid control gear.
On right: Pumpless high-voltage mutator and filter.

meet in the centre of the row of panels to form the modulated h. f. power. This layout results in a pleasing and symmetrical appearance of the row of panels (Figs. 1 and 2).

The Amplification Stages.

The amplification stages proper are located in the upper half of the front-row panels, the lower half containing the associated auxiliaries such as contacters, fuses, rheostats, transformers, rectifiers, fans, etc. A complete row of control boards bearing the operating knobs and protected by small doors separate the upper and lower panel doors. The risk of inadvertent alteration of settings by the station staff or visitors is thus reduced to a minimum without in the least inconveniencing the extremely seldom re-adjustments necessary in normal service. With the exception of the control knobs on the upper amplification panel of the h. f. preliminary stage all of the control knobs for normal service operation are concentrated on this row of control boards. The wave-length, for instance, can be changed to any desired value exclusively by means of these knobs. *Coil changing by hand*, as is frequently necessary, *is eliminated here*, the wave-length being changed *partly with change-over switches and partly with coil*

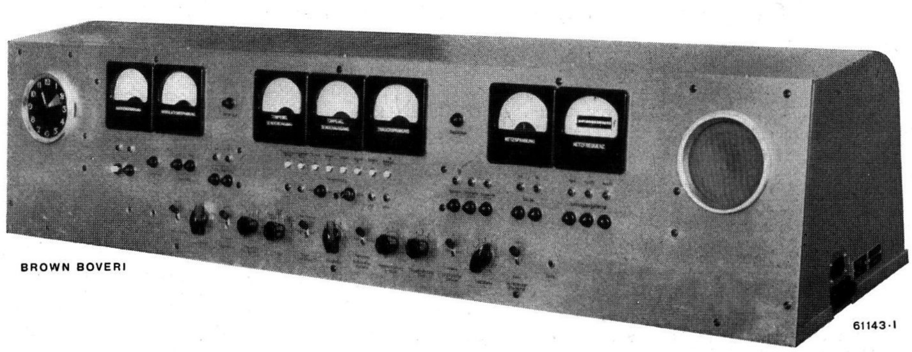


Fig. 4. — Control desk for short-wave transmitter.

Lower row: Operating knobs for programme effects and supervision, comprising knobs for the amplifier of the table microphone; audio frequency generator with plugging key for five fixed frequencies; main modulation potentiometer; the modulation checker consisting of two level meters having a linear db scale and indicating the audio frequency level of the transmitter input and, after demodulation, of the transmitter output with short operating and long release times; the modulation peak indicator with lamp adjustable to different degrees of modulation; the carrier voltage indicator; and the monitoring loudspeaker amplifier for changing over from the input to the output end of the transmitter.

Centre row: Transmitter controls with position indicating lamp; in centre fault indicating lamps.

Upper row: Checking instruments.

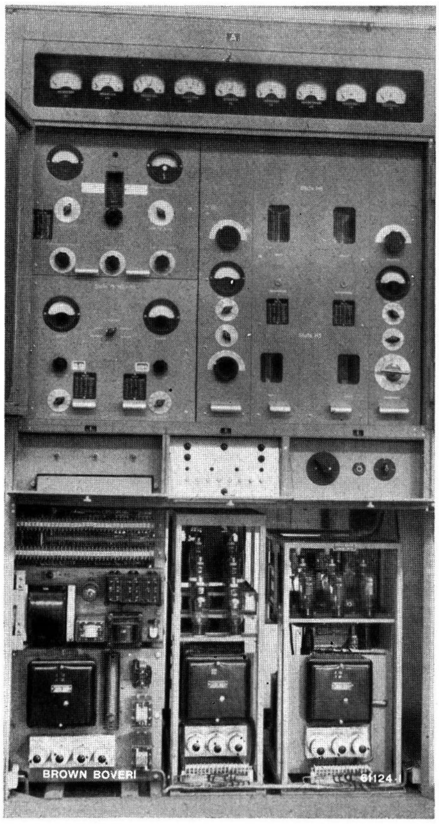


Fig. 5. — High-frequency preliminary stage panel of short-wave transmitter with door open.

Upper part of panel: Oscillator drawer capable of change over to variable wave and ten crystal-controlled waves; drawer of multiplier and 10 W stage; drawer of 100 W and 1 kW stages. (Output stage for 1 kW operation, driver stage for 10 kW operation.)
Centre part of panel: Control station; change-over switch for 1 or 10 kW operation; thermostat switch. The doors of the three adjacent panel sections are open.
Lower part of panel: Fuses, contactors, and pivoted 500 V rectifier; 2500 V rectifier; 1500/500 V rectifier.

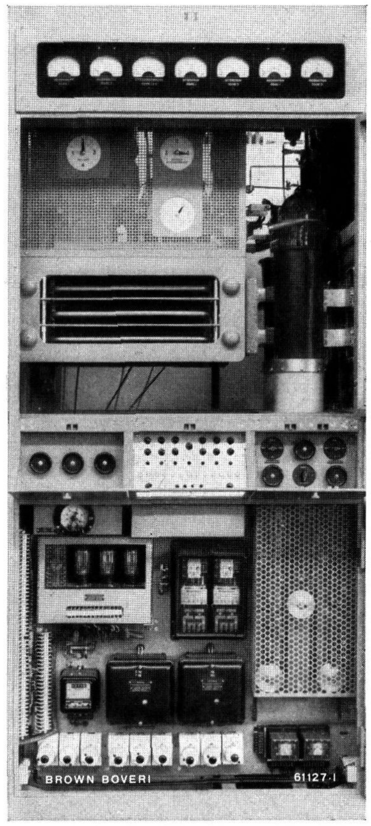


Fig. 6. — High-frequency output stage valve panel of short-wave transmitter.

At top: Grid oscillator circuit with indicator for position of coil turret, condenser switching step and variable condenser; anode auxiliary condenser controlled in two steps (together with grid condenser switching steps); Brown Boveri air-cooled transmitter valves.
In centre: Three control knobs for grid circuit, control station, filament-current rheostat with voltmeter change-over switch, grid bias regulating knob with voltmeter change-over switch.
At bottom: Grid bias rectifier, withdrawable; anode over-current relay for both valves; filament and grid-bias rheostats; operating time meter; contactors for filament heating and fan; fuses and auxiliary contactors. Behind this panel there are also the air filter and valve fan.

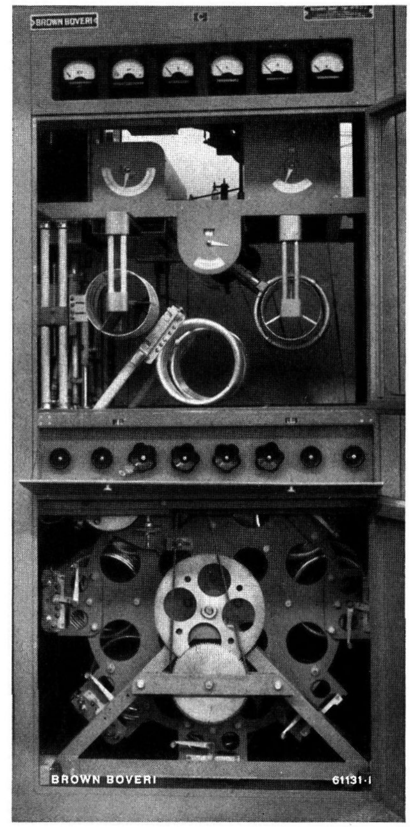


Fig. 7. — High-frequency output stage coil panel of short-wave transmitter.

At top: Neutralizing and tuning gear, tank circuit coil half lowered; coupling coils in position of rest; feeder variable condenser with switching steps.
In centre from left to right: Knobs for neutralizing, tuning, coil displacement, coil turret rotation, coupling coil selection, feeder condenser switching step, degree of coupling, feeder variable condenser.
At bottom: Coil turret for eight coils.

turrets. Each panel or set of associated panels is allotted a control station with push-buttons and pilot lamps on the row of control boards. These enable the different stages on the associated panel to be separately controlled, taken into service, and supervised. In the usual way and as long as no changes in wave-length have to be effected, however, the whole equipment is taken into service and supervised exclusively from the control desk. All of the *service instruments,* which are arranged slightly inclined along the top of the panels, can be conveniently observed from this point. They are illuminated by linolite lamps as soon as the associated panel is taken into service. The remainder of the instruments incorporated in the amplification panels are provided to check currents and voltages the magnitude of which is not of importance from an operating point of view, but is required to be known for restricting any sources of trouble which may arise.

Layout and Connection of Panels.

Ease of access to all parts of the transmitter was a point to which particular care was given in its construction. Due to the gangway between the two rows the front panels are accessible from both the front and the rear. In the case of the back row

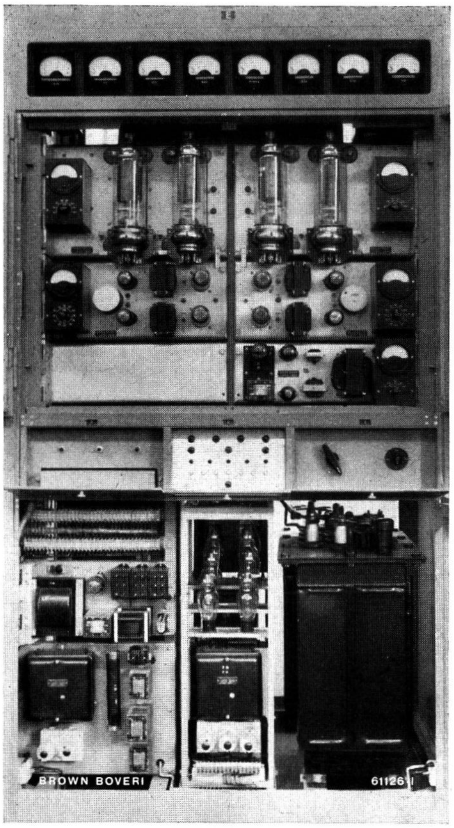


Fig. 8. — Low-frequency preliminary stage panel of short-wave transmitter.

At top: Three-stage, pivoted amplifier cascade for 1 kW operation; three-stage pivoted amplifier cascade (up to and including driver stage) for 10 kW operation; pivoted keying unit for unmodulated and modulated continuous wave telegraphy.
In centre: Control station; change-over switch for 1 or 10 kW service and switch for control rectifier.
At bottom: Fuses; contactors and pivoted 500 V rectifier; 2500 V rectifier.

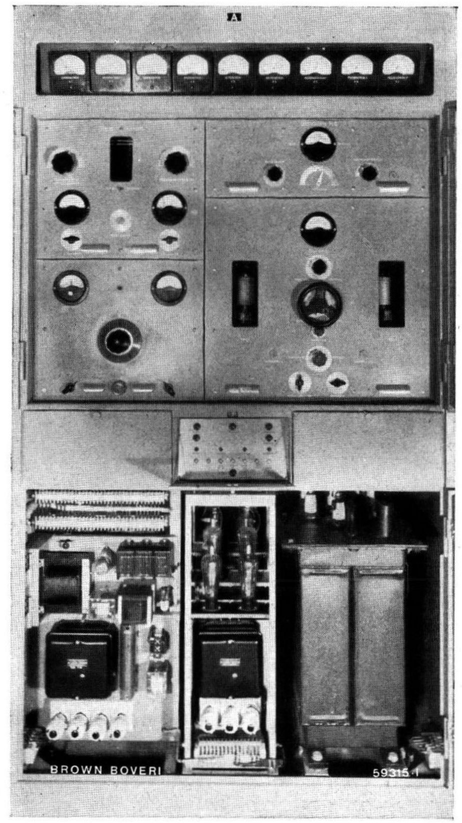


Fig. 9. — High-frequency preliminary stage of medium wave transmitter.

At top: Oscillator drawer capable of change over to variable wave and two crystal-controlled waves; drawer of 10 and 100 W stages; drawer of 1 kW stage (output stage for 1 kW service, driver unit for 10 kW service); grid circuit drawer for 10 kW stage.
At bottom: Fuses; contactors and pivoted 500 V rectifier; 2500 V rectifier with associated transformer.

access from one side only is sufficient since the equipment is not so delicate and less concentrated. Wire grids between the two rows of panels give the transmitters the appearance of a self-contained block. All mechanical and electrical connections, with the exception of those to the external power supply and control desk and the three high-voltage conductors above the panels lie within this block.

A noteworthy feature is the small number of connections between the individual panels themselves. Apart from the few power, h. f., and l. f. leads (which are in the form of short lengths of wire from panel to panel), however, there is a large number of connections between the different stages themselves which serve to ensure correct sequence of switching, order control, position indication, etc. By incorporating a separate control system, however, it was found possible to meet all of the necessary switching and control conditions in a relatively simple manner. The necessary relays, push-buttons, and lamps involved require very little space. According to experience this gives maximum reliability.

The arrangement adopted enables the control problem to be solved in a simple manner, the panels being connected by multi-pole leads with plugs. A central control set in the low-frequency preliminary stage panel may be said to form the brain of the whole transmitter, from where the individual panels are mutually connected by nineteen- and forty-pole leads with plugs. A special cable also serves to connect with the control desk. Transition from the control system to the power circuits is effected in the so-termed control set in the individual panels. This is a unit comprising the above-mentioned control station with an associated set of relays which by appropriate wiring provides the necessary control conditions and from which the control leads to the individual contactors and other points of control radiate.

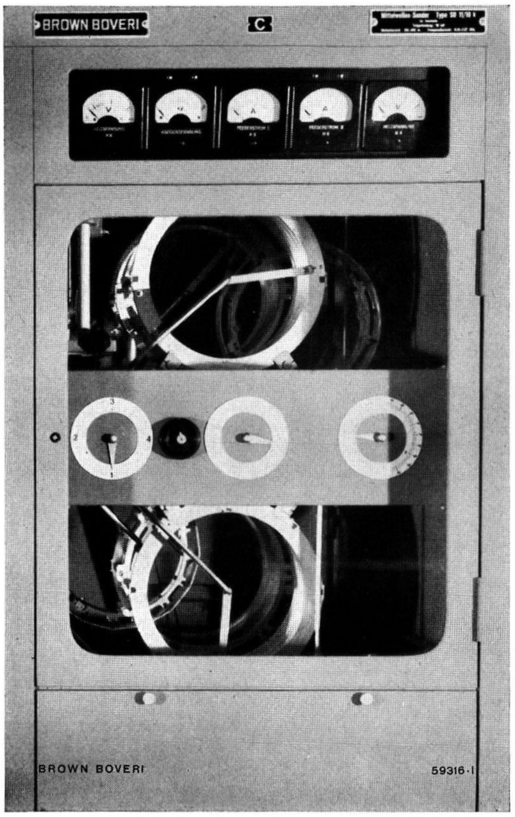


Fig. 10. — Coil panel of high-frequency output stage of medium-wave transmitter.

Tank circuit coils capable of being changed over, with pivoted coupling coils. The three indicators are for the wave range, degree of coupling, and tuning of the compressed gas condensers. The associated control knobs are immediately underneath on the control panel and the compressed gas condensers in the lower part of the panel.

Readily-manipulated Control System.

The method of control employed is interesting. In the case of the short-wave transmitter it is characterized by the following features:—

1. *Transmitter automatically switched in when "Transmitter in" push-button depressed.* A programme switch cuts in the various filament and anode voltage steps in the right sequence and at the correct intervals, i. e. the appropriate steps are always automatically switched in according as the transmitter is connected for 1 or 10 kW and telephony or unmodulated or modulated continuous-wave telegraphy.
2. *Transmitter automatically switched out by depression of "Transmitter out" push-button,* the transmitter valve fan automatically running for three minutes longer.
3. *All anode and grid bias voltages capable of being cut in and out with filament current switched on.* The pair of push-buttons provided here serves to interrupt the carrier-wave during long service intervals, but enables the transmitter to be taken into service again immediately at any time. The individual grid and anode voltages are naturally interlocked in such a manner that they can only be switched on in the correct sequence.
4. *Selection of 10 kW and 3 kW, or hand-regulation power steps.* The power step can be selected either before taking the transmitter into service (whereby the transmitter will automatically run up to the power step in question when started up) or it can be changed during operation simply by depressing a push-button. (These buttons are cut out in 1 kW service when the output stage is not employed.)
5. *Selection of telephony or unmodulated or modulated telegraphy.* The desired mode of operation is usually selected before taking the transmitter into service, by depression of the appropriate button. The change-over, however, can also be effected at any time during service. In this case the anode voltages are at first automatically cut out in succession, then various change-over operations effected, stages no longer necessary switched out, new stages required switched in step by step, and finally the anode voltages re-applied. The individual operations are naturally interlocked here, too, and faulty switching thus precluded. A further button enables the carrier-wave to be instantaneously cut in and out at any time in telephony service by h. f. blocking or it is automatically switched in by the modulating speech in the "speech controlled" position and switched out with a certain time-lag.
6. *Away from the automatic gear,* on the control desk, the anode high voltage of the output stages can be switched in and out and progressively regulated up and down from zero by further buttons. (In the power position for hand regulation.)

Pilot lamps assigned to the push-buttons both on the control desk and on the different control stations indicate the circuit conditions prevailing at any one time. A number of fault indicating lamps assigned to the various control stations and which give centralized indication at the control desk, immediately draw attention to all possible faults, including the leaving open of doors.

By operating the push-buttons on the different control stations the transmitter can be taken into service step by step and operated without calling upon the automatic gear. The system of interlocks naturally prevents all faulty operation in this case. However, this method of operation necessitates walking along the whole flight of panels.

Other Important Details.

The detailed explanations to the different engravings give an idea of the ingenious construction of the transmitters. In addition, attention is drawn to the following important details. The individual *groups of amplifiers of the h. f. preliminary stage are of the drawer-type.* This arrangement gives minimum space requirements and ready access to all components. The oscillator drawer contains the *oscillator* (which allows the control frequency to be maintained constant with a high degree of accuracy) as well as the following buffer stage. A change-over switch permits of adjustment to the continuous, variable wave with which both transmitters can cover the entire wave-range or to the various crystal-controlled waves. Whereas with the short-wave oscillator provision is made for ten crystals the medium-wave oscillator only has two. The frequency stability of the crystal-controlled waves is better than 2×10^{-6} and that of the variable wave better than 10^{-5}. Disturbing effects are eliminated by means of an accurate thermostat, together with stabilizing circuits and design features which isolate the oscillating circuit in question from all possible mechanical impulse and distortion forces. The spiral scale of the oscillator, which is calibrated directly in kilocycles, permits any wave to be adjusted with an accuracy of $0 \cdot 05\,\%$.

The control spindles in all drawers are taken through the back. This permits all h. f. preliminary stages to be synchronized for *single-knob control,* if desired. The ganging of the drawers is effected by means of couplings behind them, these being mutually connected by means of chains and gear-wheels and automatically coupled in when the drawers are pushed into the panel. Of the outfits illustrated the medium-wave transmitter is of this type, whereas the short-wave transmitter is arranged for tuning by the individual knobs. A proviso for the synchronized drive is naturally that the ranges are sub-divided in the same manner in all stages. In consequence, the wave-range of medium-wave transmitters is divided into four and that of short-wave transmitters into six sub-ranges.

With the medium-wave transmitter the four sub-ranges are continued into the h. f. output stage, whereas in the case of the short-wave transmitter the six sub-ranges are each split up into two further sub-ranges, making twelve in all. The difficult high-frequency power problems generally involved with such transmitters are normally overcome by arranging the plant for the emission of quite definite waves. In the present case, however, the problem is met in a quite general way by providing for the adjustment of any desired wave within the specified wave-range at any time exclusively by means of control knobs. This design was adopted chiefly due to the extension of the wave-range downwards to 10 m. With the correspondingly high frequencies all unnecessary inactive capacitance and inductance must be avoided in the high-power circuits, thus involving splitting up of the wave range to a high degree.

The short-wave transmitter can also be supplied in a special design for *automatic wave-change.* The driving and stopping device entailed can be mounted on the rear of the h. f. panels on the through-going control spindles. In the case of short-wave transmitters the speed with which the wave-length can be changed is of vital importance. Where continuous transoceanic traffic is concerned the wave-length has to be varied three to five times daily. Hitherto the wave-change of transmitters of the size in question took 5—15 minutes. The new design, with which the changing of swinging coils by hand is eliminated, reduces this time to about three minutes. The hand-operated change-over, however, still entails the following operations:—

1. Disconnection of the anode voltages.
2. Operation of various knobs to effect the wave-change.
3. Re-applying the anode voltage.

With automatic control the time for the completion of this cycle is reduced to 5—30 seconds. In this case *the operation of a single push-button on the control desk* suffices to enable a programme to be completely changed over from one wave-length to another. The automatic gear is designed for ten different waves, so that waves frequently required can be pre-adjusted on the stopping device. The ten waves can be distributed as desired over all of the ranges. It is therefore possible to distribute them either uniformly over the entire range or, if, necessary, to concentrate all of them in one of the twelve sub-ranges of the transmitter.

On the low-frequency preliminary stage panel the different amplifier chassis are of the *pivoted type.* They can be swung out during operation of the transmitter, so that the wiring at the rear can be checked under service conditions.

The *keying system,* which is only provided in the short-wave transmitter, differs according as the transmitter is to be keyed in telegraphy service with direct current or audio-frequency impulses or both as required. It takes care of the necessary transformation of the impulses arriving from the line for the keying of the high frequency in the preliminary stage, as well as for modulation with modulated continuous-wave telegraphy.

Such transmitters have proved very satisfactory in service both for broadcasting and for telephony and telegraphy purposes. The *auxiliary studios* necessary for broadcasting transmitters can now also be supplied.

(MS 546) *Dr. M. Dick. (E.G.W.)*

VARIOUS POSSIBLE APPLICATIONS FOR BEAM TRANSMISSION.

Decimal Index 621.396.43

Very short radio waves can be readily bunched by means of a suitable antenna, so that at optical range large distances can be covered with a small power. Since with the carrier waves used it is possible to operate with a large band width, frequency modulation and multi-channel transmission can be employed. Expensive telephone and signal cables can be replaced by this reliable and certain transmission medium. Brown Boveri have developed beam transmission apparatus with supplementary devices for various applications, this apparatus having already been successfully tried out in practice.

Fig. 1. — **Directional antenna for very short radio waves.** This simple antenna results in a tenfold increase in power in the desired direction when compared with a non-directional antenna.

CONTRARY to radio broadcasting where there is a wireless transmission of signals from a single transmitter to a large number of receivers located at different places, directional beam transmission is concerned with wireless transmission between two fixed points which are previously determined. With directional beam transmission, when compared with broadcasting, it is possible to reduce the transmitting power considerably whilst the reception field strength remains unchanged; furthermore, reflections are avoided and under favourable conditions even several connections can be established independently of each other on the same wave length without mutual interference. In principle, it is possible to build directional beam transmission antennas for any wave length; in practice, however, the space occupied by, and the cost of, the antennas for the same directive effect decreases with the wave length. Since amongst the fields which have been investigated that dealing with decimetre waves has been very thoroughly covered and technically reliable directional beam transmission sets have been built by the Company, the following observations are specially devoted to this wave range. With decimetre waves and simple reflector antennas the output is increased ten to twenty-fold when compared with non-directive radiation, and with complicated antennas even a hundredfold increase is possible (Figs. 1 and 2).

Fields of Application for Directional Beam Transmission Sets.

Decimetre wave transmission is particularly suitable where a reliable connection free from disturbance is required. Decimetre sets are suitable as a complete

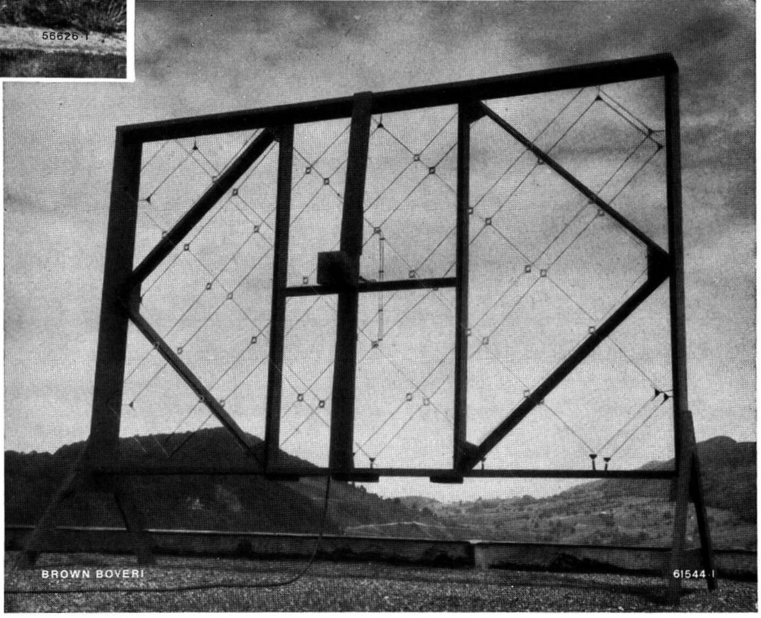


Fig. 2. — **Saw-tooth directional antenna.** Large saw-tooth directional antenna for stationary plants and decimetre waves with a hundredfold increase in power in the desired direction when compared with a non-directional antenna.

substitute for telephone cables in stationary plants, as well as a temporary substitute for telephone conductors during a short service interruption or overloading, and when new connections are required. Beam transmission is thus a valuable addition to the usual telephone service with cables. The cheapness of this cable substitute is noteworthy, because particularly in difficult hilly country wireless transmission costs only a fraction of the amount required for the laying of cables. A further advantage is that there are no interruptions due to damaged cables and a reliable connecting link can always be obtained very rapidly.

Decimetre wave transmissions are fundamentally free from atmospheric or industrial disturbances. Excellent reception is possible during a heavy thunderstorm or in factory areas with a very high noise level, without special measures having to be adopted. The roaring noises which occur with long-distance wireless communication are avoided by using frequency modulation. These properties enable the transmitting power to be reduced to a few watts or less. The experience already obtained with other frequency-modulated plants, for instance the first police radio plant with frequency modulation in Europe, has been of a great help in this respect.

Special Properties of Decimetre Waves.

Decimetre waves extend practically only as far as the optical range. In a flat district the stations are preferably located in high buildings, towers and the like, whilst in hilly country suitable elevated positions can easily be found. Since the stations are equipped for telephony, the station site can be selected as desired. The condition as regards the optical range does not prevent the use of decimetre waves.

The maximum elevation of the ground which is admissible for any particular communication link can readily be determined by means of the following graphical representation. If the field energy at the point of reception has to amount to 78 % of the maximum possible value, then an ellipsoid of revolution with the two semi-axes a = half the distance between transmitter and receiver and $b = \sqrt{\frac{\lambda}{8}} \cdot l$ (where λ = wave-length, $l = 2a$ = distance between transmitter and receiver) must be adhered to; if $b = 0$ then only 25 % of the power radiation in free space is obtained, whilst when $b = \sqrt{\frac{\lambda}{2}} \cdot l$ this increases to nearly 100 %. These calculations are based on the optical theory of wave deflections at a sharp edge. If the optical range is determined from a topographical map then the curvature of the earth must also be taken into account. Fig. 3 shows the necessary free heights compared with the straight lines connecting the transmitter and the receiver for a wave length

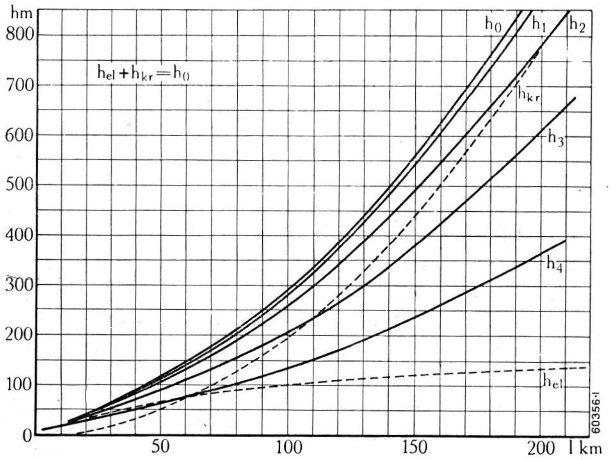


Fig. 3. — Necessary free heights for decimetre transmission.

The necessary free heights h in metres (m) for decimetre wave communication relative to the straight lines joining the transmitter and the receiver for a wave length $\lambda = 0{\cdot}75$ m as a function of the distance l in km, the field energy at the receiving site amounting to 78 % of the maximum possible value. Curve h_0 for the middle of the transmission distance is furthermore divided into the two heights h_{kr} (earth curvature) and h_{el} (optical ellipse). (See example Fig. 4.)

$\lambda = 0{\cdot}75$ m and $b = \sqrt{\frac{\lambda}{8}} \cdot l$ as a function of the distance. The semi-axis is shown in the figure by dotted lines and designated h_{el}, and the height lost due to the curvature of the earth is indicated by h_{kr}. Curves h_{0-4} represent the necessary free heights at intervals of $^1/_{10}$ of the total distance, ellipsoid and curvature of the earth being taken into account. It is interesting

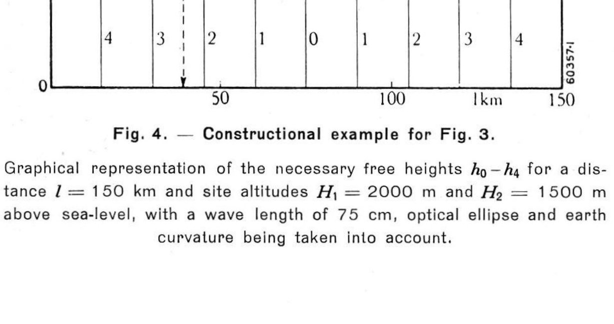


Fig. 4. — Constructional example for Fig. 3.

Graphical representation of the necessary free heights $h_0 - h_4$ for a distance $l = 150$ km and site altitudes $H_1 = 2000$ m and $H_2 = 1500$ m above sea-level, with a wave length of 75 cm, optical ellipse and earth curvature being taken into account.

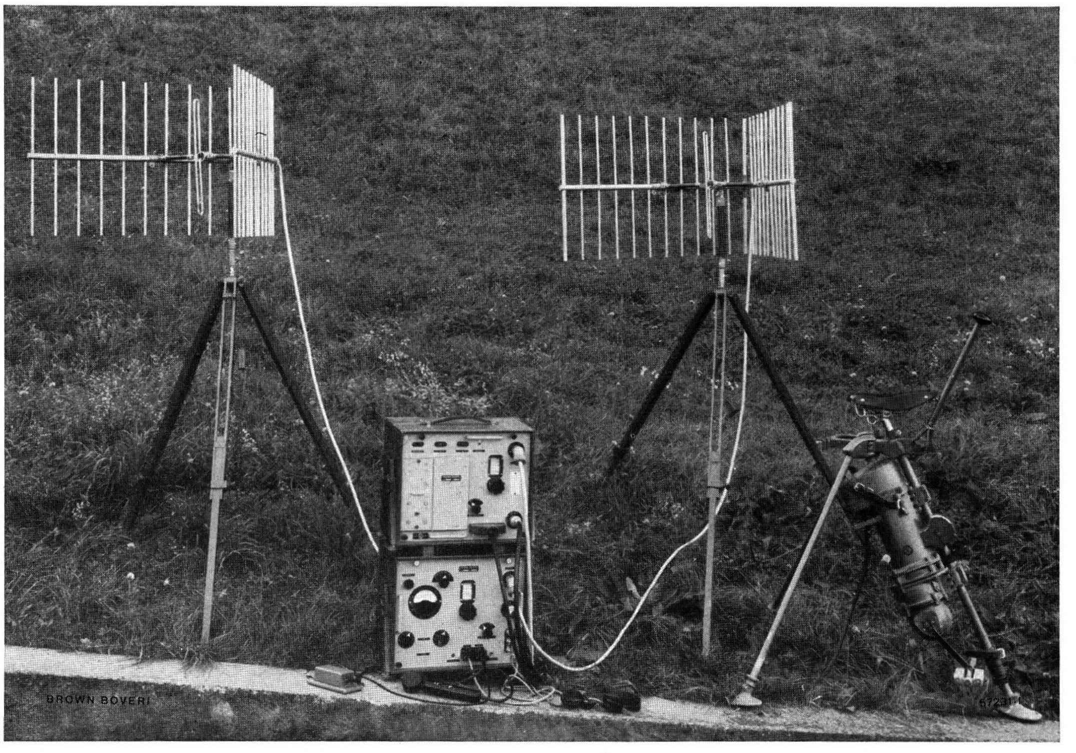


Fig. 5. — Complete beam station with frequency modulation for two-way communication over distances up to 200 km.

On the left is the transmitting antenna, on the right the receiving antenna, and in the centre the apparatus comprising a modulator and transmitter (above) and a receiver with fork connection for changing over to two-wire operation (below), together with headphones, microphone, and key. On the extreme right is the pedal-operated generator for use in the field. The apparatus can also be constructed for chassis-mounting.

to note that for short distances the ellipsoid plays an important part, whilst as the distance increases the influence of the earth curvature becomes very much more marked. An example dealing with a distance of 150 km between the sites $H_1 = 2000$ m and $H_2 = 1500$ m above sea-level, is shown in Fig. 4. From this it will be seen that for instance at the point x_1 a rise in the ground of maximum h_{x_1} above sea-level may occur.

For the distance in question it is often impossible to adhere to the condition of visual range. In such a case the total distance is divided up into individual adjoining sections and relay stations are provided, whereby the signals from a terminal station are received and transmitted on a fresh wave length in the direction of the other terminal station. By this means distances of many hundreds of kilometres can be bridged over. The relay stations do not require any attendance, except for the usual general supervision and control.

For the simultaneous operation of transmitters and receivers which are located on the same site and can be operated with different frequencies and unequal signal voltages without cross-talk, the selectivity must be very high. By this means two-way communication is guaranteed. Furthermore there is also the possibility of simultaneously operating several stations on the same site, for instance two-way communication with relay operation, at points of intersection, or at the same starting and finishing stations of several communication links.

Due to being able to use a very large modulation frequency band, decimetre wave links are not restricted to being used merely as a complete substitute for telephone cables. By the addition of multi-channel apparatus several channels for telephonic and telegraphic communication are simultaneously formed with a high-frequency connection, whereby automatic number selection or a teleprinter can be used. It is also possible to transmit television programmes, whereby the large band width and the absence of undesirable linear cable distortions due to a frequency-dependent transit time are to be noted.

In addition all supervisory control and telemetering signals can be transmitted without any disturbances and the low noise level and small distortion factor enable a high quality substitute to be provided for cables for transmitting music from the studio to the radio transmitter or for the transmission of programmes over long distances.

Brown Boveri Beam Sets.

During the past few years directional beam sets together with the aforementioned accessory apparatus have been developed in the Company's laboratories. Fig. 5 shows an example of a complete set for two-way communication over distances up to 200 km, frequency modulation being employed. On the left is the transmitting antenna, on the right the receiving antenna, and in the centre are the boxes containing the apparatus comprising a modulator and transmitter (above), receiver with fork-connection for changing over to two-wire operation (below), headphones, microphone and key. On the extreme right is a pedal-operated generator which is used for service in the field. Fig. 6 shows a superheterodyne receiver with limiter and discriminator for four and two-wire connection.

Multi-channel accessory sets are built for chassis mounting or as mobile units. Since the channels are not separated at a relay station, only the terminal stations of the wireless link require to be equipped with a multi-channel accessory set.

Accessory sets (Fig. 7) for connecting to an automatic telephone system have also been developed. They

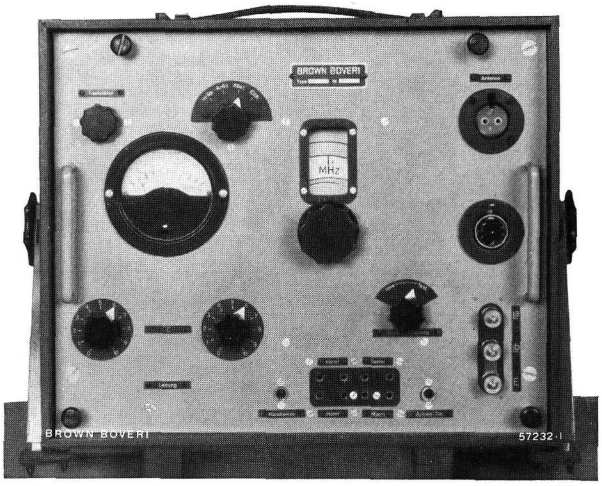


Fig. 6. — Directional receiver for the station shown in Fig. 5, in superheterodyne arrangement with limiter and discriminator, and four and two wire connections.

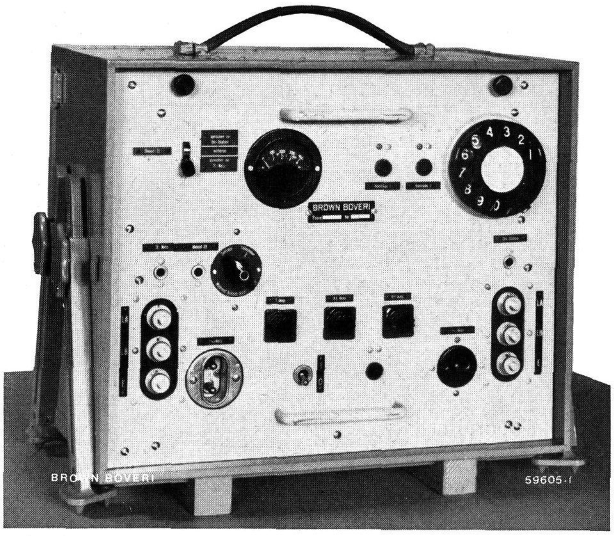


Fig. 7. — Auxiliary dialling and calling unit.

This auxiliary unit for connecting to an automatic telephone system enables dialling impulses and calls to be transmitted over the wireless connecting link with single or multi-channel operation.

enable automatic dialling and calling to be transmitted over the wireless link, it being immaterial if the connection consists of a normal relay link or if multi-channel operation is employed.

When chassis-mounted the beam and accessory sets are suitable for use in stationary plants intended for commercial telephone transmission.

For shorter distances simplified beam sets which are suitable for telephone communications and the transmission of telemetering and supervisory control signals are constructed.

Since only special tubes are used in directional beam sets, such tubes have been developed and manufactured by the Company, the experiences gathered in connection with previous development work having proved a great help in this respect.

A directional beam station with accessory apparatus thus enables several conductors having the same or different conditions to be replaced. Fields of application are telephone, teleprinter and facsimile transmission, the transmission of broadcasting and television programmes, as well as telemetering and supervisory control. It is therefore possible to speak of a comparatively cheap, efficient and safe "multi-core cable substitute".

(MS 543) *Dr. R. Schüpbach. (Op.)*

DEVELOPMENT WORK IN THE DECIMETRE WAVE FIELD.

Dezimal Index: 621.396.24.029.63

The increasing importance of the decimetre wave field for communication purposes and remote supervisory control soon resulted in intensive development work being undertaken by the Company, the object of which is the practical application of wave lengths below 1 m. The design of communication apparatus ready for service was based on a newly created technique for decimetre waves, and also on improvements in measuring methods and devices which in accordance with the peculiarities of these waves are very different from those used in normal high frequency technique.

A. MEASURING TECHNIQUE.

AT the same time as decimetre and metre wave apparatus was developed and designed the very difficult measuring methods in this wave range had also to be studied. The main difficulty is, namely, that standard conductors can no longer be used between measuring points and measuring indicator (see section B). Small leakage capacitances of a few tenths of a micri-micro farad can lead to entirely false results; furthermore the electron transit times, for instance in measuring diodes, cause an appreciable disturbance. Therefore even for a simple voltage measurement the measuring method must be considered very carefully and it may even be necessary to construct a special test layout for this purpose. A number of measuring methods are described briefly below.

For *voltage measurements* at high frequencies a diode is generally used as the rectifier, the anode and cathode being connected with the measuring points by conductors which should be as short as possible. In the decimetre wave range, however, these short conductors are — in conjunction with the tube — sufficient to cause considerable damping by radiation. The tube capacitance causes the tuning circuit where the voltage is measured to be strongly detuned, whilst the electron transit time makes itself noticeable by an inadmissible damping effect. It is, therefore, necessary to employ a calibrated potentiometer, for instance a capacitive voltage divider, with highly ohmic inlet in front of the measuring tube. The rectified alternating voltage is then passed over a direct current amplifier to the measuring instrument. Even with this measuring arrangement the calibration becomes dependent on the frequency, due to the additional conductor inductance. A division of the voltage which is independent of the frequency can be obtained by means of a tuned $\lambda/4$ parallel wire system when the measuring diode is connected in the vicinity of the short-circuited end and the voltage to be measured is connected to the other end. The measured voltage is then converted to the desired voltage, due consideration being paid to the sinusoidal voltage distribution along the Lecher system. The small error caused by the tube capacitance can be avoided by a correspondingly dimensioned parallel inductance.

A *current measurement* is very difficult because the circuit may not be opened at any point (e. g. cavity resonator) on account of the additional damping, and a shunt cannot be used as a result of current distortion. Even if a thermo-couple is employed no exact results can be obtained due to skin effect errors.

Approximate *power measurements* can be comparatively easily made by known methods with thermal or optical effects, for instance by switching in a heating wire and measuring the heat energy or switching in a lamp and measuring the light energy with a photo cell. In both cases, however, the radiation energy of the arrangement, which as regards the wave length is already comparatively large, is neglected. The power delivered by a tube can also be determined by measuring the difference of the anode temperatures when in an oscillating and non-oscillating condition by means of a pyrometer and comparing with the applied direct current powers.

The *impedance measurement* is possible with the aid of a parallel wire system of known characteristic impedance with adjustable voltage measuring device, whereby very accurate measurements can be obtained. When making a measurement the impedance is connected to one end and the measuring transmitter to the other end of the parallel wire conductor. The length of the conductor must amount to at least one wave length, but is not employed in the calculation. With the diode system arranged movably and loosely coupled to the parallel wire conductor, the relationship between the maximum and minimum voltage and the distance of an extreme value from the end of the conductor which is closed by the impedance in question, is measured. The absolute magnitude and phase position of the impedance to be measured can then be obtained from values which have previously been calculated and graphically determined. The distance between two extreme values also gives the wave length of the measuring transmitter. Fig. 1 shows the diode system of a constructed measuring device.

The *selectivity* of a tuning system is determined by recording the resonance curve and measuring the width of the curve at half the maximum value. When the inductance or capacitance can be calculated or measured, then in connection with the operating frequency it is also possible to determine the resonance resistance and thus the *loss resistance* of the system.

Another method of determining the resonance resistance is by measuring the feed-back factor over two amplifier tubes, the tuning system which is to be measured being connected in between. Such an

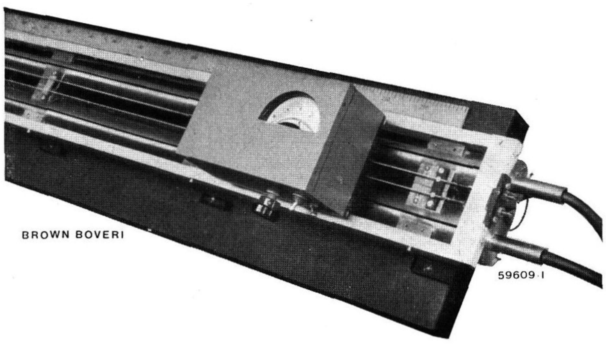


Fig. 1. — Parallel wire system for impedance measurements.

The impedance can be measured with great accuracy with the aid of a parallel wire system of known characteristic impedance and an adjustable voltage measuring device. The picture shows the two concentric systems built together to form a symmetrical parallel wire system with adjustable measuring device and the two heating leads.

apparatus which automatically enables the resonance resistance to be read off on an instrument built into the apparatus is shown in Fig. 2.

For measuring the *frequency* it is usual to use a calibrated measuring circuit to which a rectifier with d. c. amplifier is loosely coupled. For exact measurements a frequency transposition is necessary by superposing and comparing with a more accurate, for instance, crystal-controlled oscillator. Fig. 3 shows a cavity resonator wave meter for decimetre waves and Fig. 4 one for ultra-short waves.

B. CONDUCTOR AND SCREENING.

The shorter the wave length has to be, the greater becomes the comparison between the geometric dimensions of the switch elements and layout and the wave length. In practice it is namely impossible to reduce these dimensions, such as tube sizes and the like, to the same extent.

Voltages and currents at the beginning and end of a high-frequency line are therefore generally not of the same magnitude or in phase. Their *transformation properties*, which are furthermore dependent on the frequency, have to be taken into account.

Conductors such as are usually arranged in a diagram of connections as a "passive" connection between switch elements, no longer exist in decimetre wave technique.

A useful application of these conductor properties at high frequencies is facilitated by the fact that they recur periodically at intervals of half a wave length, so that certain pronounced points exist which produce identical operating conditions. In certain cases the conductors can be made into "passive" connecting links again by a mutual adjustment at the generator and consumer, whereby reflections are avoided.

It is also to be noted that the *losses* in high frequency conductors due to a poor dielectric, for-

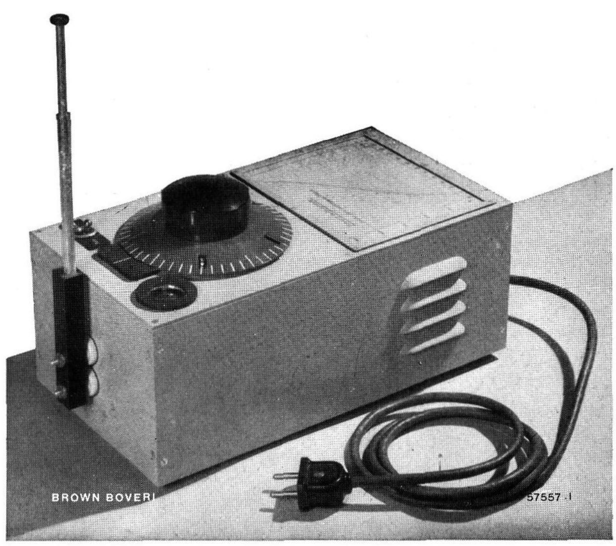


Fig. 3. — Cavity resonator wave meter for decimetre waves with receiving antenna.

A direct current amplifier is employed to increase the sensitivity. The wave length is determined by means of the self-contained electron ray tuning pointer with the aid of a calibration table.

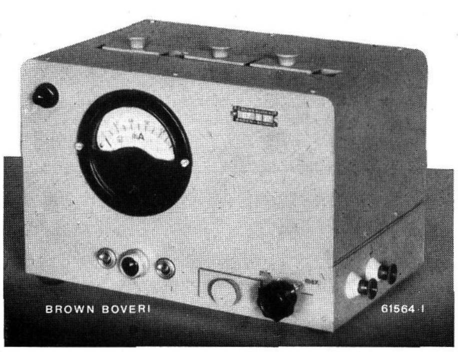


Fig. 2. — Resonance resistance measuring apparatus.

This apparatus measures the resonance resistance of an oscillation circuit connected to it. The self-contained instrument indicates the resonance resistance directly when converted with a table.

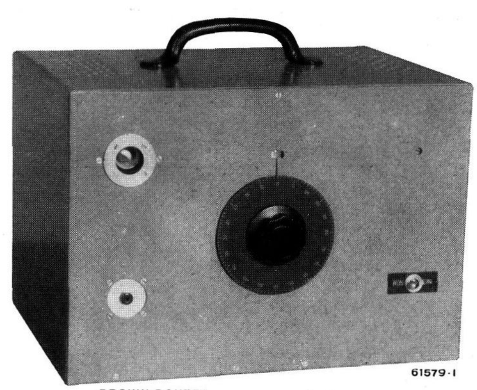


Fig. 4. — Cavity resonator wave meter for ultra short waves.

Construction as in Fig. 3, but without antenna.

mation of eddy currents in the neighbouring metallic parts, skin effect, and also as a result of energy radiation, very easily exceed an allowable value. It is therefore preferable to use screened Lecher lines or concentric conductors with a good conductive surface (a layer several hundredths of a millimetre thick is generally sufficient) and air insulation. The concentric conductor if it is constructed as a rigid tubular conductor is "self-screening", that is the tubular casing, the inside of which represents one pole of the h. f. line also acts as a screen against the parasitic high frequency fields occurring on the external surface, because if the casing is thick enough, they cannot penetrate due to the skin effect. Also certain types of cavity resonators are self-screening; fundamentally they consist of an internal and external conductor, and can be regarded as a special case of concentric conductor of $\lambda/4$ length shortened by a large capacitive end load. Like the unscreened ends of a screened conductor, the openings required for the coupling elements of a cavity resonator also form points which are likely to cause disturbances.

It is now possible by means of a skilful combination of conductors, cavity resonators or cavity resonator filters and amplifier tubes, to construct apparatus where self-screening is maintained throughout, so that there is no risk of "infection" at open points. Since the occurrence of parasitic fields and standing waves on the outside surface cannot be avoided, such a hermetic external skin provides safe protection against undesirable couplings and disturbing voltages.

C. AMPLIFIER PROBLEMS IN DECIMETRE WAVE TECHNIQUE.

In so far as electron tubes of conventional design are used, even with special constructions for ultra short waves, the tube impedances, that is grid and anode impedance, are only very small. In order to achieve a very large amplification a good mutual adjustment of the amplifier stages is necessary, because amplification no longer occurs without power. Therefore it is essential that the losses in the selection means, such as oscillation circuits and filters, should be reduced to a minimum. From this, conditions are obtained for the loss resistance and the ratio L/C of such oscillation circuits. In the decimetre wave field these conditions can only be adhered to in practice by means of cavity resonators or in part also with concentric conductors, but no longer with open oscillation circuits.

The geometric dimensions of cavity resonators can easily and quickly be derived from the necessary electrical conditions with the aid of analogous relationships obtained from tabulated values for certain "standard resonators." Fig. 5 shows a constructional form of cavity resonator for decimetre waves.

As a result of the ready calculability of cavity resonators as filter elements and the adjustment questions which occur in connection with the assembly of electron tubes and selection elements, decimetre wave technique gradually assumes a character which much more resembles that of the medium frequency amplifier and filter technique than broadcasting technique. Superficially it often has a similarity with turbo-machine model making.

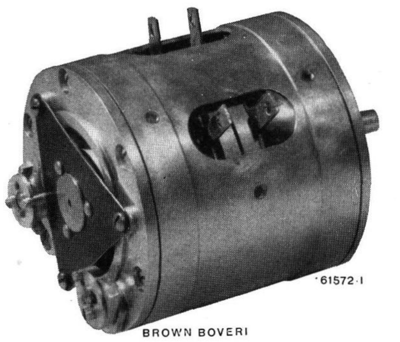


Fig. 5. — Cavity resonator for decimetre waves.

Finished cavity resonator ready for installation with symmetrical terminals. The triangular spring plate serves to fix the ceramic axis without clearance; the adjusting elements for correcting the capacitance and inductance are located at the same end. This enables the frequency characteristic to be varied so that several resonators can operate in synchronism.

D. THE CAVITY RESONATOR AS CONSTRUCTIONAL ELEMENT IN ELECTRICAL CIRCUITS.

The observations made in this section are restricted to such cases of cavity resonators where inductance and capacitance may still be calculated separately.

Fundamentally a difference is made between the inductive and capacitive coupling of the resonator. Both coupling possibilities for the open and closed or symmetrical and unsymmetrical forms of cavity resonator are shown in Fig. 6. We have restricted ourselves here to a coupling loop with *one* turn, because due to the excessive natural inductance of the loop it is very rare that more turns can be used.

With the *inductive coupling* the magnitude of the enclosed flux determines the coupling intensity. A coupling coefficient $k = 1$ is never obtained with a wire loop, because the inductance of the wire loop with the closest coupling, that is equal and maximum common turn surface, is considerably greater than the cavity resonator inductance which is small due to the current concentration. An additional capacitive coupling can be avoided by screening the coupling loop, whereby the screening casing must, however, be cut open at one point.

With the *capacitive coupling* an additional inductive coupling must be taken into account. In the extreme case of an infinitely large coupling capacitance, a direct galvanic coupling is obtained. The intensity of the coupling can in this case be altered

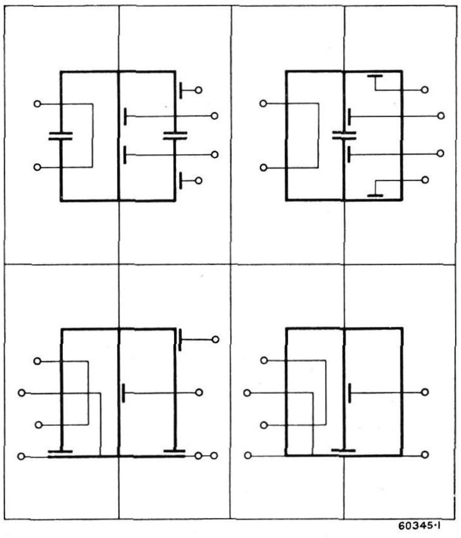


Fig. 6. — ***Cavity resonator couplings.***
The figures shows inductive and capacitive couplings for open, closed, symmetrical, and unsymmetrical forms of resonators.

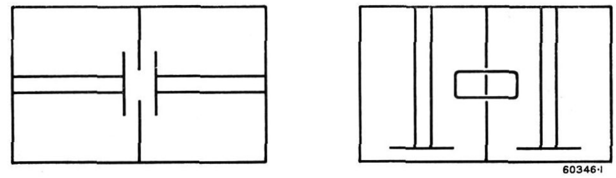
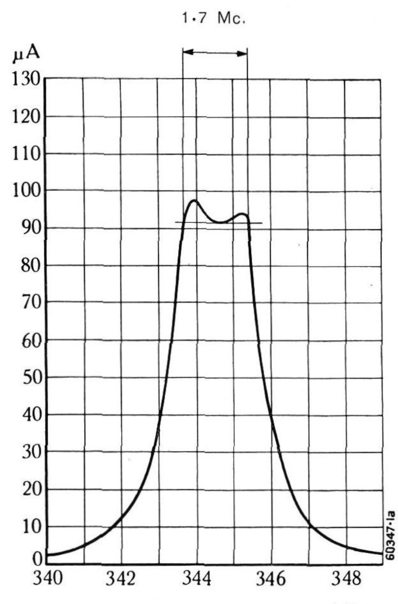
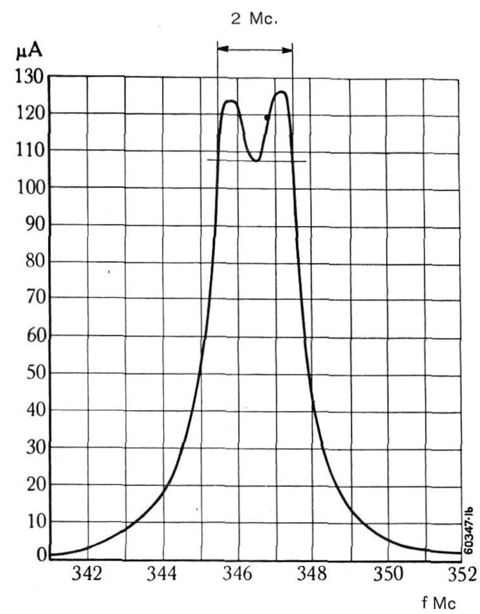


Fig. 7. — Cavity resonator band filters.
With the aid of suitable capacitive (left) and inductive (right) couplings, cavity resonators can be built together to form band filters, whereby synchronism of the resonators at frequency variations with a high degree of accuracy of the mechanical construction is possible.

coupling with a common coupling loop. Filters such as have been built by the Company for instance for decimetre waves, are shown in Figs. 8 and 9. The filter in Fig. 9 consists of two separate cavity resonators with a resonance sharpness Q of about 5000, each of which are artificially damped with 8500 ohms, so that Q drops to 350. The coupling of the circuits is inductive with a coupling factor of 0·51 %. Since the natural inductance of cavity resonators in the decimetre





Figs. 8 and 9. — Measured band filter curves with inductively coupled cavity resonators for decimetre waves.
Fig. 8 shows a filter with strong damping, whilst the curve in Fig. 9 represents a weaker damping with a higher coupling factor, for a band width of 2 Mc and a mean passing frequency of 345·5 Mc.

by displacing the connections, whereby it is to be noted, however, that large voltage variations occur at the central axis due to its large inductance, whilst at the casing there are only small changes in voltage.

The magnitude and nature of the coupling naturally also play an important part in cavity resonators which are designed as *band filters*. In this case it is also possible to employ a capacitive or inductive coupling, as is shown in the two-part filter according to Fig. 7. The capacitive coupling is achieved by one or more holes in the dividing wall at the electrical field, and the inductive coupling by means of holes in the dividing wall at the magnetic field or by a closer

wave range is already very small, the opposing inductance in this case only amounts to $0{\cdot}58 \times 10^{-10}$ henry.

Transition time effects, the influence of distributed capacitances and inductances, the pronounced skin effect, in brief the reality of the electromagnetic field which at these high frequencies is much more tangible and obvious, compel everybody who approaches the decimetre wave field for the first time to revise those fixed ideas which are based on omissions which in industrial alternating current technique, even sometimes partly extending up to the field of broadcasting frequencies, are quite permissible.

(MS 542) *Dr. R. Schüpbach / Dr. A. de Quervain. (Op.)*

THE MUTUAL EFFECT OF TWO FREQUENCY MODULATED WAVES IN LIMITERS.

Decimal Index 621.396.619.31

It is well-known that in the reception of frequency-modulated waves only the transmission with the greatest reception field strength (h. f. amplitude) is comprehensible. The weaker stations are noticeable merely in the form of crackling noises, i. e. beat notes at predominantly higher frequencies. What is remarkable is that of two frequency-modulated transmitters, for instance, the weaker is suppressed by the stronger even when the difference in amplitude is only slight.

Hereafter the frequency spectrum delivered by the f. m. detector is computed for the case of a limiter with two f. m. voltages of any amplitude ratio. From the result it is clear that only the audio frequency of the one transmitter remains, whereas all trace of the other disappears. Overtones and beat notes of the audio frequencies are also absent. The disturbing element consists chiefly of beat notes due to the difference between the carrier frequencies and their harmonics.

LET it be assumed that a voltage of the form

$$U = A_1 \sin y_1 + A_2 \cdot \sin y_2 \ldots\ldots (1)$$

exists at the input to the limiter. For the sake of brevity put:

$$y_1 = \Omega_1 t + m_1 \cdot \sin \omega_1 t \ldots\ldots (2)$$

and

$$y_2 = \Omega_2 t + m_2 \cdot \sin \omega_2 t \ldots\ldots (3)$$

By an elementary trigonometrical transformation we obtain from (1)

$$U = \bar{A} \cdot \sin \left\{ \frac{y_1 + y_2}{2} + \varphi(t) \right\} \ldots\ldots (4)$$

with $\bar{A}^2 = A_1^2 + A_2^2 + 2 A_1 A_2 \cos(y_1 - y_2) \ldots (5)$

$$\tan \varphi = \frac{A_1 - A_2}{A_1 + A_2} \cdot \tan \left[\frac{y_1 - y_2}{2} \right] \ldots (6)$$

If A_1, A_2, and $\cos(y_1 - y_2)$ are assumed to be time functions of a low-frequency character an ideal limiter will keep constant the amplitude $\bar{A}$ which is variable as a function of time.

The *discriminator* supplies an output voltage proportional to the expression

$$E = \frac{d}{dt} \left[\frac{y_1 + y_2}{2} + \varphi(t) \right] \ldots\ldots (7)$$

From (7)

$$E = \frac{1}{2}(\Omega_1 + \Omega_2) + \frac{m_1 \omega_1}{2} \cos \omega_1 t + $$
$$+ \frac{m_2 \omega_2}{2} \cos \omega_2 t + \frac{d\varphi}{dt} \ldots\ldots (8)$$

is obtained. It is now chiefly a question of determining the term $\frac{d\varphi}{dt}$. To this end and for the sake of simplicity assume A_1 and A_2 to be constant from the point of view of time. We then obtain:

$$\frac{d\varphi}{dt} = \frac{A_1^2 - A_2^2}{2 A_1 A_2} \cdot \frac{d\xi / dt}{\frac{A_1^2 + A_2^2}{2 A_1 A_2} + \cos 2\xi} \quad (9)$$

with the abbreviation:

$$\xi = \frac{1}{2}(\Omega_1 - \Omega_2) \cdot t + \frac{m_1}{2} \sin \omega_1 t - \frac{m_2}{2} \sin \omega_2 t \ (10)$$

The next step is to develop the quantity

$$f(\xi) = \frac{1}{\frac{A_1^2 + A_2^2}{2 A_1 A_2} + \cos 2\xi} = \frac{1}{\sigma + \cos 2\xi}$$

into a Fourier series of the form

$$f(\xi) = \frac{a_0}{2} + \sum_{\nu=1}^{\infty} \left\{ a_\nu \cdot \cos \nu \xi + b_\nu \cdot \sin \nu \xi \right\} \quad (11)$$

All b_ν and all a_ν with uneven ν immediately disappear so that solely:

$$a_{2N} = \frac{1}{\pi} \int_{-\pi}^{+\pi} \frac{\cos 2 N \xi \cdot d\xi}{\sigma + \cos 2\xi} \ldots\ldots (12)$$

has to be solved. The calculation is effected by means of a complex integration with the substitution

$$z = e^{2 i \xi} \ldots\ldots\ldots\ldots (13)$$

The result of this transformation is:

$$a_{2N} = \frac{1}{\pi i} \int \frac{(z^N + z^{-N})}{z^2 + 2 \sigma z + 1} \cdot dz \ldots (14)$$

The path of integration is the unit circle $|Z| = 1$. The poles of the integrand are:

$$z_0 = 0,\ z_1 = -\frac{A_2}{A_1} \text{ and } z_2 = -\frac{A_1}{A_2}$$

Here it is assumed that

$$|A_1| > |A_2| \ldots\ldots\ldots\ldots (15)$$

Then only z_0 and z_1 lie inside the unit circle. The residue for $z = o$ is then $\frac{z_1^N - z_2^N}{z_1 - z_2}$, while that for $z = z_1$ is $\frac{z_1^N + z_2^N}{z_1 - z_2}$

Hence the value of the coefficient (14)

$$a_{2N} = \frac{4 A_1 A_2}{A_1^2 - A_2^2} \left(-\frac{A_2}{A_1} \right)^N \ldots\ldots (16)$$

Reverting to equation (9) we finally find that

$$\frac{d\varphi}{dt} = [\Delta \Omega + m_1 \omega_1 \cos \omega_1 t - m_2 \omega_2 \cos \omega_2 t] \times$$
$$\times \left\{ \frac{1}{2} + \sum_{N=1}^{\infty} \left(-\frac{A_2}{A_1} \right)^N \cdot \cos 2 N \xi \right\} \ldots (17)$$

with the abbreviation $\Delta \Omega = \Omega_1 - \Omega_2$ (difference between carrier frequencies). Substituting the value obtained in this way for $\frac{d\varphi}{d}$ in (8) we obtain

$$E = \text{const.} + m_1 \omega_1 \cos \omega_1 t + \quad (18)$$
$$+ [\Delta \Omega + m_1 \omega_1 \cos \omega_1 t - m_2 \omega_2 \cos \omega_2 t] \times$$
$$\times \sum_{N=1}^{\infty} \left(-\frac{A_2}{A_1} \right)^N \cdot \cos N \left\{ \Delta \Omega t + m_1 \sin \omega_1 t - m_2 \sin \omega_2 t \right\}$$

By employing the well-known relations for Bessel functions

$$e^{i x \sin \omega t} = \sum_{K=-\infty}^{+\infty} J_K(x)\, e^{i K \omega t} \quad \ldots\ldots (19)$$

and $$J_{K-1}(x) + J_{K+1}(x) = \frac{2K}{x} \cdot J_K(x) \quad \ldots . (20)$$

we finally obtain

$$E = \text{const.} + m_1\, \omega_1 \cos \omega_1\, t +$$
$$+ \sum_{N=1}^{+\infty} \sum_{K=-\infty}^{+\infty} \sum_{K=-\infty}^{+\infty} (-1)^{N+L} \left(\frac{A_2}{A_1}\right)^N \cdot \frac{\omega_{N;K,L}}{N} \cdot$$
$$\cdot J_K(m_1 N) \cdot J_L(m_2 N) \cdot \cos(\omega_{N;K,L} \cdot t) \quad (21)$$

with $\omega_{N;K,L} = N \cdot \Delta\Omega + K\omega_1 + L\omega_2$

From the result (21) *it is clear that of the audio frequencies only* ω_1 *remains. Every trace of* ω_2 *has disappeared, while no overtones of* ω_1 *or* ω_2 *nor beat notes* ($K\omega_1 \mp L\omega_2$) *occur whatsoever.* The triple sum over N, K, and L, comprising beat notes due to the difference $\Delta\Omega$ between the carrier frequencies and their harmonics is to be looked upon as the disturbing element.

When $\Delta\Omega$ is over 10 kc these beat notes have hardly any disturbing effect. It can be proved by means of numerical examples, however, that even when $\Delta\Omega$ is small the disturbing noise is not very great. The special case of $\Delta\Omega = 0$ is also interesting, but due to the involved calculations entailed, space will not allow of its mathematical treatment here.

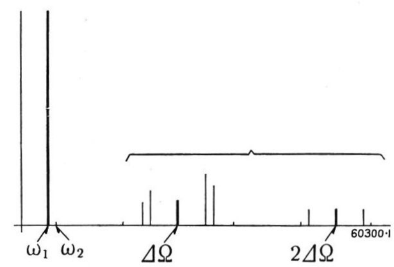


Fig. 1. — Frequency spectrum of output voltage of a f. m. detector with two f. m. voltages of slightly different frequency at the input.

ω_1. Carrier frequency.
ω_2. Audio frequency of disturbing transmitter.
$\Delta\Omega$. Difference between carrier frequencies.

The audio frequency of the disturbing transmitter does not come within the spectrum. As will be clear the disturbing component comprises beat notes due to the difference between the carrier frequencies.

To illustrate more clearly our result the frequency spectrum of the output voltage E (formula 21) is shown in the accompanying figure.

(MS 527) *Dr. P. Güttinger. (E. G. W.)*

Wiring a small frequency-modulated ultra-short-wave transmitter.

Such sets have proved their worth in police radio applications.

Industrial Applications of High-frequency Engineering.

Decimal Index 621.364.15

High-frequency power is finding ever-increasing application for industrial purposes and in particular for the drying and hardening of insulating materials and plastics as well as for the surface heat treatment of metals. Where the hardening of plastics is concerned the high-frequency method has the advantage that due to the dielectric loss phenomenon the processing heat is developed in the interior of the material and has not to penetrate into it by conduction. In point of fact, when thermo-setting plastics are subjected to a high-frequency field the temperature is highest and in consequence the hardening process the most advanced in the very interior, so that the reaction products can be freely expelled from the beginning to the end of the hardening process. Further merits are shorter hardening times, lower power consumption, and more uniformly hardened finished product than with any other process. These advantages are the most striking when hardening thick-walled parts. The first article hereafter: "Hardening of Synthetic Resins by High-frequency Power" describes tests in connection with synthetic resins of the phenol-formaldehyde class.

The surface hardening of metals, and in particular steel, by high-frequency power also affords quite definite advantages. In contradistinction to insulating materials such as synthetic resins, etc., only the peripheral layer of metallic conductors is heated. This takes place so rapidly and to such a slight depth that the heat does not penetrate into the core and thus cause distortion of the work. As a result the part in question can be completely machined before hardening, while it is also unnecessary to heat the whole length of unwieldy parts when only a small portion of their surface is required to be hardened. Gradual transition from the hardened layer to the core, accurate regulation of the process temperature, cleanliness of the whole process, and the possibility of hardening lengthy parts by the progressive method are other advantages of high-frequency surface hardening. A second article entitled "The Surface Treatment of Metals and in particular the Surface Hardening of Steel by High-frequency Currents" deals with a number of fundamental questions in connection with this process. Based on test results with the high-frequency electronic generator developed by Brown Boveri for industrial purposes a number of interesting applications are finally given.

I. HARDENING OF PLASTICS BY HIGH-FREQUENCY POWER.

ALTHOUGH details of the requisite equipment and conditions for the hardening of plastics by high-frequency power are to be found in technical literature nothing has so far been published on the subject of the fundamental processes taking place in connection with the hardening of synthetic resins. In order to understand the following notes, therefore, an idea of the different stages through which thermo-setting plastics pass during the hardening process must be obtained.

Modern organic materials are divided into two large classes, thermo-plastic and thermo-setting. By hardening is understood a non-reversible chemical process through which with rising temperature the originally soluble and fusible initial condensation products are rendered insoluble and infusible. Inasmuch as the thermo-plastics cannot be hardened they will not be dealt with here. Fig. 1 shows the relation between the individual hardening stages of phenol-formaldehyde poly-condensation products. It will be noticed that the change from one state to the other is not spontaneous, but that it takes place in a succession of stages.

Thermo-setting resins are rarely used pure, being combined with additional substances to obtain the desired material. The additions may be of anorganic (asbestos, talc, stone flour) or organic (wood flour, organic fibres) nature, and in powder (wood or asbestos flour), fibrous (cellulose or asbestos fibres) or lumpy (chopped fabric) form. The mixture of resin and addition is converted into mouldings in entirely closed, heated moulds. Apart from these materials there are also laminated sheets, comprising anorganic or organic papers and fabrics or wood veneers impregnated with resins, which are produced, chiefly in the form of semi-finished manufactures, in hot-plate presses. Laminated plastics are generally machined to obtain the desired shape.

HARDENING OF PLASTICS IN HOT-PLATE PRESSES.

Hitherto, plastic mouldings and laminated sheets have generally been hardened in presses with heated platens, i. e. the heat was applied to the exterior of the material. The heat conduction of the mixture to be moulded, however, is generally so poor, that the applied heat can penetrate only very slowly into the

$C_6H_5OH + CH_2O$

↓

Phenol alcohols

↓ $-H_2O$

Initial condensation products

↓ slight polycondensation

Resols	→	**Resitols**	→	**Resites**
Low molecular weights, fusible, soluble	$-H_2O$ $-$ Low-molecular polycondensation products, Increasing polycondensation	Larger molecules, fusible, practically insoluble	Polymerization, Polycondensation, Formation of three-dimensional linkage	Macromolecules, infusible insoluble

Effect of heat, Hardening →

Fig. 1.

interior. In order, therefore, to cut down the moulding times, especially where thick, laminated sheets are concerned, the material must be subjected to a subsequent hardening process in a stove without the application of pressure. The heat required to be applied to a moulding, as well as the heat distribution in it is dependent to a large extent on its dimensions.

For hardening purposes it has been found that quite definite temperature ranges are required according to the composition of the moulding mixture, whereby, however, with increasing thickness very much longer heating times have to be allowed to attain the mean temperature.

Since the hardening process of plastics is highly dependent on temperature it goes without saying that the outer layers will have reached the final hardened state long before the interior. The differences are all the greater the thicker the walls of the moulding happen to be. This non-uniform hardening process gives rise to non-uniform mechanical properties within the moulding.

As will be clear from Fig. 1 intermediate reaction products (water, formaldehyde, low molecular condensation products) are formed during the hardening process and the rate of diffusion of these decreases as hardening progresses. In other words — particularly where laminated sheets above certain dimensions are concerned — the intermediate reaction products can no longer escape from the interior because the outer, directly heated zones are the first to become completely hard. This may lead to the formation of internal fissures, either during the pressing process or in service. From these considerations it is obvious that the present conventional method of hardening synthetic resins has big technical shortcomings.

The chemical changes taking place during the hardening process can be traced by determination of the volatile intermediate condensation products. The *losses in weight* observed for a given hardening temperature with different hardening times enable a hardening curve to be plotted. As will be seen from Fig. 1 the originally fusible and soluble resols change to infusible, insoluble resites during the hardening process. *The variation in the solubility of the material in alcohol or acetone,* therefore, also enables the hardening process to be followed indirectly. Whereas the losses in weight increase with the hardening time until a steady value is reached, the percentage of soluble constituents diminishes as a function of the hardening time. The foregoing changes thus enable the hardening process to be followed, if not qualitatively, at least quantitatively.

HARDENING OF PLASTICS BY HIGH-FREQUENCY POWER.

Here the conditions are fundamentally different in various respects. In their original condition the more common plastics have quite high power factors and corresponding dielectric constants, which are chiefly caused by the mobility of the polar groups in the high-frequency field to which, however, must be added the mobility of the chain units. The power factor governs, inter alia, the power input and the heat produced in the electrical alternating field. During the hardening process the macromolecules of plastics become three-dimensionally linked, especially in the resite formation stage, which considerably diminishes the mobility of the chain. Due to intermediate reactions or steric hindrance, however, the mobility of the polar groups is more and more impaired. In consequence the power factor becomes lower and lower in the course of the hardening process, with a corresponding variation in the dielectric constants. The reaction characteristic for the formation of phenol plastics shown in Fig. 1 leads one to assume that this class of plastics is particularly suitable for hardening by high-frequency power, due to the large number of highly mobile polar (OH) groups occurring in the macro-

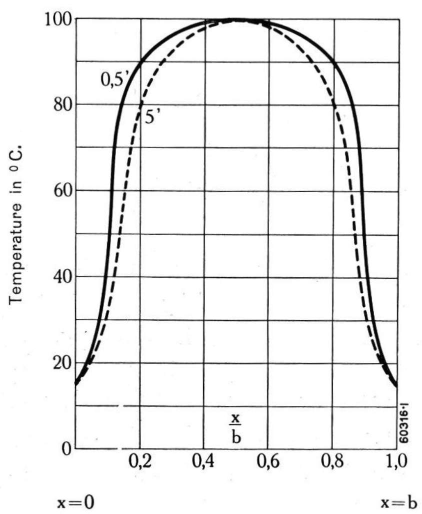


Fig. 2. — Temperature distribution through a laminated sheet heated with high-frequency power, after 0·5 and 5 minutes, respectively.

(According to data given by George H. Brown). In contrast to the hot-plate method of heating, the temperature here is greatest in the centre, where, in consequence, the hardening process is the furthest advanced. This permits the intermediate reaction products to escape. As a result high-frequency hardening gives a more uniform and better result.

Assumption: Maximum temperature = 100° C.
Thickness of sheet = 1·27 cm.

molecule. With the high-frequency method of heating, governed as it is by the dielectric constant and power factor, quite a different temperature distribution is to be expected in the moulding than when hardening in presses with heated plates. The heat is uniformly distributed, especially in the interior of the moulding. Due to the radiation of the heat from the outer zones it is not necessary for the temperature to be uniform throughout, but a temperature gradient from the centre to the exterior may occur. These conditions are shown in Fig. 2. In the first place it will be noticed that the heating time is substantially shorter than when heat is applied externally. As a result, it is also to be assumed that high-frequency heating will allow of larger cross-sections being heated up to the requisite hardening temperature in a shorter time than with external heating. Moreover, the intermediate reaction products can readily attain the surface due to the fact that the hardening process cannot be further advanced than in the interior.

TESTS UNDERTAKEN.

(a) *General.*

The following tests were undertaken by Brown, Boveri & Co., Ltd., Baden, and the Industrial Research Department of the Technical Physics Institute at the Swiss Federal Institute of Technology, Zurich, in collaboration. Since to the authors' knowledge, as already stated at the beginning of these notes, nothing has so far been published on the mechanism of the hardening of synthetic resins, the tests about to be described will serve to fill, although somewhat imperfectly, this gap in the literature on the subject.

Inasmuch as the phenol class of thermo-setting plastics is by far the most frequently employed in the manufacture of modern organic materials, the investigations were confined to this type of material. Moreover, as pointed out in the foregoing notes, phenol resins appear to be particularly suitable for such tests on account of their chemical nature.

In order to enable the hardening process to be correctly followed pure resins must at first be investigated to ensure other effects being precluded. The pure resins were made especially for the tests. The power factor was made particularly high by the addition of salt in the proportions shown in Table I, where the associated dielectric constants and power factors are also given.

Since various, involved effects render tests on moulded materials difficult several other series of tests were carried out on laminated materials, i. e. plywood of beech and oak veneers 0·6 and 0·8 mm thick, respectively. Due to the wood veneers and the fact that the dielectric properties of wood depend on the moisture content the board varies in behaviour during the heating process. Moreover, the interaction between the bonding agent and basic material may be different according to the previous treatment, type of bonding agent, and degree of compression. Microscopic examination clearly shows up the difference in structure (see Fig. 3). In the case of compressed laminated wood with which pre-impregnated wood veneers are subjected to high pressure the bonding agent has a predominating effect.

A self-excited 1·5 kW high-frequency electronic generator giving a frequency of about 23 megacycles was employed for the tests. To begin with the hardening process was followed on the pure resins in Table I. The test specimens were cylindrical in form, their diameter being 26 mm and their height 28 mm. Preliminary tests with compressed laminated wood were carried out on cube-shaped test-pieces of 30×30×30 mm. The test set-up is shown in Fig. 4.

In a further series of tests, specimens of plywood of 80 × 80 × 80 mm were investigated. Since the escaping intermediate reaction products are liable to cause the wood to split, especially when heating rapidly, a corresponding counter-pressure must be applied. As is shown in Fig. 5 the tests were carried out in a 20-ton press.

(b) *Test Results.*

Since for general considerations the hardening process cycle in connection with pure resins is the most important, the tests made on such materials will be discussed first of all. In order to enable the variation in temperature in the interior of the pure resin test specimens to be followed during the hardening process a fine hole was bored in the centre and a copper-constantan thermo-couple introduced, while a further temperature measuring point was arranged 0·5 cm below the surface. If a symmetrical field is employed a thermo-couple can be used without risking any additional heating effect. The *relation between the applied voltage and the temperature* in a pure resin test specimen with the addition of sodium lye as conducting medium is shown in Fig. 6. At a given voltage the temperature rises to a definite value and then drops off slowly again. This fact signifies that after the maximum temperature has been reached the corresponding hardening state is attained in a very short time and the temperature must then drop due to the condensation of the intermediate reaction products and the consequent reduction in power factor. This observation is of significance for the fundamental elucidation of the mechanism of hardening of phenol resins. With the conventional methods of hardening in stoves or hot-plate presses it is impossible to keep a check on such phenomena.

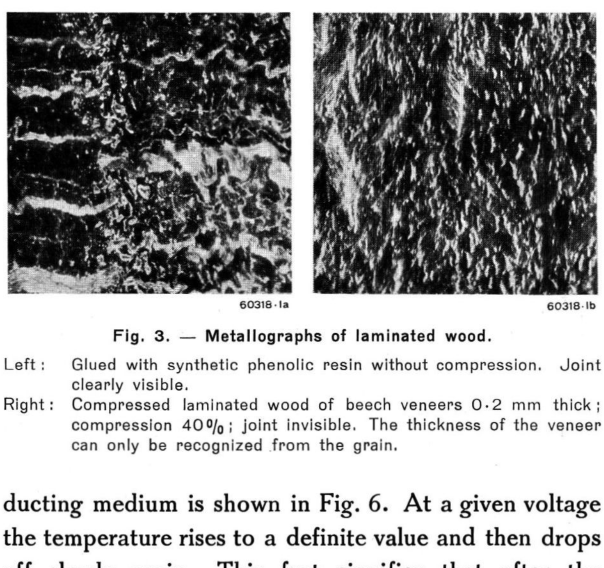


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Fig. 3. — Metallographs of laminated wood.

Left: Glued with synthetic phenolic resin without compression. Joint clearly visible.
Right: Compressed laminated wood of beech veneers 0·2 mm thick; compression 40%; joint invisible. The thickness of the veneer can only be recognized from the grain.

TABLE I.

Dielectric constant ε and power factor $\tan\delta$ of pure resins with salt additions.

Pure resin with salt additions	Frequency in cycles									
	5×10^4		2×10^5		1×10^6		4×10^6		$1{\cdot}3\times10^7$	
	$\tan\delta$	ε	$\tan\delta$	ε	$\tan\delta$	ε	$\tan\delta$	ε	$\tan\delta$	ε
2% $NaOH$	0·167	12·0	0·118	11·3	0·098	10·2	0·094	9·2	0·089	8·1
1% $NaCl$	0·057	9·2	0·059	8·8	0·060	8·2	0·065	7·7	0·069	7·1
0·1% $AlCl_3$	0·074	9·9	0·066	9·2	0·064	8·6	0·069	8·2	0·069	7·4

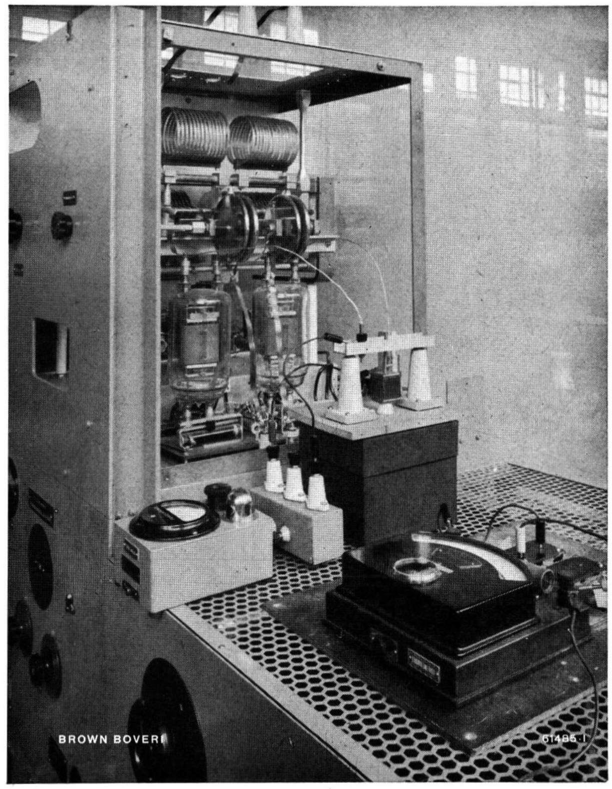


Fig. 4. — Test set-up for hardening of pure resins.

The self-excited high-frequency electronic generator has a power of 1·5 kW at about 23 megacycles.

Similar test specimens were hardened in a high-frequency field and in a conventional electrically heated stove. The *heating times* required to reach the desired hardening temperatures differ very widely for the two cases, as will be seen from Fig. 7.

The *hardening process* itself was followed (as already mentioned) by determining the losses in weight and the reduction in solubility in acetone. In order to obtain comparable hardening conditions the same heating conditions, as shown in Fig. 8, were selected for both the high-frequency and stove method of hardening.

An example will now be taken to show the mechanism of hardening under the given conditions. For this pur-

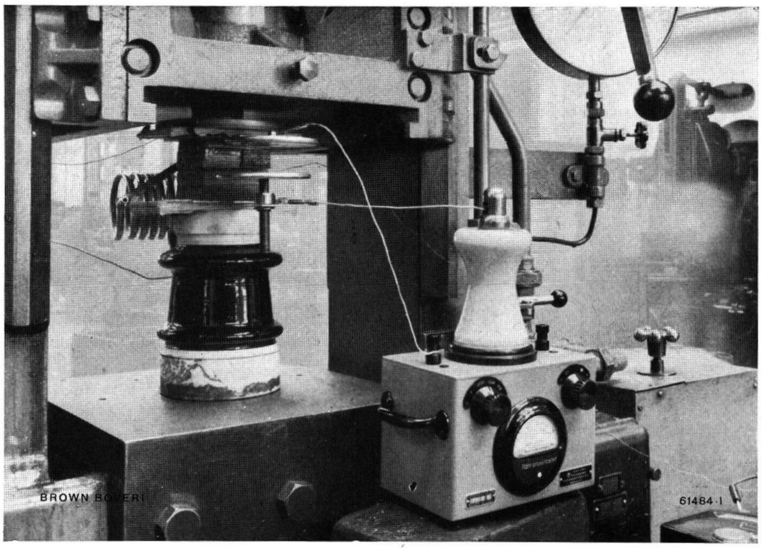


Fig. 5. — Test set-up for hardening laminated materials (compressed laminated wood) in a 20 t press.

The high-frequency power was supplied by the electronic generator depicted in Fig. 4.

pose the *results of the tests on pure resins with the addition of natrium hydroxide as conducting medium* will be taken. From Fig. 9 it will be seen that the hardening losses are greater with the high-frequency method than with the conventional stoves. As already inferred this is due to the fact that the hardening of the outer zones is not sufficiently advanced to prevent the intermediate condensation products escaping. The temperature characteristic is also of special significance in the hardening process. The hardening effect is the most complete in the hardening stage II due to the short heating period. If the specimen is heated up slowly similar hardening losses occur with both high-frequency and stove heating. For practical purposes it is to be concluded therefore that as rapid a rise in temperature — such as is obtained with high-frequency power — gives the best hardening effect. This is corroborated by Fig. 10, where the hardening cycle for pure resins is plotted as a function of the applied voltage.

The material must become less soluble in acetone as hardening progresses. Absolute insolubility was not aimed at in the tests, so that only definite hardening times were arranged for. After the selected hardening times had expired pure resin in powdered form was tested in the Soxhlet apparatus up to the point of exhaustive extraction to determine its solubility in acetone. Fig. 11 gives the test results of the three hardening stages shown in Fig. 8. The facts resulting from the lost weight method of checking up the hardening process are corroborated by the change in solubility, which is at a minimum in hardening stage II, and attains a value practically corresponding to the completely hardened condition. In hardening stage III the differences are again smallest. The previous conclusions regarding the practical application of high-frequency power to the hardening of phenol resins are thus again fully corroborated.

Where *laminated materials* are concerned conditions are different in that apart from the change in the resin itself the changes in the basic material have also to be taken into consideration. With the conventional arrangements these two

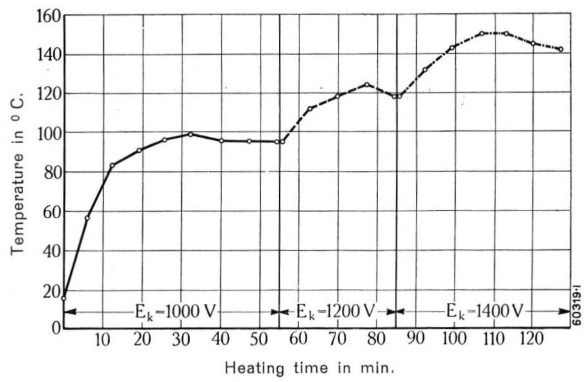


Fig. 6. — Relation between hardening temperature in laminated materials and voltage of high-frequency field.

For a given voltage the temperature rises to a maximum, then drops slightly, and does not attain another maximum value unless the voltage is increased.

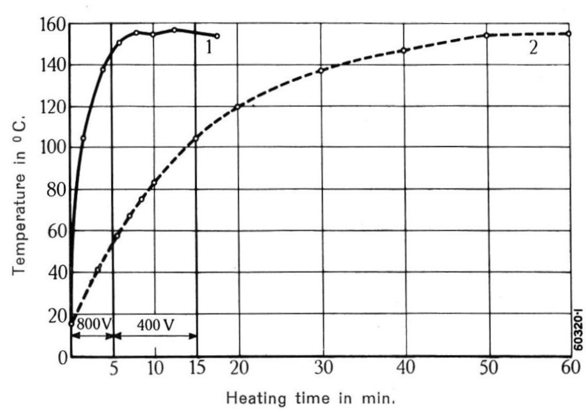


Fig. 7. — Comparison between temperature rise in pure phenolic resins in high-frequency field and in electrically-heated stove.

The heating process takes place far more quickly in the high-frequency field (curve 1) than in the stove (curve 2).

influences are inherently connected. The dielectric properties of the resin vary in the course of the hardening process due to the associated chemical changes, whereas those of the basic material (in the present case wood veneers) can vary with changing moisture content.

At the outset cubes of laminated wood (30 × 30 × 30 mm) were employed, but due to the low resin content compared to the total volume no practical result could be obtained. Thereupon larger cubes (80 × 80 × 80 mm) were used and the tests carried out in a 20 t press (see Fig. 5). Since with external heating cubes of such size already involve long hardening times the tests were at first confined to high-frequency heating. The temperature across the section was again measured by means of thermo-couples. A big advantage of the high-frequency method of hardening is that by regulating the voltage it is possible to maintain the tem-

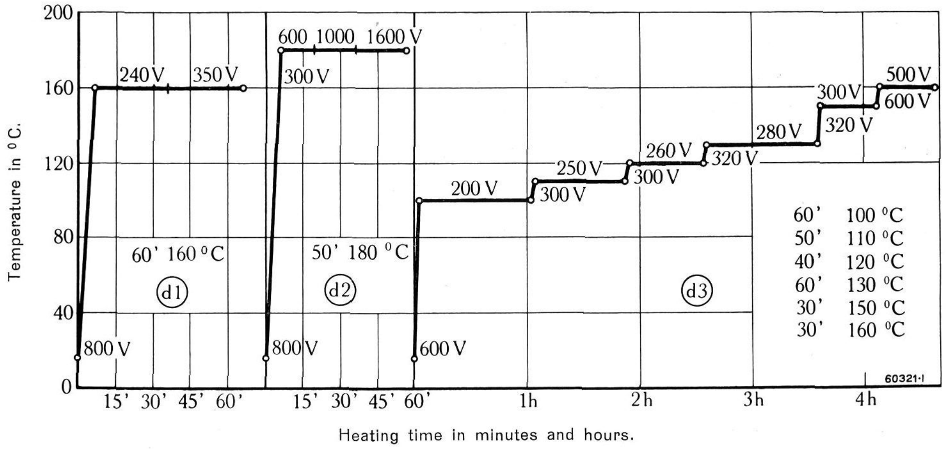


Fig. 8. — Characteristics of three different hardening conditions employed during tests, in relation to time, temperature, and voltage.

In order to obtain comparable results these hardening conditions were employed both for stove and high-frequency heating.

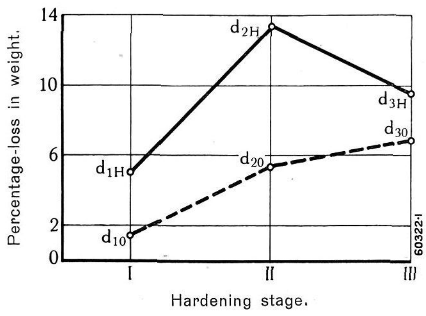


Fig. 9. — Relation between hardening losses of pure phenolic resins and hardening conditions shown in Fig. 8. Test series d.

Dash line: Stove hardening.
Full line: Hardening by high-frequency power.

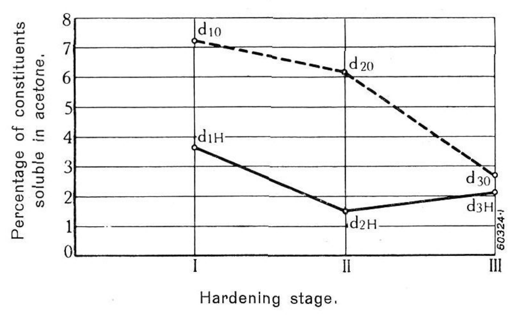


Fig. 11. — Percentage of pure phenolic resins soluble in acetone in various hardening stages.

The stove-hardened resins contain a greater percentage of constituents soluble in acetone and are thus less uniformly hardened.

Dash line: Stove hardening.
Full line: High-frequency hardening.

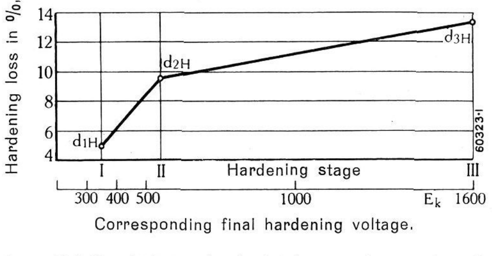


Fig. 10. — Relation between hardening losses of pure phenolic resins and applied voltage.

perature constant at different points in the laminated material. For the tests the same hardening stages were employed as before, these being as shown in Fig. 12. In order to be able to follow the mechanism of the heating process in the desired part of the specimen test-pieces were taken from the outside and inside zones as well as from the centre (see Fig. 13). The small cubes taken from the specimen were treated with acetone in the usual way to determine their solubility and in consequence the state of hardening. The results of these investigations are shown in Fig. 14. In the case of laminated wood the acetone extracts contain

The earlier observations that with the high-frequency method of hardening phenol resins, hardening takes place from the interior to the exterior, due to the radiation of heat from the outer zones, were thus fully corroborated by the described tests. Since the power

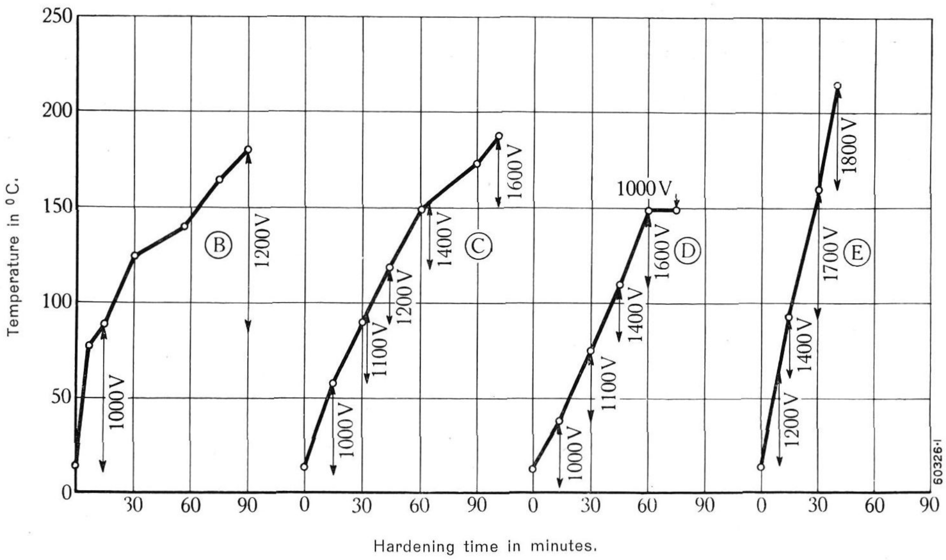


Fig. 12. — Characteristics of hardening conditions of compressed laminated wood investigated, in relation to temperature, time, and voltage.

(apart from the soluble constituents of the thermo-setting bonding agent) also the soluble elements of the basic material (in the present case the wood veneers). In the case of specimen E (Fig. 14) an increase in the constituents soluble in acetone was observed. This was due to the fact that by reason of a too rapid voltage rise in the centre the wood veneer was already slightly damaged and even to some extent carbonized. This condition is shown in Fig. 15 (top and centre). A section through the centre zone shows only traces of carbonization (Fig. 15, below).

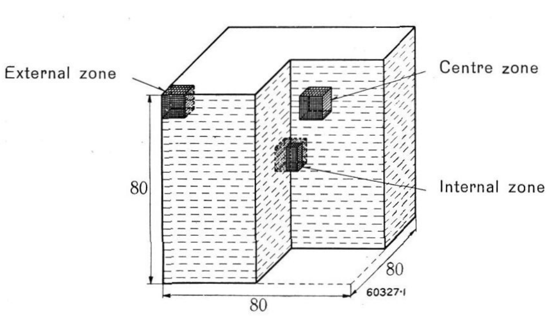


Fig. 13. — Points from which specimens of compressed laminated wood taken for tests.

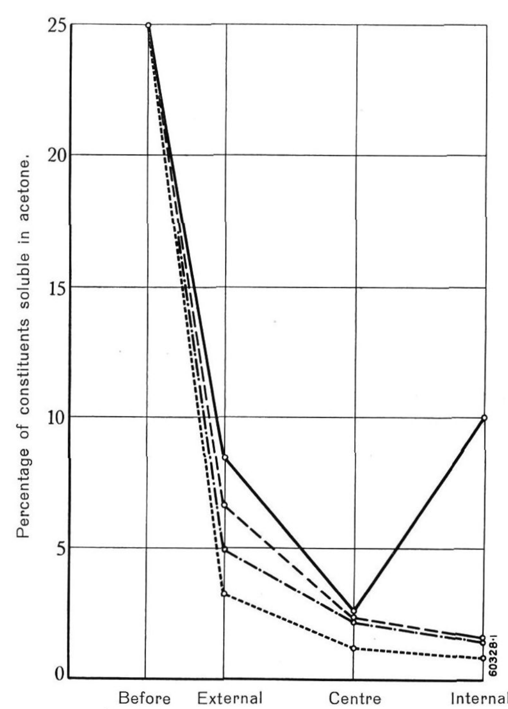
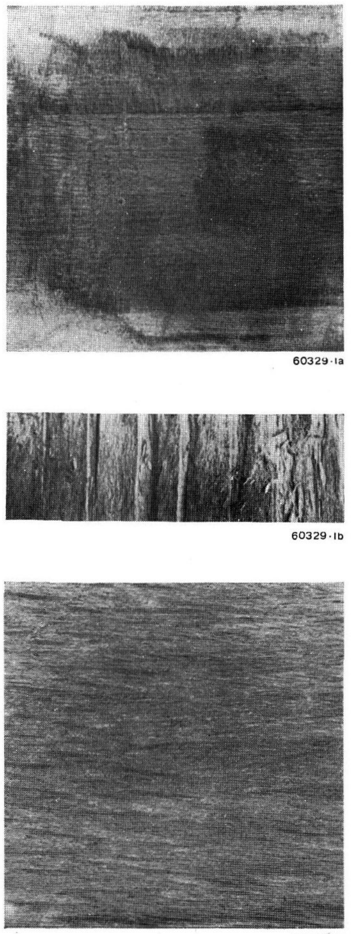


Fig. 14. — Percentage of constituents of various specimens soluble in acetone (internal, centre, and external zones).

——— Specimen E – – – – – Specimen C
—·—·— Specimen B - - - - - - - Specimen D



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60329-Ib

60329-Ic

Fig. 15. — Effect of incorrect temperature characteristic (too rapid voltage rise at beginning of hardening process) on compressed laminated wood with heating by high-frequency power.

Top = Section through core zone of specimen E; heavy carbonization.

Centre = Section through core zone of specimen E; heavy carbonization and destruction of fibres.

Below = Section through centre zone of specimen E; carbonization only slight, thus proving that the temperature in the core zone was highest, i. e. that there was a temperature gradient towards the exterior.

Such deterioration of the material can be avoided and first-class results obtained by correct regulation of the voltage.

factor is highly dependent on the temperature, while a relatively slight temperature rise in the interior causes an increase in power factor, an additional temperature rise can occur with the result that given unsuitable heating conditions, i. e. incorrect rate of voltage rise, damage can be caused to the interior of the laminated material. As the hardening process advances the power factor decreases and the temperature drops back to a value which is no longer excessive. These phenomena only occur when the initial voltage rise is too rapid. By proper temperature control and corresponding voltage regulation the foregoing harmful effects are out of the question, as proved by Fig. 14, specimen D.

CONCLUSIONS.

As a practical result of these tests it is proved that the hardening of phenol resins by high-frequency power from the interior can be achieved in a much shorter time than with stoves or hot-plate presses from the exterior. This is of particular advantage where thick work-pieces are concerned. The comparison between the conventional

TABLE II.

Comparison of high-frequency and hot-plate methods of heating when gluing plywood boards with phenolic resins, according to George H. Brown.

	High-frequency method (HF)			Hot-plate method (HP)			
Thickness b in cm	Power into wood in watts	Gluing time in min	Energy in watt-min	Gluing time in min	Energy in watt-min	HP time / HF time	HP energy / HF energy
1·27	6,670	1·1	7,350	2·375	8,520	2·15	1·16
2·6	17,900	2	35,800	8·32	40,700	4·16	1·138
15·4	37,500	8·2	307,500	342	40,900	41·7	1·33

process with hot plates and the new process with high-frequency heating in Table II shows up clearly the advantages of the latter method. Following on this excerpt from the results of tests on phenol resins the next task for industrial investigators is to examine the fundamental mechanism of hardening of other thermo-setting plastics and to collect the necessary data for their technical application.

(MS 532) *Dr. H. Stäger / Dr. F. Held. (E. G. W.)*

II. THE SURFACE TREATMENT OF METALS AND IN PARTICULAR THE SURFACE HARDENING OF STEEL BY HIGH-FREQUENCY CURRENTS.

It is a well-known fact that the eddy currents produced by strong high-frequency fields cause an intense heating effect even in good conductors. This phenomenon is employed to advantage in various industrial applications. For instance, harmful adsorbed gas can be removed from metallic parts by heating with high-frequency currents under vacuum. Another important use is the instantaneous inductive heating of thin peripheral layers of steel, which is practically only possible with high frequency currents. The thin heated layer is subsequently hardened by subjecting it to a rapid cooling effect. Surface hardening has great advantages over all other hardening methods. Inasmuch as the surface only is heated a minimum amount of energy is dissipated during the hardening process. For the same reason the heating time is extremely short, which is of importance where mass production is concerned. Since the inner core is not affected by the hardening process the warping produced with other methods is kept within very narrow limits and in consequence the machining of the work can generally be completely finished in a relatively simple manner before hardening, thus avoiding subsequent costly grinding operations. A further noteworthy advantage is the cleanliness of the process.

The mechanism of the heating up of thin layers of steel parts will be readily visualized from the following notes:—

Due to the skin effect the currents induced in a piece of steel by a high-frequency field are concentrated on the surface, so that heat is only produced in the peripheral layers. If the applied power is sufficiently large to permit the hardening temperature to be attained and the part is thereupon immediately cooled, heat dissipation has no time to take place; hardening to depths of less than 1 mm is therefore possible. The three chief factors determining the thickness of the hardened zone are: the depth of penetration ε of the eddy currents, the necessary power required per unit surface area, and the heating time. The depth of penetration is a criterion for the thickness of the current-conducting peripheral

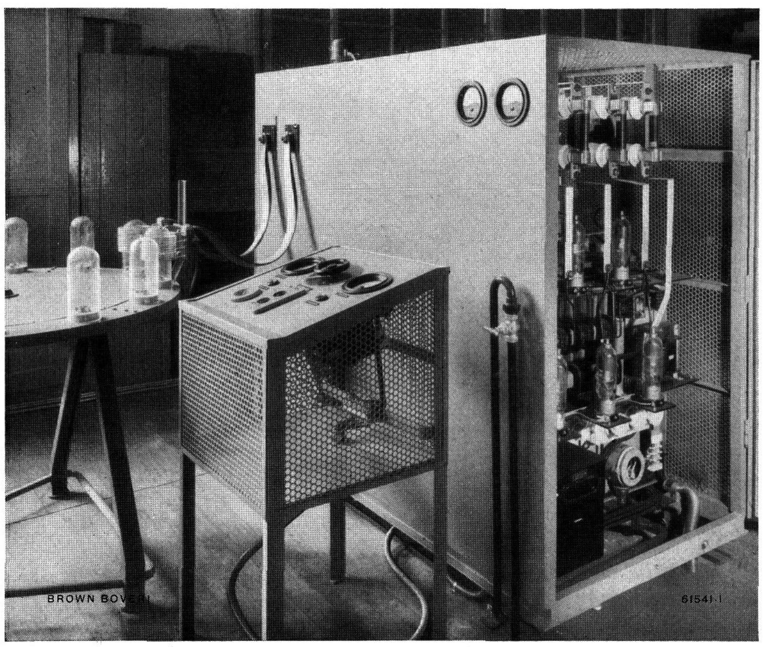


Fig. 1. — High-frequency generator for industrial purposes.
Universally applicable for hardening, outgassing of metal parts under vacuum, drying, etc.

layer. It is proportional to $1/\sqrt{f}$ (f= frequency), whereas the power input to the work is proportional to $\sqrt{f}$. This explains why surface hardening in the real sense of the word is only possible with high-frequency currents.

For carbon steel ε is about 0·9 mm at 850° C, 500 kc, and the heating time roughly 1 second.

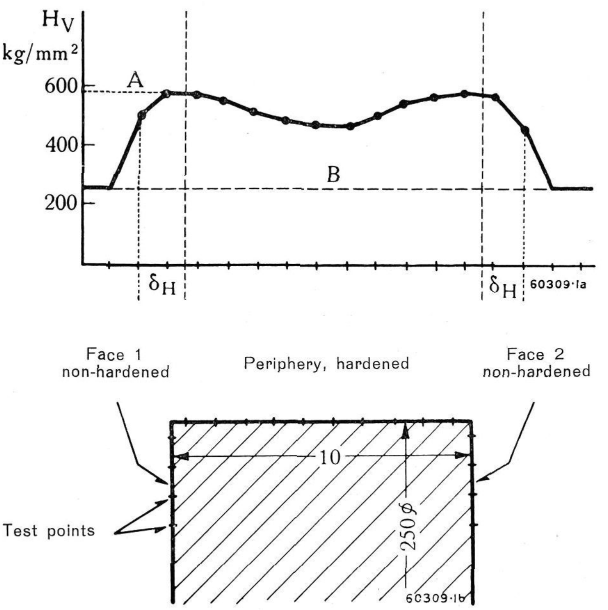


Fig. 2. — Measured hardness values of a chromium-steel disc in Vickers units.

Below, the distribution of the hardness test points over the faces and periphery of the disc ; above the corresponding test results developed.

A. Maximum hardness.
B. Initial hardness.
δ_H. Thickness of hardened layer determined metallographically.

The heating time is of importance because the shorter it is the less intense will be the heat dissipation from the heated layer and, in consequence, the thinner the hardened layer. To denote this the ratio of 1:5 between the metallographically determined thickness of the hardened layer δ_H and the depth of penetration ε is employed. It follows that heating times of less than 1 second are necessary to give a hardened layer 1 mm thick.

The design of the heating coil or inductor which produces the high-frequency electromagnetic field must be given particular care. In order to permit the requisite powers to be transmitted the coupling between the inductor and charge should be as high as possible, i. e. the inductor must be adapted to the contour of the part to be treated. Where particularly large specimens are concerned with which the available power is too low to heat the entire surface in one operation, or in the case ot complicated contours, the method of progressive hardening may be employed to advantage. In this way long shafts, for instance, can be hardened with a relatively small expenditure of power, the shafts being advanced at a certain rate or possibly even rotated during the process.

Notable applications are the surface hardening of gear wheels of up to 200 mm in diameter, spindles and shafts (by the progressive process), tools, highly stressed parts of small apparatus, and needles, as well as the internal hardening of annular parts, tubes, etc.

Fig. 1 depicts a high-frequency generator for industrial purposes, ready for service.

Hereafter, the *results of tests* carried out in connection with surface hardening by high-frequency currents are given.

(a) *Hardening a Circular Chromium Steel Disc.*

The periphery of the disc (250 mm in diameter, 10 mm thick) was heated up to its hardening temperature of 950° C and hardened by quenching in cold water.

Material: Chromium steel with a 15% Cr content.
Heating period: 3 seconds.
Power input to disc: roughly 50 kW.
Frequency: about 250 kc.

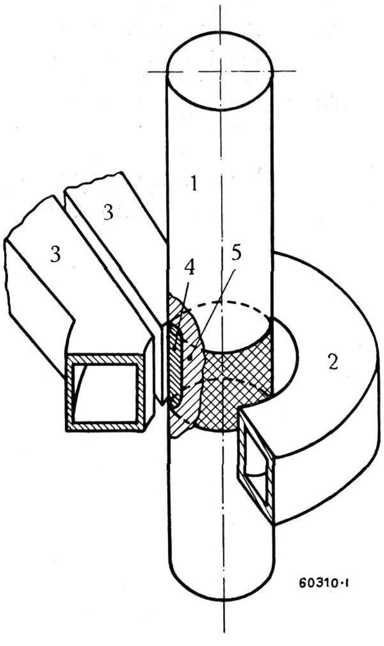


Fig. 3. — Diagram of arrangement of inductor and shaft for the surface hardening of various annular zones.

1. Shaft.
2. Inductor, one turn, water-cooled.
3. Connections.
4. Hardened surface layer.
5. Non-hardened.

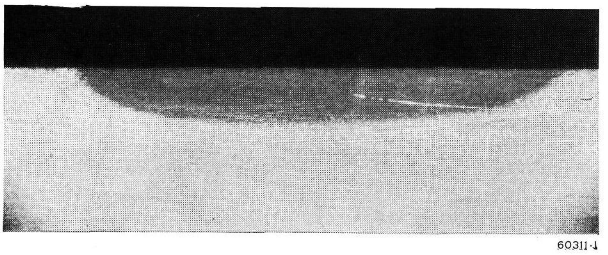


Fig. 4. — Surface ground specimen showing hardened zone 4. Magnification 10 ×.

Longitudinal section of shaft through hardening zone 4. The darker, semi-elliptical portion represents the hardened material, the remainder the non-hardened core. The arrangement of the inductor around the shaft is shown in Fig. 3.

Fig. 5. — Microstructure of treated steel. Magnification 200 ×.

On left, original material.
On right, hardened material.

Particularly noteworthy is the smooth transition from the core to the hardened layer. No distinct transition zone is to be observed.

Fig. 2 gives the hardness values H_v, in Vickers units, of the faces and periphery of the disc, as found by test.

The maximum measured hardness H_v was 585 kg/mm², corresponding to a Brinell hardness of 536 kg/mm². The thickness δ_H of the hardened layer, as determined metallographically, was 1·5 mm. These values fully meet practical requirements, e. g. such as are imposed by the hardening of gear-wheels.

(b) *Hardening a Carbon-steel Shaft.*

In order to determine the influence of the heating period for a given power, four annular zones of the surface of a shaft 30 mm in diameter were heated up successively to the hardening temperature of 860° C and quenched in cold water.

Material: Steel with 0·5 % C content (tensile strength 75 kg/mm²).
Initial hardness H_v: 325 kg/mm².
Input: roughly 10 kW.
Frequency: about 415 kc.

The following table gives the measured maximum hardness values of the four hardening zones:—

	Zone 1	Zone 2	Zone 3	Zone 4
Heating time t_a in seconds	0·5	0·75	1	1·5
Vickers hardness H_v in kg/mm²	782	752	803	782
Thickness of hardened layer δ_H in mm . .	0·5	0·86	1·05	1·45

Hardening of a carbon-steel shaft.

The hardness values and the thickness of the hardened layer are given as a function of the heating time for a given power.

Fig. 3 depicts the arrangement of the single-turn inductor and the shaft within it.

Fig. 4 is a magnified, ground, longitudinal section through zone 3. The hardened steel is represented by the dark semi-elliptical area, the non-hardened core by the remaining white part.

Fig. 5 shows the microstructure of the treated steel, whereby the smooth transition from the martensite to the core structure is particularly noteworthy.

The test proves in the main that increasing the heating period alone only affects the thickness of the hardened layer δ_H, but not the maximum degree of hardness, as, for that matter, is to be inferred from the foregoing table. Variation of the heating time thus forms a means of fixing the thickness of the hardened layer without affecting the degree of hardening. These simple conditions, together with the advantages already alluded to, are further proof of the importance of the surface hardening of steel by high-frequency currents. The method is particularly suitable for the mass production of articles of all kinds, such as gear-wheels, needles, shafts, tubes, etc., since after the plant has been set up it ensures absolutely uniform and faultless quality of the hardening effect, whereby either the whole surface or only parts of it can be hardened to the depth imposed by the purpose in view.

(MS 530) *R. Casti. (E. G. W.)*

Tube Manufacture:

SEALED-OFF TRANSMITTING TUBES AND THEIR PRODUCTION.

Decimal Index 621.396.615.1

Nowadays high-vacuum tubes are the most important constructional element in all transmitting and receiving plants. Their supply is thus one of the main problems in high frequency technique and their further development must be aided by all possible means. Our firm has therefore taken up the manufacture of transmitting tubes. We develop and manufacture sealed-off tubes for our new high-frequency sets and high-power transmitters and as replacements in existing plants. In addition research work on micro-wave tubes is being continued, because these are presumably destined to play an important part in the future.

The production of high-quality transmitting tubes, concerning which this article gives a general description, makes great demands on the scientist and handicraftsman and also requires widespread experience. Very special tools, and manufacturing and checking devices, are necessary for the various constructional components, their assembly, and the treatment of the complete tubes. The various special devices required for tube manufacture, such as vacuum furnaces, smelting devices, pump and test stands, have been made in our workshops from our own designs.

SEALED-OFF tubes consist of an evacuated, or in special cases a gas-filled envelope, internal elements, current bushings and external parts. A fundamental condition for the good operation of normal transmitting and amplifying tubes is an adequate vacuum because the electron streams which flow inside the tubes must not be disturbed by gas ions.

The *tube envelope* is made of a material which does not allow air to pass through and which must fulfil very strict requirements as regards vacuum-tightness. Glass is most suitable for this purpose, it being at the same time a good insulator. Furthermore, due to its transparency, glass enables the internal tube assembly to be controlled and is also favourable as regards the radiation of heat developed inside the tube. The glass must have properties in accordance with the requirements which it must satisfy, i. e. it must be selected with a special chemical composition. For transmitting tubes glasses with a high melting point and of great mechanical and thermal strength are preferably used. These kinds of glasses have low heat expansion coefficients and are not sensitive to rapid changes in temperature. For ultra-short-wave tubes and for such tubes which operate at high temperatures and voltages, special high-quality glasses have to be used which have low dielectric losses also at high temperatures, a small thermal expansion and high softening temperatures. These glasses are expensive to produce and work and can therefore only be used for special tubes. The best glass is pure quartz which, due to the very great difficulty of working it, however, is very seldom used as a vacuum container. In certain cases ceramic material is also suitable for the construction of vacuum vessels.

The current-conducting *internal parts* such as cathodes, grids and anodes, are made of metal or in exceptional cases of metallized insulators or graphite. These materials have also to fulfil very strict requirements so that by no means all metals used as conductors in electrical engineering are suitable for this purpose. Since the materials are in a vacuum it is of primary importance that they can be easily, quickly and as nearly as possible completely degassed. Furthermore, since no heat can be conducted through the vacuum (only heat radiation being possible), practically all electrodes attain high temperatures and must therefore consist of a material with a high melting point.

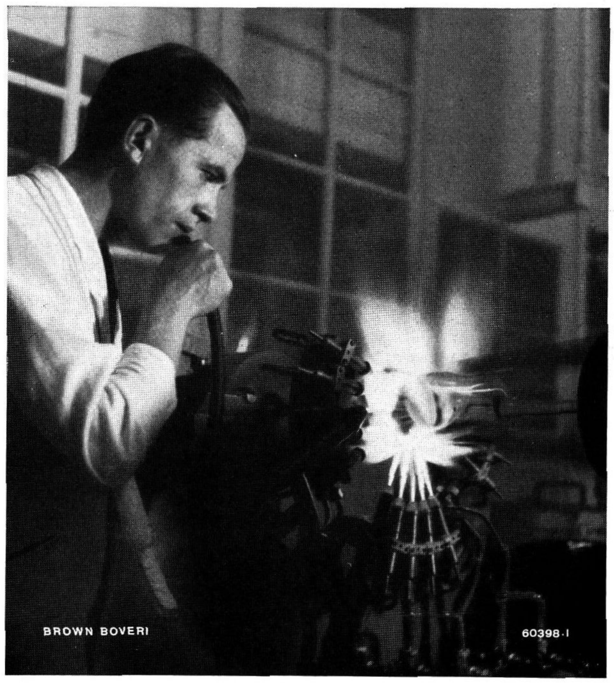


Fig. 1. — Glass lathe for fusing metal anodes to the glass envelopes of high-power transmitting tubes.

In a vacuum all substances vaporize more readily than when under an external gas pressure (atmospheric pressure) so that it is also important that materials with a low vapour pressure should be used. It is also desirable that the selected materials should be good electrical conductors, especially at ultra-high frequencies, good heat radiators, readily capable of being mechanically formed, highly non-oxidizing, and also not too expensive. For parts which are continuously subjected to very high temperatures, for instance above about 2000° C, only pure tungsten, and for somewhat lower temperatures, also tantalum, graphite, molybdenum, platinum, nickel, iron, copper and silver

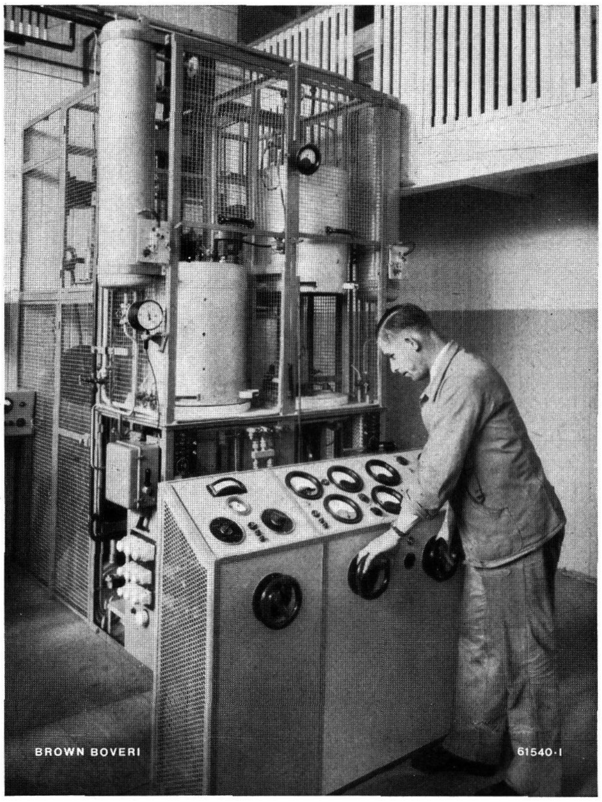


Fig. 2. — Pump stand for high-power transmitting tubes.

In order to degas the individual parts completely, the tubes which are connected to a high-vacuum pump are subjected to various thermal and electrical processes. After degassing, the tube is fused off from the connecting pipe.

are used. All other metals are generally unsuitable for this purpose. Only in special cases can aluminium, zinc, cadmium, tin or lead be used in vacuum technique on account of their low melting points and high vapour pressures.

Numerous metal alloys are also employed. For the wire leads of small tubes made of soft glass, for instance nickel-iron with a copper covering or chrome-iron is used. For hard glasses fused caps of nickel-iron-cobalt alloy in sheet form are provided, and nickel-cobalt alloys have proved to be very suitable as cathode bands for oxide-cathode rectifiers.

Furthermore, materials have been developed which when sprayed on to metallic elements have a very high electron emission at relatively low temperatures. Other substances, so-called getters, have a strong chemical attraction for small gas residues and are therefore introduced into the tubes in order to maintain a good vacuum. In addition there are also various kinds of surface layers which can be provided on the internal elements by spraying, vaporization or burning, these serving to lower or increase the total or partial radiation capacity, or to increase or prevent natural and secondary emissions.

The internal assembly of the tubes also comprises *non-conductors* in addition to the electrical conductors. These serve to support those electrodes which have different electrical potentials. There are only a few really good insulators suitable for use in a vacuum. In small tubes in addition to glass, mica and ceramic materials are always used practically without exception, whilst for large tubes porcelain and quartz are employed.

The *external elements* of the tubes, such as sockets, caps and all kinds of connections, can consist of materials used in electrical engineering, provided their properties are suitable. They do not require to be

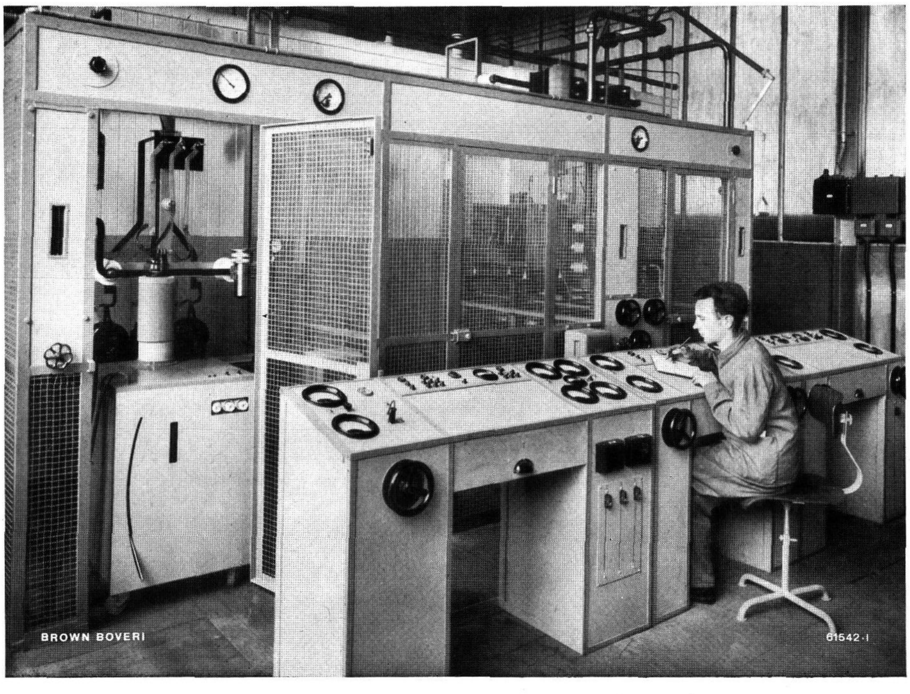


Fig. 3. — Test plant for the static and dynamic testing of high-power transmitting tubes.

Completely evacuated tubes are tested as regards their characteristics and then put into operation in a test transmitter. They are as far as possible tested under the same conditions as in service.

Fig. 4. — Brown Boveri transmitting tube ATL 5—2.

Modern transmitting triode with three phase heating and anode with forced air cooling. Special type for short and ultra-short wave operation. Maximum anode power loss 5 kW. To meet practical requirements the development and manufacture of both small and large tube types have been taken up.

specially suitable for vacuum operation. Socket pins are generally made of copper, brass or aluminium, the insulating layers of bakelite, ceramic material or porcelain, and the sleeves of aluminium, brass or iron.

Great differences exist as regards the *manufacture of transmitting tubes.* For smaller powers the tubes are produced in batches.[1] For this purpose all raw materials as well as all components have always to be examined very carefully during the production process, so that the finished products diverge as little as possible from each other and rejects are reduced to a minimum. Transmitting tubes for larger powers are manufactured more individually, that is by hand. This affords various different preparations and treatment of the raw materials and the manufacturing devices.

Whilst the components for receiving and small transmitting tubes are mass produced by punching and pressing, those required for large transmitting tubes are obtained by careful hand work. The working of the parts provides certain difficulties, especially in the case of tungsten, molybdenum and tantalum.

The *assembly* of the components, i. e. the construction of the tubes, also varies according to the type of tube. With large tubes the individual elements are as far as possible connected together by screwing, riveting, or soldering, in order to be able to salvage expensive components when the tubes become defect.

For the construction of transmitting tubes, especially short wave and high power tubes, it is necessary to employ materials which can stand a high thermal and electrical load so that the power which can be accommodated in a small space is as high as possible. An exact knowledge of the stress limits of the materials is thus particularly important. When new tube types are being planned, attention has to be paid to good electrical and mechanical properties. At the same time the tube must be so designed that it can be easily, simply, safely, and well constructed. When constructing the tube due regard must be paid to the glass and vacuum technique concerned. The glass technique involved is particularly difficult. At all events it must always be possible for the glass operation to be performed without any damage occurring to the metal parts connected to the glass or to metal elements or sensitive materials in the vicinity of the glass. Special care must be taken in this respect with thick glasses. On the other hand high temperature fusing glass and thick glasses can be more highly stressed both thermally and mechanically.

Similar requirements have to be fulfilled by the tubes as regards *vacuum technique.* In order to obtain a high vacuum in the tubes it is not sufficient to evacuate them with a high-vacuum pump. All the various components of the tubes must also be heated during evacuation to enable the tubes to be degassed as completely as possible. To prevent the escape of gas from enclosed pockets during operation it is necessary that the various parts should be heated up to temperatures which are considerably above the operating temperature. The pumping process can be considerably shortened if the individual components are subjected to a preliminary evacuation in special vacuum and hydrogen ovens. The tube components can be heated by various methods, such as direct heating by current conduction, eddy-current heating with high frequency, electron or ion bombardment, by using supersonic waves or also heat radiation. It is, however, not always easy to reach the desired temperature for the various parts without at the same time endangering others. Components which become deformed due to unequal expansion and subsequently cause short circuits in the tubes are to be avoided. Also parts which are near glass or fused to the glass must be carefully heated in order to prevent the glass from cracking or becoming soft. Furthermore care must be taken that the cathode is not spoilt due to a too sudden escape of certain gases from the

[1] The article "Some facts concerning the construction of Brown Boveri small tubes" on page 313 of this number gives a detailed account of this.

Characteristic curves of a transmitting triode ATL 5—2.

The characteristics show that this tube is suitable for most practical applications, both in high and low frequency preliminary and end stages, as well as in industrial plants. The special construction guarantees perfect operation, even with very short waves down to 5 metres.

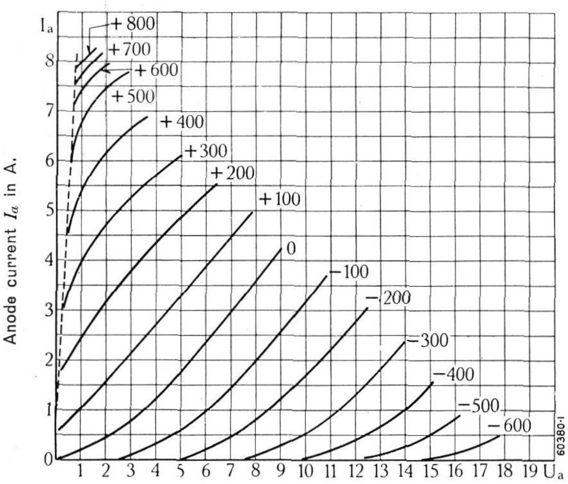


Fig. 5.

Anode current I_a as a function of the anode voltage U_a for various grid voltages U_g.

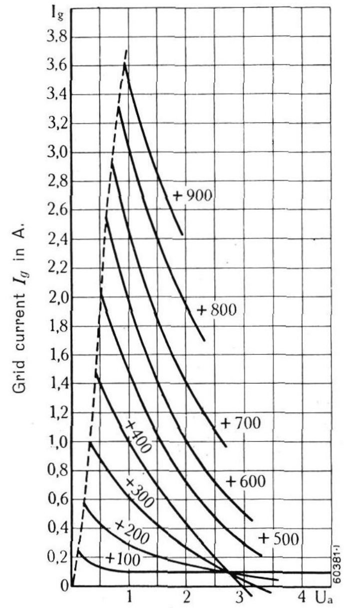


Fig. 6.

Grid current I_g as a function of the anode voltage U_a for various grid voltages U_g.

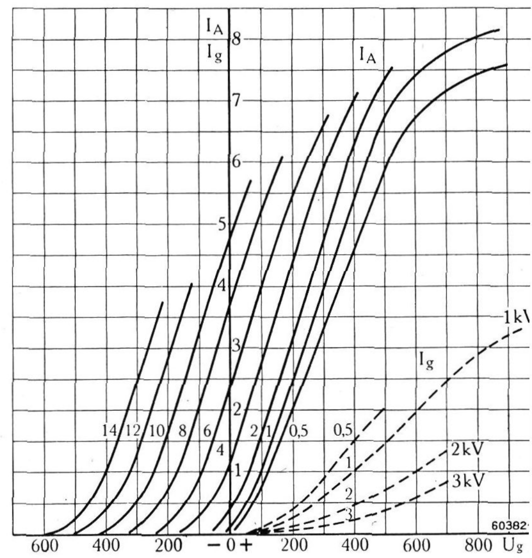


Fig. 7.

Anode current I_a, grid current I_g as a function of the grid voltage U_g for various anode voltages U_a.

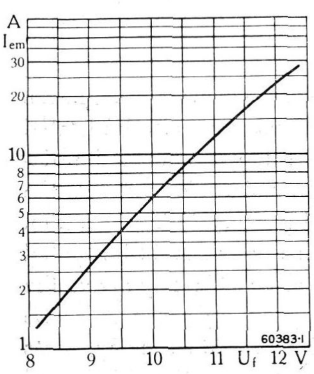


Fig. 8.

Emission current I_{em} as a function of the heating voltage U_f.

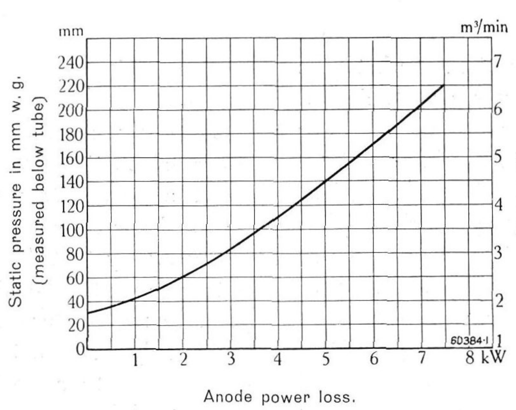


Fig. 9.

Static pressure and air quantity as a function of the anode power loss.

Maximum air inlet temperature = 30° C.
Maximum anode flange temperature = 160° C.

various parts. After the tube has been evacuated it is 90 % finished. The time required for evacuation is an important factor as regards costs. With large transmitting tubes the evacuation is a fairly costly and lengthy operation to which great care must be devoted. The apparatus required for this purpose is expensive.

After evacuation the tubes are fitted with the necessary sockets and leads; thereupon according to the kind of cathode, the cathode is activated, the tube being subsequently stabilized and tested. Activation is necessary in the case of tubes with oxide or tungsten-thorium cathodes, because with these the full cathode emission is only reached after being subjected to certain burning-in or activation processes. The tubes must be subsequently loaded, this occurring mostly in a state of oscillation, in order for them to become stabilized, that is finally to attain their characteristics.

The foregoing description is only intended to indicate how manifold are the operations involved in the construction of tubes and what difficulties may be encountered unless proper measures are adopted to avoid them.

The air-cooled triode ATL 5—2 shown in Fig. 4 is an example of a transmitting tube built by us, its anode power loss amounting to 5 kW. Its characteristic curves are shown in Figs. 5—9.

(MS 560) *F. Jenny. (Op.)*

SOME FACTS CONCERNING THE CONSTRUCTION OF BROWN BOVERI SMALL TUBES.

Decimal Index 621.385.1

We have greatly extended the manufacture of small tubes in view of the great importance of suitable tubes for our wireless sets. After a brief description of our new manufacturing apparatus, the special features of several tube types which are series produced for our directional beam sets are explained.

THE development and manufacture of small transmitting and receiving tubes was taken up in order to be able to equip our sets with tubes of the highest

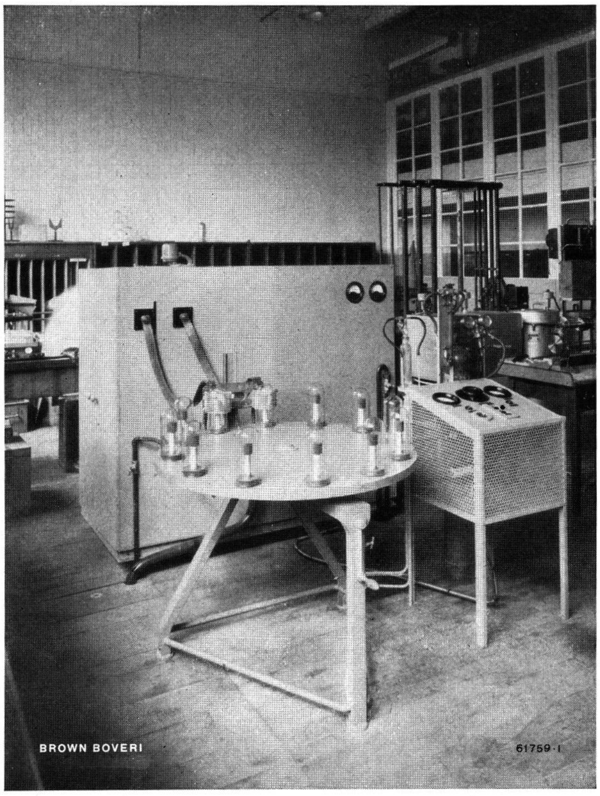


Fig. 1. — Socket sintering machine for manufacturing thin discs from glass powder and unglazed bushings.

The sintering forms are heated with high frequency power.

quality, such tubes being hardly obtainable nowadays. This has enabled us to construct plants of the most modern type, the operation of which requires special novel tubes to enable the requirements as regards amplification, wave length, freedom from noise and the like, to be fulfilled.

In order to accomplish our development and manufacturing programme many different technical and technological problems had to be solved. Special devices required for the manufacturing process, such as filament covering machines, socket sintering machines, grid winding machines, pump stands, high-vacuum ovens, burning frames, measuring apparatus, etc., had to be built according to our own designs. After overcoming considerable difficulties it has become possible to manufacture tubes having first class mechanical and electrical properties. The following account gives some details concerning some manufacturing devices and a few of the tube types constructed nowadays.

Special machines have been constructed for producing the cathodes. Very thin layers which have to be uniformly insulated are applied to the metal by electro-chemical means. The heating filament is electrically insulated from the cathode in such a manner that a good heat transfer between both electrodes is still ensured. Furthermore, by this means the cathode attains the necessary high specific saturation current of 1 A/cm^2 after the tube has been evacuated and formed.

An interesting apparatus is the socket sintering machine for manufacturing thin glass discs which contain the tube leads. The glass discs are made from glass powder by a semi-automatic sintering method which has been evolved in our laboratories. By means of a special arrangement of the electrode terminals in these glass discs with their excellent high frequency insulating properties the capacitances between the electrodes and the inherent inductances of the electrodes are kept low.

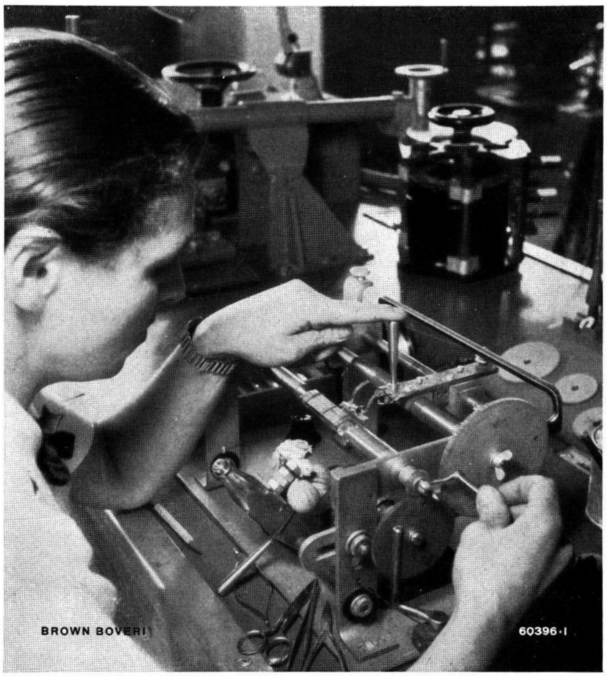


Fig. 2. — Grid winding machine.

This machine permits of the series production of grids from winding wire with a minimum diameter of 0·025 mm.

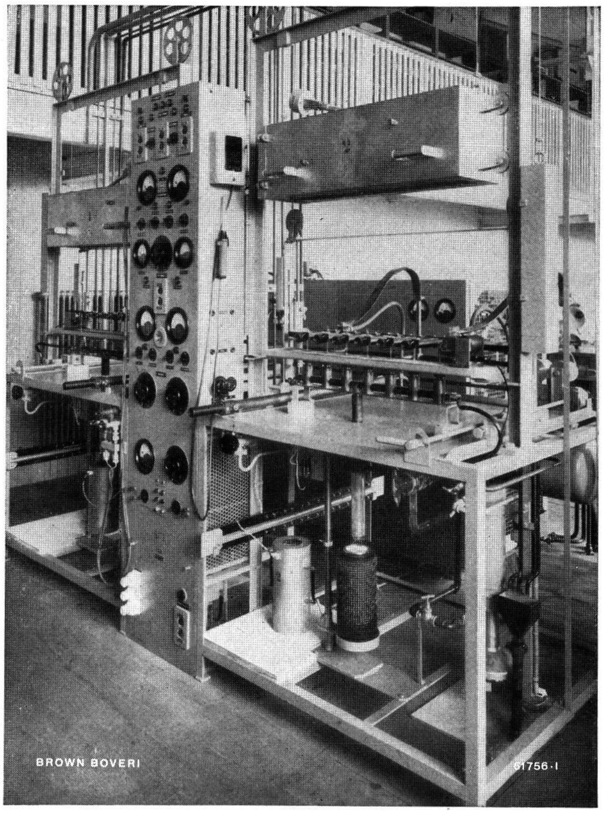


Fig. 3. — Double pump stand for small tubes.

Six tubes are simultaneously formed on each side. The glass formation takes place in resistance ovens; for the metal formation high frequency currents and for the cathode formation low frequency heating is employed.

The production of the internal elements of the tubes and the assembly of the tubes requires special machines and high-precision tools. These enable minimum distances of the order of only 0·1 mm, with tolerances of $\pm$ 0·02 mm, to be maintained between hot electrodes.

The tubes are evacuated and the cathodes activated in accordance with very strict rules, because these operations have a definite effect on the characteristics and life of the tubes.

Very special attention is devoted to the electrical testing of the finished tubes. A series of semi-automatic test stands enable differences in the characteristics of individual tubes to be checked. They also enable each tube to be tested dynamically and mechanically. Those tubes which do not have the required mechanical, static and dynamic properties, are rejected at this stage. The whole check on these test stands involves twenty different testing operations.

In addition to these semi-automatic tests a certain percentage of the tubes are also subjected to exhaustive tests in the laboratory.

There are also endurance tests whereby tubes are subjected to very unfavourable conditions and are kept under observation for very long periods.

In conclusion, some of the small tubes developed in our laboratories are described; these are employed in our decimetre wave sets. It is, however, to be emphasized that in addition to these types other kinds of tubes have also been developed by us. The machines described above are for instance also suitable for the production of our "turbator" tubes.

The particulars given below refer especially to the following tube types: single-plate diode (D 1), decimetre wave transmitting triode (T 3), low and high frequency amplifier pentode (P 2). These three tubes have an indirectly heated cathode for direct or alternating current heating, a series connection also being possible.

A notable feature of these special tubes is the comparatively low cathode heating power required, this being due to the high saturation characteristic of the cathode and the favourable static and dynamic properties of the tube. Another special feature of these tubes is the low capacitance between the electrodes and the small inherent inductance of these latter. It is also of importance that there are no electrode bushings in the tube head, whereby the installation of the tube is simplified. The small overall dimensions of the tubes have a favourable effect on the size of the sets, and the tube socket with its guide pin ensures that it is firmly mounted and can be easily replaced.

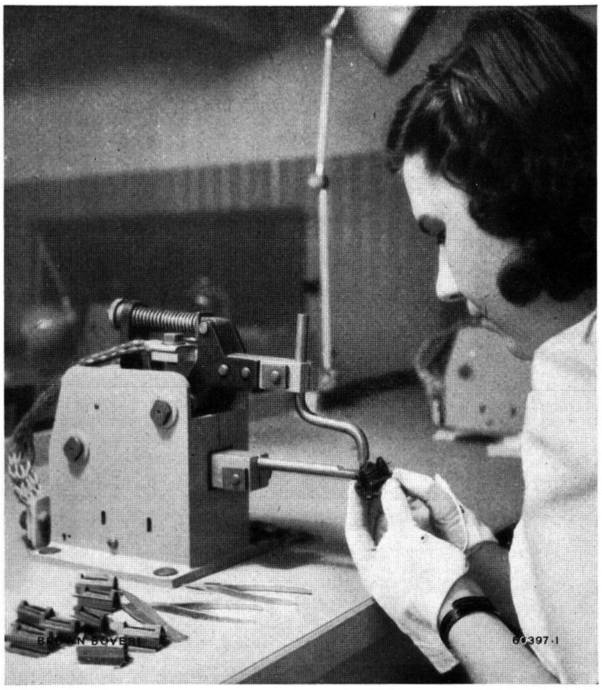
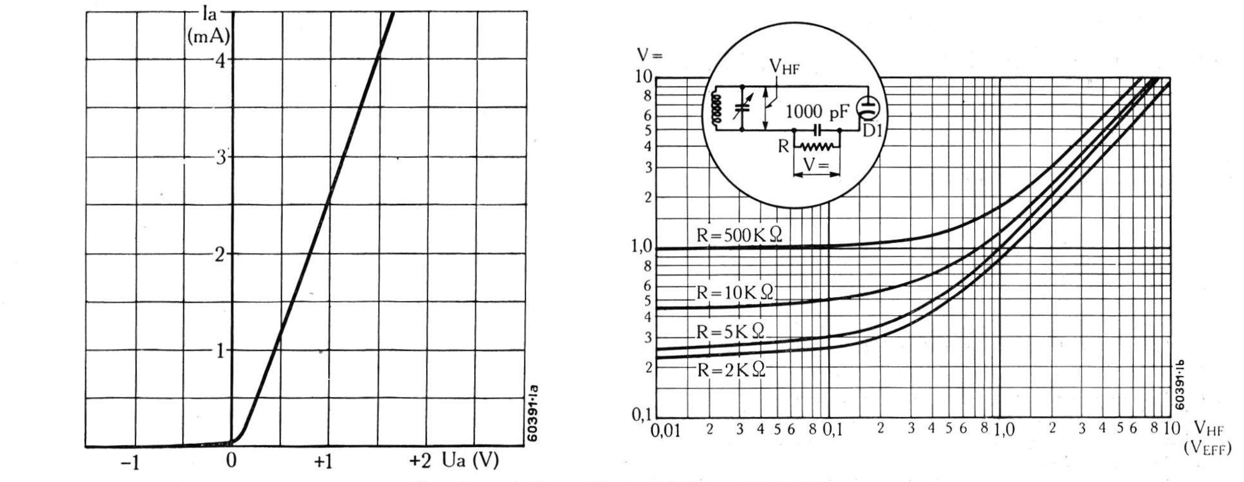


Fig. 4. — Welding machine.

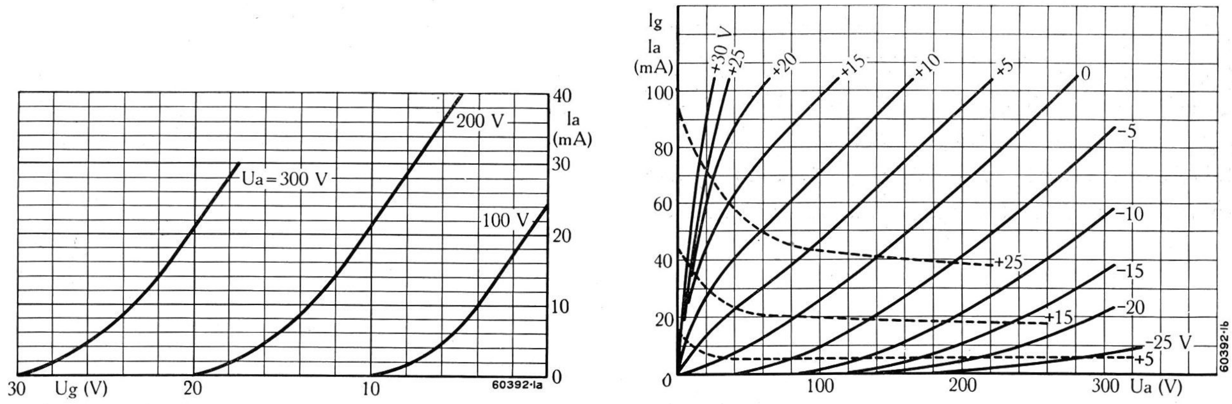
This machine serves for assembling the tube components. The welding currents are automatically interrupted after a very short time.



Figs. 5a and 5b. — Characteristics of diode D 1.

Left: Diode current I_a as a function of the applied d. c. voltage U_a.

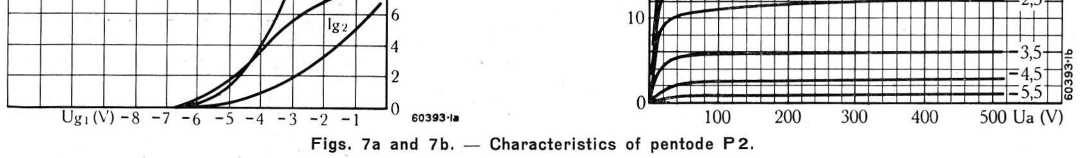
Right: D. C. voltage V as a function of the applied high frequency voltage V_{HF} with various external resistances R.



Figs. 6a and 6b. — Characteristics of triode T 3.

Left: Anode current I_a as a function of the negative grid voltage U_g with various anode voltages U_a.

Right: Anode current I_a and grid current I_g as a function of the anode voltage U_a with various grid voltages U_g.



Figs. 7a and 7b. — Characteristics of pentode P 2.

Left: Anode current I_a, mutual conductance S and screen grid current I_{g_2} as a function of the negative grid voltage U_{g_1} for a given anode voltage U_a and screen grid voltage U_{g_2}.

Right: Anode current I_a as a function of the anode voltage U_a with various grid voltages U_{g_1} and constant screen grid voltage U_{g_2}.

Present-day manufacturing devices enable us to produce high quality tubes for our own transmitting and receiving sets, as well as special apparatus for radio communication and remote supervisory control plants.

The development of our tube construction is being continued systematically to enable all future requirements, especially in the ultra-short wave range, to be fulfilled.

(MS 555)

A. Bertschinger. (Op.)

Radio Engineering Design Elements:

PIEZO-ELECTRIC CRYSTALS AND THEIR USE IN ELECTRICAL WAVE FILTERS.

Decimal Index 537.228.1
621.318.7

The nature of the piezo-electric effect is explained. It is shown by a simple model that piezo-electric crystals can take the place of resonant circuits and thus be employed in filters. Crystals specially adapted to this use may be artificially grown, they even having certain advantages over the quartz used until now; for example, an increased band width can be obtained.

Following the experiments made at the Physical Institute of the Swiss Federal Institute of Technology at Zurich, Brown Boveri cultivate crystals with which filters are built. They are used in the firm's privacy maintaining device, carrier telephone system and for many other purposes.

1. *Introduction.*

UNDER the effect of mechanical strain, electric charges appear on opposite surfaces of a crystal; this is called the piezo-electric effect. It is common to many crystals and was first discovered in 1880, on a tourmaline crystal, by the brothers *J.* and *P. Curie.* They gave a correct explanation of the phenomenon. Later on other piezo-electric crystals, such as sugar, quartz, acetic acid and chlorate of potassium, were discovered. The technical importance attained in recent years by applications of piezo-electricity has led to the systematic research of all substances possessing such properties. For this purpose the method developed by *Giebe* and *Scheibe* in 1925, which does not require well-formed crystals, and may even be used with powdered substances, is mostly employed. Nowadays, piezo-electric crystals have many applications of which a few are: microphones, gramophone pick-ups, standard frequency oscillators (i.e. quartz clocks), pressure measuring apparatus, and in particular as filter elements and supersonic wave generators for submarine telephony.

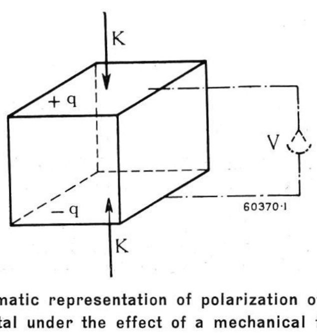


Fig. 1. — Schematic representation of polarization of a piezo-electric crystal under the effect of a mechanical force.

q. Surface charge. K. Force. V. Potential difference.

Under the effect of the force K charges of opposite sign occur on two opposite surfaces.

J. and *P. Curie* found that electric surface charges $\pm q$ appear on electrodes (tinfoil) covering opposite surfaces of a suitably cut crystal parallelepiped, when it is subjected to a uniformly applied mechanical force K (see Fig. 1). q is independent of the crystal dimensions and is proportional to K, $q = d \cdot K$, in which d is a constant, representing the *direct piezo-electric modulus* of the crystal. It is usual to refer the charges and pressure to the unit surface, the charge density thus being expressed by:

$$\sigma = \frac{q}{F} = d \cdot \frac{K}{F} = d \cdot X$$

This density is independent of the length of the crystal as the piezo-electric polarization is a *volume effect.* The crystal is polarized throughout, each crystalline element becoming an electric dipole of moment p, because of the displacement of its ions. If there are n dipoles per cm^3, the specific polarization of the crystal becomes $P = n \cdot p$. According to a fundamental law of electrostatics, however, electrical polarization results in so-called free surface charges of the surface density $\sigma = P$, so that our first equation for the piezo-effect can also be written:

$$P = d \cdot X \quad \ldots\ldots\ldots \quad (1)$$

The polarization produced by an elastic deformation of the crystal, and which is due to the ion displacement in each elemental crystal, may be exactly evaluated in the case of simple structures by atomic considerations.

The piezo-electric effects are very small. An idea of their order of magnitude may be obtained if we consider a cube with sides 1 cm long, suitably cut from a crystal of *potassium phosphate* (KH_2PO_4) and subjected to a pressure of 1 kg. The charge density

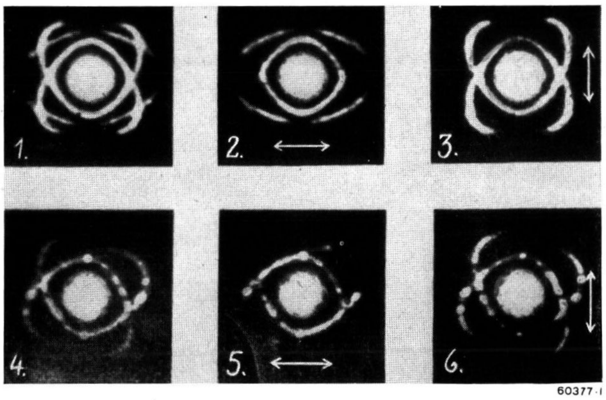


Fig. 2. — Diffraction of quartz in the x and y axes in unpolarized (1 and 4) and linear polarized light (2, 3, 5, and 6), according to Bergmann-Schäfer.

$$V = \frac{\sigma F}{C + C'} = 7{\cdot}85 \text{ V}$$

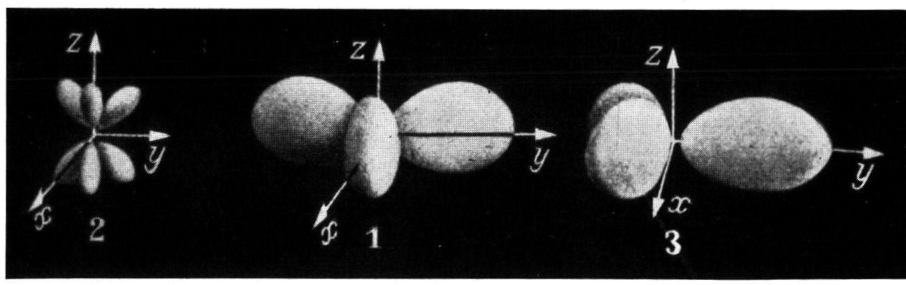
In 1881 *Lippmann* discovered the so-termed *converse piezo-effect,* i. e. that a piezo-electric crystal became deformed when placed in an electrical field. For thermo-dynamic reasons the specific deformation $x = \frac{\Delta l}{l}$ is connected with the strength of the electrical field E by the same coefficient d, i. e. the *piezo modulus* connecting P and X in formula (1):

$$x = d \cdot E \quad . \quad . \quad . \quad . \quad . \quad (2)$$

The converse piezo effects are extremely small, but nevertheless play an important part technically. The very small inherent deformations can be greatly amplified by *resonance* in the case of oscillations or the deformation can be made directly evident by gluing together two thin plates of crystal, like the well known bi-metal strips, with the electrical axes in opposite directions. With the above-mentioned unit cube of

Fig. 3. — Piezo-electric moduli of quartz in space co-ordination.

The charges produced in different directions are measured by the radial vector. 1, 2, 3 refer to different "cuts" of the crystal according to Günther.



which then appears on the electrodes is $\pm 1 \cdot 10^{-10}$ Clb/cm². From this we can deduce the piezo-electric modulus of the substance:

$$d = \frac{\sigma}{X} = \frac{1 \times 10^{-10} \text{Clb}}{\frac{1}{10{\cdot}2} \frac{\text{Joule}}{\text{cm}}} = 1{\cdot}02 \times 10^{-9} \, \frac{\text{cm}}{\text{V}}$$

(The factor $1/10{\cdot}2$ is taken from the relation:

$$1 \text{ kg weight} = \frac{1}{10{\cdot}2} \frac{\text{Joule}}{\text{cm}}).$$

The maximum piezo-electric modulus of quartz is about five times smaller. The voltage between the electrodes of the cube, which are supposed insulated, is $V = \frac{\sigma F}{C}$, C being the capacitance between the electrodes. The high dielectric constant $\varepsilon = 31$ of KH_2PO_4 causes a capacitance

$$C = \varepsilon \cdot \varepsilon_0 \frac{F}{d} = 31 \times 8{\cdot}86 \times 10^{-14} \times \frac{1}{1} = 2{\cdot}74 \, \text{pF}$$

therefore the voltage is:

$$V = \frac{10^{-10} \text{ Clb}}{2{\cdot}74 \times 10^{-12} \text{ Farad}} = 36{\cdot}4 \text{ V}$$

If measured with an electrometer, this voltage would drop considerably; supposing the meter capacitance to be $C' = 10$ pF, the voltage would be reduced to:

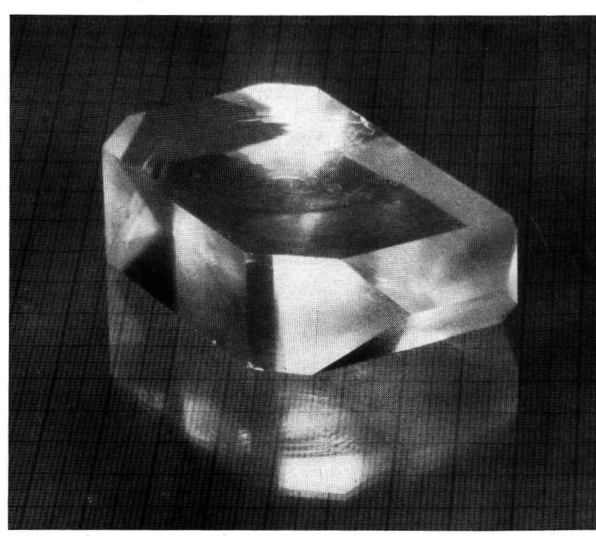


Fig. 4. — Synthetically cultivated potassium phosphate crystal.

Such crystals are used in Brown Boveri crystal filters instead of natural quartz. Since the cultivated crystals have better properties in various respects, filters can be made which would be impossible with quartz.

KH_2PO_4, for instance, the longitudinal deformation produced by a field strength of $E = 1000$ V/cm would only be

$$x = \frac{\Delta l}{l} = d\,E = 1{\cdot}02 \times 10^{-9} \times 10^{-3} = \frac{1{\cdot}02 \times 10^{-6}\ \text{cm}}{1\ \text{cm}}$$

i. e. not even one light wave-length per cm. This slight elongation x could be produced by an elastic stress X on the crystal having (from Hooke's law $X = \xi \cdot x$ and Young's modulus $\xi = 2 \times 10^5 \frac{\text{kg Weight}}{\text{cm}^2}$) a value of only $0{\cdot}2$ kg/cm^2.

The piezo-effect is restricted to crystals with polar axes and deprived of a centre of symmetry. The phenomenological theory of the effect was first worked out by *Duhem, Pockels,* and *Voigt.* In order to investigate these more closely, it must be remembered that the specific electrical polarization P and the applied voltage E are *vectors,* whereas the elastic forces X and mechanical deformations x are *tensors.* Referred to an $x\,y\,z$ coordinate system, therefore, the *three* components $P_x\,P_y\,P_z$ must be given. A condition of stress, however, is described by six components: the three normal stresses $X_x\,Y_y\,Z_z$ perpendicular to the three coordinate planes and the three shear stresses $Y_z\,Z_x\,X_y$ parallel to the axes of the coordinates.

To enable us to fix general equations we must realize that our equations (1) and (2) are incomplete. Although the presence of an elastic force X on a piezo-electric crystal produces an electric polarization P, the crystal is, as any other substance, and in proportion to its electrical susceptibility $\varkappa$, liable to polarization by an electrical field E. Therefore the complete equation is:

$$P = d \cdot X + \varkappa \cdot \varepsilon_0 \cdot E \quad . \quad . \quad . \quad (1')$$

Analogously, in the case of the converse piezo effect, a *deformation* x can be caused not only by an applied voltage E, but also by ordinary mechanical stresses X. Denoting the so-called elastic coefficient, i. e. the reciprocal of Young's modulus s, the general equation of (2) becomes

$$x = s \cdot X + d \cdot E \quad . \quad . \quad . \quad . \quad (2')$$

Given a constant temperature, therefore, a piezo-electric crystal forms a system with four variables. In the present case X and E have been chosen as independent variables from which, when known, the dependent variables P and x can be computed.

In view of the fact that E and P are vectors and x and X tensors the equations for electrical polarization and mechanical deformation unfortunately become fairly complicated:

$$P_x = d_{11}X_x + d_{12}Y_y + d_{13}Z_z + d_{14}Y_z + d_{15}Z_x + d_{16}X_y + \varepsilon_0 \cdot (\varkappa_{11}E_x + \varkappa_{12}E_y + \varkappa_{13}E_z) \quad . \quad . \quad . \quad (1'')$$

$$P_y = \ldots\ldots$$

$$P_z = \ldots\ldots$$

$$x_x = s_{11}X_x + s_{12}Y_y + s_{13}Z_z + s_{14}Y_z + s_{15}Z_x + s_{16}X_y + d_{11}E_x + d_{12}E_y + d_{13}E_z \quad . \quad . \quad . \quad (2'')$$

$$y_y = \ldots\ldots$$

$$z_z = \ldots\ldots$$

$$y_z = \ldots\ldots$$

$$z_x = \ldots\ldots$$

$$x_y = \ldots\ldots$$

In the most general case a crystal is characterized by eighteen piezo moduli d_{ik}, six susceptibilities $\varkappa_{ik}$, and twenty-one elastic coefficients s_{ik}, for which reason the phenomena of piezo-electricity are often difficult to follow. Happily, with most piezo-electric substances this large number of constants is considerably reduced due to the high degree of symmetry of the crystals.

For instance, in the case of

Rochelle salt . . .	only 3 d_{ik}	9 s_{ik}	3 $\varkappa_{ik}$
Potassium phosphate	2 ,,	7 ,,	2 ,,
Quartz	2 ,,	6 ,,	2 ,,

differ from zero.

In simple cases, as for instance with crystal pick-ups, where it is mainly a question of obtaining a high voltage by means of deformation, the mode of operation of the crystal is easy to follow, practically without recourse having to be had to mathematics. In other cases involving, for instance, the excitation of crystals to pre-determined inherent frequencies, evaluation of equations (1″) and (2″) cannot be avoided. It is for this reason that big efforts have been made to measure the constants d_{ik}, s_{ik}, and $\varkappa_{ik}$ for quartz and other crystals. A particularly elegant method of measuring the elastic coefficients is that due to Schäfer and Bergmann, where use is made of the phenomenon of the dispersal of ultra-high frequency sound in crystals to refract light, like a screen. With only three photographs of such refraction spectra it is possible to determine all of the twenty-one moduli of elasticity, the measurement of which would have entailed months of laborious work formerly (Fig. 2).

The category of substances represented chiefly by Rochelle salt and KH_2PO_4 have abnormal high piezo-electric effects. These substances form the electrical counterpart of ferro-magnetic materials. Within certain temperature limits they give rise to a spontaneous electrical polarization, and can be electrically saturated with relatively weak fields, while the phenomenon of hysteresis encountered in the magnetization curve of ferro-magnetic materials is also present in their electrization curve $P\ (E)$.

All known substances of this category have very marked piezo-electric properties; their piezo-electric moduli and dielectric constants depend greatly upon temperature. Whilst the maximum piezo-electric modulus attains $2{\cdot}2 \times 10^{-10}$ cm/V, values up to 800×10^{-10} cm/V have been measured for Rochelle salt and $20{,}000 \times 10^{-10}$ cm/V for KH_2PO_4. In many applications the advantage of a high piezo-electric modulus is partly offset by the fact that the dielectric constant increases with it. However, for use in electrical filters, KH_2PO_4 and $(NH_4)H_2PO_4$ (ammonium phosphate) crystals are eminently suitable. Their quality factor Q is of the same order of magnitude as that of quartz over which they have the great advantage that they can be cultivated artificially.

2. *The Piezo-electric Crystal as Resonator.*

Langevin was the first to produce self-oscillations in a quartz parallelepiped under the influence of an alternating field supplied by a condenser. The mechanical vibrations of the quartz give rise to electric charges on its surfaces, which react in turn on the current taken by the condenser. We thus have *an electrical system coupled to a mechanical one,* which has very numerous technical applications. The impedance of such a resonator is capacitive for certain frequency intervals and inductive for others. The transition point from capacitive to inductive impedance corresponds to a resonance. Compared with purely electrical systems, the piezo-electric resonator has the advantage of extremely sharp resonance, in other words, its damping is very much less than that of equivalent circuits using best quality coils and condensers, and operating at the same frequencies.

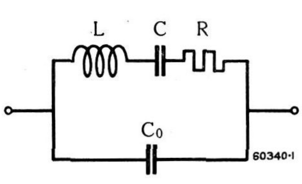


Fig. 5. — Equivalent network of a piezo-electric crystal oscillator.

L. Series inductance. *R.* Resistance.
C. Series capacitance. C_0. Shunt capacitance.

Of interest to an electrical engineer desiring to employ a quartz resonator in connection with other electrical systems is the equivalent electrical network. This is shown in Fig. 5 and is made up of a capacitance C_o connected in parallel to a resistance R, inductance L, and capacitance C in series. For such a circuit there are two characteristic frequencies. At a certain frequency ω_R *voltage resonance* occurs between L and C, when the current will be high. At a second frequency ω_A, *current resonance* arises in the circuit incorporating L, C, and C_o, and it then acts as blocking circuit with the result that the current is low.

A simple model will show the way in which the mechanical vibrations of the quartz are coupled to the external electrical circuit. Consider a condenser between the electrodes of which a sphere of mass m and charge $+q$ is elastically fixed by a spring of stiffness f (Fig. 6). Applying an alternating voltage $U_0 e^{j\omega t}$ to the condenser electrodes will cause forced oscillation of the sphere. When the latter moves towards the right at velocity v, the external circuit is traversed by a current:

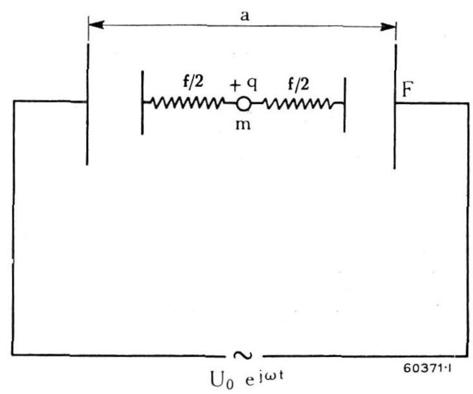


Fig. 6. — Mechanical model of a piezo-electric crystal.

A charged sphere held in a state of equilibrium by springs is free to move in the field of a condenser.

F. Surface of condenser plate.
f. Coefficients of stiffness of springs.
q. Charge of sphere.
m. Mass of sphere.
a. Distance between condenser plates.
$U_0 e^{j\omega t}$. Applied a.c. voltage.

$$I = \frac{q\,v}{a} \quad (a = \text{distance between electrodes})$$

The differential equation of the sphere movement:

$m\ddot{r} + f\,r = \frac{q}{a} U_0\, e^{j\omega t}$ (in which r = displacement)

leads immediately to v:

$$v = j\left[\frac{q}{m\,a}\cdot\frac{\omega}{\omega_0{}^2-\omega^2}\cdot U_0\cdot e^{j\omega t}\right]\cdot \quad \Bigg| \quad \omega_0{}^2 = f/m$$

The current taken by the condenser consists of the current through the unloaded condenser (without the sphere) of capacitance C_o, with the additional current produced by the oscillating sphere

$$J_0 = j\,\omega\left[C_0 + \frac{q^2/m\,a^2}{\omega_0{}^2-\omega^2}\right] U_0 \;\ldots\ldots (3)$$

It will be seen in this case, in which we assume undamped oscillations, that a *maximum of current* is produced for $\omega_R = \omega_0\sqrt{f/m}$, whereas *the current cancels itself* when $\omega = \omega_A$, ω_A satisfying the equation:

$$C_0 + \frac{q^2/m\,a^2}{\omega_0{}^2-\omega_A{}^2} = 0 \quad \text{or} \quad \omega_A{}^2 = \omega_0{}^2 + \frac{q^2}{m\,a^2\,C_0}$$

The difference $\Delta\omega$ between the frequencies at resonance ω_R and at anti-resonance ω_A is calculated by:

$$\frac{\Delta\omega}{\omega_R} = \frac{1}{2}\,\frac{q^2}{a^2\cdot f\cdot C_0}$$

The quantities q, a, f, and C_0 of our model can easily be made to correspond to the characteristic values of a piezo-electric resonator. For a crystal, only the modulus of elasticity ξ, the piezo-electric modulus d, and dielectric constant ε go into the formula giving $\frac{\Delta\omega}{\omega_R}$.

The piezo-electric modulus is evidently:

$$d = \frac{P}{X} = \frac{P\cdot F}{X\cdot F} = \frac{q\cdot x/a}{K} = \frac{q}{a\cdot f}$$

and Young's modulus for our system would be:

$$\xi = \frac{K/F}{x/a} = \frac{K/F}{K/f\cdot a} = \frac{f\cdot a}{F}$$

The dielectric constant ε results from the definition of the capacitance $C_0 = \varepsilon\cdot\varepsilon_0\cdot F/a$. Replacing q, a, f and C_0 in our equations by their values in function of d, ξ and ε, the difference between series and parallel resonances of our model will be given by:

$$\frac{\Delta\omega}{\omega_R} = \frac{1}{2}\,\frac{q^2}{a^2\cdot f\cdot C_0} = \frac{1}{2\,\varepsilon_0}\cdot\frac{d^2\cdot\xi}{\varepsilon}\;*$$

* The constant $\frac{1}{2\,\varepsilon_0}$ characterizing our model depends in the case of an actual crystal resonator on the mode of vibration of the piezo crystal. For an ordinary thickness vibrator, for instance, its value is $\frac{4}{\pi^2\,\varepsilon_0}$ which differs very little from $\frac{1}{2\,\varepsilon_0}$.

The difference $\omega\Delta$ is very important for the use of piezo-electric crystals in electrical filter circuits as it determines the width of the transmitted band.

The impedance value is derived from equation (3).

$$Z = \frac{U_0}{J_0} = \frac{-j}{\left(C_0 + \frac{q^2}{m\,a^2}\cdot\frac{1}{\omega_R{}^2-\omega^2}\right)} \;\ldots (4)$$

Such a condenser incorporating a crystal oscillator has therefore a *capacitive* characteristic between $\omega = 0$ and $\omega = \omega_R$. In the range above the resonance frequency of the crystal, however, the exciting field and elastic waves are opposed in phase, thus resulting in an *inductive* electrical characteristic, although only as far as $\omega = \omega_A$, where the amplitude of the crystal oscillations has again become so small with increasing distance from the point of resonance that the capacitive effect of the condenser again predominates. The reactance characteristic is therefore as illustrated in Fig. 7.

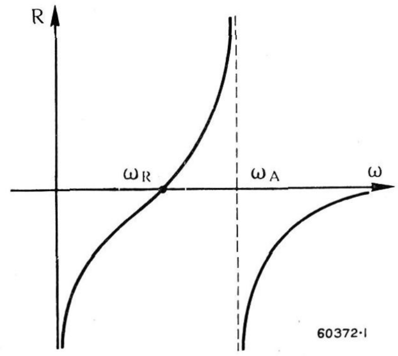


Fig. 7. — Reactance characteristic of a piezo-electric crystal in the vicinity of the fundamental resonance frequency.

The reactance is zero at ω_R and becomes infinite at ω_A. The crystal behaves as an inductance between ω_R and ω_A, otherwise as a capacitance.

R. Reactance. ω_R. Resonant frequency.
ω. Circle frequency. ω_A. Anti-resonant frequency.

3. *Lattice (Cauer) Filter with Four Crystals.*

This filter (Fig. 8) can always be considered as a bridge circuit comprising four impedances, equal two by two (Fig. 9). We shall examine this circuit with regard

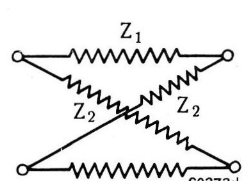


Fig. 8. — Schematic representation of a lattice filter.

Z_1 and Z_2 are each two identical impedances in the series and diagonal arms.

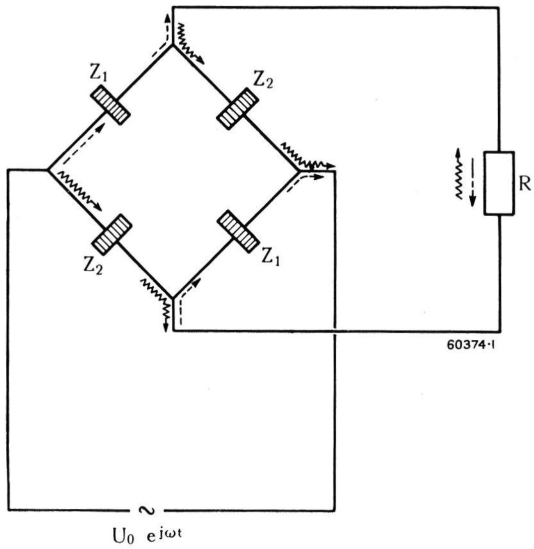


Fig. 9. — Simplified diagram of a bridge filter.

The current through the element R is formed by the two currents flowing through the impedances Z_1 and Z_2, respectively. When Z_1 and Z_2 are of equal magnitude the currents offset each other in the element R.

$- - - -\succ I_1$. $\sim\sim\sim\succ I_2$.

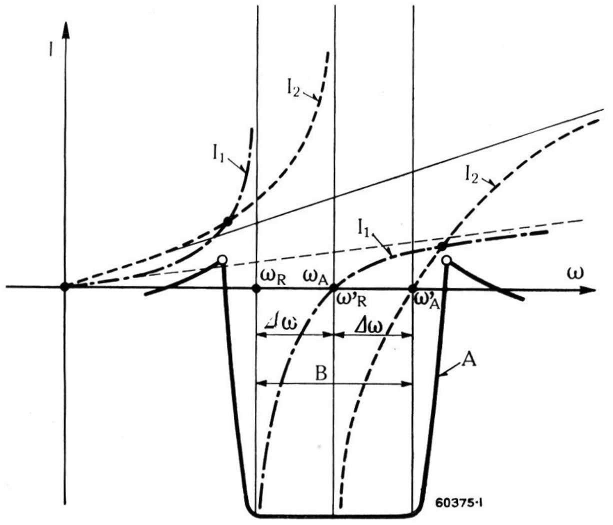


Fig. 10. — Attenuation characteristic of filter.

This is the resultant of the difference between the currents in the impedances Z_1 and Z_2. Each of these currents is given as a function of the frequency by

$$I = \omega\left(C_0 + \frac{q^2 / m\, a^2}{\omega_0^2 - \omega^2}\right)\frac{U_0}{2} \text{ (for } Z \gg R).$$

A. Attenuation characteristic. B. Pass band.

to the current through the resistance R. This current is the difference between currents I_1 and I_2; I_1 passes through the two impedances Z_1 and I_2 through the two impedances Z_2. When Z_1 and Z_2 are of the same sign and have only slightly differing constants, currents I_1 and I_2 have nearly the same value and cancel themselves in R. This corresponds to the attenuated band of the filter. When Z_1 is capacitive and Z_2 inductive, currents I_1 and I_2 are of opposite sign and their difference is large, causing a high current to circulate through R; this is the transmitted band of the filter.

It is evident, therefore, that there is a possibility of designing a filter with a transmitted band extending from ω_R to $\omega_R + 2\,\Delta\,\omega$; for this, Z_1 should be a crystal resonating at ω_R and Z_2 one whose resonance occurs for $\omega_R' = \omega_R + \Delta\,\omega$ corresponding to the anti-resonance of the first one. This condition is easily complied with by the suitable choice of crystal dimensions. The attenuation curve is obtained by subtraction of the currents as shown in Fig. 10. It will be seen that the *maximum band width* of this filter is $2\,\Delta\,\omega$.

The relative difference $\Delta\,\omega / \omega_R$, as we have already seen, is entirely dependent on the properties of the piezo-electric crystal used:

$$\frac{\Delta\,\omega}{\omega_R} = \frac{1}{2\,\varepsilon_0} \cdot \frac{d^2 \cdot \xi}{\varepsilon}$$

For the more commonly used piezo-electric crystals, the values of $\Delta\,\omega / \omega_R$ are:

Quartz	0·36 %
KH_2PO_4 . . .	0·5 %
$(NH_4)\,H_2PO_4$.	4·0 %
Rochelle salt . .	4·0 %

This shows that the two latter substances enable a much wider transmission band to be obtained than the two former. Unfortunately Rochelle salt has a high hysteresis loss and cannot therefore be efficiently used in filters.

We have not so far considered the losses in oscillating crystals, which are represented by resistance R in the equivalent circuit of Fig. 5. These losses are often referred to the quality factor

$$Q = \frac{1}{R}\sqrt{\frac{L}{C}}$$

of the equivalent circuit, and attain quite high figures for Rochelle salt. Whilst quartz and phosphates may have a $Q = 50{,}000$, Rochelle salt rarely exceeds a $Q = 3000$. As normal ambient temperature is quite close to the Curie temperature point for this salt, its constants are very dependent upon temperature. Also, the low mechanical solidity of the crystals and their

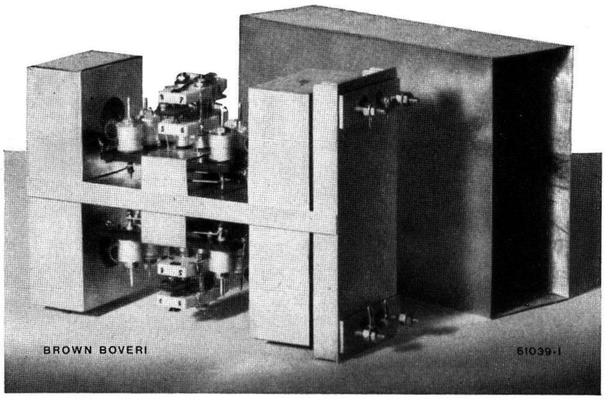


Fig. 11. — Electrical wave filter with potassium phosphate crystals.

Such crystal filters give a better attenuation characteristic than filters built up exclusively of electrical circuit elements, at relatively low cost. They can be constructed for various band widths and for band pass frequencies of roughly 10—100 kc.

Characteristics of filter illustrated:

Width of pass band = 3·4 kc.
Mid-frequency of pass band = 89·5 kc.
Matching resistance = 500 ohms.

lack of stability render Rochelle salt unsuitable for use in filters, where great consistency is required. These difficulties are not found with the phosphates because their Curie temperature point is under — 100° C.

4. *Practical Realisation of Filters.*

Detailed study of the mechanical, dielectric, and piezo-electric properties of potassium and ammonium phosphates has proved that these crystals are eminently suited for use as selective systems. Their very high piezo-electric modulus and moderate value of dielectric coefficient enables the design of wide band filters presenting very high factors of quality.

If the band width obtained is still insufficient, it can be increased by the appropriate adjunction of inductances. With a well chosen circuit this even allows a better cut-off to be obtained, without decreasing the quality factor.

Fig. 11 shows one form of such a filter; its measured insertion loss curve is given in Fig. 12. It is also possible to design filters having a band width of approximately 13 %, for a mean transmitted frequency of 90 kc/s *.

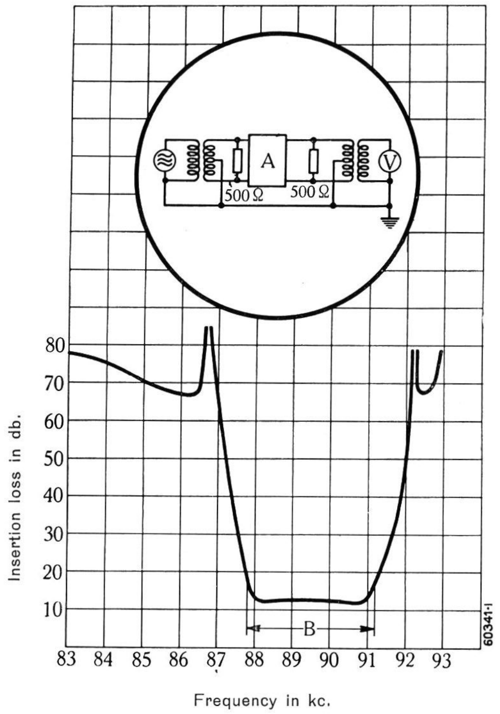


Fig. 12. — Measured insertion loss characteristic of crystal filter illustrated in Fig. 11.

Noteworthy features are the relatively big difference between the insertion loss in the pass band and the attenuated regions, as well as the uniform insertion loss characteristic throughout the pass band.

B. Pass band.

Relative minimum insertion loss = 55 db. on either side of pass region.

In circle: Connection of test set-up with filter *A*.

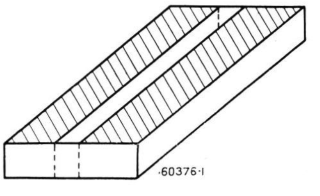


Fig. 13. — Crystal with divided electrodes.

Here we have one mechanical system, but two electrical systems with exactly the same resonance frequency.

The two opposite branches of a bridge filter must have the same reactance curve, which is obtained by using identical oscillating crystals, two by two. A known procedure permits the use of a single crystal with equally divided electrodes, in two branches of the filter, instead of two identical crystals. This gives exact coincidence of ω_R and ω_A in both branches without the necessity of adjusting two crystals, Fig. 13 shows a subdivided electrode crystal. The filter shown has rather special mountings for the crystals; the latter are kept between very fine points, exactly at a nodal point, in such a way that no damping of the mechanical resonance occurs.

(MS 563) *Prof. Dr. P. Scherrer / Dr. B. Matthias.*

* B. Matthias and P. Scherrer, Helv. phys. Acta 16, p. 432.

MODULATION TRANSFORMERS FOR BROADCASTING TRANSMITTERS.

Decimal Index 621.396.619.22

Amplitude modulation of a transmitter may be obtained by various means. For broadcasting transmitters Brown Boveri have adopted anode modulation of the final radio-frequency stage, by a modulation transformer, fed from a Class B audio-frequency amplifier. This system has enabled a high quality of transmission combined with high transmitter efficiency to be obtained. The design of the modulation transformer, which differs from that of a power transformer, set an interesting technical problem which has been favourably solved.

A transmitter is nothing other than a radio-frequency generator, feeding an aerial. The aerial radiates an electromagnetic wave, which travels through space and can be used for conveying a signal. The wave serves as a support for the signal and for this reason is called the *carrier wave.* In order to transmit a signal it is necessary to vary one of the characteristics (amplitude, frequency or phase) of the carrier wave. This operation constitutes the *modulation.*

Amplitude modulation is mostly used. Fig. 1 shows a carrier wave, amplitude-modulated by a sine wave. The following notes give some indications relating to the amplitude and power variations during modulation.

A broadcasting station has to transmit speech and music. The sound vibrations correspond to frequencies between 20 and 20,000 cycles per second, but experience has shown that a musical programme is not adversely affected if the frequency band transmitted is restricted to 30 to 10,000 cycles.

One of the most important criteria of the quality of a broadcast transmission is the faithful reproduction, from the studio microphone to the loudspeaker of the receiver, of the frequency band between 30 and 10,000 cycles. The extraordinary development of broadcasting is marked by the unceasing attempt to attain this quality by improving reception and transmission technique. [1]

Since the early days of broadcasting (around 1920), modulation methods have considerably changed. Primitive systems, such as modulation by absorption, were rapidly displaced by more flexible methods acting on the grid or anode voltage of a transmitting tube. The initially low powers and degrees of modulation were gradually increased. The demand for still greater powers led to the use of a low power modulation stage, followed by one or more Class B [2] linear high frequency amplifier stages, in order to increase the modulated wave to the required level.

Class B amplification is, however, inefficient for low degrees of modulation and this is a great disadvantage with high power transmitters. Nevertheless,

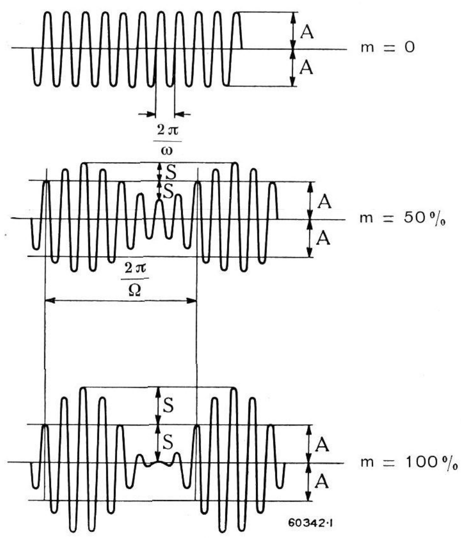


Fig. 1. — Amplitude modulation of a carrier wave by sinusoidal signals.

Amplitude modulation applies to factor A of the expression $a = A \sin \omega t$, representing the carrier wave radiated by the aerial. Superposing the audio frequency sinusoidal signal to be transmitted $s = S \sin \Omega t$ (with $\Omega \ll \omega$ and $S \leq A$) we have now the amplitude:

$$A + s = A + S \sin \Omega t = A\,(1 + m \sin \Omega t)$$

in which $m = S/A$ is the degree of modulation.

Thus the modulated wave corresponds to

$$a = A\,(1 + m \sin \Omega t) \sin \omega t$$

and is represented in the Figure for $m = 0\%$, 50% and 100%. Taking the carrier power as unity, the instantaneous power during modulation is $(1 + m \sin \Omega t)^2$ and varies between $(1 - m)^2$ and $(1 + m)^2$.
The mean power integrated throughout the duration $T = 2\pi/\Omega$ of a modulation cycle is given by

$$\frac{1}{T}\int_0^T (1 + m \sin \Omega t)^2 \, dt = 1 + \frac{m^2}{2}$$

Thus the effective amplitude is:

$$A\sqrt{1 + \frac{m^2}{2}}\,.$$

a large number of transmitters with low level modulation have been built with carrier powers up to over 100 kW, but efficiencies not much above 20 %.

During the last ten years, far more efficient modulating systems have been designed. To illustrate the considerable progress made in this direction, it is worth

[1] The actual importance of broadcasting may be judged from the fact that the annual world consumption of electrical energy for transmitters and receivers, is estimated at 15 thousand million kWh.

[2] Class B operation of an electronic tube is the state in which the grid is negatively biassed to such a value that anode current flows only during the positive half-cycle of the input grid voltage.

mentioning that for the Company's large power transmitters overall efficiencies of up to 40 % are obtained.[1]

It is generally recognized that for large degrees of modulation, the highest quality is given by a modulating system acting on the anode voltage of the final radio-frequency stage (high level modulation). For 100 % modulation, a large power, representing one half of the d. c. anode feed of this stage or approximately 75 % of the carrier power, is required. Such power may now be obtained, economically and with good fidelity, by means of a symmetrical class B audio amplifier. The overall efficiency of the transmitter is then at least as high as with other known high efficiency modulating systems.

This modulating method has been adopted for the broadcasting transmitters marketed by the Company, because it ensures high quality, good efficiency, ease of operation, and stable adjustment. Moreover, since it does not complicate the radio frequency circuits, it is applicable to short wave as well as medium and long wave transmitters.

Fig. 2 shows the theoretical circuit for *anode modulation* of the final transmitter stage, by a class *B* modulator. The radio-frequency stage 2, feeds the aerial 3; it receives its grid excitation from the preceding oscillator and amplifier stages marked 1. The symmetrical modulator stage 6 comprises tubes *a* and *b*. The alternating audio-frequency voltage V_g applied between the grids of these tubes is provided, through an appropriate chain of audio amplifiers 5, by the telephone line 4, connected to the studio microphone. Tubes *a* and *b* feed, in turn, their respective half-primary of modulation-transformer *T*. The secondary winding of *T* applies the audio-frequency voltage *V* in series with the d. c. anode voltage V_a of modulated stage 2. The d. c. anode current of this stage is supplied through choke-coil *L* and is prevented from passing through the secondary winding of *T* by the condenser *C*, which closes the circuit for the modulation frequencies[2].

[1] For a 100 kW transmitter operating 15 hours daily, the efficiency increase from 20 to 40 % results in an annual saving of approximately 1·5 million kWh. The installed power also decreases from 500 to 250 kW.

[2] Without this precaution, the d.c. current would saturate the magnetic circuit of the transformer and seriously reduce its effective permeability, causing a corresponding reduction of the frequency band correctly transmitted. This effect can be reduced if an air-gap is provided in the core, but, even with the optimum gap, the reduction is noticeable and such a transformer would be unsuitable for a broadcasting transmitter. On the other hand, if we wish to transmit a restricted frequency range (for example 200 – 3000 cycles for commercial telephony) it can be used to advantage.

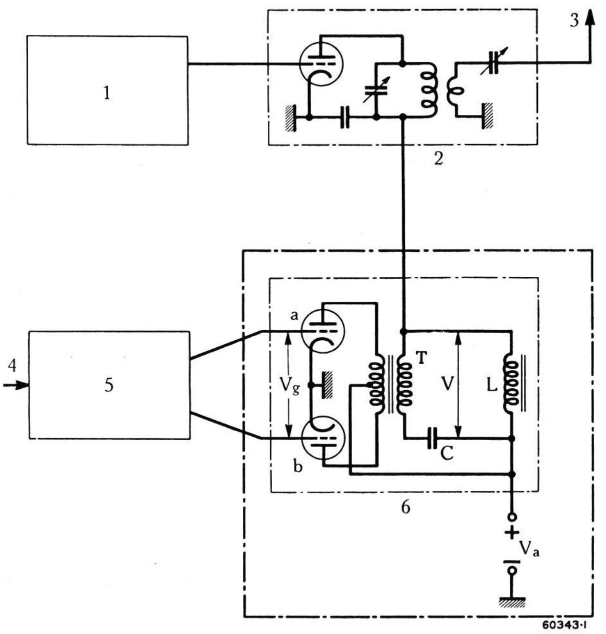


Fig. 2. — Theoretical circuit for anode modulation of a final transmitter stage by a class B modulator.

This is the system used for Brown Boveri transmitters. It gives high efficiency and excellent quality up to the highest degree of modulation. Excited from preceding stages 1, the h. f. output stage 2 feeds the aerial 3. The audio-frequency voltage V_g, coming from input line 4, through the amplifier 5, drives the grids of modulation stage 6. This stage feeds the modulation transformer *T*, the secondary winding of which adds its audio frequency voltage *V* to the anode d. c. voltage V_a of the output stage 2.

1. H. F. preliminary stage.
2. H. F. output stage.
3. Antenna.
4. A. F. input.
5. A. F. amplifier.
6. Modulator stage.

a, b. Tubes of class B modulator stage.
T. Modulation transformer.
L. Iron choke-coil for by-passing anode d. c. current of h. f. output stage.
C. Isolating condenser for d. c. voltage, which closes the circuit for A. F.
V_a. Anode d. c. voltage of h. f. final stage and modulator.
V_g. A. F. voltage between modulator grids.
V. A. F. output voltage of the modulator.

The modulator stage must satisfy the following requirements: It must supply its maximum power *P*, for 100 % modulation, with an a. c. output voltage *V* equal in amplitude to the d. c. anode voltage V_a of the modulated stage 2. For a modulation degree *m*, the a. c. voltage will be $V = m V_a$ and the corresponding power $= m^2 P$. For power levels between zero and *P*, and frequencies between 30 and 10,000 cycles, the modulator stage must ensure faithful amplification of the input voltage V_g.

In fact, due to the non-linear characteristics of the amplifier tubes and of the iron magnetization curve, a certain amount of *amplitude distortion* is introduced. For a perfectly sinusoidal input voltage V_g of frequency *f*, this distortion is characterized by the appearance, in the output voltage *V*, of harmonic frequencies $2f$, $3f$, . . . nf. The percentage of the

n^{th} harmonic gives the corresponding distortion d_n, whilst the total distortion is given by:

$$D = \sqrt{\sum_{2}^{n} d_n^2}$$

For a given input voltage V_g of varying frequency, the output voltage V will change in sympathy with the variation of impedance of the various circuit elements. The curve relating V to f is the *response curve* of the modulator stage. It is usual to plot

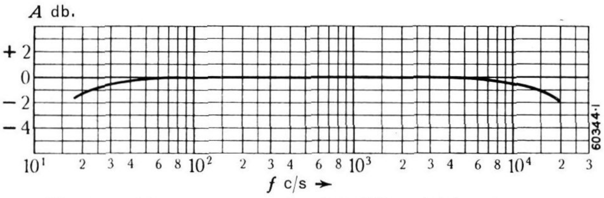


Fig. 3. — Response curve of a 8·5 kW modulator stage.

This curve shows, for a given input voltage V_g of variable frequency f, the relative value of the output voltage given in decibels

$$A = 20 \log V/V_{400}$$

V_{400} being the output voltage at 400 cycles. This curve indicates a loss of only 0·5 db (6 %) at 30 and 10,000 cycles and may be considered as very good. The theoretical circuit of the modulator stage is given in Fig. 2, while Fig. 4 shows the modulation transformer.

this curve with a logarithmic scale for f, as horizontal axis, and the relative value of V expressed in decibels, that is 20 log V/V_{400} as ordinate, V_{400} being the output voltage at 400 cycles (see Fig. 3). The phase displacement between input voltage V_g and output voltage V is closely related to the shape of the response curve. Correct transmission of phase relationship is essential to achieve satisfactory reproduction of transients.

Values for the various circuit elements are determined by a survey of the mode of operation of the modulator stage. This leads, particularly for the *modulation transformer*[1], to the following considerations:—

It must have a sufficiently large primary inductance in order to maintain the magnetizing current, and hence the voltage drop, at a low value for the lowest frequency it is desired to transmit. Leakage inductance and distributed capacitance must also be small in order to prevent excessive losses at high frequencies. Finally, to limit amplitude distortion, the maximum flux density should remain moderate; since for a given power, the voltage and therefore the product "flux × frequency" remain nearly constant, maximum flux density is produced at the lowest frequency for which full power is still applied.

Operating conditions of a modulation transformer compared with those of a normal power transformer, show a marked difference. A power transformer operates at fixed voltage and frequency and is fed from a source of very low internal resistance; it delivers a power quite close to its nominal power and of which the variations, generally quite small, are entirely due to those of its load resistance. A modulation transformer, on the other hand, is fed from a source of very high internal resistance at constantly changing frequency and voltage. With a constant secondary load resistance its output varies very greatly, the mean value being of the order of 10 % of the maximum value. Heating problems which are important in the first case are only of a subordinate nature here.

The construction of a modulation transformer capable of transmitting all frequencies between 30 and 10,000 cycles without distortion, is a very different and more complicated problem, than that of a power transformer operating at a fixed frequency.

The law of induction, which as physical principle applies in both cases, is not sufficient to determine the dimensions of a transformer. Where power transformers are concerned the design data are dictated by economic considerations (proportioning of iron and copper losses and prices). In the case of modulation transformers, transmission requirements, i. e. desired response curve and maximum allowable distortion, are determinant for the values of inductance, electric and magnetic leakages and flux density, with the result that the design is entirely dependant on technical considerations.

Some relations of proportionality existing between the different characteristic sizes of a transformer are shown, for both cases, in the following table; in which:

P = Power.
B = Flux density at frequency f_n and power P.
f_n = Nominal frequency for flux density B and power P.
μ = Permeability.
σ = Current density.
f_{max} = Highest operating frequency.
f_{min} = Lowest operating frequency.

The values shown in the table are the exponents for the quantities figuring at the head of each column; i. e.:

$$G_f \sim P^{3/4} B^{-9/4} f_n^{-3/2} \sigma^{3/4}$$

for a power transformer, and:

$$G_f \sim P B^{-2} f_n^{-2} f_{max}$$

for a modulation transformer.

[1] Our remarks and conclusions may be applied to audio-frequency transformers in general.

TABLE.

Laws of growth of power and modulation transformers.

	Power transformers				Modulation transformers				
	P	B	f_n	σ	P	B	f_n	σ	$f_{\max}$ or $\mu f_{\min}$
Weight of iron G_f	$^3/_4$	$-\,^9/_4$	$-\,^3/_2$	$^3/_4$	1	-2	-2	0	1
Weight of copper G_c	$^3/_4$	$-\,^1/_4$	$-\,^1/_2$	$-\,^5/_4$	$^2/_3$	$-\,^1/_3$	$-\,^1/_3$	-1	$-\,^1/_3$
Ratio G_f/G_c	0	-2	-1	2	$^1/_3$	$-\,^5/_3$	$-\,^5/_3$	1	$^4/_3$
Iron losses %	$-\,^1/_4$	$-\,^1/_4$	$-\,^1/_2$	$^3/_4$	0	0	-1	0	1
Copper losses %					$-\,^1/_3$	$-\,^1/_3$	$-\,^1/_3$	1	$-\,^1/_3$
Turns per volt	$-\,^1/_2$	$-\,^1/_2$	0	$-\,^1/_2$	$-\,^2/_3$	$^1/_3$	$^1/_3$	0	$-\,^2/_3$

Concerning the part of the table relating to modulation transformers, the following should be noted: f_n is a frequency situated in the lower part of the transmitted range ($f_n \geq f_{\min}$), for which the flux density B, fixed to correspond to power P, is limited by the maximum allowable distortion. Minimum and maximum frequencies f_{min} and f_{max} are fixed by the desired response curve. The ratio $f_{\max}/f_{\min}$ is directly proportional to the permeability μ of the magnetic circuit.

From the several interesting conclusions that can be drawn from the table, we will only examine the following two:—

1. The ratio G_f/G_c between the iron and copper weights, which is independent of the power for power transformers, increases with the cubic root of the power for modulation transformers. For large powers, the latter have only a small proportion of copper. For a power of 100 kW the ratio G_f/G_c may become 100. The magnetic circuit of such a transformer, compared to that of an ordinary one[1] will have quite unusual proportions.
2. Minimum iron weight is obtained by choosing for $f_{\max}$ the lowest and for f_n the highest value possible. As the lowest frequencies are rarely present and are usually of small amplitude, f_n may be taken higher than $f_{\min}$ without fear of introducing objectionable distortion. Raising f_{max} entails a proportional increase in the weight of the core. $f_{\min}$ increases at the same rate, unless higher permeability stampings can be used, because $f_{\max}/f_{\min}$ is proportional to μ.

The design of high-performance modulation transformers calls for consideration of many other factors, such as: symmetry of both halves of the primary winding, amount of leakage inductance and distributed capacitance, operation during the transition period between the two modulator valves, iron and copper losses at high frequencies, etc. . . .

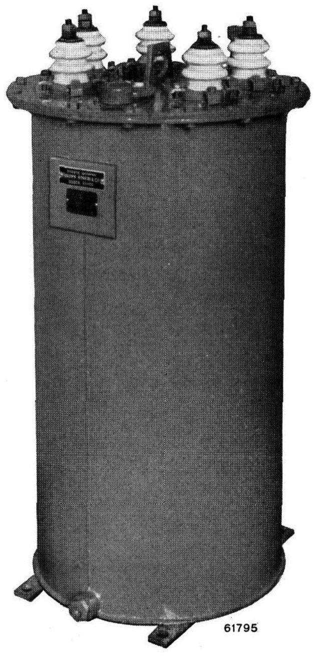


Fig. 4. — 8·5 kW modulation transformer.

This transformer gives a uniform transmission of the frequency band between 30 and 10,000 cycles (see Fig. 3). Despite its small size (casing diameter 520 mm and height 880 mm) it has an efficiency of 95 %, and gives only 1·5 % distortion (at 50 cycles and full power).

The Company's great experience in the design of power transformers has enabled various modulation transformers, to be developed which satisfy the foregoing conditions, and have given excellent operating results. By a new shape of magnetic circuit and a particularly favourable winding disposition, perfect symmetry has been realized without the use of an electrostatic screen between windings, and a great reduction in size obtained.

Fig. 3 shows the response curve of an 8·5 kW modulator stage using a Brown Boveri transformer. The curve indicates a loss of only 0·5 db at 30 and 10,000 cycles and may be considered very good. Fig. 4 is a photograph of the transformer; despite its small size (casing diameter 520 mm and height 880 mm) it has an efficiency of 95 %; the distortion at 50 cycles and full power is only 1·5 % and this decreases rapidly for higher frequencies.

Research, checked and completed by tests, has provided full data concerning modulation transformer theory and technique. These transformers can now also be supplied for all applications, from low to very high power. Naturally, the choke-coil L which diverts the d. c. current can also be supplied.

Complete modulator stages, for the Company's own transmitters or for modernizing existing low efficiency transmitters, are available.

(MS 536)

M. G. Favre.

[1] For which G_f/G_c is between 3 to 5.

NEW METHOD OF IMPEDANCE MATCHING IN RADIO-FREQUENCY CIRCUITS.

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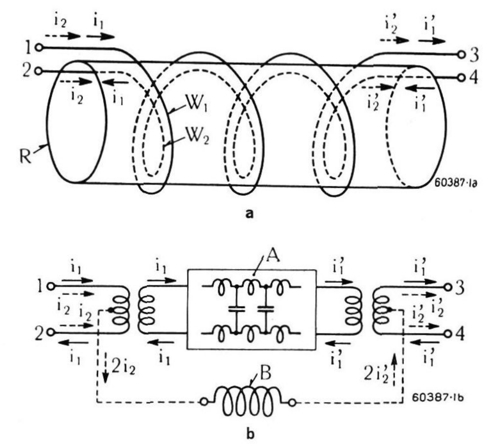
A new transformer method is described which is suitable both for matching circuits of unequal impedance and coupling symmetrical and unsymmetrical radio-frequency circuits. In contradistinction to conventional methods of impedance matching the frequency of the oscillations being transmitted can be varied over a wide range without the necessity of re-tuning.

THE impedances of the individual circuits of radio-frequency equipment are frequently unequal. In order to obviate the reflections and losses involved by mismatching, special matching devices have to be inserted between such dissimilar circuits for the transmission of energy. For instance, matching is necessary between the tubes of a transmitter output stage with high load resistance and the low-impedance antenna transmission line or feeder system. In the case of low frequencies transformers with a corresponding turns ratio can be employed. By reason of the unavoidable leakage inductance of the coupled transformer coils, high frequencies generally involve tuning by means of additional condensers, and should the working frequency be varied, corresponding re-tuning is therefore entailed.

For impedance matching purposes a quarter-wave Lecher wire system having a surge impedance which is the geometric mean between the two impedances to be matched can likewise be employed. Such matching sections must naturally also be re-tuned in the event of the frequency being altered, to correspond to the changed wave-length. Small frequency deviations are, however, permissible when the impedance transformation takes place in several steps adjusted to the mean frequency. — Another method of matching, the line with exponential taper, permits large frequency variations without re-tuning, but has amongst other things the drawback of taking up a large amount of space.

Special couplers are also necessary for transition from symmetrical to unsymmetrical circuits, e. g. between the symmetrical output of a push-pull transmitter stage and a coaxial antenna cable with earthed sheathing. Here, too, variation of the frequency generally involves re-tuning.

A new coupler which obviates re-tuning is shown in Fig. 1a. It comprises two superposed windings W_1 and W_2 separated by an insulating tube R. Given symmetrical currents i_1 (full-lined arrows) the magnetic fields produced by two closely-spaced superposed sections of conductor practically neutralize each other, i. e. the mutual inductance of two successive turns of a coil can be neglected, while it is possible to replace the two windings by two straight conductors having the same cross-section, length, and spacing as the two developed windings. This Lecher wire system is represented in the equivalent diagram (Fig. 1 b) by the equivalent line A.



Figs. 1a and 1b. — Double-wire coil system with equivalent diagram.

(a) The coil system comprises two superposed windings W_1 and W_2 separated by an insulating tube R.

(b) According to this equivalent diagram, where symmetrical currents i_1 are concerned, the coil has the effect of a Lecher wire system A, but with unsymmetrical currents i_2 the nature of a choke coil B. The symmetrical and unsymmetrical currents are segregated by ideal centre-tapped transformers.

On the other hand, with unsymmetrical currents i_2 (dotted arrows), the field vectors produced by two superposed sections of the conductors are added together, with the result that the mutual inductance between the individual turns of the coil becomes an important factor. The double-wire coil system behaves here like a conventional choke coil, represented in the equivalent diagram by B. In this diagram the symmetrical and unsymmetrical currents i_1 and i_2, respectively, are segregated by centre-tapped ideal transformers T. Given an adequate number of turns on the windings W_1 and W_2 the impedance of the equivalent choke coil B becomes so high that, even assuming unequal potentials between the centre tappings of the input and output coils, the unsymmetrical current i_2 can be neglected. In this case the described coil system forms an ideal transformer combined with an ideal line.

In view of the effect of this ideal transformer such a system S can now be employed, as shown for example in Fig. 2a, to couple a physically symmetrical circuit (connected to terminals 1 and 2) to a load resistance R_a having one pole earthed. By making the coil of suitable dimensions the surge impedance Z_o of the matching line (A in the equivalent diagram Fig. 1 b) represented by the coil system can be adapted to the pure load resistance R_a. In this case the input impedance R_e occurring between terminals 1 and 2 is equal to the surge impedance Z_0 and in consequence also to the load resistance R_a, immaterial of the actual working frequency.

The curves in Fig. 3 give the input impedance computed from the coil dimensions for conditions of short circuit and no-load. The measured impedance values are also given and agree with the curves to a high degree. These measurements, which demand great care, were made by a method specially developed for the purpose (cf. Fig. 1, page 293). The characteristic surge impedance can be determined from the geometric mean of the measured or computed short-circuit and no-load input impedances. In the present case it is about 240 Ω. Fig. 4 gives the curve of the input impedance for a load impedance of about 53 Ω. From the test points it is clear that the desired impedance transformation in the ratio of 1 : 16 is actually possible over a very wide frequency range. The deviation of the plotted mean-value curve R_e from the theoretical curve 1 is due to the load

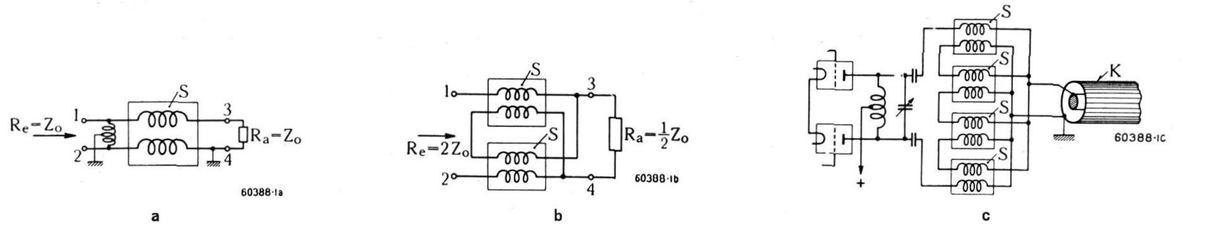


Fig. 2 a, b, and c. — **Employment of double-wire coil systems for coupling and impedance matching purposes.**

(a) Due to the suppression of the unsymmetrical currents by the series inductance of the coils such units can be used for coupling physically symmetrical circuits (connected to terminals 1 and 2) to circuits having one pole earthed (connected to terminals 3 and 4).

(b) By series-parallel connection of two coil systems S the load resistance $R_a = {}^1/_2\, Z_o$ is transformed to the input impedance $R_e = 2\, Z_o$ (Z_o = surge impedance of a coil system).

(c) The antenna cable K and output stage are "matched" by the four coil systems S.

Surge impedance of coil systems = 240 Ω.
Surge impedance of cable = 240 Ω : 4 = 60 Ω.
Load impedance of output stage = 240 $\Omega \times 4$ = 960 Ω.

By series-parallel connection of two or more coil systems impedance matching is now also possible in a simple manner, independent of the frequency. Fig. 2b shows by way of example the input terminals of two systems of coils S connected in series and the output terminals in parallel. No objections can be raised to this practice provided the series inductance (B in the equivalent diagram in Fig. 1 b) is large enough. The load resistance $R_a = {}^1/_2\, Z_o$ is thus transformed to the input impedance $2\, Z_o$. Analogously, with n coil systems impedance transformation in the ratio $1 : n^2$ can be achieved.

In Fig. 2 c, for instance, four coil systems are shown connected between a transmitter output stage and the high-frequency antenna cable K, the resulting impedance transformation being in the ratio $4^2 : 1 = 16 : 1$. With a coil system having a surge impedance $Z_0 = 240\ \Omega$, for example, a transmitter output stage with a load impedance of $4 \times Z_0 = 960\ \Omega$ can be coupled to an antenna cable of $Z_0 : 4 = 60\ \Omega$. The coupled coil systems have the same effect as a transformer with separate windings, i. e. the symmetry of the anode circuit at the input end is not affected by single-pole earthing of the cable connected to the other end. Furthermore, the coupled coil systems behave like a Lecher wire system, i. e. the input impedance must follow a tangential function of the frequency when the terminals at the other end are open or short-circuited.

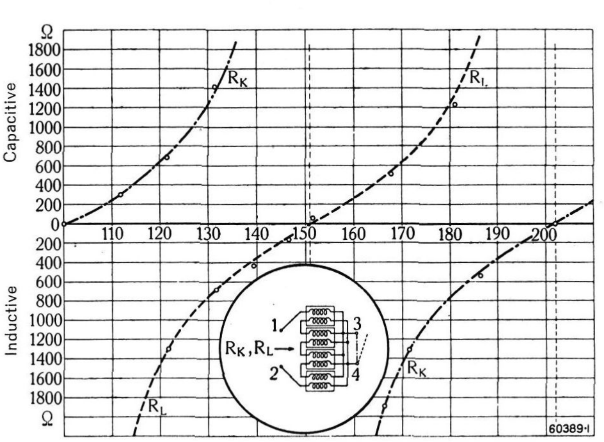


Fig. 3. — **Input impedance of matching unit when output terminals short-circuited or open.**

The matching unit comprises four double wire coils in series-parallel connection. The computed and measured primary impedances are plotted as a function of the frequency with the secondary terminals open and short-circuited.

—·—·— / — — — } Computed curves. o. Test points.

R_K. Input impedance with secondary terminals short-circuited.
R_L. Input impedance with secondary terminals open.

impedance being slightly lower than the theoretical value, as well as to the inherent capacitance of the circuit.

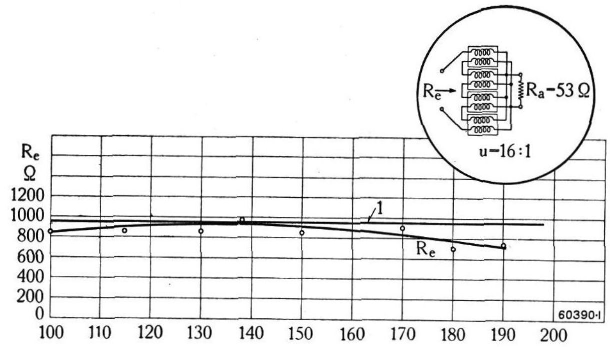


Fig. 4. — Theoretical and measured input impedance of a matching unit with a pure resistive load.

The matching unit comprises four double-wire coils in series-parallel connection. A surge impedance of 240 Ω was computed from the coil data and the measurements in Fig. 3, whence, assuming a pure resistive load of 60 Ω, the theoretical value of the input impedance is 960 Ω. The measured values of the input impedance are somewhat lower owing to the load impedance having been somewhat lower than theoretically assumed.

Theoretical value of $R_a = Z_o$. o. Test points for $R_a = 53$ Ω.
Impedance transformation ratio 16:1.

The described method of matching is particularly suitable for application in the ultra-short-wave field, where it represents a big simplification compared to conventional tuned matching devices. Fig. 5 shows the external appearance of an impedance transformer with four coils, employed as antenna coupler in a

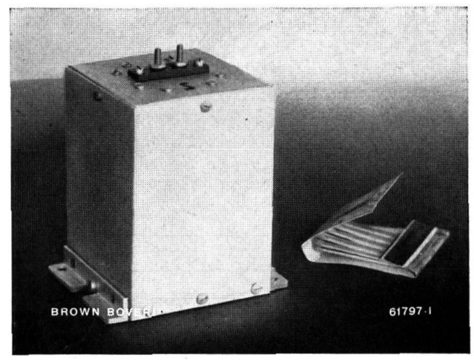


Fig. 5. — Matching unit with double-wire coils.

The system contains four double-wire coils for impedance transformation from 60 Ω to about 1000 Ω in the case of metre waves. With a power of over 100 W the losses are negligible.

medium-power transmitter. It requires little space and its losses are very low. This new component greatly simplifies the construction and operation of the equipment marketed by the Company.

(MS 564) *G. Guanella. (E. G. W.)*

Electric Filters built up from Choke Coils and Condensers for Frequencies up to 60 kc.

Decimal Index 621.318.7

ELECTRIC filters are employed for the discrimination or segregation of individual frequencies or frequency bands. For frequencies over 60 kc crystal filters are now finding favour, whereas in the range below this figure coils and condensers are practically exclusively used as filter elements. The most commonly employed types are the low, high, and band-pass, so-called from the different positions of the pass and rejection regions in the frequency range. On the score of selectivity present-day demands are very exacting, that is:

1. In the pass frequency band as uniform a response to pass frequencies as possible with minimum attenuation.
2. Virtually complete rejection of the frequencies outside the pass region.
3. Transition from pass to rejection regions with as short a frequency interval as possible (sides of attenuation curve very steep, see Fig. 2).

High-class filters can only be built up from good quality components, i. e. coils and condensers with small loss angles and large time stability factors. Since suitable coils could not be found on the market, the Company took up the manufacture of moulded coil-cores of powdered iron[1]. As a result filters of first-rate quality are

Fig. 1. — Band-pass filter for 2800—3200 cycle range.

Dimensions 11 × 17 × 5 cm. Such filters are built up from coils with Brown Boveri powdered-iron cores (μ = 50) and mica condensers. The attenuation curve in Fig. 2 testifies to their high quality.

[1] See page 331.

now available. Fig. 1 shows a filter built up from condensers and coils with Brown Boveri powdered iron cores and Fig. 2 the attenuation curve of the filter.

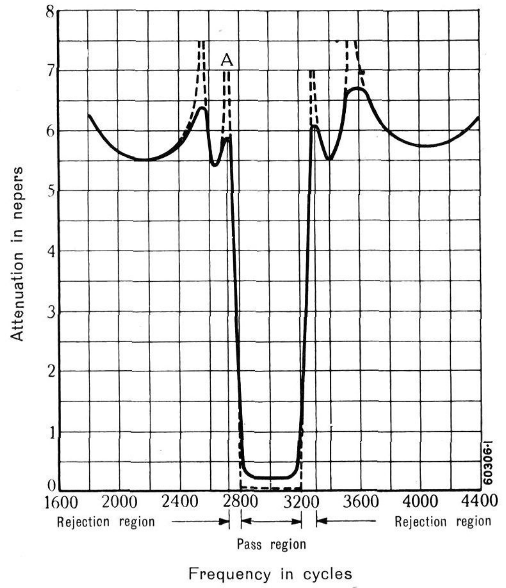


Fig. 2.— Attenuation characteristic (working attenuation) of band-pass filter with coils having Brown Boveri powdered iron cores, depicted in Fig. 1.

Band-pass range 2800—3200 cycles. The dash-lined curve shows the attenuation characteristic of an ideal (loss-free) filter. Due to the use of high-class components the drift of the actual curve from the ideal characteristic is relatively slight. A = Attenuation pole.

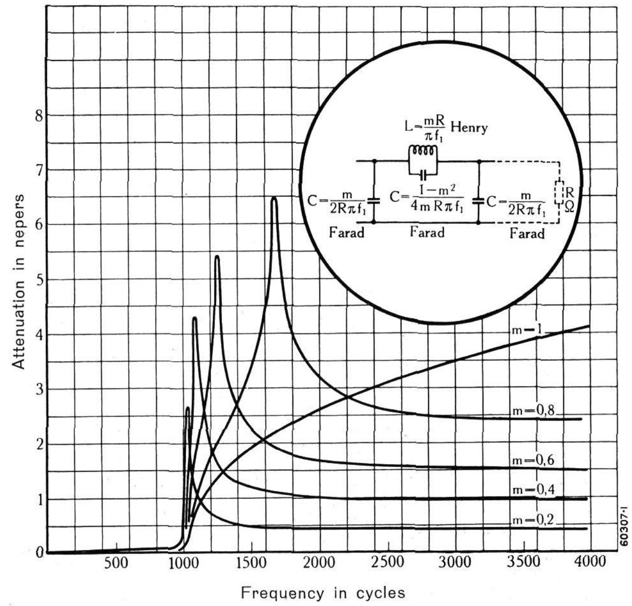


Fig. 3. — Computed attenuation curves for low-pass filters.

Similar attenuation curves have also been plotted for the more common types of high- and band-pass filters.

Pass range of low-pass filter 0—1000 cycles.
Coil loss angle tan $\delta = 0{\cdot}01$.

The filter elements can be calculated from the above formulæ.

f_1 = Cut-off frequency (1000 cycles).
R = Load resistance.
m = Parameter (according to the value of m the different attenuation curves are obtained).

The *computation of filters* presents no particular difficulties, but is very laborious, especially when the actual or working attenuation is required to be known, i. e. the coil losses have to be taken into consideration. The drift of the actual attenuation curve from the ideal characteristic of a loss-free filter circuit in the pass region, at the extremities of the pass range and at the attenuation poles, can generally not be neglected (see Fig. 2). To facilitate the design of filters charts in the form of attenuation curves have been compiled from the results of lengthy calculations, for the more common low-, high-, and band-pass filter structures. An example of such a chart is shown in Fig. 3 for a low-pass filter. Chains of filter cells can thus be readily assembled to obtain the desired attenuation characteristic.

In order to select the most satisfactory filter from the multiplicity of filter circuits available for the solution of any one attenuation problem the following points should be observed:—

In the first place the coils and condensers must have reasonable constants, that is, constants with which they can be manufactured in a good quality and economically. When choosing *coils* care should be taken that the main resonant frequencies occurring in the filter lie around the maximum factor of merit $\frac{\omega L}{R}$ of the coils[1], which can be obtained by suitably selecting the permeability of the core of the coil.

[1] See Fig. 4, p. 331.

Another point to be observed is not to allow the flux density in the coil cores to be too great, since the hysteresis loss angle increases approximately linearly with the coil voltage. In the case of high coil voltages this results in a poor filter attenuation curve and distortion of transmission through the pass region. With certain types of filters the voltage across the coils can attain 1000 V or more when even only a few watts are being transmitted.

The hysteresis constant of coils with powdered iron cores drops with diminishing permeability of the core. In consequence, coils with cores having a low permeability should be employed for filters through which power is to be passed.

Since the Company now disposes of powdered iron cores with permeabilities $\mu = 10 - 60$ the most favourable coils for any frequency and voltage can be selected.

The capacitance of the *condensers* should not be too high since good (mica and styroflex) condensers of more than $0{\cdot}1\ \mu$ F are both costly and bulky. Other factors determining the types of filters to be employed are surge impedance characteristic, time constancy of elements, and economy.

The foregoing filters are employed in the equipment for multi-channel telephony, apparatus for ensuring the privacy of speech, remote supervisory control gear, etc. marketed by the Company.

(MS 534)

K. Ehrat. (E. G. W.)

Brown Boveri Powdered-iron Cores for Filter and Tuned Coils in Communications Engineering.

Decimal Index 621.3.042.15

CHOKE coils, piezo-electric crystals, and condensers are employed in modern communications engineering filter circuits. At the present stage of development of the art, however, the factor-of-merit or Q of coils is still

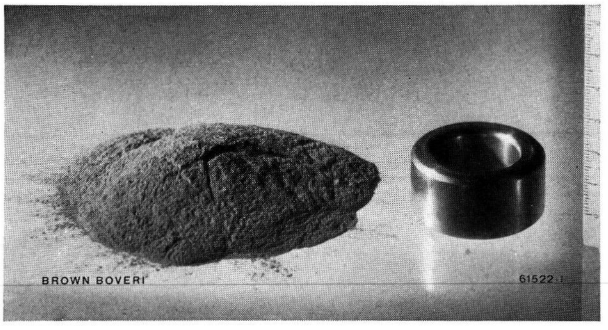
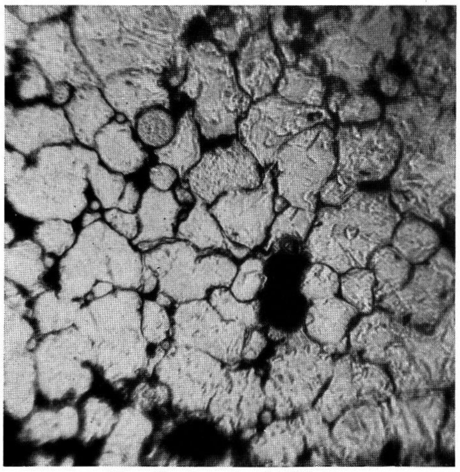


Fig. 1. — The moulded core.

Left: The "raw" material: special iron powder of a definite grain size and of which the individual particles are electrically insulated from one another by a coating of special material. Right: The finished core moulding suitable for a frequency range of 1—100 kc according to the material and mixing ratio employed.



Magnification about 750 ×. 60303·I

Fig. 2. — Metallograph of ground-surface specimen of an annular core.

The iron powder referred to in Fig. 1 is moulded under high pressure. As a result the particles become heavily deformed, but the layer of insulation around them remains unimpaired. Permeability $\mu = 40$.

far below that of condensers and crystals. Nevertheless, choke coil filters can be built to satisfy exacting requirements if high-class powdered-iron cores, such as have been developed by Brown Boveri, are employed. The accompanying illustrations testify to the high quality of the types manufactured by the Company on mass production lines. A large number of powdered-iron cores have been embodied in the firm's multi-channel telephone equipment, apparatus for ensuring the privacy of speech, and remote supervisory control gear, with excellent results.

(MS 528) *E. Ganz. (E. G. W.)*

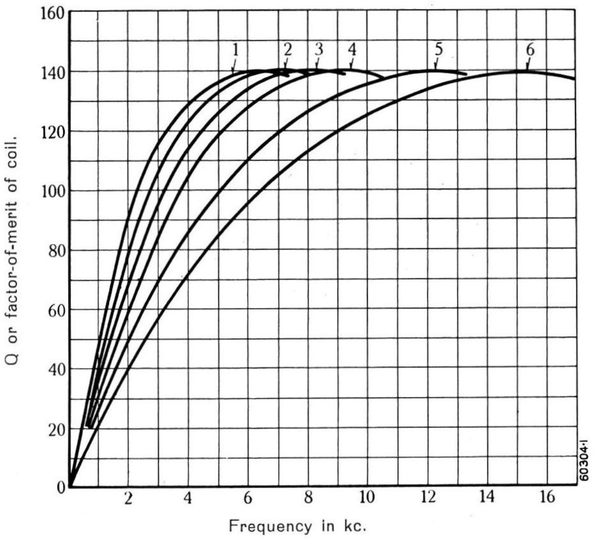


Fig. 3. — Factor-of-merit or Q of various annular types of cores.

The Q of the coils is plotted as a function of the frequency with a current of 1 mA through the coil and a core of 16 cm^3 volume and 120 g weight, the number of turns remaining constant. A notable feature is that even for such low frequencies as in the 3000—20,000 cycle range coils of extremely small dimensions have a high Q.

Curve	Self-inductance in mH	Permeability μ of annular core
1	105	50
2	80	45
3	70	40
4	57	30
5	42	22
6	34	15

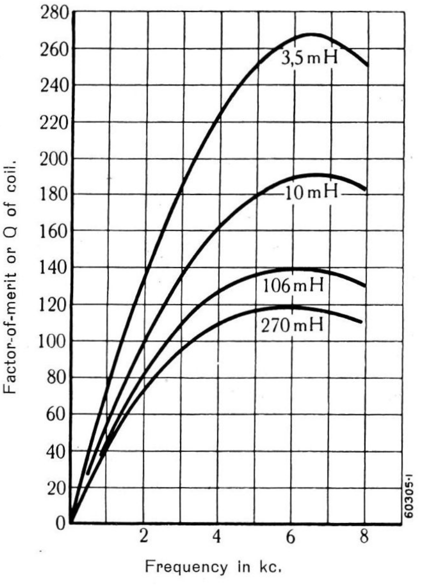


Fig. 4. — Factor-of-merit or Q for different numbers of turns.

The curves show the influence of the number of turns and, in consequence, of the self-inductance on the Q of an annular core of only 16 cm^3 volume with a permeability $\mu = 50$. Current through coil 1 mA. In order to improve the factor-of-merit Q for higher self-inductances a higher permeability or a larger core volume must be selected.

BROWN BOVERI

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