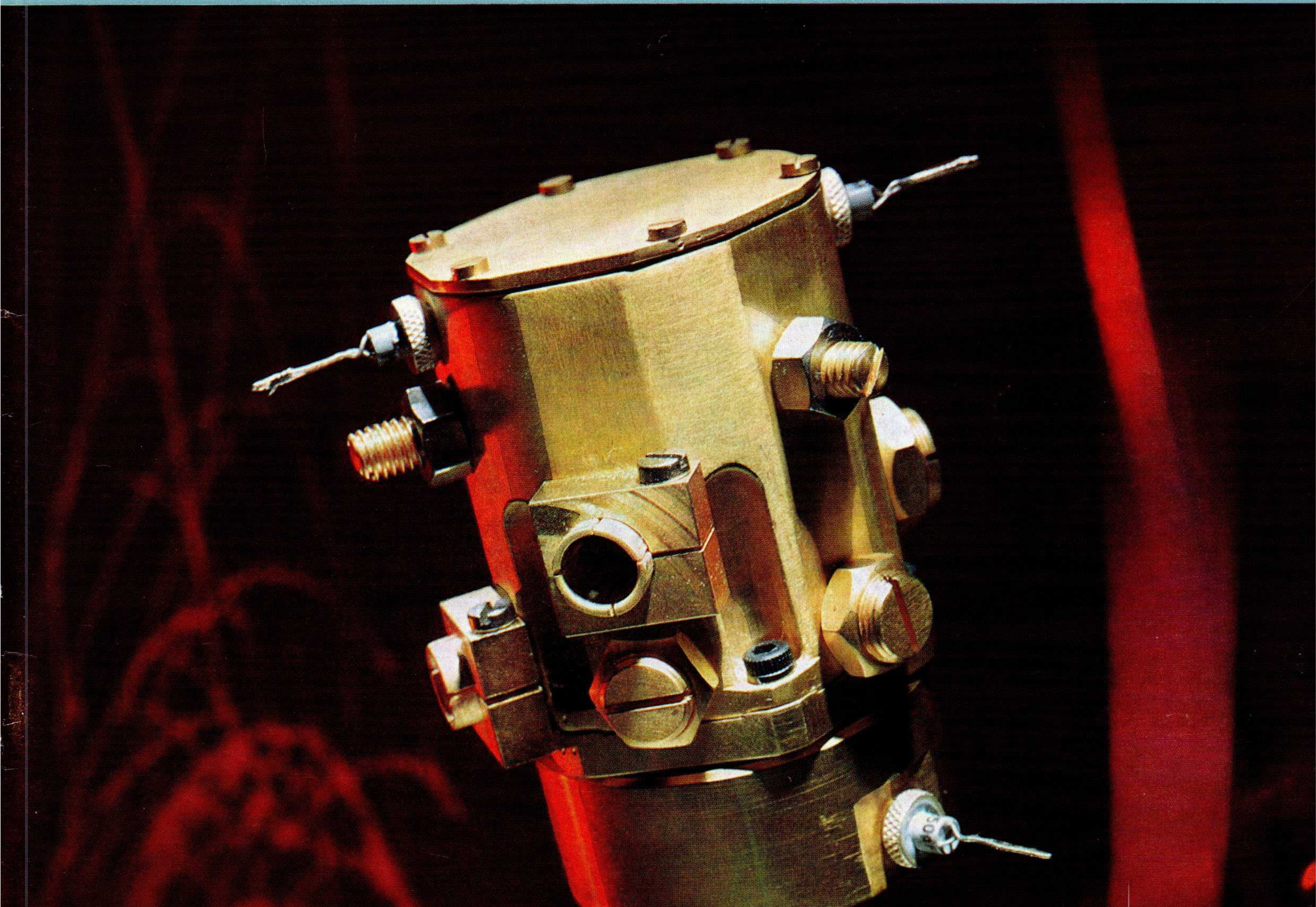
SIEMENS

Siemens **ELECTRONIC COMPONENTS** Bulletin



Vol. I - DECEMBER 1966

5-66

News at a Glance

The 1966 edition of the "Transistor Circuit Manual" is the latest issue of the "Green Book" series already well known for a number of years to many circuit designers. An extensive chapter on a-f amplifiers includes a typical application of the 3-stage monolithic circuit a-f amplifier TAA 111. In addition to a large number of control and regulating circuits with photodiodes, thermistors, PTC resistors and Hall-effect devices as signal sources, the chapter on radio and tv circuits also deserves special mention. This contains a low-noise uhf tuner equipped with the new germanium mesa transistor AF 239, a diode-tuned vhf tuner, an a-m/f-m i-f amplifier without neutralization and many other useful circuits for radio and television applications.

The new phototransistor BPY 61 considerably extends the range of applications covered by Siemens photoelectric devices. This particular device is a silicon planar transistor system with a miniature glass encapsulation designed for frontal illumination. With no connection to the base of the transistor system, only the collector and emitter wires are led out. The main advantage of this phototransistor is a sensitivity almost 100 times higher than that of photodiodes or photoelements. Such a high sensitivity is, however, associated with wider spreads in light current and a lower cutoff frequency. Typical applications for the phototransistor BPY 61 are photoelectric readout from binary-coded storage disks and in conventional light barriers.

Typesetting computer for the Netherlands

A Siemens-Hell typesetting computer was placed into operation by Gelderlander Pers N. V. Nijmegen in the fall of 1966. The syllabication program for the Dutch language was developed by Siemens Den Haag. This typesetting computer simplifies typesetting work in the same manner as the systems used by German publishing houses: insertion of display type, quadding, automatic centering when setting up advertisement type, optional use of up to 16 type fonts and automatic line justification for all desired column widths.

Multiple diodes simplify circuit design, especially when many diodes of the same type are used, for example in computer gate circuits. The Siemens diode BAX 28 contains 3 high-speed silicon planar diodes with a common cathode. The electrical characteristics of each diode approximately correspond to those of our catalogued diode BAY 60.

The new silicon planar transistors BC 147, BC 148 and BC 149 have electrical characteristics similar to the a-f preamplifier types BC 107, BC 108 and BC 109. The new devices differ from the latter types in that they have a plastic encapsulation in place of the metal case. The connecting lead geometry conforms to the standard T0 5 outline. The leads are flat wires designed in such a way that the transistor can be inserted easily into printed circuits and remains in position without requiring additional measures until dip soldered. This design not only permits economic manufacture but easy and cost-saving circuit assembly as well.

From the French CCT (Comité de Coordination des Télécommunications), the Siemens Components Division has received type approval for the SIFERRIT pot core types B 65 541 (14 × 8), B 65 651 (18 × 11), B 65 661 (22 × 13) and B 65 671 (26 × 16) in material N 22. The certificates are registered under the numbers 66-61 to 66-63 and 66-65. The approved SIFERRIT pot cores in the standard material N 22 which conform to the German standard DIN 41 293 and French specification CCTU 06.04 are used primarily for filter coils in communications equipment. They are currently available at short notice from stock.

"Expo 67" in Montreal. For the "Expo 67", which is to be held in Montreal, Canada, from April 28 to October 27, 1967, Siemens has received a contract for the installation and lighting equipment in the German Pavillion, for floodlighting the stadium (capacity 50,000 spectators) and for the floating fountain system in the "Lac des Nations."

Contents

SIFERRIT, SIMIBLOCK are trademarks

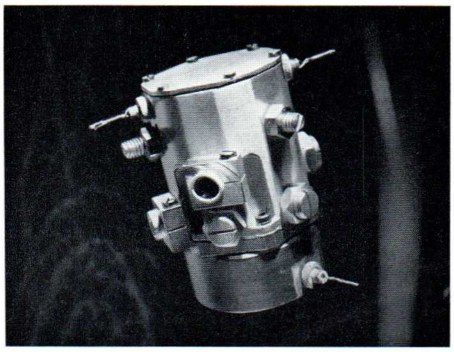
Published by the Siemens Aktiengesellschaft · Berlin · München

Responsible for contents: D. Rost, München

Editors: E. Pollak, Erlangen
and R. Grumbt, München

Inquiries should be addressed to: Siemens Aktiengesellschaft
Schriftleitung "Siemens-Bauteile-Informationen",
8000 München 25, Hofmannstrasse 43, Germany
Telex 05-24721

Printed by Süddeutscher Verlag, München, Germany



Cover: This 20-watt single-stage amplifier embodying a disk-seal triode RH 5 C was developed with design emphasis placed on low weight. A power/weight ratio of 10 gm/watt with a total of 200 gm for the amplifier (gold-plated aluminum body 45 mm in diameter and 70 mm in height) was here realized. The gain at 2.3 Gc is 13 db, the bandwidth 20 Mc and the efficiency about 35 %. The RH 5 C emerged from the RH 7 C-c used in Mariner IV. Having almost twice as large a cathode surface, however, it is capable of handling a higher load, has a higher output power, and its service life is about twice as long.

The Electron Tube – A Component with a Future

In many important branches of electronics there is no substitute for the electron tube, and progress here is probably dependent upon further tube development.

Prof. Fritz Paschke, Tube Development Manager of the Components Division of Siemens Aktiengesellschaft, was questioned by Dr. Metschl about the prospects for the future of electron tubes. The problem is first considered from a general viewpoint and subsequently a few questions pertaining to physics are discussed.

Prof. Paschke, one frequently hears it said that electron tubes will soon be seen in museums only. Is this really the case?

That is by no means correct. According to an American market investigation, an annual growth rate of 5 to 7 % is anticipated for the electron tube industry at least until 1969. This forecast envisages that even in the field of the classical grid-controlled amplifier tube for radio and television very attractive turnovers will be achieved. This particularly applies to exports. For instance, due to economic considerations, the transistorization of television receivers has not progressed so far in the United States as it has in West Germany.

What is the situation in the special field of color television?

For color television the power required for the deflection is about twice that necessary in black-and-white receivers. The tubes necessary for this purpose have already been developed and they will help to stimulate tube sales.

In addition, there is no doubt a range of other tube types with an assured future, isn't there?

I would like to divide the tubes with an assured future into three large groups: the first includes all power tubes, i. e. transmitter tubes for communications and industrial transmitting tubes; the second can be combined under the designation "electron-beam scanning tubes", and the third group can be referred to as "special components using tube technology".

You have mentioned transmitter tubes. Is it not true that, generally speaking, a certain degree of saturation has been reached in installing new transmitters, and that no further substantial stimulation of tube sales by transmitter tubes is to be expected?

That is not the case. When speaking of transmitters one should not forget industrial transmitters. Think of their use in particle accelerators. You will no doubt remember the discussion on the "Ebersberger Forst" project in the daily newspapers. According to planning at the present time, the transmitting tube requirement for this single accelerator project is comparable with the total requirement for the uhf transmitters of the German Television Service.

You referred to special components using tube technology as electron tubes with future prospects. What do you mean by that?

According to the classical definition, an electron tube is a device whose operation is based on the transport of electrons in vacuum or in gases. I would like to extend the definition and call every device that can be produced with the aid of tube technology a tube. I am thinking here, for instance, of the gas laser. If you look at the design of an Argon laser in our laboratory, you will find that there is an amazing technical similarity to a modern transmitter tube: you have the classical application of metal-ceramic techniques, you have the same methods of cooling as are used with tubes. This is equally true of the previously mentioned particle accelerator, in which the tuned circuits have to be built using metal-ceramic techniques and the resonant cavities have to be evacuated—once again the techniques of classical electron tubes.

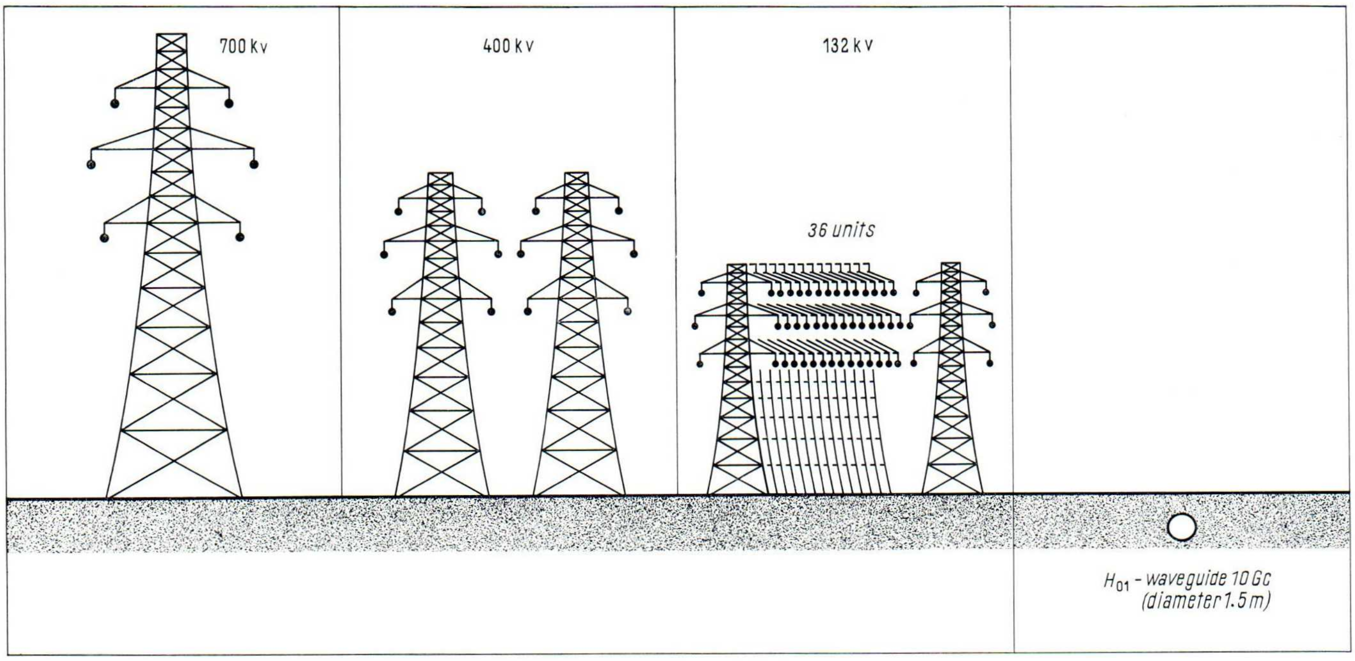
You have now taken a big step. The start of the departure from the classical tube was surely the change to the transit-time tube, which occurred at the beginning of the thirties. I am now thinking back to the time when, studying for my doctor's degree, I worked with the Barkhausen tube, the first practical transit-time tube, and to my subsequent professional connection with the magnetron.

What possibilities do transit-time tubes offer today?

The most important transit-time tubes today are the traveling-wave tube, the klystron and the magnetron.

It was primarily traveling-wave tubes and reflex klystrons that made communications at microwave frequencies possible. In the future also, transit-time tubes will retain their importance for communications engineering above certain power levels. In addition, the use of microwave industrial transmitters will no doubt gain in importance. Just recently, microwave kitchen ranges for large restaurants have been actively discussed and initial commercial successes have been achieved.

In connection with power transmission by means of microwaves there are projects which are utopian in a sense, but on the other hand are not so unreasonable as not to be discussed by engineers. I am now thinking of the possibility of large-scale electric power transmission by means of microwaves in waveguides laid in the ground, instead of using open-wire lines and 50 cps as has been done hitherto.



Comparison
of transmission possibilities
for a power of 4,000 Mw
via open-wire lines (50 cps)
and an
underground H_{01} waveguide (10 Gc)

In one of your recent publications [1] you spoke of the future significance of scanning tubes. Apart from the well-known camera tubes and picture tubes, what else is there in this field?

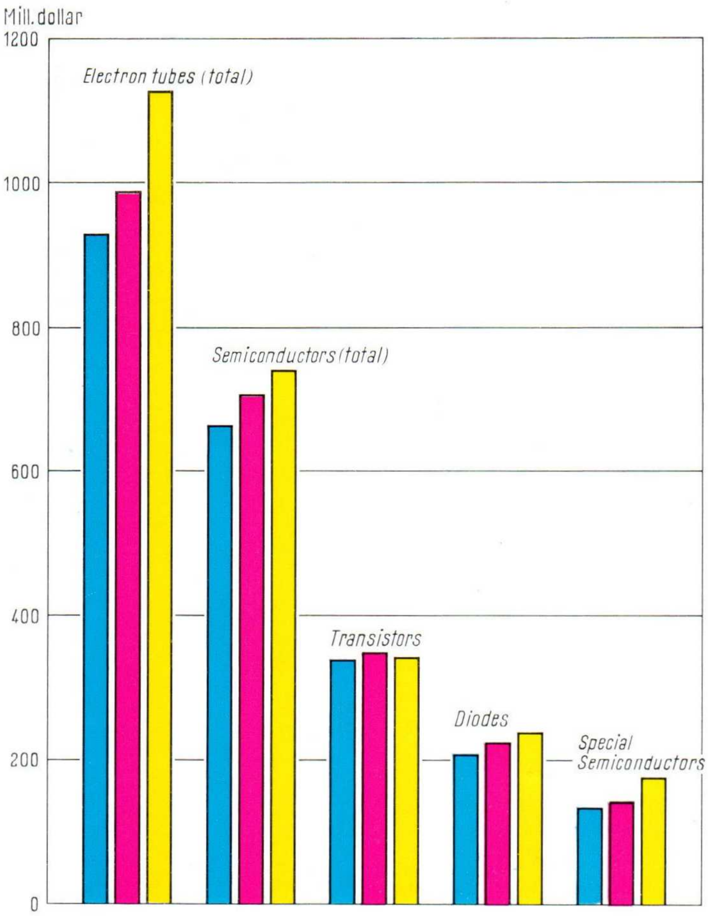
An electron-beam scanning tube is a tube in which an electron beam is deflected transversely for the transmission or conversion of information. Amongst these tubes are the color and black-and-white television picture tubes, the oscilloscope tube, the vidicon and also the coding tube, which we have developed here and which can be used for the conversion of an analog signal into a digital signal. There is already a larger turnover in the market for electron-beam scanning tubes than there is in that for transistors. According to the American forecast that I have already mentioned, this turnover will continue to increase.

At the meeting in Ulm [2], the possibility of replacing the three-beam shadow-mask tube by a single-beam tube was one of the subjects broached. Will the shadow-mask tube continue to dominate, or is there any possibility that this system, which is very complex due to the necessary deflection mechanism, can be replaced by a simpler tube?

I am convinced that it would be possible to replace the shadow-mask tube by a single-beam tube. I believe, however, that economic considerations at present prevent this. A very high standard has been reached in the production of the shadow-mask tube and it would be absolutely uneconomical to abandon it now. In addition, the receiver circuit in all single-beam systems known to me would be more complex than the circuit using a shadow-mask tube.

Up to now our reflections on the progress in electron tubes and the possibilities that it offers have been of a more general nature. What do you consider to be the possibilities and the limits of the electron tube from the point of view of physics?

In the case of the klystron, the magnetron, the traveling-wave tube and the grid-controlled transmitter tube, the upper power limit, at least for continuous operation, is set by the fact that only a certain amount of the heat resulting from power dissipation in the tube can be exchanged. However, this upper limit is not fixed: the electron optics and the heat transfer techniques are being improved with a resulting approach to a theoretical upper limit. This limit is set by the maximum current that can be focused in a tube and the maximum field strength that can be established between two electrodes. This maximum field strength is limited by the onset of field emission and the discharges resulting from it. If one considers it quantitatively, one finds that at frequencies below about 10 Gc the fundamental upper



Future prospects for electron tubes and semiconductors, based on an American market forecast

Blue: 1965
Red: 1966
Yellow: 1969

[1] Paschke, F.: Zukünftige Entwicklung der Elektronenröhren. Siemens-Z. **39** (1965) pp. 1161 to 1167

[2] "State of the Art and Development in the Field of Electron Tubes" Meeting of Professional Group 4 (Tubes) in the Nachrichtentechnische Gesellschaft, September 29 to October 1, 1965. Metschl, E. C.: The Tube Is Not Dead. Siemens Electronic Components Bull. **I** (1966) pp. 74 to 76

Used in uhf transmitters, particle accelerators and other industrial r-f generators, power tubes have an assured future power limit for klystrons and traveling-wave tubes with solid beams lies at about 50 Mw. For frequencies above about 10 Gc this power limit falls rapidly.

In my opinion one can gain by a factor of 10 if one changes to electron tubes with hollow beams, since then about ten times the current can be focused. Below about 500 Mc, grid-controlled transmitter tubes are theoretically suitable for power values greater than 50 Mw. In the case of the magnetron the maximum obtainable power will probably always be limited by the dissipation, since the electron current impinges on the intricate r-f structure.

You mentioned theoretical limits and have indicated these in your previously mentioned published work. Is it possible to reach these limits?

Yes, I would say that they have almost been reached, in particular with pulse-modulated tubes. There is a klystron used in the large linear electron accelerator at Stanford University which produces a maximum power of 32 Mw. At the frequency used one is thus just below the theoretical limit.

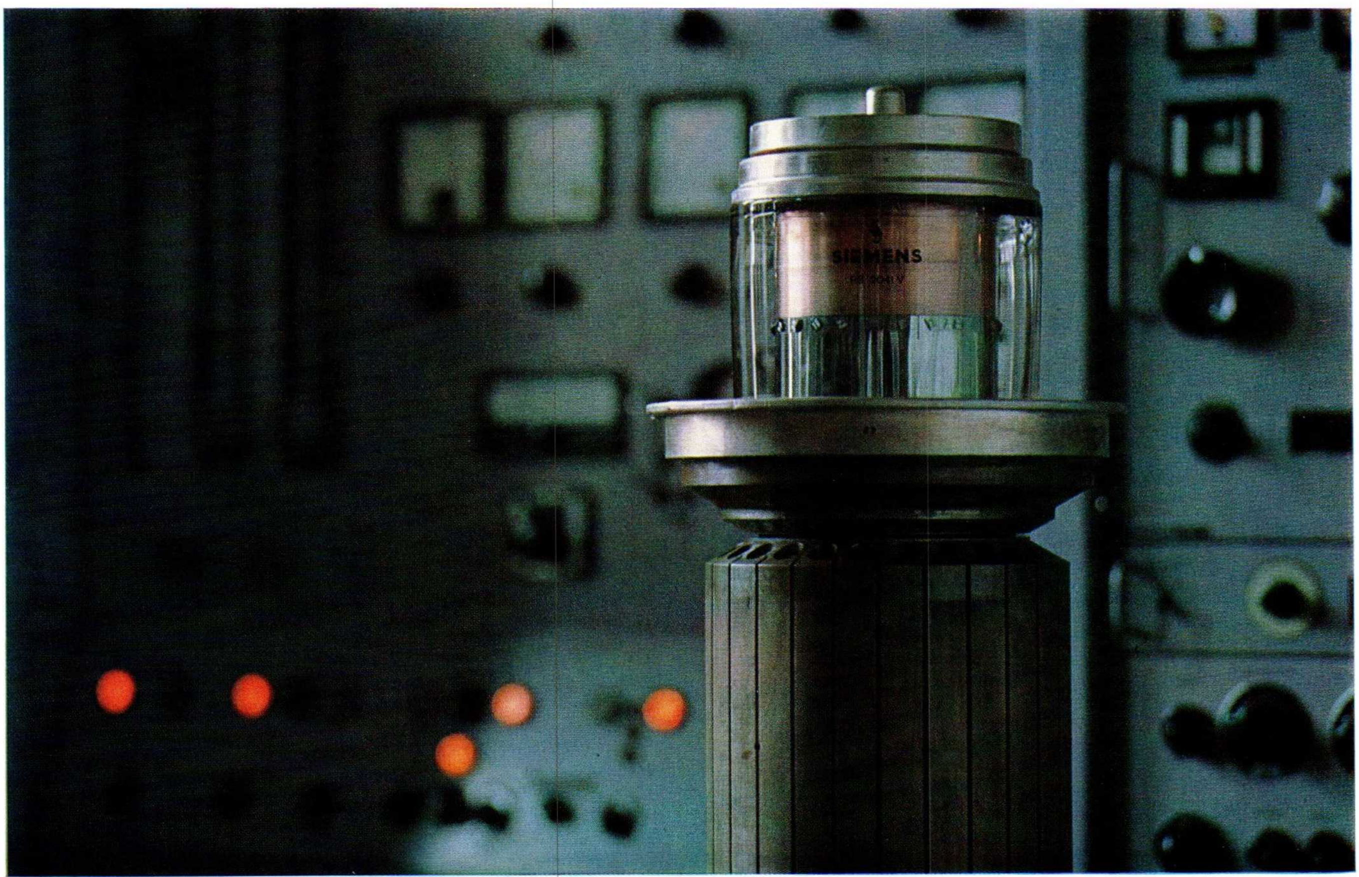
Before, you touched on the problem of power transmission. Is it possible to generate energy in the microwave band at power ratings normally transmitted today?

As I have already said, I am convinced that it is theoretically possible to generate a microwave power of 500 Mw in a *single* tube with a hollow beam. However, in order to achieve a c-w power of this order much development work with extremely high capital investment must still be carried out. In this connection it is interesting to note that the well-known Russian physicist and Nobel-prize winner *Kapitsa* favors such a development in the Soviet Union and has made many suggestions.

Is the development in the field of waveguides already so far advanced that this mode of transmission is comparable, on the basis of losses, with the normal transmission paths?

It so happens that much progress has already been made in microwave transmission line techniques. In order to be able to compete with classical methods, an attenuation of about 1/1000 db per kilometer is necessary, and this actually appears to be attainable with modern microwave techniques. There are also no corona losses in a waveguide, which can be regarded as a further advantage. The laying of the microwave transmission line beneath the earth's surface could be significant, for instance, for the interruption-free supply of power to military installations.

The consumers will largely require the conventional type of current, so that a conversion of the microwave power will be necessary. Do suitable rectifiers exist?



In theory every microwave tube can be operated as a rectifier. Up to now the efficiency ratings obtained have fallen a long way short of those obtained with generators. Only recently has optimism arisen here. We at Siemens are at present investigating a microwave rectifier which promises efficiency ratings of about 90%.

Low-noise receivers are receiving much attention nowadays, particularly in connection with space travel. Is it possible with electron tubes to approach the low noise factors of masers, and what, if anything, favors further development in this direction?

Dr. Metschl, you are now touching an open sore. Not a very high opinion is held of low-noise amplifier tubes — in my view unjustly. And yet there are now traveling-wave tubes which operate at 3 Gc and have a noise temperature of only 57 °K, and that with the enormous bandwidth and amplification of a traveling-wave tube. One could further show that this noise temperature is inversely proportional to the focusing magnetic field. Development was stopped at 57 °K because it was not possible to increase the magnetic field beyond 5,000 gs with the available simple coils. However, I am of the opinion that with modern engineering techniques it would be possible to realize a focusing field of 20,000 to 30,000 gs. One would then obtain a noise temperature of 10 to 20 °K. I can well imagine that a broadband amplifier tube with such a noise temperature would be attractive for radio links via satellites and for radio astronomy.

Prof. Paschke, you were engaged in tube development in America for many years. How do you view the state of development in Germany from this perspective?

We must here differentiate between the requirements of military and commercial engineering. It is my opinion that the Americans are far ahead of us in the field of military electronics. But I am not convinced that we cannot keep up with them in the field of commercial engineering. In communications engineering in particular, due to the large number of channels required, problems such as noise, non-linear distortion and constancy of amplification over a wide frequency band are of decisive importance. In addition, in commercial applications, much interest is attached to the question of price and it is here that we have an advantage over the American manufacturers. Naturally, it is a matter of great importance that the American industry—I am now speaking of the whole American industry—is in a position to draw from the government year after year a sum of 15,000 million dollars, as a grant for development and research. The American government expenditure on research and development amounts to about 3% of the gross national product. Here in West Germany we should aim at one day receiving a similar proportion. Only in that way can we keep up with American industry in the long run, and prevent our becoming economically dependent.

The Siemens Helium-Neon Gas Laser

By Hans Golser and Otmar Hintringer

The operation of the laser is based mainly on the combination of two principles. These are simple energy feedback, and the discovery by A. Einstein that with atoms or molecules, the phenomena of spontaneous emission and absorption are inadequate to describe the interaction between matter and an electromagnetic field, but must be supplemented by a third phenomena, the so-called induced or stimulated emission. Contrary to spontaneous radiation, induced radiation is coherent, and thus in phase with the stimulating field. It was first utilized in 1954 in the maser (*m*icrowave *a*mplification by *s*timulated *e*mission of *r*adiation). Shortly afterwards it was suggested that the maser principle could be extended into the light frequency band but it was not until 1960 that such a practical laser (*l*ight *a*mplification by *s*timulated *e*mission of *r*adiation) could be realized.

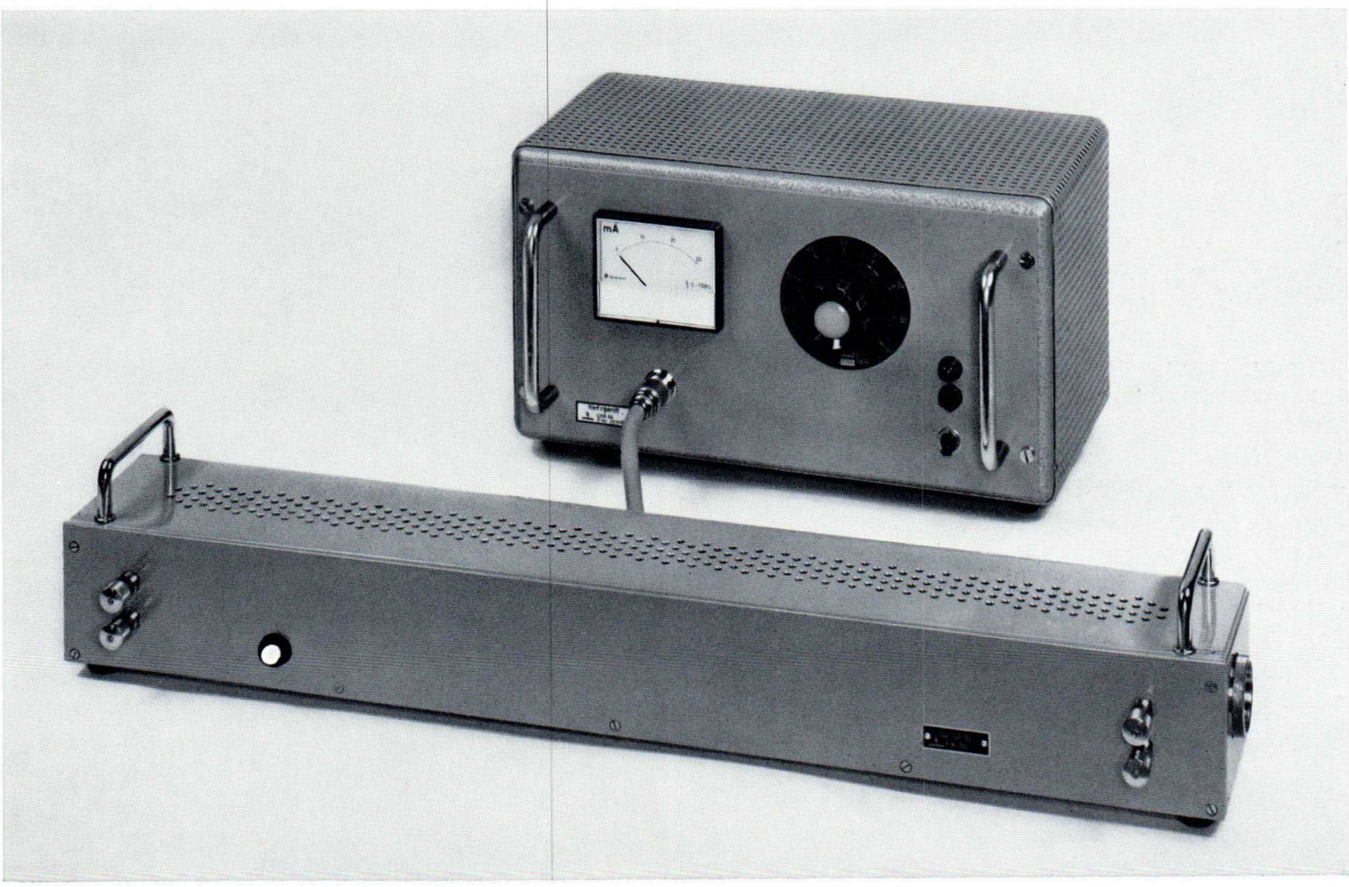
Laser principle

In order to understand the generation of a laser oscillation, it should first be remembered that atoms and molecules can only exist in certain energy levels. The difference in energy between two such levels is given by $W_2 - W_1 = hf$, where h is *Planck's* constant and f the frequency of the absorbed or emitted radiation during the transition from one energy level to the other. When considering a large number of similar atoms, not all of them are at the same energy level, but are statistically distributed over the individual possible levels such that in the state of equilibrium the higher energy levels are less densely populated than the lower ones (*Boltzmann* distribution). However, induced emission requires an upper energy level to be more heavily populated than a lower level; the necessary reversal of population, called inversion, must be produced artificially. The longest possible life of the inverted (upper) energy level is a prerequisite for successful inversion, since a short life prevents sufficient atoms collecting in the upper level. Once the inversion is obtained, the strongest possible induced emission must then occur in the *active* medium thus created. For this purpose use is made of the feedback principle: The active medium is introduced into an optical resonator consisting of two mirrors arranged in parallel (*Fabry-Perot* interferometer). If it is assumed that an atom falling back to a lower energy level radiates the energy difference in the form of a light wave, this wave propagates in the active medium and stimulates other excited atoms in its path to the same energy transitions, and the wave already starts to grow. It is then reflected at the mirror, the process of induced emission repeats itself along the return path and the wave continues to grow. Repeated reflection between the two mirrors produces a standing wave if the distance between the two mirrors is a whole multiple of the half wavelength. If one mirror is partially transparent, part of the stored energy can leave the resonator to form the laser beam.

This beam has a low divergence, since only such beams which have been built up by frequent reflection within the resonator can leave it. Waves deviating from the axis have no chance of growing and leaving the resonator to contribute to the beam, since after only a few reflections they emerge laterally and are thus lost.

Furthermore this radiation is spatially coherent, i. e. when considering the beam divided into individual bunches, these bunches will have a fixed phase relationship to each other. The combination of induced emission and feedback produces an oscillator having an extremely monochromatic radiation also coherent with respect to time. Another notable feature of the laser radiation is its high spectral intensity yielded by the synchronous energy emission of a large number of atoms.

Fig. 1 The operational helium-neon laser LG 64 with power supply unit LGN 64



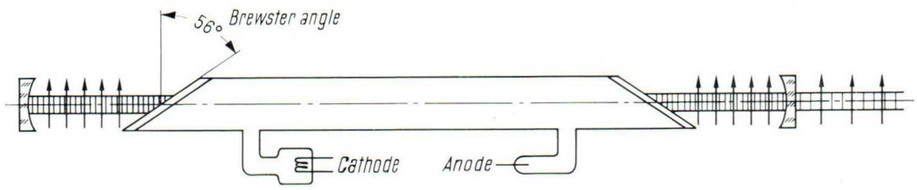


Fig. 2 Laser schematic. The arrows indicate the direction of the electric field vector of the linearly polarized laser light

The gas laser

There are already many substances known which are suited for generating laser radiation. Among these the gas-type laser offers considerable advantages in some respects over other forms. One advantage is that the homogeneous active medium results in very low losses and thus high-quality optical resonators. For this reason, and because of the narrow line widths of the transitions in gases, both monochromatism and coherence are several orders of magnitude better than with other types of lasers. Yet another advantage is that the atoms can be excited relatively simply by applying electrical energy (gas discharge), thereby producing inversion. Furthermore, with few exceptions gas lasers operate continuously.

Design of the Siemens He-Ne laser LG 64

The helium-neon gas laser is now a technically proven device. Such a laser, the Siemens Type LG 64, is illustrated in its operational state in Fig. 1, and its operation explained by Fig. 2. A plasma tube with an internal diameter of 2.5 mm and 61 cm long filled with a mixture of five parts helium to one part neon at a total pressure of 1 torr is vacuum-sealed at both ends with high-grade optical flats inclined at the *Brewster* angle. The tube is fired by a high-voltage pulse. The *Brewster* plates ensures that the laser only starts to oscillate in one direction of polarization, namely in the plane of incidence provided by them, since in this direction no reflection losses occur. The resulting laser radiation is linearly polarized.

The movable plasma tube is supported in a frame consisting of three Invar rods. Two flanges fix the ends of the Invar rods and hold the two interchangeable mirrors. The basic fittings of the laser consist of a plane mirror and a pair of spherical mirrors arranged almost confocally, i.e. the distance between the mirrors corresponds approximately to the radius of curvature of the mirrors. These have been coated with multiple dielectric layers to minimize the losses at a wavelength of 632.8 nm. One mirror has a reflection factor of about 99.9%, the other of approximately 98%. The spherically-curved outer surfaces of both mirrors are dimensioned such that the angle of divergence of the emerging laser beam is determined only by diffraction, in other words by the spot diameter at the mirror.

With the Siemens He-Ne laser, the mirrors are preset and fixed in their optimum position before leaving the plant. Maximum output is obtained by adjusting the position of the plasma tube with four micrometer screws, which is a less critical method than the more conventional mirror adjustment.

Frequencies of the Siemens He-Ne laser LG 64

A helium-neon laser can oscillate at several different frequencies. Which of the laser transitions will be stimulated depends on the selective reflection properties of the mirrors used in the optical resonator. The Siemens He-Ne laser can also be supplied with mirrors as optional accessories for the infrared lines 1,152.3 nm and 3,391.2 nm. In one aspect the optical resonator of the laser resembles the conventional microwave cavity resonator; self-oscillations, or modes, also occur. In the optical resonator they are almost transversal electromagnetic (TEM) modes, and thus have negligible field strength components in the axial direction. There are two basic types of TEM mode in the optical resonator, the transverse modes exhibiting points of zero field amplitude and phase reversals in planes normal to the direction of the beam, and the longitudinal modes differing by their number of nodes in an axial direction with the same transverse field distribution.

Without taking particular measures many modes will occur simultaneously (multi-mode operation), resulting in a beam cross section appearing as an almost uniformly bright spot. If thin wires are introduced into the beam in the optical resonator, certain transverse modes can be filtered out as shown in Fig. 4. The wire produces nodes in the transverse field, suppressing many modes and leaving one particular mode predominant. The seven photographs reproduced in Fig. 4 represent certain transverse modes, and were taken in the Siemens research laboratories. Many applications require single-mode (TEM_{00} mode) operation. In this case the field strength distribution over the beam cross section is gaussian with uniform phase over the complete cross section. The only method of producing single-mode operation is to design the optical resonator such that the diffraction losses for all higher modules are too high and thus prevent buildup of oscillation. One solution is to use the spherical resonator, which consists of a plane mirror and a hemispherically-curved mirror with a spacing equal to the radius of curvature of the spherical mirror. In the Siemens He-Ne laser only one mirror of the confocal resonator need be replaced by a plane mirror.

Single-mode operation does not exclude the possibility of the laser simultaneously oscillating in several longi-

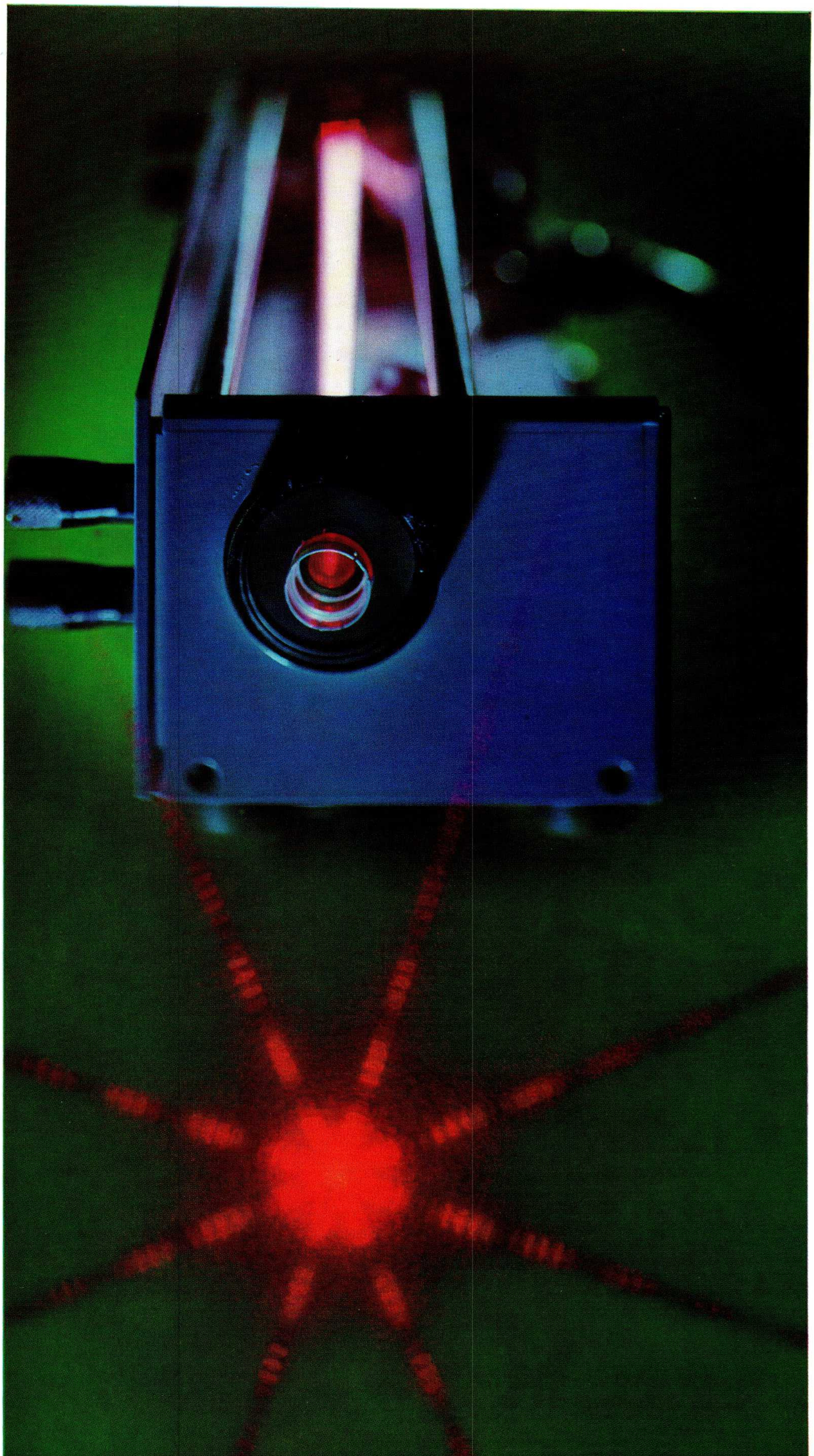
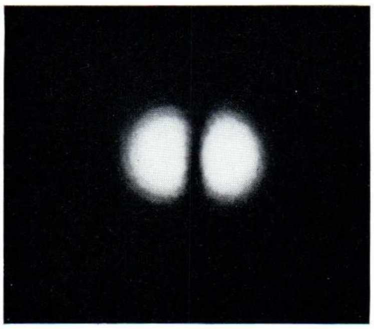


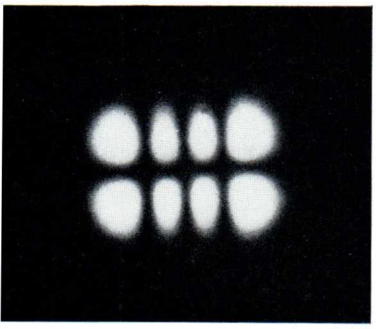
Fig. 3 High-order diffraction patterns obtained due to the excellent coherence of the laser radiation

Fig. 4
Various transverse modes

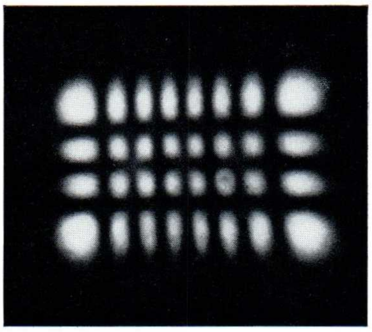
a. Single (fundamental) mode TEM_{00}



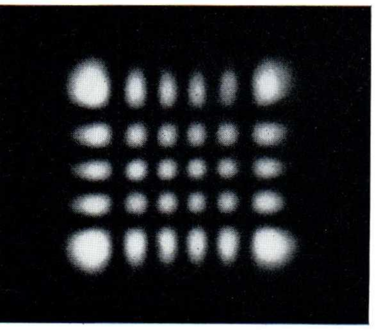
b. TEM_{01}



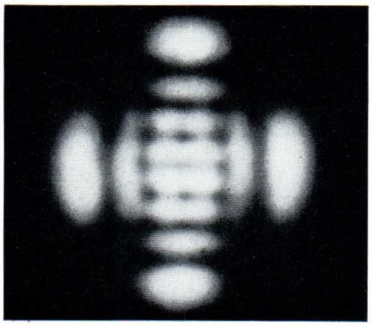
c. TEM_{31}



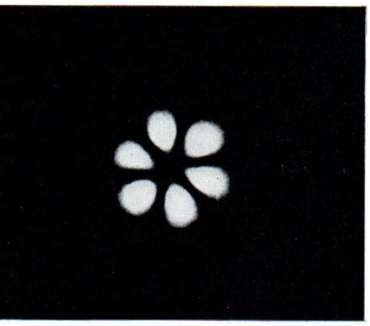
d. TEM_{73}



e. TEM_{54}



f. $TEM_{50} + TEM_{07}$



g. TEM_{16}

tudinal modes with a frequency separation $c/2l$, where c is the velocity of light and l is the length of the optical resonator. In the LG 64 the resonator length is about 63 cm, giving a frequency separation of 240 Mc. At 632.8 nm this corresponds to a wavelength difference of $3 \cdot 10^{-4}$ nm. Evidence of longitudinal modes can be obtained by mixing them with the aid of a photomultiplier tube. The cathode current of a photomultiplier tube is directly proportional to the intensity of the incident light and thus to the square of the field strength. As a result of this square-law characteristic the optical frequencies corresponding to the individual longitudinal modes are mixed. This produces the difference frequency, in this instance 240 Mc, which can be indicated by a suitable receiver.

Because of its very high coherence, both spatially and with respect to time, and its high spectral intensity, the laser has become a useful piece of auxiliary equipment in many branches of physics, chemistry, medicine and engineering. The versatile applicability is based primarily on the refinement of hitherto known methods. A few examples are range finding and bearing, precision measurement of length (interferometry), navigation, investigations on material surfaces, optical tracking, flying-spot scanning, analog computers for example when used in determining antenna polar diagrams, holography (three-dimensional photography without lenses), telecommunications, working of materials, spectral analysis, bloodless sterile surgery, *Raman* spectroscopy and the selective excitation of chemical reactions. This list could almost be supplemented daily.

Apart from such specific applications, the progress resulting from studies of the laser itself has contributed considerably to the improvement in knowledge on the structure and properties of matter. Thus one of the oldest branches of physics, optics, has received new stimulus from quantum electronics since the invention of the maser and laser.

Data on the Siemens He-Ne laser

Wavelength:	Standard	632.8 nm
	Optional	1,152.3 nm
Power output at 632.8 nm		3,351.2 nm
Confocal (multi-mode) operation		10 mw
Hemispherical (single-mode) operation		1 mw
Beam diameter		2.5 mm
Beam divergence in single-mode operation with collimating output reflector		2′

The Use of Industrial Generator Tubes

A model calculation for high-power transmitter triode RS 2041

By Hans Pirich

The growing importance being attached to the use of r-f power for the heating and welding of workpieces has led to the development of suitable generating tubes. The powers required in r-f installations are growing continuously. During the development of transmitter tube RS 2041 (Fig. 1), its industrial use was specially taken into account in addition to its employment in shortwave transmitters and for pulse operation. The RS 2041 can be used, for instance, in welding generators for the r-f welding of steel tubes, which calls for powers considerably in excess of those required in other fields of application. A simple method of calculating the transmitter tube operating conditions in the case of industrial use will be shown with a detailed example.

Requirements for industrial generator tubes

Since tubes in industrial r-f equipments operate under much less favorable conditions than those in broadcast transmitters, the following essential requirements must be met:

robust mechanical construction,
insensitiveness to fluctuations of load,
ability to operate with relatively low anode voltages,
long life,
must withstand switching on and off,
low operating costs.

The mesh grid and mesh cathode techniques developed by Siemens contributed most to the construction of a high-power triode with the characteristics referred to.

Construction of the RS 2041

Concentric tubes fitted one in the other are used instead of the rod-shaped heater leads which were normal hitherto. In this way considerably smaller lead inductances and good decoupling of the inner cathode lead are achieved. In addition, coupling between the input and output circuits via the common cathode inductance is substantially reduced, so that the tendency to parasitic oscillations is much diminished.

The tube characteristics at high frequencies are considerably improved by the mesh grid, which is made in a similar manner to the cathode. This is particularly important when the loading is capacitive, e.g. in plastic welding and foundry core-drying installations. The grid of triode RS 2041 is designed to yield an amplification factor μ of 35, the grid disk being constructed by means of the metal-ceramic technique. The grid-plate insulation consists of hard glass; connections to the electrodes are coaxial. The tube is delivered with water or evaporation cooling (versions W and V, respectively). The maximum rated plate dissipation is 180 kw (RS 2041 W) or 220 kw (RS 2041 V). As a class-C r-f amplifier, the tube can deliver an output power of 660 kw with a plate voltage of 18 kv.

Fig. 1 High-power transmitter triode RS 2041 with mesh grid and ready-to-assemble cathode and grid construction

Hints for the use of the tube

When the transmitter tube is used for industrial purposes, attention has to be paid to the following points, which do not have to be considered in radio transmitter operation:

1. The line voltage is generally not automatically stabilized against variations.

Care must therefore be taken to ensure that the maximum ratings are not exceeded in case of line voltage fluctuations. From the viewpoint of cathode life, it is an advantage to at least stabilize the heater voltage, particularly if the maximum rated cathode current is used.

2. In contrast to transmitter operation, the loading is not constant. As a result, the output r-f power and the plate and grid dissipations vary.

For this reason the plate and grid dissipations should not be fully utilized. While with increasing load it is mainly the plate dissipation that goes up, the grid a-c voltage and, as a result, the grid current rise rapidly when the load decreases. The current loads the grid, and the grid dissipation, i.e. grid heating, goes up. Overloading of the grid can be prevented by suitable measures, such as automatic control of the grid drive voltage as a function of the match (loading), by the inclusion in the circuit of tungsten-filament lamps or a resistor with a large positive temperature coefficient, or by switchable grid resistors.

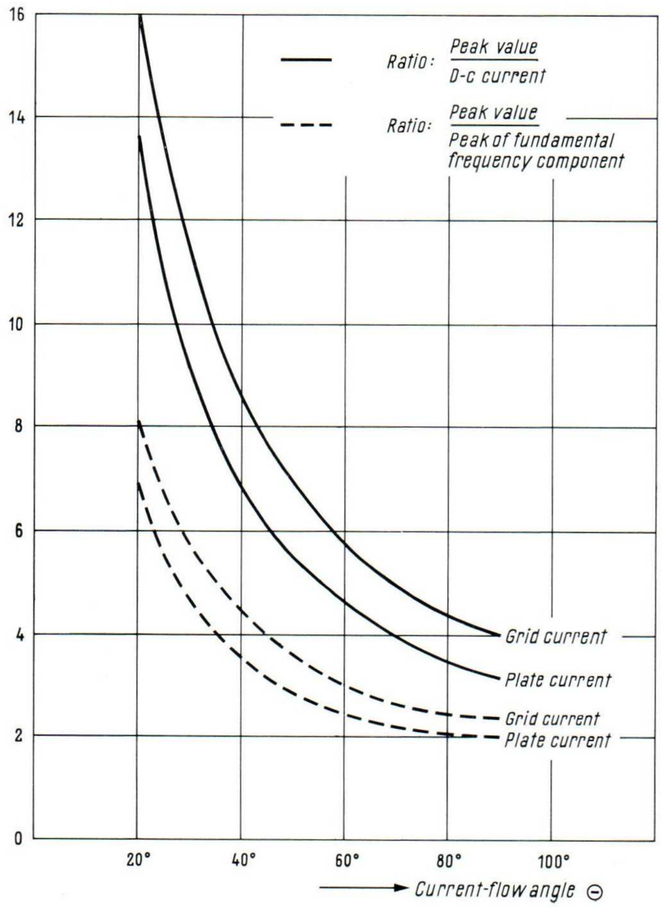
3. The tube plate voltage is generally not filtered.

In order to save expensive filter capacitors and chokes, tubes in industrial generators are usually operated with unfiltered plate voltages. With a view to obtaining better tube utilization, threephase bridge rectification is used almost exclusively for high-power transmitter tubes. With this mode of operation, only a very small residual ripple results.

Calculating the operating conditions of industrial oscillators

A simple method for the calculation of an oscillator circuit with the aid of the maximum tube ratings is discussed in the following. The calculations are based on the data for transmitter triode RS 2041, for class-C operation with smoothed plate-supply voltage.

In principle, each of the three possible sets of characteristic curves, $I_b = f(E_c)$, $I_b = f(E_b)$, $E_c = f(E_b)$, can be used for the calculation of the operating point of a tube. The E_c/E_b characteristics (constant current characteristics) are to be preferred to the others, since with them the oscillator operating conditions can be most easily calculated. The use of these characteristics is particularly advantageous where tuned circuits are used which result



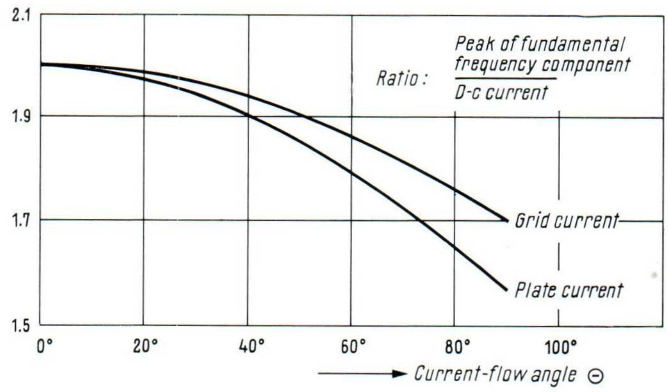


Fig. 2 The main components of simple pulses (according to Wagener)

in sinusoidal a-c drive voltages $E_{c\sim}$ and a-c output voltages $E_{b\sim}$ with 180° phase shift. The drive voltage appears as a straight line on the characteristics graph; the plate current curve as a function of time can be very easily determined. The individual operating parameters are found by graphical means. The method of obtaining the tube current waveforms and calculating the magnitude of the fundamental frequency component will be shown for an assumed case. With the aid of these quantities, drive power, grid dissipation, output power, output impedance and plate dissipation can then be determined.

The curves given in Fig. 2 show the ratios

$$\frac{I_p}{I_\sim}, \frac{I_p}{I}, \frac{I_\sim}{I}$$

as functions of the angle of current flow. (The angle of current flow is defined as that corresponding to half the current-flow period.)

The plate current curve is used for the plate current and the grid current curve for the grid current. The following calculations can be equally applied to oscillator operation and to class-C amplifiers in r-f transmitters. In the case of oscillator operation, however, the output power is smaller by an amount equal to the drive power component.

Model calculation

An r-f generator which is to deliver a nominal power of 300 kw to the workpiece is assumed. Allowing for the tank circuit losses, the transmission losses and the required drive power, an a-c plate power $P_{b\sim}$ of about 400 kw is necessary. The insulation sets a limit of 13 kv to the d-c plate voltage E_b.

The end points are determined for the construction of the load line on the constant-current diagram (Fig. 3 a). End point A of the load line is fixed by the values I_{bp} and $\hat{E}_{b\sim}$. For the peak plate current I_{bp}, one obtains from Fig. 2 for a current-flow angle $\Theta_b = 65°$ (optimum value with regard to efficiency and peak cathode current), $I_{bp} = 2.4 \cdot \hat{I}_{b\sim} \cdot \hat{I}_{b\sim}$ is calculated from the equation

$$\hat{I}_{b\sim} = \frac{2 \cdot P_{b\sim}}{E_{b\sim}}$$

$E_{b\sim}$ is the peak value of the plate a-c voltage, and is assumed constant at 0.9 times the d-c plate voltage, $\hat{E}_{b\sim} = 0.9 \cdot 13 = 11.7$ kv.

Summarizing:

Point A lies at the intersection of the plate current line

$$I_{bp} = \frac{2.4 \cdot 2 \cdot P_{b\sim}}{\hat{E}_{b\sim}} = \frac{2.4 \cdot 2 \cdot 400 \cdot 10^3}{0.9 \cdot 13 \cdot 10^3} = 164 \text{ amps}$$

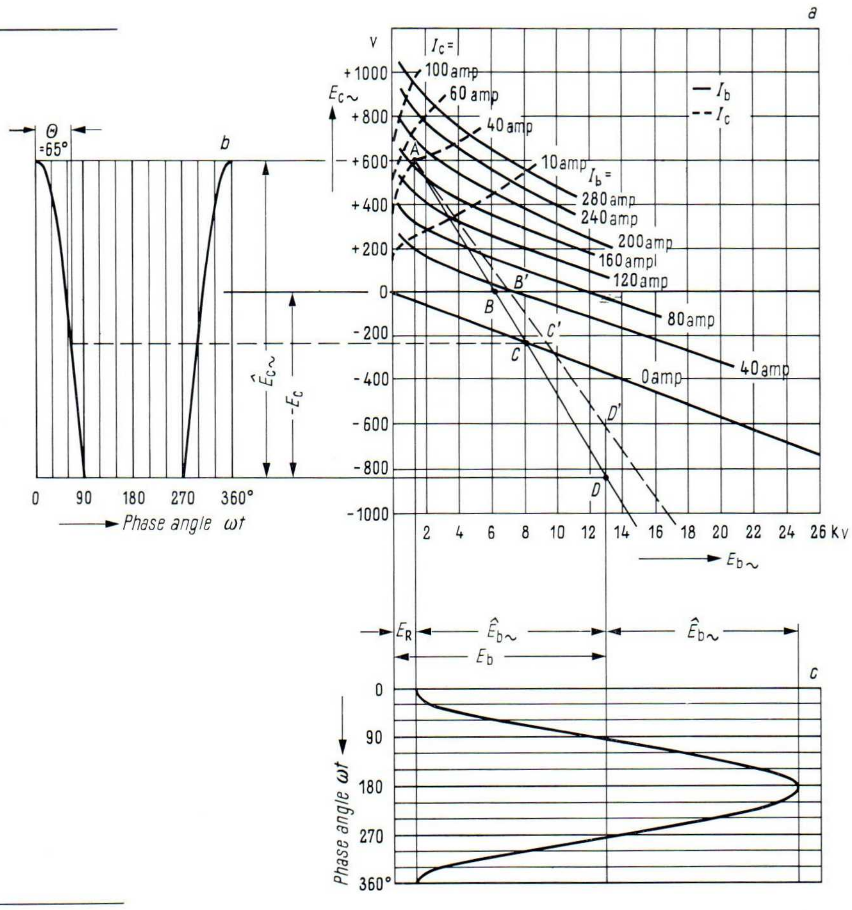


Fig. 3

Determination of the end points for the construction of the load line

a Constant current characteristics $E_c = f(E_b)$ of RS 2041 with load line ABC

b A-c drive voltage waveform $E_{c\sim}$

c Plate a-c voltage waveform $E_{b\sim}$

with the residual plate voltage $0.1 \cdot 13{,}000 = 1{,}300$ v. (This point also determines the peak grid current $I_{cp} = 42$ amps.)

When point A has been fixed, an arbitrary preliminary load line, which is shown dashed in Fig. 3a, is drawn through it.

Point B lies at the intersection of the coordinate for $E_c = 0$ with the preliminary load line.

Point C lies at the intersection of the curve for $I_b = 0$ with the preliminary load line.

Point D lies on the coordinate for $E_b = 13$ kv and the preliminary load line.

The final load line is obtained by rotating the preliminary load line about point A until the equation

$$\frac{\overline{CD}}{\overline{AD}} = \cos \Theta_b = 0.42 \text{ for } \Theta_b = 65^\circ$$

is satisfied.

Determination of values

The projection of points A and D of the load line on to the E_c coordinate gives the peak value of the grid a-c voltage $E_{c\sim} = 600\text{ v} + 840\text{ v} = 1{,}440$ v. Calculations then yield the feedback factor C from

$$C = \frac{\hat{E}_{b\sim}}{\hat{E}_{c\sim}} = \frac{11{,}700}{1{,}440} = 8.1$$

and the peak cathode current

$I_{catp} = I_{bp} + I_{cp} = 164 \text{ amps} + 42 \text{ amps} = 206$ amps.

Since the peak grid current $I_{cp} = 42$ amps (from point A) and the grid current-flow angle given by the ratio of $\overline{BD}$ to $\overline{AD}$ in Fig. 3a,

$$\cos \Theta_c = \frac{\overline{BD}}{\overline{AD}} = \frac{65}{112} = 0.58,\ \Theta_c = 54.5^\circ,$$

are both known, one can obtain with the aid of Fig. 2 the a-c grid current and the d-c grid current. The current-flow angle function is determined from the grid current curve. For the example one obtains

$$\frac{I_{cp}}{I_c} = 6.3 \text{ for } \Theta_c = 54.5^\circ \text{ and } \frac{I_{cp}}{I_{c\sim}}\ 3.4;$$

$$I_c = \frac{42}{6.3} = 6.7 \text{ amps; } \hat{I}_{c\sim} = \frac{42}{3.4} = 12.3 \text{ amps.}$$

The drive power is obtained from the a-c quantities

$$P_d = \frac{\hat{E}_{c\sim} \cdot \hat{I}_{c\sim}}{2} = \frac{1{,}440 \cdot 12.3}{2} = 8{,}850 \text{ watts.}$$

The drive power includes the grid dissipation Q_c and the power $E_c \cdot I_c$ dissipated in the grid resistor. The actual grid dissipation is thus

$Q_c = P_d - E_c I_c = 8{,}850 - 840 \cdot 6.7 = 3{,}220$ watts.

The grid resistor required to generate the negative grid bias is

$$R_c = \frac{E_c}{I_c} = \frac{840}{6.7} = 125\ \Omega.$$

The peak value of the plate current fundamental frequency component is

$$\hat{I}_{b\sim} = \frac{2 \cdot P_{b\sim}}{\hat{E}_{b\sim}}\ 68.5 \text{ amps.}$$

The d-c plate current is obtained by again using the current-flow angle function according to Fig. 2.

Accordingly,

$$\frac{\hat{I}_{b\sim}}{I_b} = 1.76, \text{ for } \Theta_b = 65^\circ\ I_b = \frac{68.5}{1.76} = 39 \text{ amps.}$$

The required load impedance is

$$R_b = \frac{\hat{E}_{b\sim}}{\hat{I}_{b\sim}} = \frac{11{,}700}{68.5}\ 171\ \Omega.$$

The d-c power consumption is given by $P_b = E_b\ I_b = 13{,}000 \cdot 39 = 507$ kw, the plate dissipation $Q_b = P_b - P_{b\sim} = 107$ kw, and the efficiency is determined from the ratio

$$\eta = \frac{P_{b\sim}}{P_b} = \frac{400}{507} = 79\,\%.$$

A comparison of the operating parameters and the maximum ratings shows that there is such a wide margin between actual load and maximum load that the maximum ratings will hardly be exceeded due to load and voltage fluctuations, and additional control circuits are thus not necessary.

Bibliography

Wagener, W. G.: Performance of Transmitting Tubes. Proc. IRE **25** (1937) Part I, pp. 47 to 77

Shift Register and Ring Counter with Transistors and Toroidal Cores

By Otto Macek and Hans Tropper

This paper describes an arrangement incorporating toroidal switching cores of Type B64520-A5-X405 (2 mm) or B64530-A5-X405 (3 mm) and BC 123 transistors which can be used as shift register or as ring counter; it operates at clock frequencies (computing speeds) up to 300 Kc over a temperature range between −10 and +60 °C. A drive circuit developed simultaneously generates the two required clock pulses from a sinewave of 3 v rms taken off across a 150-Ω resistor. The shift frequency of the register is determined by these pulses. Among the advantages of this register are its high clock frequency and the fact that only two clock pulses are needed for operation. Due to its low power dissipation, this shift register lends itself particularly well to SIMIBLOCK applications.

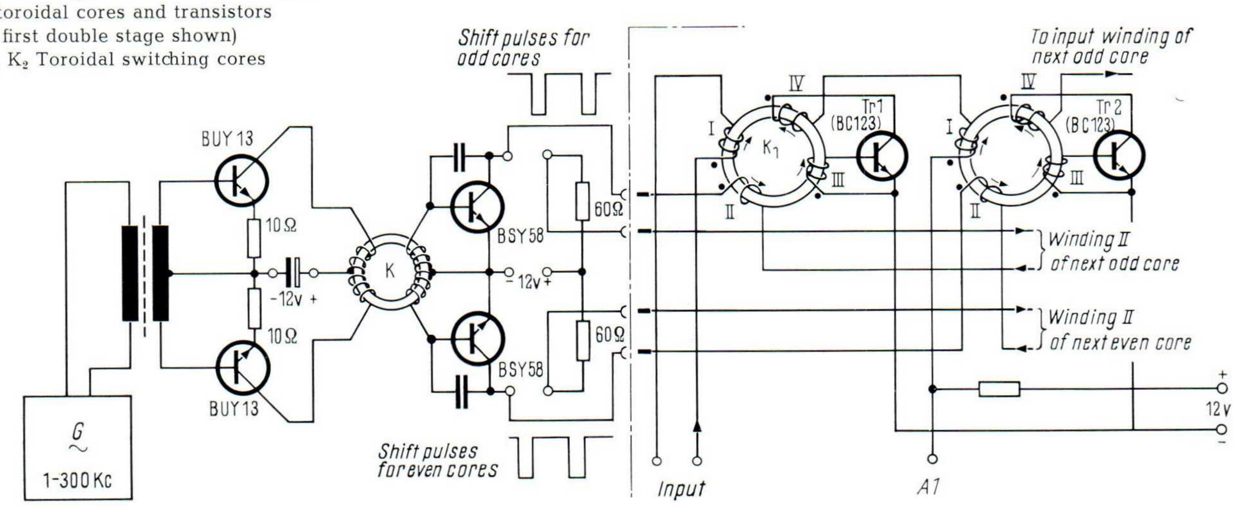
Shift registers assembled from toroidal cores and transistors have the advantage of operating up to high clock frequencies. Also, two clock pulses suffice for shifting the information by one step, and less power is required for driving these circuits than shift registers with saturable reactor type cores [1].

When toroidal switching cores of Type B64520-A5-X405 or B64530-A5-X405 are used, the rise time of the driving current pulses should be between 0.2 and 0.5 μsec. Such short rise times can only be achieved with high-frequency transistors. For this reason, a BC 123 transistor (with white dot) was chosen, a planar transistor in a plastic case. Its maximum collector current is 50 ma, maximum collector voltage 45 v, and the switching time is below 0.1 μsec corresponding to a cutoff frequency of 250 Mc. The basic shift register circuit diagram is shown in Figs. 1 and 2.

For ring counter applications, i.e. when a *single* "1" is passed from one stage to the next while all other stages are set to "0" (code: "one-out-of-ten"), 10 single stages with one core and one transistor each may be adequate for representing one counter decade. However, if a shift register for any binary numbers, such as *1111,* is to be

[1] Schmitt, R.: Ferrite-Core Shift Registers. Siemens Electronic Components Bull. I (1966) pp. 81 to 85

Fig. 1 Pulse generator and shift register with toroidal cores and transistors (only first double stage shown)
K, K_1, K_2 Toroidal switching cores



Version A: 2-mm core B64520-A5-X405

Winding	I	10 turns
	II	10 turns
	III	10 turns
	IV	10 turns

R = 300 Ω

Version B: 3-mm core B64530-A5-X405

Winding	I	30 turns
	II	10 turns
	III	6 turns
	IV	12 turns

R = 560 Ω

implemented, one double stage containing two cores and two transistors must be provided for each bit of information. When the information is to be shifted, the "1" stored in a certain stage must first be read out and transferred to the next stage before the subsequent "1" can be written in. Therefore, "storage stages" have to be interposed between the "counter stages." In other words, $2n$ single stages are needed for a "word" of n bits, i.e. a binary number of arbitrary length n. A tetrade, the preferred way of binary coding of decimal digits, requires 8 single or 4 double stages. The simple drive circuit described later makes it possible to reach shift frequencies[1] up to 300 Kc.

Shift register and drive circuit

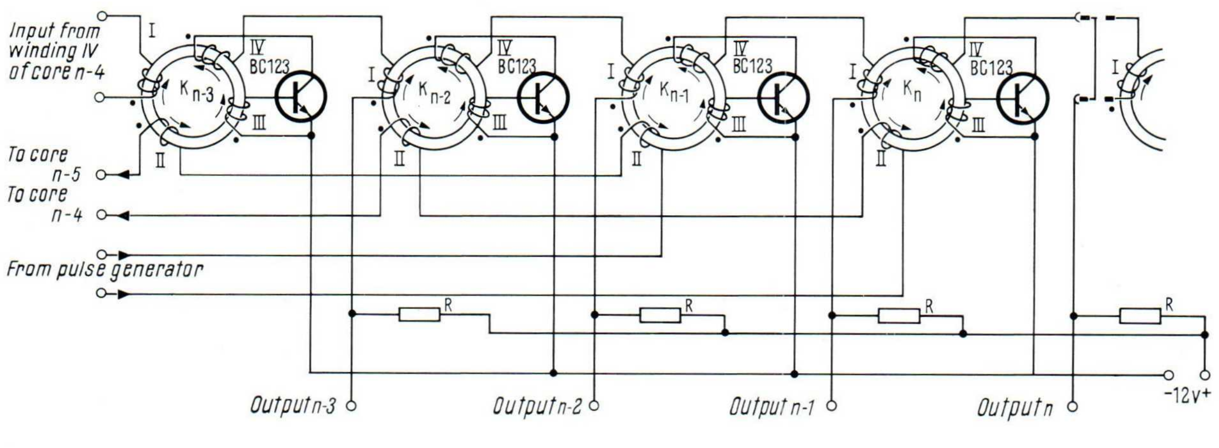
2-mm toroidal switching cores

The upper cutoff frequency of shift registers with toroidal cores and transistors depends essentially on the core properties. Type B64520-A5-X530 is one of the fastest switching cores which still have reasonable dimensions; but the current needed for attaining a field strength above the coercive field is relatively high. Cores of X405 material are better in this respect; however, they are not as well suited for very high frequency registers as X530 cores. The shift pulses (clock pulses) have current amplitudes of about 400 ma, a rise time of 0.5 μsec and a pulse width of about 2 μsec. It can be seen from the characteristics of the toroidal core used that a rise time of 0.5 μsec is a good value for the two switching pulses and also for the information pulses representing a binary "1." The output pulse width is 2 μsec. If the number of turns of the current control winding is 10 with this core size, the abovementioned current amplitude is adequate for the *two* switching pulses. The most favorable number of turns and the current amplitude were found experimentally.

The number of turns for the various windings were chosen such that the triggered blocking oscillator formed by the core/transistor circuit is triggered securely by the 400-ma shift pulses, and a "0" is transferred to the next stage when a "1" bit is removed, i.e. a blocking oscillation cannot occur. All windings (Figs. 1 and 2) have 10 turns each. The feedback winding connected to the base is wound in a sense opposite from the collector winding, as is customary with blocking oscillators. During a "1", the collector current flowing through resistor R is only about 40 ma. As most of the heat dissipation takes place in this resistor, the current had to be kept at a low value. In the manufacturing process of magnetic components, core winding is an operation contribut-

[1] The shift frequency shall be defined as the frequency of shifting one bit of information by one step (two cores).

Fig. 2 Shift register with toroidal cores and transistors (next to last and last double stage)
K_{n-3}, K_{n-2}, K_{n-1}, K_n Toroidal switching cores



Version A: 2-mm core B64520-A5-X405
Winding I 10 turns
II 10 turns
III 10 turns
IV 10 turns
$R = 300\ \Omega$

Version B: 3-mm core B64530-A5-X405
Winding I 30 turns
II 10 turns
III 6 turns
IV 12 turns
$R = 560\ \Omega$

ing considerably to the final price; therefore, great care was taken to obtain the simplest possible winding instructions. It was tried for this reason to achieve identical numbers of turns for all windings. The output pulses which can be taken off each shift register stage have amplitudes of about 10 v; their rise time is approximately 0.1 μsec, and their width about 2 μsec.

3-mm toroidal switching cores

If one attempts to minimize power dissipation by increasing the number of turns and reducing the current —while the product of both retains its required constant value—one finds that a limit is reached very soon.

Firstly, the necessary number of turns cannot be placed on 2-mm cores if lacquered copper (CuLL) wires of reasonable diameter are to be used. One would have to change to the next size, i.e. to Type B64530-A5-X405 cores (3-mm cores). Secondly, it will be noticed that a shift register with identical numbers of turns will no longer work at larger numbers. The numbers indicated in Fig. 1 for version B were found experimentally. If 3-mm cores are equipped with these windings, the collector current pulses through resistors *R* of a shift register decrease to about 22 ma. Hence considerably less heat is dissipated here than in the shift registers with 2-mm cores.

Drive circuit or shift pulse generator

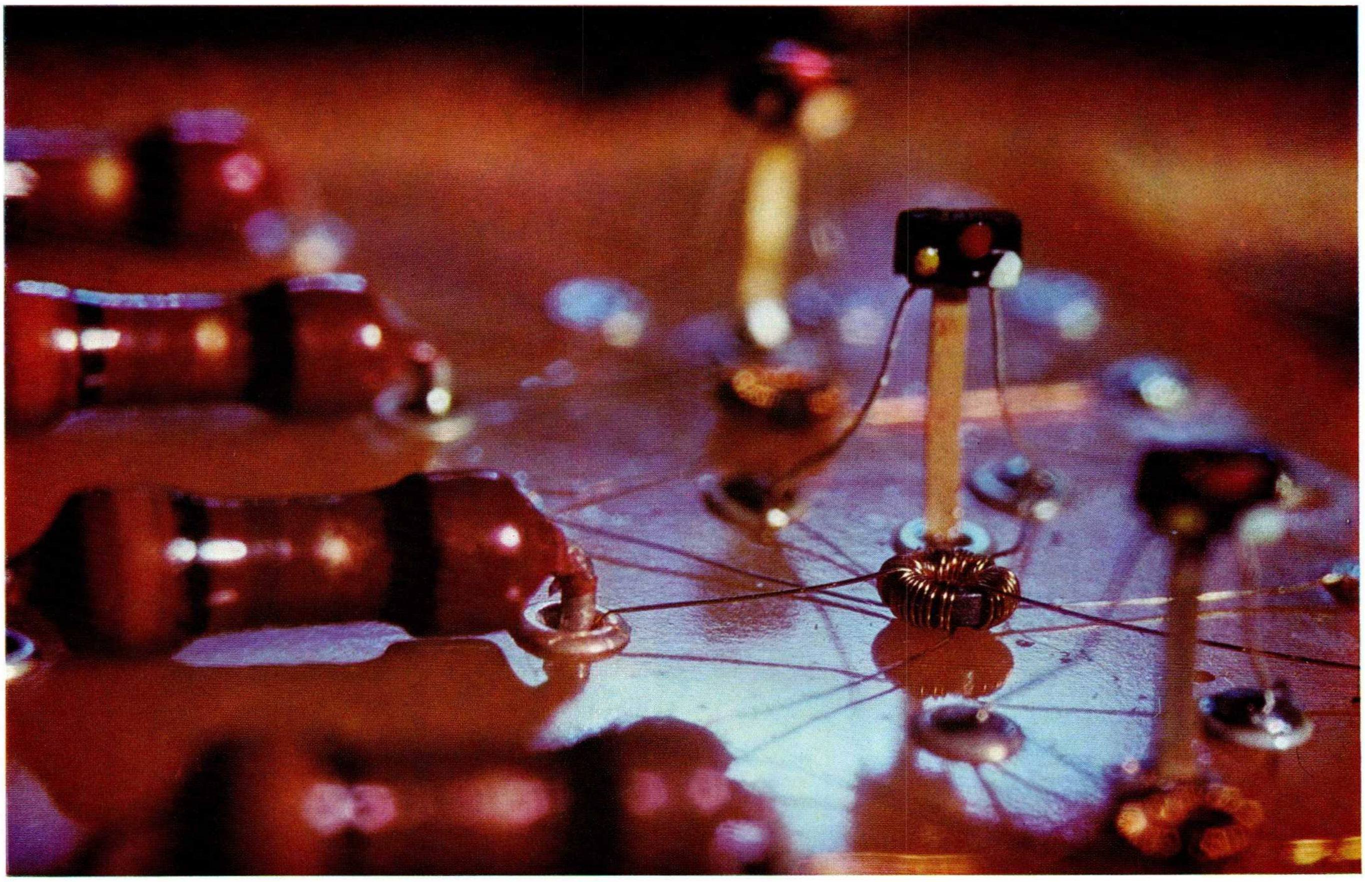
A sinewave generator produces a voltage of 3 v rms across a 150-Ω resistor. This voltage is then amplified by a pushpull transistor amplifier and used to drive a toroidal core of Type B64550-A5-X405. The core is saturated during the better part of each half-cycle, and supplies two narrow output pulses per cycle of the driving sinewave; the pulses are shaped and amplified by two following transistors. All coupling windings for the clock or shift frequency are connected in series in the collector circuits of these output transistors. In shift registers, one output feeds all odd cores, the other all even cores.

Mode of operation

The circuit contains 2 cores and 2 transistors per bit. The cores have the function of blocking oscillator transformers; they are driven into saturation. For simple types of ring counters, *one* core and *one* transistor per stage are sufficient.

Through the winding identified as "input" in Fig. 1, a "1" is written into the first core causing it to be saturated in a clockwise sense. Winding II on the first core (K 1) receives the first shift pulse; a counterclockwise magnetic field is generated, and the impressed flux is flipped over into the opposite direction. This flux reversal induces a

Fig. 3 This detail of an experimental laboratory setup shows toroidal cores, transistors and resistors connected



voltage in the base winding of transistor Tr 1 which is positive with respect to the base. A voltage is applied to the collector of this transistor through a resistor *R* of 300 Ω (for shift registers with 2-mm cores) or 560 Ω (for shift registers with 3-mm cores), a collector winding on the subsequent core (K 2) and a collector winding on the first core.

As a certain positive threshold voltage at the base is required for silicon transistors to become conducting, no collector current can flow with zero volts across the base winding. A positive base voltage results in a collector current flowing through the collector winding of the first core and increasing the counterclockwise field. The collector winding on the second core is arranged such that the collector current generates a clockwise field in this core. Hence this winding has the same function as the input winding of the first core, and sets the second core. At the same time, a voltage pulse appears across resistor *R* which can be taken off at point A_1. The "1" bit has been transferred from the first to the second core. When the second shift pulse passes through winding II of the second core, the flux is reversed and the same processes as before are repeated for the second core: transistor Tr 2 becomes conducting, and its collector current flowing through a winding of the third core sets the flux in this core in a clockwise sense. The "1" bit has now been transferred from the second to the third core. The next shift pulse transfers the "1" bit to the fourth core, and so on.

If a "0" is written into the first core, the shift pulse does not change the flux direction; no voltage is induced in the base winding; no collector current can flow, and the counterclockwise flux direction previously written into the second core (by the shift pulse) is maintained.

A "1" fed into the front end is moved one step by each clock pulse. Two cores per bit are required in order to obtain the necessary delay and to avoid any influence of the subsequent on the preceding core. The core which is to receive the information must first be set to zero. Before it can be set to zero, the information stored so far ("1") must be transferred to the next core. Hence it is obvious that a new bit of information cannot be fed in with each successive clock pulse: It is impossible to transfer a "1" from one core to the next and to write in a new "1" simultaneously.

The double stages with 2 cores and 2 transistors need 2 clock pulses, one for the odd and one for the even cores. The shift frequency by which a bit is moved one step is equal to the frequency of the basic generator (here a sinewave oscillator). The clock pulses producing the output pulses of the respective cores have the same period as the sinewave, but are offset with respect to each other by half a period.

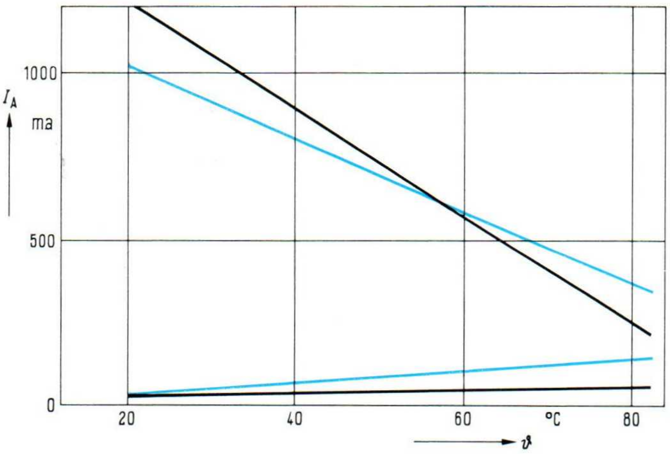


Fig. 4 Amplitude I_A of driving pulses as function of temperature for a 10-bit shift register with transistors and toroidal cores

Frequency of driving pulses 200 Kc
Pulse width about 2.5 μsec
Noise ratio 1 : 2

Black: 15-v collector voltage
Blue: 12-v collector voltage

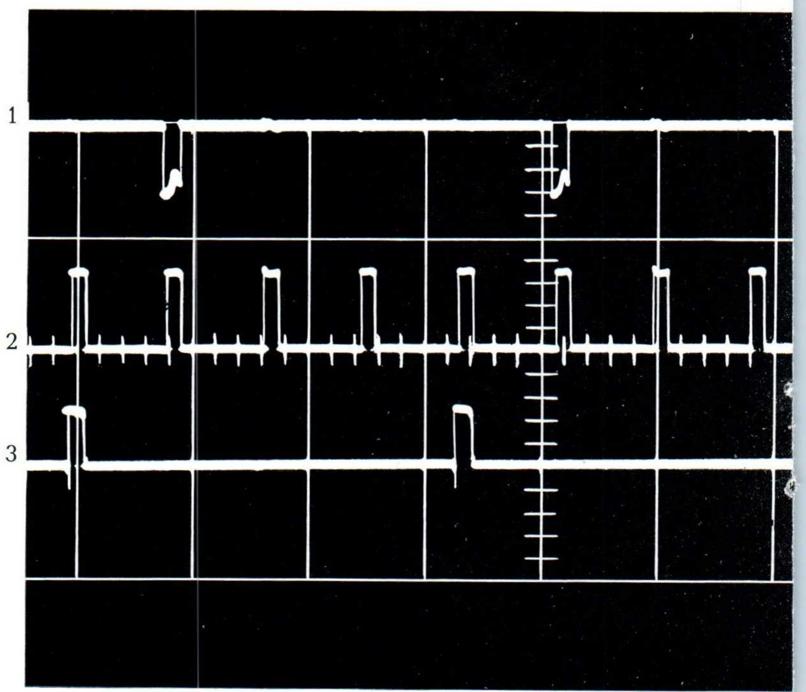


Fig. 5 Pulse shapes in a shift register with 4-mm toroidal cores and transistors

1 Information input 1, 0, 1, 0, 1, . . .	Vertical scale	1 cm ≙ 20 v
2 Clock pulses 30 Kc	Vertical scale	1 cm ≙ 0.2 amp
3 Output voltage	Vertical scale	1 cm ≙ 10 v
	Horizontal scale	1 cm ≙ 10 μsec

The size of the resistor in the collector circuit must be chosen smaller for an 8-core register than for a 4-core register. As the clock pulse generator falls short of an ideal current generator, pulse amplitudes are lower with 8 cores. Hence the collector current must be increased (by reducing the collector circuit resistance) unless the amplitude of the clock pulses fed in from the drive circuit can be further increased.

In Fig. 4, the shift pulse amplitude is plotted as function of temperature for a 10-bit shift register consisting of 10 double stages with 20 cores. The shift register operates smoothly at a supply voltage of 12 v if the amplitude is kept between the solid blue lines. The measurements were taken over a temperature range from 20 to 80° C. The amplitude tolerance is even greater at lower temperatures. Fig. 4 also demonstrates that rather large supply voltage variations are permissible.

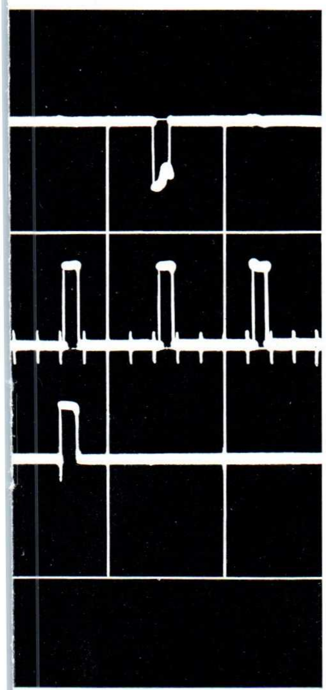
The curves were measured at a noise ratio of 2:1, i.e. the shift pulse amplitudes were such that a voltage pulse representing a "1" at the last shift register output was just twice as large as the worst noise pulse. Interference pulses are generated by mutual coupling between windings which does not, however, cause any flux reversals. As most practical applications do not require a temperature range beyond −10 to +80 °C, no temperature compensation has to be provided. For a shift register, the amplitudes of the information pulses ("1") may be varied from 100 to 900 ma; the shift pulse amplitudes can change between ±20 to ±80% according to Fig. 4 (depending on the ambient temperature).

Fig. 5 represents oscillograms of the input pulses (1—0—1—0, etc.), the shift pulses (both shift pulses shown superimposed), and the output pulses after the 4th transistor, all for a shift register with 4-mm cores. The shift frequency was 30 Kc; the noise ratio, i.e. the ratio of "1"-pulse amplitude to largest noise pulse amplitude was 10:1.

Fig. 6 shows the pulse shapes obtained from a 10-bit shift register with 2-mm cores: from top to bottom there are shift pulses (1 cm ≙ 0.5 amp), information input 0—1—0—1, etc. (1 cm ≙ 0.5 amp), the pulses at the base of the 20th transistor (1 cm ≙ 2 v), and the pulses at the collector of the 20th transistor (1 cm ≙ 20 v).

Applications

The shift register described can be put to many uses such as word memory, accumulator for computing operations, interim storage, series input and parallel output or vice versa, ring counter, and frequency divider. It also proved to be a good generator of the 4 clock pulses required for an all-magnetic shift register with saturable reactors.



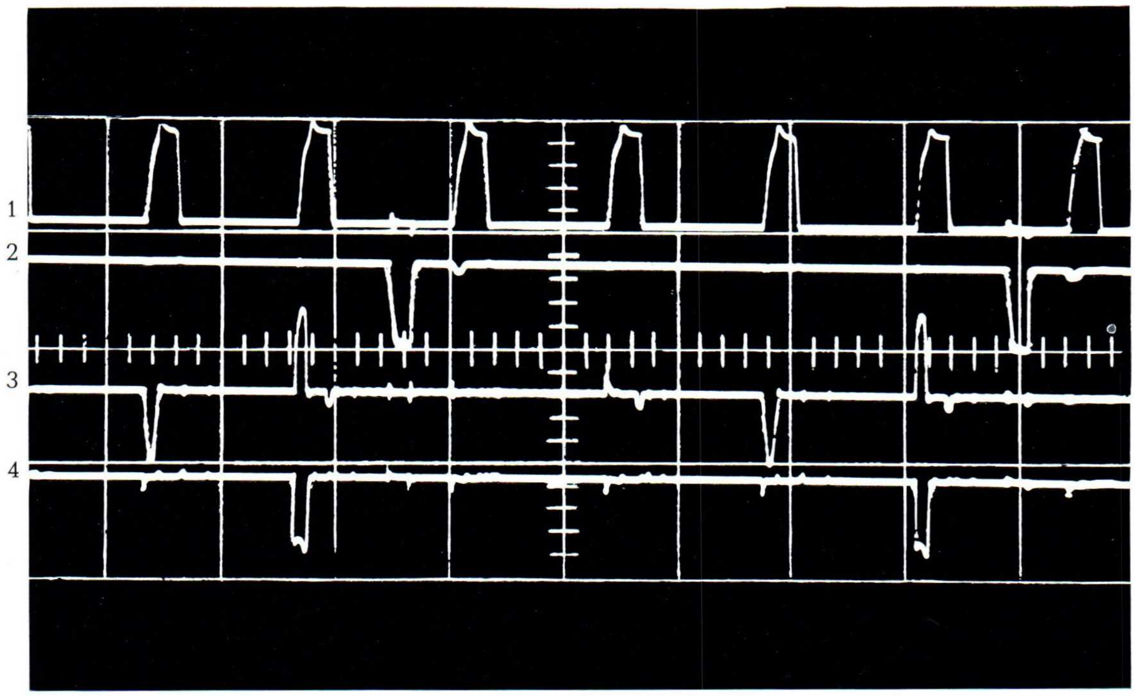


Fig. 6 Pulse shapes in a shift register with 2-mm toroidal cores and transistors

1 Clock pulses	Vertical scale	1 cm ≙ 0.5 amp
2 Information input	Vertical scale	1 cm ≙ 0.5 amp
3 Pulses at base of 20th transistor	Vertical scale	1 cm ≙ 2 v
4 Collector pulses	Vertical scale	1 cm ≙ 20 v
	Horizontal scale	1 cm ≙ 10 μsec

K 273 — A New Thermistor-Temperature Feeler for Liquids

By Hans Kaiser

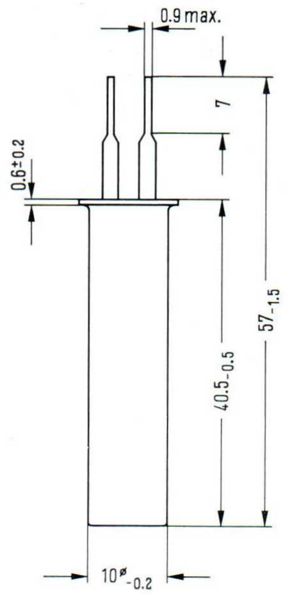
Thermistors are gaining more and more importance in the field of temperature regulation and measurement. With their high temperature coefficients and large intelligence signals, they are superior to thermoelements and metallic resistor thermometers. In comparison to expansion-type temperature feelers, thermistors offer a higher temperature accuracy and steady temperature settings over the entire temperature range. In constant contact with equipment producers, the Semiconductor Factory has developed a special thermistor-temperature feeler for liquids. It is especially suitable for installation in washing machines, dish-washing machines and hot water heating systems.

In practice it is especially often the case that liquid temperatures between 20 °C and 100 °C have to be regulated. Most of the time it concerns liquids whose chief component is water. The temperature feeler is fastened to the wall of the receptacle and should project sufficiently far enough into the liquid to insure that no measurement errors are caused by the receptacle wall or by the layer of liquid at the receptacle wall which is somewhat cooler than the rest.

Generally it is required that the temperature accuracy be about 2 °C at both ends of the temperature range of interest.

The temperature feeler developed by Siemens fulfils the requirements described above. Fig. 1 shows the dimensions of the temperature feeler. A thermistor disk is inserted in a 40 mm long feeler case and then cast in plastic to produce a sturdy component (Fig. 2). The thermistor leads are shaped into two flat terminals (2.8 mm × 0.8 mm) to be plugged into AMP-Faston-type plugs. At the plug end, the feeler has a flange which makes a tight installation possible. Fig. 3 shows a possible method for installing the K 273.

The tolerance on the rated resistance value is approximately ±20 % in batch production. In order to achieve the required temperature accuracy, the entire production tolerance of the K 273's resistance value of ± 20.5 % is divided into 10 tolerance groups (Table 1). The technical



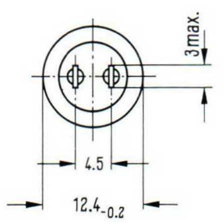


Fig. 1 Scale drawing of temperature feeler K 273

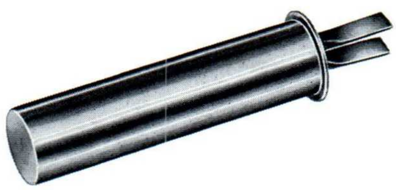


Fig. 2 K 273 life size

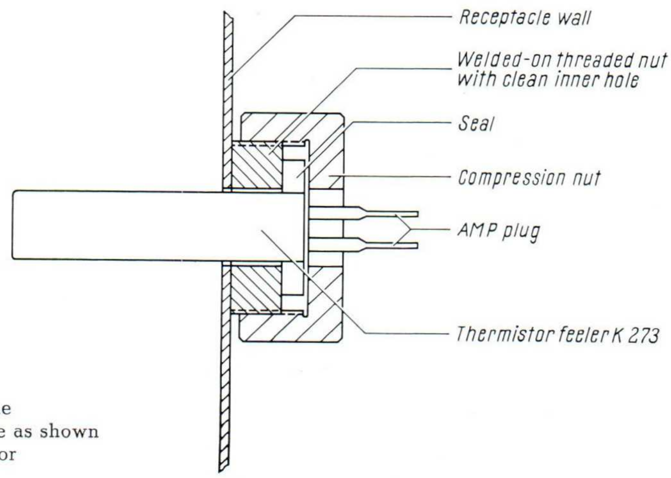


Fig. 3 Tight installation is made possible by flange on feeler case as shown in drawing of installed thermistor

Fig. 4 The 10 tolerance groups are indicated by different colored dots as shown in Table 1

Table 1
R_{60} tolerance groups

Group	Tolerance	Color
1	−20.5 ... −15.5 %	brown
2	−16.5 ... −11.5 %	red
3	−12.5 ... − 7.5 %	orange
4	− 8.5 ... − 3.5 %	yellow
5	− 4.5 ... + 0.5 %	green
6	− 0.5 ... + 4.5 %	blue
7	+ 3.5 ... + 8.5 %	purple
8	+ 7.5 ... +12.5 %	gray
9	+11.5 ... +16.5 %	silver
10	+15.5 ... +20.5 %	black

Table 2
K 273 technical data

Maximum ratings			
Maximum working temperature ϑ		100	°C
Characteristic data			
B value[1]	B	3930	°K
B value tolerance	B tol.	±3	%
Thermal conduction constant $G_{th\ amb}$		<50	mw/°C
Resistance value ($\vartheta_{Th} = 60$ °C) R_{60}		1.25	kΩ

[1] Obtained by measuring at 20 °C and 100 °C

data are given in Table 2. In Table 3 the corresponding maximum and minimum resistance values are given for the various thermistor temperatures.

The resistance which is usually measured at 20 °C or 25 °C is measured in this case at 60 °C so that the dispersion of the resistance values is practically the same at both ends of the temperature range. In addition, the tolerance of the B value has been reduced to ±3 %.

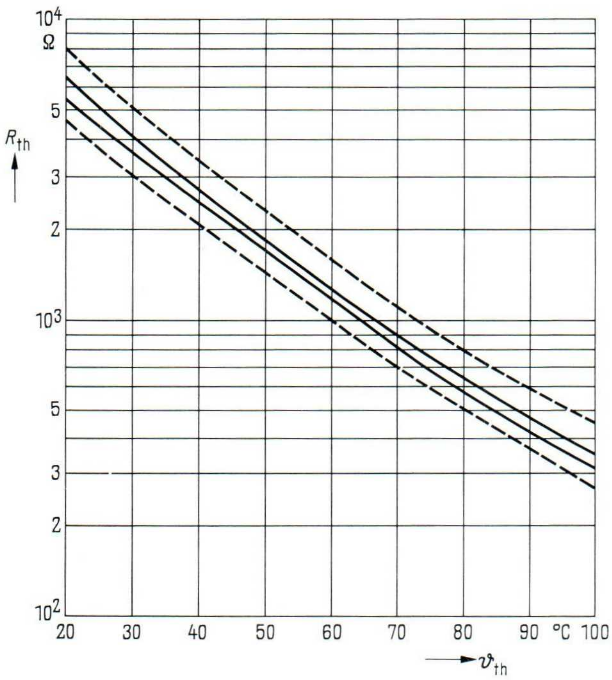


Fig. 5 Tolerance range of characteristic at an R_{60} tolerance of ± 20.5 % (dashed curve) and tolerance range of a single tolerance group from Table 3 (solid curve)

Fig. 5 shows the tolerance range of the $R = f(\vartheta)$ characteristic at a resistance tolerance of ±20.5 % in comparison with the tolerance range of the $R = f(\vartheta)$ characteristic of a single tolerance group.

As a consequence of the thermistor's large resistance change, only low amplification is needed for temperature regulation. In addition to this, the switch amplifier can in many cases, such as in washing machines and dishwashing machines, be used for other tasks (e.g. monitoring liquid levels) during those intervals where no temperature regulation takes place.

Fig. 6 shows a suggested circuit for liquid temperature regulation. The input circuit consists of a d-c bridge. The bridge resistors are modified by the user according to which tolerance group is installed. The thermistor-temperature feeler K 273 is found in the first bridge arm. The second bridge arm contains three switchable resistors which correspond to three desired temperature values. A silicon diode BA 103 is in series to these resistors. Because of its temperature characteristic, it performs the function of temperature compensation for the first transistor stage. The three switchable resistors can be replaced by a potentiometer if continuous value setting is desired. Each of the third and fourth bridge arms contains a 500 Ω resistor. The bridge circuit is followed by three BC 107's which form the switch amplifier proper. The last stage is a relay which switches the heating system on and off. In washing machines, the clutch magnet of the program switching mechanism for the relay can be substituted. A switch amplifier with only two transistors is also possible.

An overall curve of the temperature accuracy which can be obtained by using the K 273 is plotted in Fig. 7.

Thermistors and expansion-type feelers

Up to now, expansion-type feelers have been used a great deal for temperature regulation and measurement. A mechanically actuated needle is used as an indicator during the measurement process. For temperature regulation, the expansion-type feeler actuates a switch

Table 3

Maximum and minimum resistance values of the individual tolerance groups for different desired temperatures

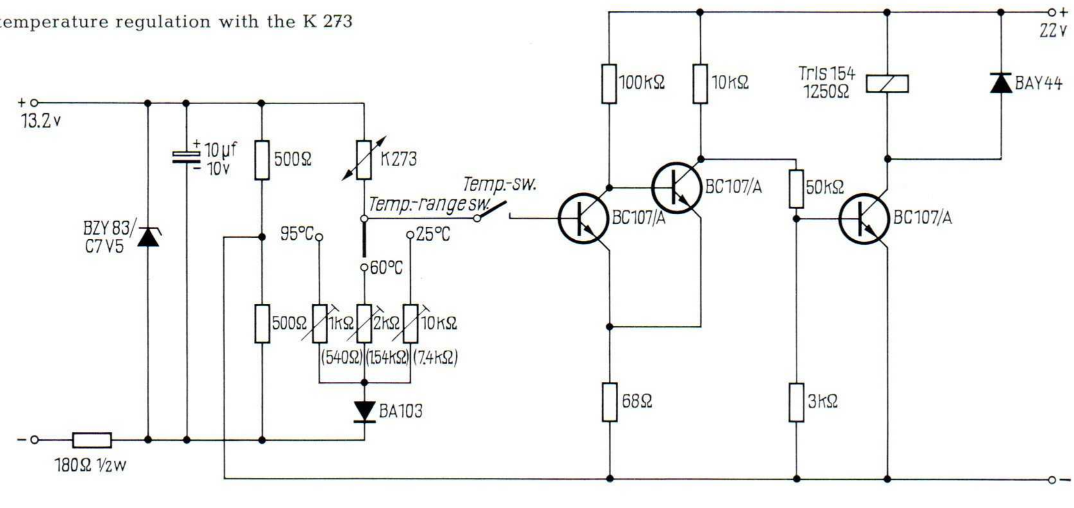
Resistance values in kΩ

ϑ_{Th} (°C)	Group 1		Group 2		Group 3		Group 4		Group 5		Group 6		Group 7		Group 8		Group 9		Group 10	
20	4.53	5.40	4.75	5.65	4.98	5.90	5.20	6.16	5.44	6.42	3.82	6.68	5.89	6.93	6.11	7.19	6.34	7.44	6.57	7.70
30	3.02	3.47	3.20	3.64	3.35	3.80	3.51	3.97	3.66	4.13	2.57	4.30	3.96	4.46	4.11	4.63	4.27	4.79	4.42	4.96
40	2.05	2.28	2.16	2.39	2.26	2.50	2.36	2.61	2.47	2.73	1.77	2.82	2.67	2.93	2.78	3.04	2.88	3.15	2.99	3.23
50	1.415	1.536	1.484	1.610	1.560	1.68	1.63	1.755	1.700	1.83	5.66	1.90	1.84	1.97	1.915	2.025	1.985	2.115	2.05	2.19
60	0.994	1.056	1.043	1.107	1.093	1.156	1.142	1.207	1.193	1.256	1.243	1.307	1.293	1.356	1.343	1.406	1.391	1.455	1.441	1.509
70	0.696	0.766	0.731	0.804	0.766	0.840	0.800	0.876	0.837	0.912	0.871	0.949	0.906	0.985	0.941	1.021	0.975	1.057	1.010	1.096
80	0.496	0.550	0.521	0.577	0.546	0.603	0.571	0.629	0.597	0.655	0.622	0.681	0.647	0.706	0.672	0.733	0.696	0.759	0.721	0.786
90	0.361	0.406	0.379	0.426	0.397	0.445	0.415	0.465	0.434	0.484	0.451	0.503	0.470	0.522	0.488	0.541	0.505	0.560	0.524	0.581
100	0.266	0.305	0.279	0.320	0.292	0.334	0.305	0.348	0.318	0.363	0.332	0.377	0.346	0.392	0.359	0.406	0.372	0.421	0.385	0.436

mechanically which operates the heating system. Since expansion-type feelers operate mechanically, they are relatively unreliable and inaccurate.

With thermistor-temperature feelers a substantially higher degree of accuracy and reliability is achieved. Today the artificial preaging of thermistors can be done so well that the operational aging of the components is practically negligible. Because of this fact, reproducible measurement results and regulation settings are assured. It is true that for temperature regulation with thermistors a switch amplifier is necessary, and for temperature measurements a milliammeter. But due to the economical way in which semiconductors are produced nowadays, the combined costs of the thermistor-temperature feeler and switch amplifier are not higher than those of an expansion-type feeler.

Fig. 6 Suggested circuit for temperature regulation with the K 273



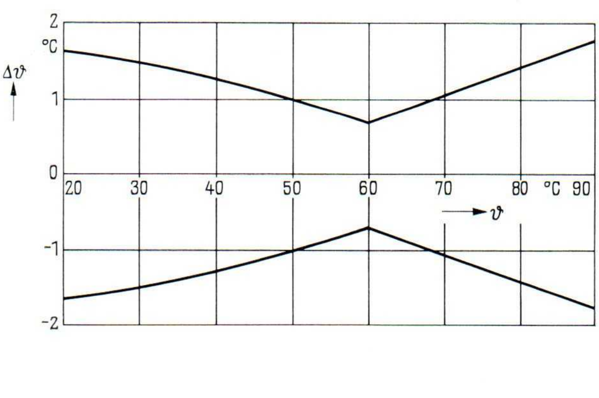


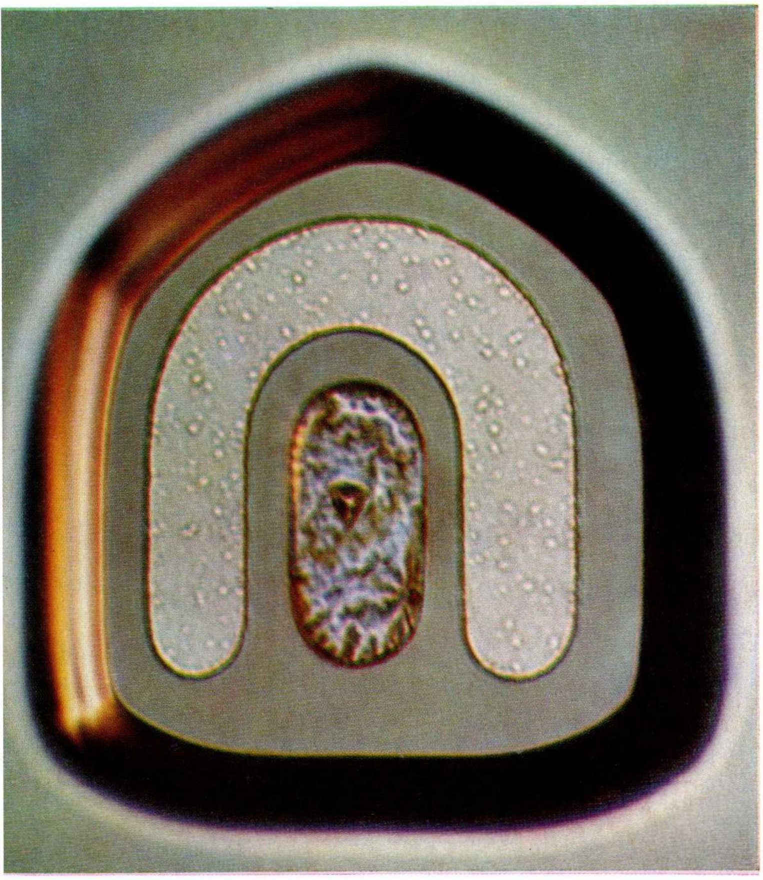
Fig. 7 Overall curve for temperature accuracy obtainable with K 273 as a function of desired temperature

A Low-Noise UHF Antenna Amplifier Using the AFY 42 Transistor

By Josef Schelle

The incessant sophistication of modern semiconductor technologies led to the manufacture of transistors with remarkably favorable high-frequency characteristics. The mesa II method used by Siemens for the first time permits the manufacture of uhf transistors combining an extremely low noise figure with high power gain. The electrical properties of the new Siemens AFY 42 transistor are illustrated by means of a typical circuit for a single-stage antenna amplifier.

Fig. 1 The AFY 42 is a mesa II germanium transistor



In areas where the receiving field strengths are low the quality of the television picture tends to be deteriorated by disturbing picture noise. A measure of the picture quality is the signal-to-noise ratio, which is the ratio between the useful voltage offered to the receiver and the noise voltage generated mainly in the receiver itself. If the local field strength, the antenna gain, and the attenuation of the antenna cable are assumed to be given quantities, the signal-to-noise ratio may be markedly improved by connecting an extremely low-noise amplifier between the antenna and the antenna cable. Ideally, the power gain of this amplifier should be so high that the input stages of the receiver have no degrading influence on its noise characteristics. The development of the new AFY 42 transistor meant decisive progress in this respect. Uhf antenna amplifiers using the AFY 42 reach noise figures which are between 3.5 and 4.5 db on the average. This means that a signal-to-noise ratio of 100 : 1 (40 db), commonly regarded as absolutely necessary for a good television picture, can be realized if signal voltages as low as 160 to 180 μv (across 60 Ω) are picked up. Moreover, a transistor amplifier may be designed to be so compact as to fit into the junction box of a television antenna. This arrangement, allowing the antenna voltage to reach the amplifier without attenuation, secures optimum signal-to-noise ratio and picture quality.

The AFY 42 transistor

The AFY 42 is a mesa II p-n-p germanium transistor (Fig. 1) in a standard TO-18 case. The system is electrically isolated from the case. On account of its high oscillation frequency, the low collector junction capacitance and the low noise figure, this transistor is particularly well suited for uhf circuits.

Circuitry of the single-stage antenna amplifier

The electrical and mechanical design features were determined mainly with a view to obtaining a low noise figure, a simple circuit layout, and small outside dimensions (Fig. 2). As cross-modulation is not very likely to occur, the amplifier input has been given wideband character in the uhf range. Interfering signals of low frequency are suppressed by the highpass filter between the antenna input and the emitter. In addition, this highpass protects the base/emitter diode of the transistor from dangerous overvoltages likely to occur during thunderstorms. The AFY 42 is operated in a grounded-base configuration with the operating point of $-I_C = 1.6$ ma. The transistor case is connected to the chassis through a reliably contacting support. This ensures very stable operating conditions throughout the frequency range. A $\lambda/4$ resonant line circuit, which can be tuned to the desired channel in band IV or V with a 7-pf trimmer, is connected to the collector. Besides, attenuation is provided by the 8.2-kΩ resistance of the base divider, securing sufficient bandwidth even in the lowermost frequency region. The amplified input signal is fed to a 60-Ω load through one tap of the resonant line circuit. Provided that suitable baluns are used, the amplifier is also suitable for 240-Ω systems.

Electrical characteristics

Fig. 3 shows the noise figure, the power gain, and the reflection coefficient of the amplifier measured at the input and output as a function of frequency. The usual fluctuations of temperature and operating voltage change these values only slightly. A loss in gain of only 2 db has been measured at 860 Mc and an ambient 83 °C. As the heat generated in the transistor raises the local temperature by only 7 °C, this ambient temperature is permissible in extreme cases. The bandwidth of the amplifier is between 16 and 48 Mc.

Fig. 4 shows the cross-modulation encountered. U_{int} is half the EMF of a 100 % sinewave-modulated television carrier which causes the useful carrier to be 1 % amplitude-modulated. The permissible interference voltage was determined for the useful frequencies of 500 and 800 Mc.

Let us also mention the very good stability of the amplifier against self-sustained oscillations owing to the low internal reverse transfer of this transistor. For the reverse power gain of the amplifier, values between −37 and −42 db were measured.

Significant characteristics of the AFY 42 transistor

Dynamic characteristics (ϑ_U = 25 °C)

Transition frequency ($-I_C$ = 2 ma; $-U_{CE}$ = 10 v; f = 100 Mc)	f_T	650	Mc
Collector-base time constant ($-I_C$ = 2 ma; $-U_{CE}$ = 10 v; f = 30 Mc)	$r_{bb'}, c_{b'c}$	2	psec
Maximum frequency of oscillation	f_{max}	3.6	Gc
Short-circuit reverse transfer capacitance (I_C = 2 ma; $-U_{CE}$ = 10 v)	$-C_{12e}$	0.25	pf
Common-base power gain ($-I_C$ = 2 ma; $-U_{CE}$ = 10 v)			
f = 900 Mc; R_L = 500 Ω	V_{pb}	11	db
f = 900 Mc; R_L = 2 kΩ	V_{pb}	13	db
Noise figure ($-I_C$ = 2 ma; $-U_{CE}$ = 10 v) (f = 900 Mc; R_g = 60 Ω)	F	<7	db

D-c characteristics (ϑ_U = 25 °C)

Current gain ($-I_C$ = 2 ma; $-U_{CE}$ = 10 v)	B	33 >10	
Collector cutoff current (U_{CBO} = 20 v)	$-I_{CBO}$	0.5 (<3)	μamp

Thermal resistance

Collector junction — air	$R_{th\,U}$	<750	°C/w
Collector junction — Transistor case	$R_{th\,G}$	<400	°C/w

Maximum ratings

Collector/emitter voltage	$-U_{CEO}$	25	v
Collector/base voltage	$-U_{CBO}$	30	v
Emitter/base voltage	$-U_{EBO}$	0.3	v
Collector current	$-I_C$	10	ma
Junction temperature	ϑ_j	90	°C

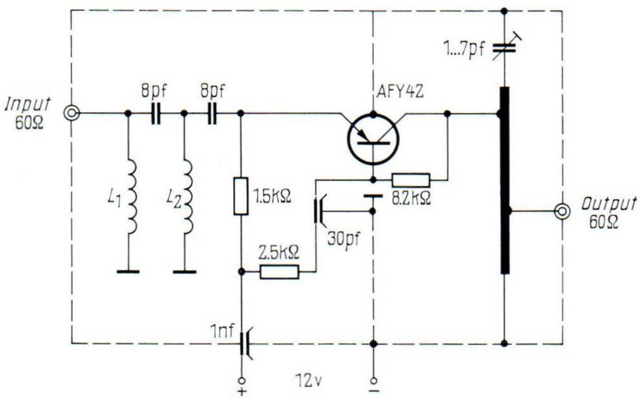


Fig. 2 Circuit layout of a uhf antenna amplifier
Frequency range 470 to 860 Mc
Operating voltage 12 v, input current 2.7 ma
Operating point of the transistor $-I_C$ = 1.6 ma; U_{CE} = 9.6 v
Resonant line circuit: width 10 mm, depth 15 mm, length 45 mm
Inner conductor: length 35 mm (1 mm diam. silver-coated copper)
Collector tap: 9 mm from hot end
Output tap: 5 mm from cold end
Input reactance coil L_1: 8 self-supporting windings, 3.5 mm diam.
Highpass coil L_2: 2 self-supporting windings, 3.5 mm diam.
Dimensions of amplifier: 45 mm × 20 mm × 15 mm

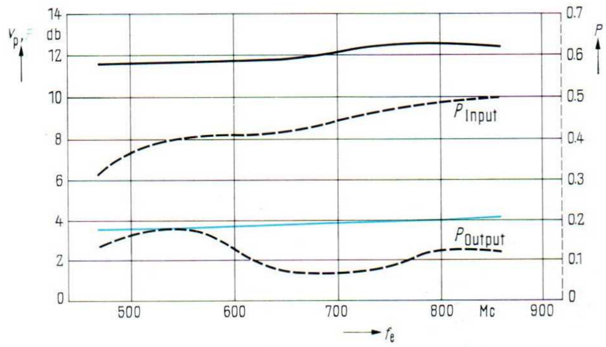


Fig. 3 The usual fluctuations of temperature and operating voltage will cause only minor changes in the values of noise figure, power gain, and reflection coefficient

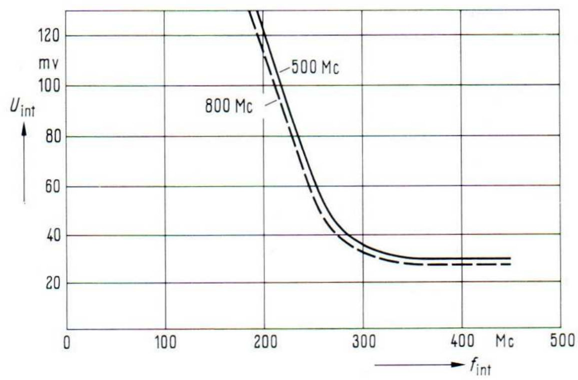


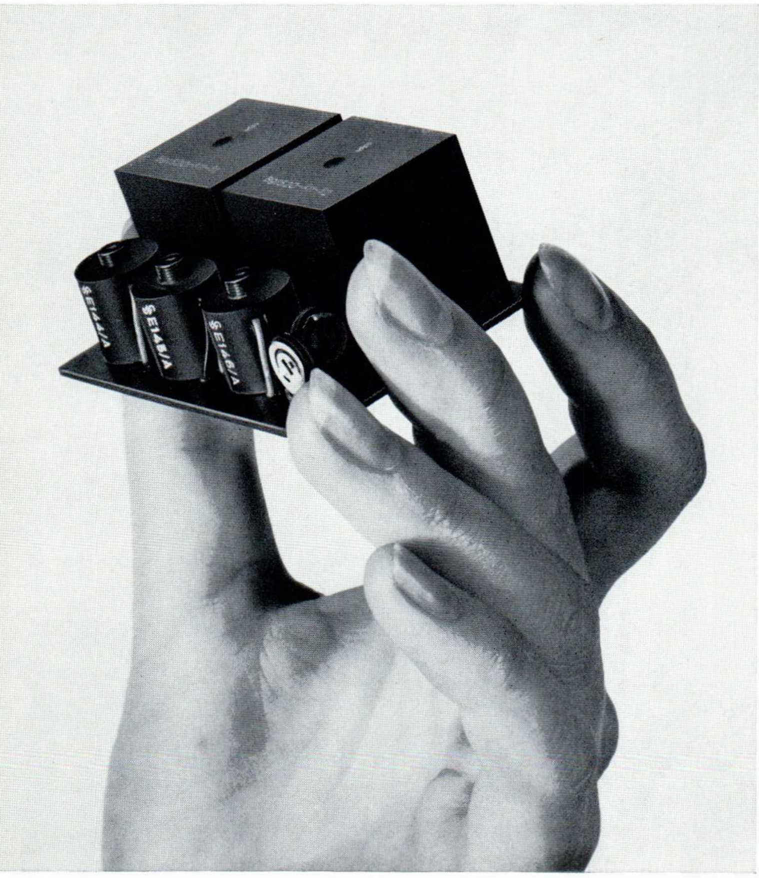
Fig. 4 Interference voltage for 1 % cross-modulation
Useful voltage: 1 mv
Useful frequencies: 500 and 800 Mc

Stereo Decoder Developed as Siemens Module "R"

By Karl Langecker

Universal modules are units suitable for use in a great variety of applications. They must be designed to operate under very different conditions without requiring a more complex layout. A typical universal module of this kind is the stereo decoder (Fig. 1) described below.

Fig. 1 Complete stereo decoder developed as Siemens module "R"



When talking about the use of modules in communications engineering, even the expert will think mainly of electronic equipment with a great number of identical building blocks, as used in data processing systems. In home-entertainment sets they are hardly known as yet or are not regarded as modules in the current sense of the word. On the other hand, more or less complex functional units, prefabricated from components, have been in use in sound and television broadcast sets for over 30 years, and are well known in the radio industry. A typical module is the power supply unit in a receiver set, which is prefabricated and mounted as a complete unit. Many other components, such as coil sets, wavetraps or 9-Kc filters, are among the modules which have seen long use. The introduction of vhf and television broadcasting added some further modules recently, above all the vhf tuners, the pushbutton sets with the integrated bandswitch coils, and the vhf and uhf tuners in television sets.

The introduction of printed circuits about ten years ago changed module design radically. The manufacturers provided units that were completely equipped, wired and pretested. As the industrial development and design departments improved their command of circuit board engineering, these originally small modules were gradually combined to form large printed plates, a technique which cut down the wiring during the completion phase of the set. This development was aided by the widely used semi-automatic equipping process for conductor boards, which requires less capital expenditure for equipment when large plates are involved. However, the large plate is not without disadvantages, because a conductor board carries a great number of components. There are undesired side effects to this method: the plate is more likely to be fitted with wrong components, and it has to run through several stations before it is completed. This already outlines the manufacturing process to be aimed at: less components to be fitted, i.e. modules that are completed and tested when leaving the factory. Strictly speaking, this in a way means returning to the modules of the small conductor boards already mentioned, except that the modules are of a different form and plug into a common conductor board like the usual components. But this method will be of interest to the equipment manufacturer only if he may expect further considerable advantages. The high packing density of modern components, i.e. the low space requirements alone, will only in a few cases be a sufficient stimulus (in the case of portables, for instance).

As integrated circuits, such as thin-film and solid-state circuits, are not yet available on a large scale for the home-entertainment sector, modules consisting chiefly or exclusively of inexpensive conventional components may be of increasing interest to the equipment manufacturer. This leads to the question of how to combine components into modules.

Siemens module "R"

Siemens developed a compact-style unit under the designation Siemens module "R", fulfilling these requirements [1]. The method used also provides for the step-by-step inclusion of future thin-film and solid-state circuits, so that the modules can always be updated to embody the latest state of the art. A comparison alone of their price with that of the hitherto used components and the associated equipment costs hardly reveals the advantage provided by the new module. It is rather the indirect costs the equipment manufacturer saves through simplified stockkeeping and ordering and through the possible reduction of the number of testing procedures and workshop places required for equipping the sets that are of paramount importance. Even the capital expenditure for development and construction can be slashed. In addition, the reliability and hence the quality of the ultimate product are improved.

The design of the Siemens module "R" permits of two possibilities: building customer-tested circuits in the form of modules equivalent to those circuits (i.e. the modules are tailored to fill the special requirements of the particular customer concerned), or developing universal modules suitable for a large number of customers. As such universal modules must under no circumstances impair the "individualization" of the set, they must meet a number of extra requirements on top of the pure functional tasks without noticeably increasing the complexity of layout, and thus the costs of the set. That it is indeed possible to construct modules for universal use in these conditions will be shown by way of the stereo decoder.

[1] Still, H.: Compact Circuits, Siemens R-Type Assemblies. Siemens Electronic Components Bull. **I** (1966) pp. 54 to 58

The stereo decoder as a universal module

The function of the stereo decoder is to convert the stereo multiplex signal (MPX signal) from the f-m detector to the audio signals that correspond to the two stereo channels. The crosstalk between the two channels should be negligible, and the transmission must not be impaired by any audible noise voltages. The following requirements must always be fulfilled:

1. The stereo decoder must have such a high crosstalk attenuation over a sufficiently wide audio range that the total crosstalk attenuation of the set is chiefly determined by the transmitted stereo signals and by the envelope delay of the receiver.

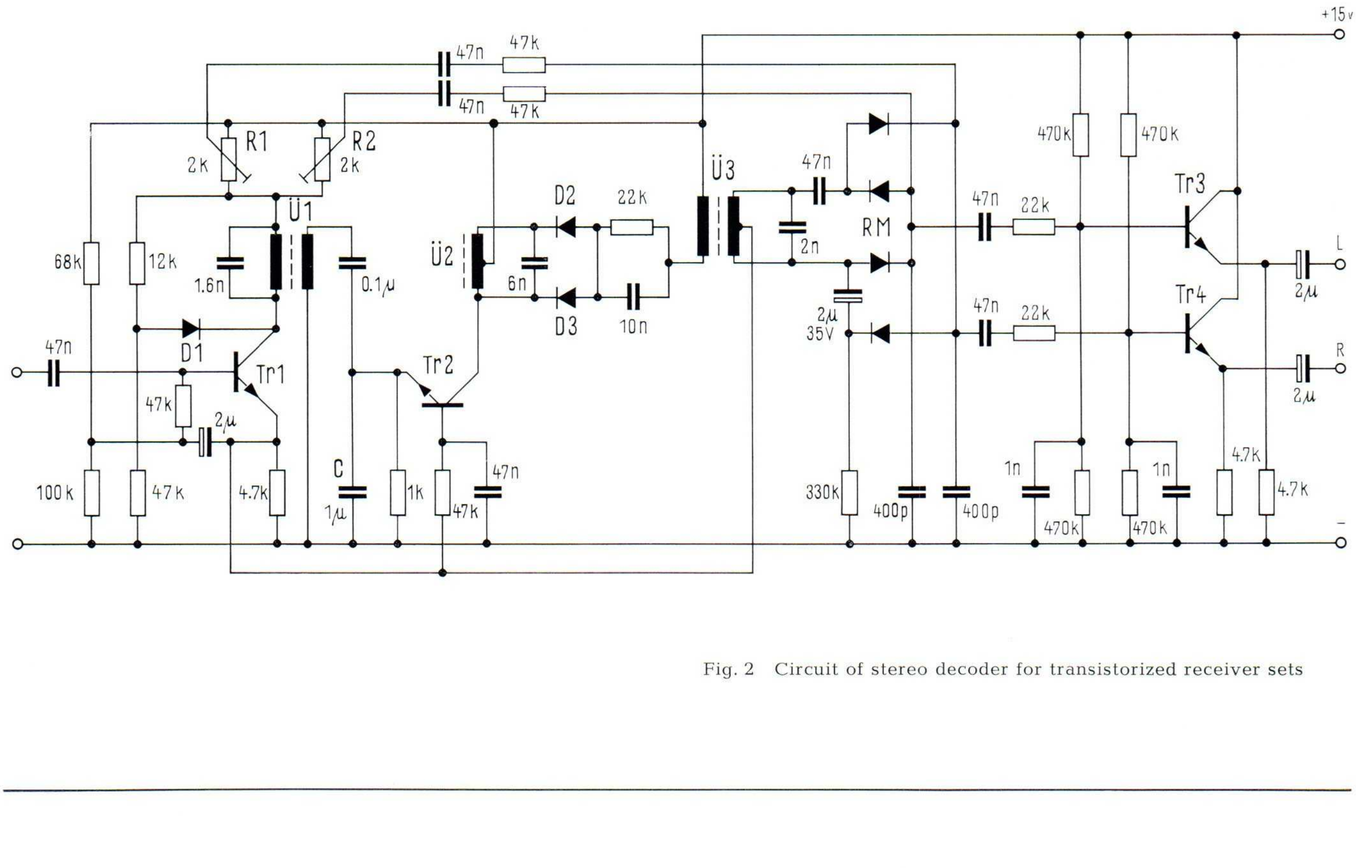


Fig. 2 Circuit of stereo decoder for transistorized receiver sets

2. No combination frequencies apt to deteriorate the transmission quality (such as distortion or difference frequencies through interaction with the pilot or the locally supplied carrier) must be generated in the stereo decoder.
3. Switching from stereophonic to monophonic transmission, and vice versa, shall be automatic with a negligible level difference.

From the condition of universal applicability it follows that:

4. The input impedance must not present any disturbing load on the f-m demodulator; it must be sufficiently high even when compared to high-impedance demodulator circuits.
5. The decoder must have a wide range of permissible input voltages because the MPX signal voltages of the f-m demodulators are very different in different broadcast receivers and signal voltage changes depending on the antenna voltage may occur, the amount of change depending on the efficiency of the limiter stages.
6. The decoder must fulfil all the requirements mentioned over a wide range of operating voltages, because the available operating voltages differ from set to set. Moreover, it must not be impaired by variations of the operating voltage to eliminate the need for stabilizing the latter.

The developed decoder meets all these requirements to a very wide extent.

Layout of the stereo decoder

An impedance transforming stage at the decoder input (Fig. 2) secures an input impedance of about 200 kΩ, which places hardly any load on the usual f-m demodulators. The stereo multiplex signal (MPX signal) is applied from the emitter of transistor Tr 1 of this stage directly to ring modulator RM at the decoder output through the secondary of transformer T 3. To the pilot tone the input stage is a strongly degenerative grounded emitter configuration. The gain of this stage is given more or less by the ratio of the resonant impedance of the collector circuit to the emitter resistance and is thus little affected by variations of the transistor parameters. The input impedance to the pilot tone is negligibly reduced. A biased diode (D 1) attenuates the resonant circuit output so as to deliver a constant-amplitude signal to eliminate variations of the input voltage and thus of the pilot tone. Although in these frequency ranges a grounded-base circuit provides less gain than a grounded-emitter circuit, it has been selected for the second pilot-tone amplifier stage (Tr 2). Its advantage lies in the fact that its gain is essentially independent of variations in the transistor parameters, because it is only determined by the transadmittance which depends upon the emitter current and is approximately the same for all transistors customarily used in these frequency ranges. The input is fed to the emitter of this stage through a very low-impedance secondary of the preceding resonant circuit. Capacitor C, connected in parallel to the emitter, shifts the phase of the pilot tone by 90° as required. The bias to the base of the transistor is taken from the emitter of the transistor in the first stage through a resistor. A two-diode doubler circuit (D 2, D 3) converts the 19-Kc pilot tone at the second-stage collector to the 38-Kc frequency of the local carrier. The local carrier passes through the following *R-C* network, which improves the efficiency of the doubler circuit, the third transformer T 3, whose secondary is tuned to resonance, and through the ring modulator.

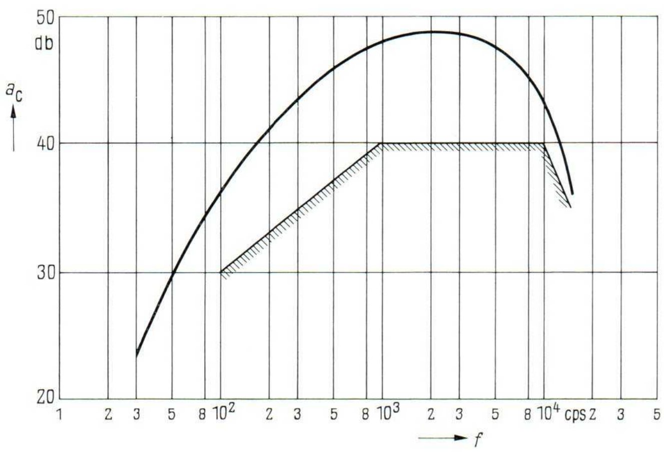


Fig. 3 Crosstalk attenuation a_c of the stereo decoder for transistorized broadcast receivers as a function of frequency

Upper curve: Mean attenuation
Lower curve: Guaranteed minimum attenuation

Operating voltage range 10 to 20 v,
MPX input voltage range 50 to 500 mv rms

The ring modulator works as a peak-type rectifier. The applied local carrier voltage is only so high that low-distortion demodulation is secured even when the MPX signal is at the permissible maximum, reducing the additional noise depending on the amplitude of the local carrier to an optimally low value. A d-c voltage applied in forward direction boosts the efficiency especially when low signals are involved; in monophonic broadcasts it also through-connects the audio signal through the diodes, so that no special monophonic/stereo switching arrangement is required.

When sideband information is detected, the message content of the sidebands is decreased to a value corresponding to the demodulation efficiency. However, as the relation between the sum signal and the difference signal contained in the sideband must be retained to secure maximum crosstalk attenuation, a proportional decrease of the sum signal is necessary. To this end, a 180° out-of-phase voltage of the sum signal is taken from the two adjustable potentiometers R 1 and R 2 in the collector circuit of the first stage for properly dosed application to the output of the ring modulator. It is always possible in the range of a constant demodulation efficiency to obtain maximum crosstalk attenuation (Fig. 3). As the efficiency depends on the magnitude of the local carrier voltage, the latter must be optimally stabilized. As already mentioned, the limiter action of diode D 1, connected in parallel with the collector circuit of the first stage, keeps it adequately independent of the voltage of the pilot tone at the decoder input. Hence the MPX input voltage, whose ratio with the pilot tone voltage is invariable, may change by a factor of about 10 before crosstalk attenuation exceeds the tolerance range.

The crosstalk attenuation remains optimum up to operating voltage fluctuations of 1:2. Increasing the operating voltage means an increase in the bias voltage to the limiter diode and in the pilot tone voltage at the resonant circuit of the input stage. Since the emitter current of the second amplifier stage increases also, the input impedance of this stage goes down, reducing the input voltage. A rise in the voltage to the frequency doubler increases the load on the collector circuit of the second stage and also reduces the increase in voltage. Nevertheless, the local carrier voltage to the ring modulator increases, although the rise is slowed down considerably, and improves the demodulation efficiency. To achieve optimum crosstalk attenuation, the component of the 180° out-of-phase sum voltage supplied by the two adjustable potentiometers at the output of the ring modulator must be decreased. The effective voltage is a function of the voltage divider ratio composed of the series resistance between the adjustable potentiometers and the ring modulator output, as well as the resistance of the ring modulator in the direction of the very low output impedance of the impedance transformer. The resistance of the ring modulator decreases as the local carrier voltage increases, which results in a relative decrease of the 180° out-of-phase sum voltage at the ring modulator output with respect to that across the adjustable potentiometers. This decrease of the 180° out-of-phase sum voltage can be controlled by selecting suitable series resistances so that crosstalk attenuation remains an optimum although the local carrier voltage and thus the demodulation efficiency increase. A major change of the voltage-dependent resistance of the ring modulator, apt to occur if the following audio amplifier were overloaded, would interfere with this careful design. The load resistance should therefore be at least 200 kΩ.

In cases where the input impedance of the following stereo audio amplifier is lower, a dual impedance transformer module is available providing standardized deemphasis. Its output impedance of about 100 Ω is adequate for all loads encountered.

This example of a stereo decoder has shown that a sophisticated broadcast circuit arrangement can be designed as a universal module without any compromise. It does not call for more components, standard or special, than would be necessary anyhow. As compared to some other decoders on the market, the described module is rather less complex.

The stereo decoder presented here is chiefly intended for use in partially or fully transistorized broadcast receiver sets where the signal voltages of the f-m demodulator are relatively low. A second version will be available in the near future, designed with conventional tube sets in mind. With the exception of the MPX input voltage rating, covering about 0.5 to 2.5 v rms, the technical data for the second model remain the same as for the first model listed in the table, as do layout and dimensions.

Electrical data of the stereo decoder for transistorized broadcast sets

Operating voltage E_0	15 ± 5 v
Input current at E_0 = 15 v	about 10 ma
Range of input voltages (UPX signal) E_{in}	about 50 to 500 mv rms
Minimum values of crosstalk attenuation in the ranges of	
100 cps to 1 Kc	> 30 db
1 Kc to 10 Kc	> 40 db
10 Kc to 15 Kc	> 35 db
Distortion factor at E_{in} = 500 mv rms (Fig. 3)	
Stereo	< 0.6 %
Monophonic	< 0.5 %
Interference introduced by local oscillator	> 45 db down
Residual local carrier (38 Kc)	< 5 mv
Transmission loss	
Stereo	2 db
Monophonic	3.5 db
Input impedance	200 kΩ shunted by 15 pf
Output impedance with impedance transformer	100 Ω in series with 2 µf
Lowest permissible load without impedance transformer	200 kΩ

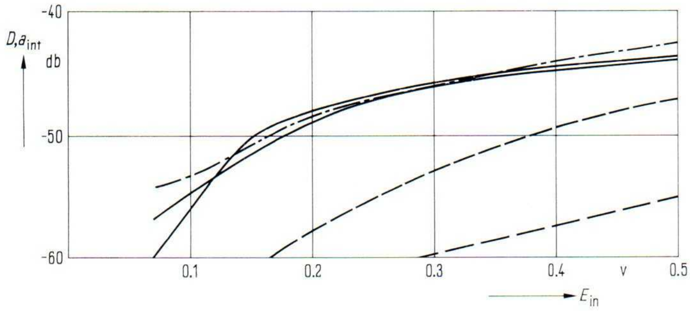


Fig. 4 Signal-to-distortion ratio D and intermodulation a_{int} of the stereo decoder for transistorized broadcast receivers as a function of the MPX input voltage E_{in} for a modulation frequency or an intermodulation tone of 1 Kc

Dashed-dotted curve: $a_{int} = f(E_{in})$

Solid curves:
Upper curve: $D_2 = f(E_{in})$ } Stereo
Lower curve: $D_3 = f(E_{in})$

Dashed curves:
Upper curve: $D_2 = f(E_{in})$ } Monophonic
Lower curve: $D_3 = f(E_{in})$

Subscripts 2 and 3: 2nd and 3rd harmonics

Ferrites with Initial Permeabilities from 3,000 to more than 20,000

Used in Miniature Transformers for Nanosecond-Pulse Techniques

By Erich Röß

Recent progress in the development of high-permeability ferrites covers many areas. As one of the most important improvements, the initial permeability could be increased considerably beyond customary values. For example, the permeability values of the new SIFERRITES N 30, T 35 and T 38 are $\mu_i = 3{,}500$; $\mu_i = 5{,}000$; and $\mu_i = 10{,}000$, respectively. Cores with permeabilities above 10,000 are already available for certain applications. Initial permeability values above 30,000 have been reached in the laboratory, and hence it is to be expected that in the future users can also be supplied with similar ferrites. In the following, the magnetic properties of these ferrites will be discussed briefly, and it will be shown by means of examples how these materials have opened up new possibilities. We will first summarize the physical and technical requirements for attaining such a permeability increase.

Among magnetic materials, ferrites are distinguished by a high resistivity as compared to metals, an advantage due to the semiconductor properties of these ceramics. Up to now, however, ferrites had not only a lower magnetization at saturation than metals, but above all a lower permeability which could not be increased over a value of $\mu_i = 2{,}000$ for a long time.

The permeability of a magnetic material is determined by the magnetization at saturation as well as by the

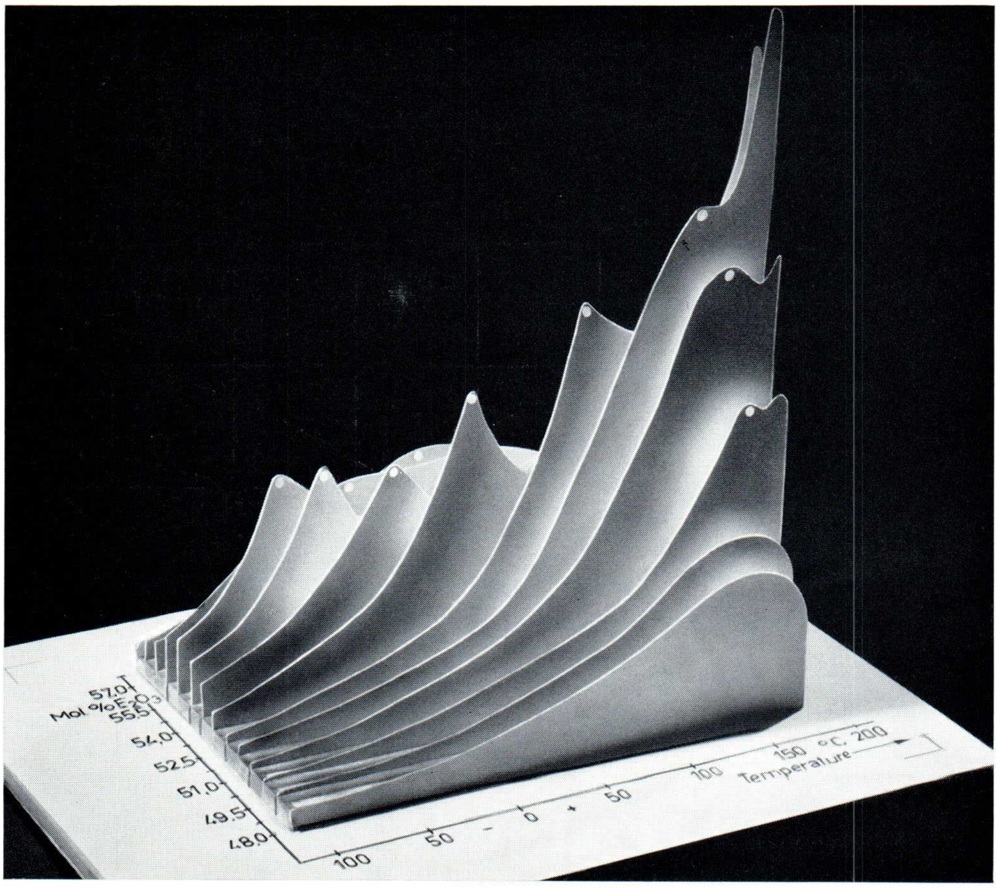


Fig. 1
Model of "permeability mountains" for a small range of manganese-zinc ferrites

various energies influencing the magnetization process, mainly the energy of crystal anisotropy which expresses the relationship between direction of magnetization and crystal lattice, and the energy of stress anisotropy which results from imperfections of the crystal structure. The permeability is the higher the larger the magnetization at saturation and the lower the crystal and the stress anisotropy. As the saturation flux of ferrites cannot be materially increased due to basic limitations, a permeability gain must be achieved by reducing the crystal anisotropy and the stress anisotropy. Both forms of energy depend in a complicated manner on the ferrite's chemical composition, on the manufacturing process, and also on temperature. In other words, ferrites of different composition exhibit widely differing permeability values which, in turn, may be quite different functions of temperature.

Fig. 1 shows a three-dimensional model of the "permeability mountains" for a small range of manganese-zinc ferrite compounds. The permeability-temperature curves of 14 cores with the same Curie temperature of 170° are represented in the range from 48 to 58 mole% Fe_2O_3.

It can be seen that so-called abnormal curves with a secondary maximum exist alongside the normal permeability-temperature curves in the foreground where the permeability rises steadily to a maximum located close to the Curie temperature.

The secondary maxima occurring in a number of manganese-zinc ferrites constitute one of the reasons for the particularly high initial permeability of these materials, i.e., if the composition is chosen such that this maximum is located near room temperature, much higher permeability values are obtained there than for a normal curve, regardless of the Curie temperature. This maximum, which does not have to be very distinct—as will be shown later—is caused by the fact that the crystal anisotropy goes through zero at the respective temperature. If one desires to increase the initial permeability even further, the stress anisotropy must also be kept as small as possible. This can be accomplished under certain conditions, for instance by using raw materials of high purity and by controlling the manufacturing process in such a way that the crystal structure is as uniform as possible. It is apparent from Fig. 2 that the permeability increases as the mean grain diameter of the polycrystalline ferrite grows larger.

High permeability values at room temperature

If all these factors are taken into account, cores can be obtained with an initial permeability exceeding 20,000 at room temperature. The permeability-temperature curves of N 30, T 35 and T 38 ferrites as well as of several ferrites with higher permeability are plotted in Figs. 3

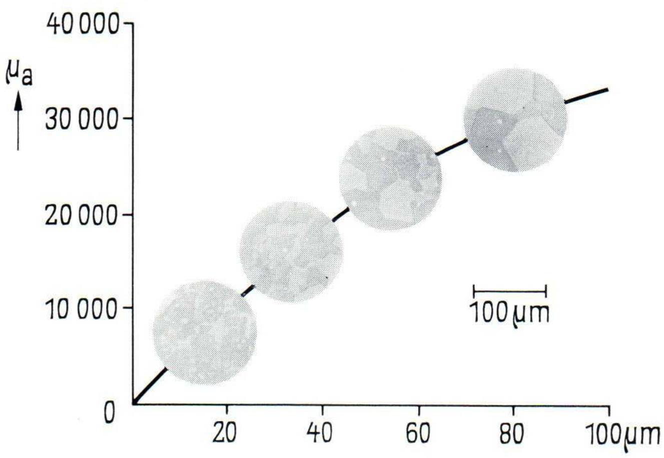


Fig. 2 Increase of initial permeability with mean crystal diameter
Micro photographs of polycrystalline manganese-zinc ferrite structures

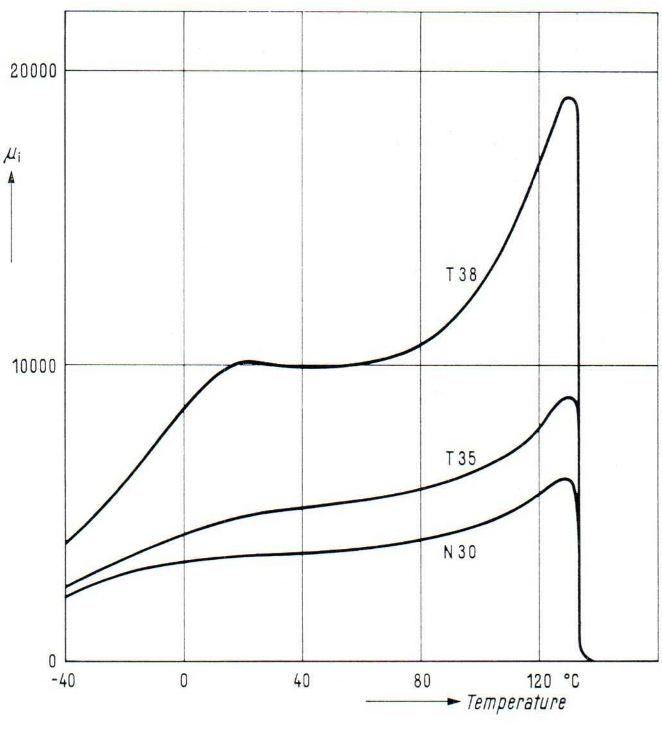


Fig. 3 Permeability-temperature curves for N 30, T 35, and T 38 ferrites

and 4 over a temperature range from -40 to $+140$ °C. The secondary maximum near room temperature becomes clearly distinct only at extremely high permeabilities. Cores with permeabilities from 3,000 to 10,000 have temperature coefficients of correspondingly small positive or negative values in the range above room temperature.

Permeability changes but little with time. The relative disaccommodation $\frac{i_z}{\mu_i}$ of a core ranges from $-0.5 \cdot 10^{-6}$ to $-2 \cdot 10^{-6}$ between 2 and 20 hours after demagnetization.

As the attempt of achieving a relatively high Curie temperature was successful, i.e. the temperature at which magnetism vanishes is above 130 °C, the values of the saturation magnetization J_s at room temperature are large for high-permeability ferrites. It can be seen from the hysteresis loops shown in Fig. 5 that for T 38 ferrite a flux of 4,000 gauss is reached already at a field strength of about 1 amp/cm. The very low coercive field strength is important for a number of applications. It drops below 50 ma/cm for the cores of highest permeability.

The shape of the permeability-frequency and of the loss-frequency curves is partly determined by the comparatively low resistivity of these ferrites. Cores with a permeability of 20,000 have values around 1 Ωcm.

Frequency effects are further governed by gyromagnetic resonance. The high permeability causes the gyromagnetic cutoff frequency

$$f_g = \frac{c \cdot J_s}{\mu_i} \left(c \approx \frac{2\ \text{Mc}}{\text{Oe}} \right)$$

to shift to relatively low frequencies.

For $\mu_i = 10{,}000$, f_g is about 800 Kc. Fig. 6 represents the complex permeability μ' and μ'' as a function of frequency which was measured in toroidal cores of N 30, T 35 and T 38 ferrites, having an outer diameter of about 6 mm. Depending on the core cross section, whether core walls are thick or thin, either eddy current or gyromagnetic resonance effects will have a predominant influence on the frequency curves. These relations are illustrated by Fig. 7 showing permeability and relative loss angle tan δ/μ_i as a function of frequency for a thin-wall and a thick-wall core. Both cores have an initial permeability around 10,000. High-permeability ferrites have been developed primarily for miniaturization purposes; hence cores made from these materials will in general have small dimensions and the eddy current influence mentioned above will be of little significance.

No air gaps in toroidal cores

In most cases, ferrites with a permeability of 5,000 and more are shaped into toroidal cores since in forms con-

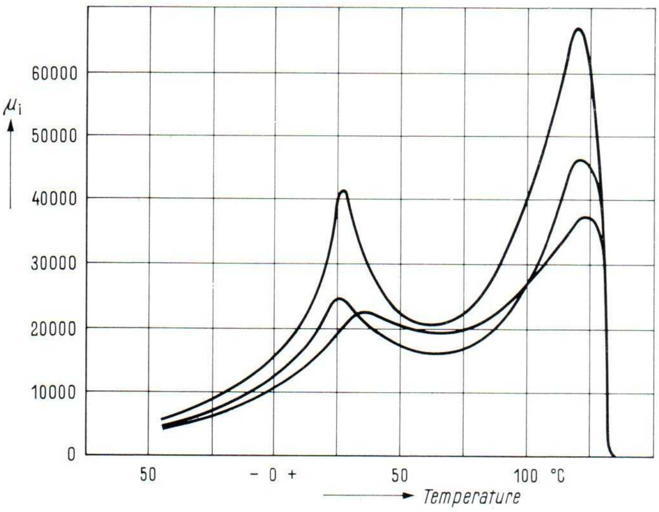


Fig. 4 Permeability-temperature curves for manganese-zinc ferrites having an initial permeability around 20,000

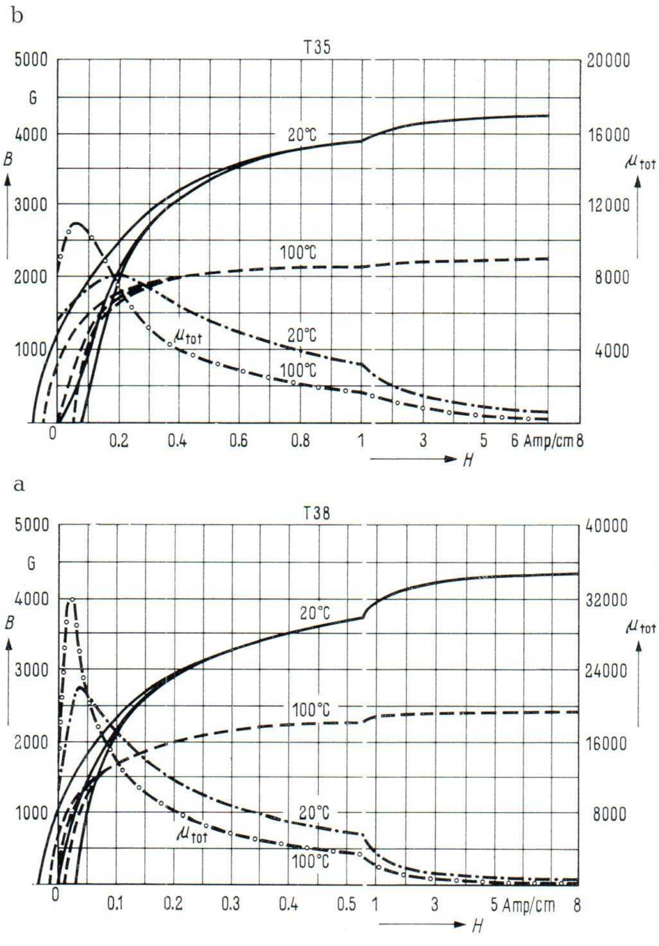


Fig. 5 Hysteresis loops $B(H)$ and total permeability μ_{tot}

a of T 35 ferrites at 20 and 100 °C

b of T 38 ferrites at 20 and 100 °C

sisting of several parts, such as pot cores, air gaps cannot be completely avoided even with the best grinding techniques. However, even very narrow air gaps reduce the permeability considerably—especially when core dimensions are small. Winding machines for toroidal cores facilitate this task which would otherwise be more difficult than for pot cores. Besides, only a few turns are generally required due to the properties of these ferrites. It is a further advantage of toroidal cores that no parts have to be cemented or pressed together.

Ferrite cores replace tape-wound cores

High-permeability ferrites should be used whenever their application opens up new avenues of technical approach or permits a reduction in cost or size of existing equipment. The latter is the case if customary cores built up of thin tapes are replaced by the much cheaper and often also smaller ferrite cores. One such possible application would be subscriber identification in Siemens large-size PABX's. A certain telephone number is found by means of an evaluation loop in a so-called toroidal core evaluation field. If the evaluation field covers 2,000 subscribers, 2,000 normal wires must be threaded through a total of 40 cores in such a way that 200 different wires pass through a group of 4 cores. The toroidal cores must satisfy the following requirements:

1. the geometry must be such that a sufficient number of wires can be threaded through the core and it still does not become too large;
2. the initial permeability must be as high as possible as only very small driving fields are available for saturation;
3. the permeability must drop as little as possible up to a certain frequency.

In this special case, it is an advantage if the relatively large toroidal cores, being less sensitive to air gaps, can be taken apart in order to avoid threading. So far, only high-permeability tape-wound cores could satisfy all of these requirements. However, the newly developed high-permeability ferrites can also meet these demands. In some respects they are even better than the much more expensive tape-wound cores. For instance, frequency effects are negligible up to 20 Kc and more—in contrast to the tape-wound cores in use. Impedance factors are considerably below those of toroidal tape-wound cores, while inductances may even be higher due to a more favorable filling factor. Ferrites have the further advantage of low mechanical sensitivity. It is to be expected that high-permeability ferrite cores will lead to savings and simplifications in a number of other cases, just as was shown for this special example.

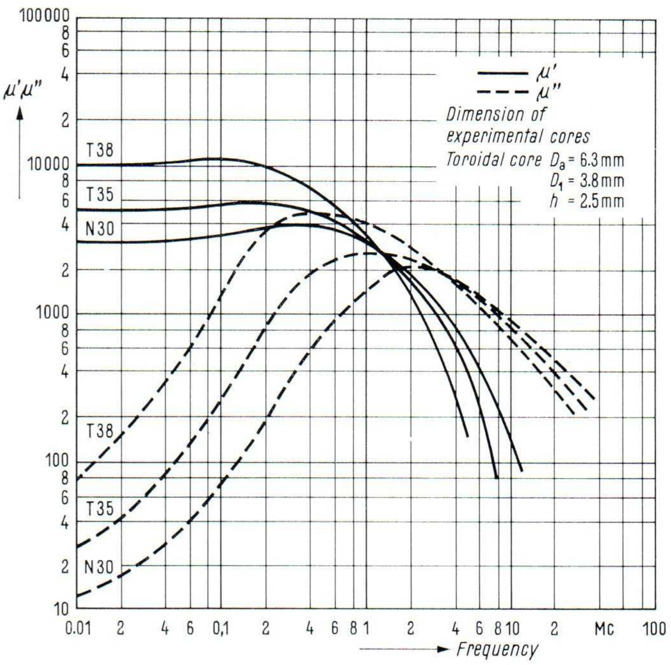


Fig. 6 Complex permeability μ' and μ'' of N 30, T 35, and T 38 ferrites as function of frequency

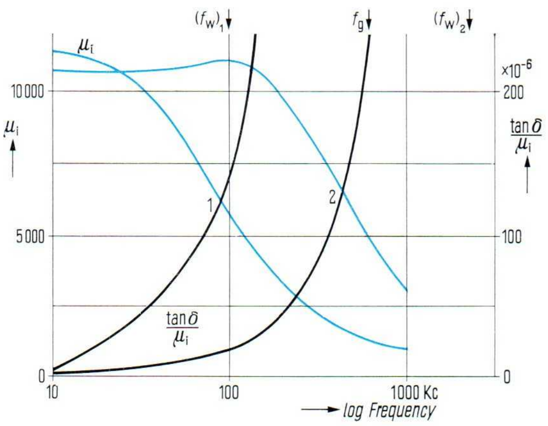


Fig. 7 Permeability and loss factor versus frequency,
(1) for a toroidal core of 20 mm² core cross section,
(2) for a toroidal core of 1 mm² core cross section
f_W = eddy current cutoff frequency
f_g = gyromagnetic cutoff frequency

Miniature transformers based on high-permeability ferrites

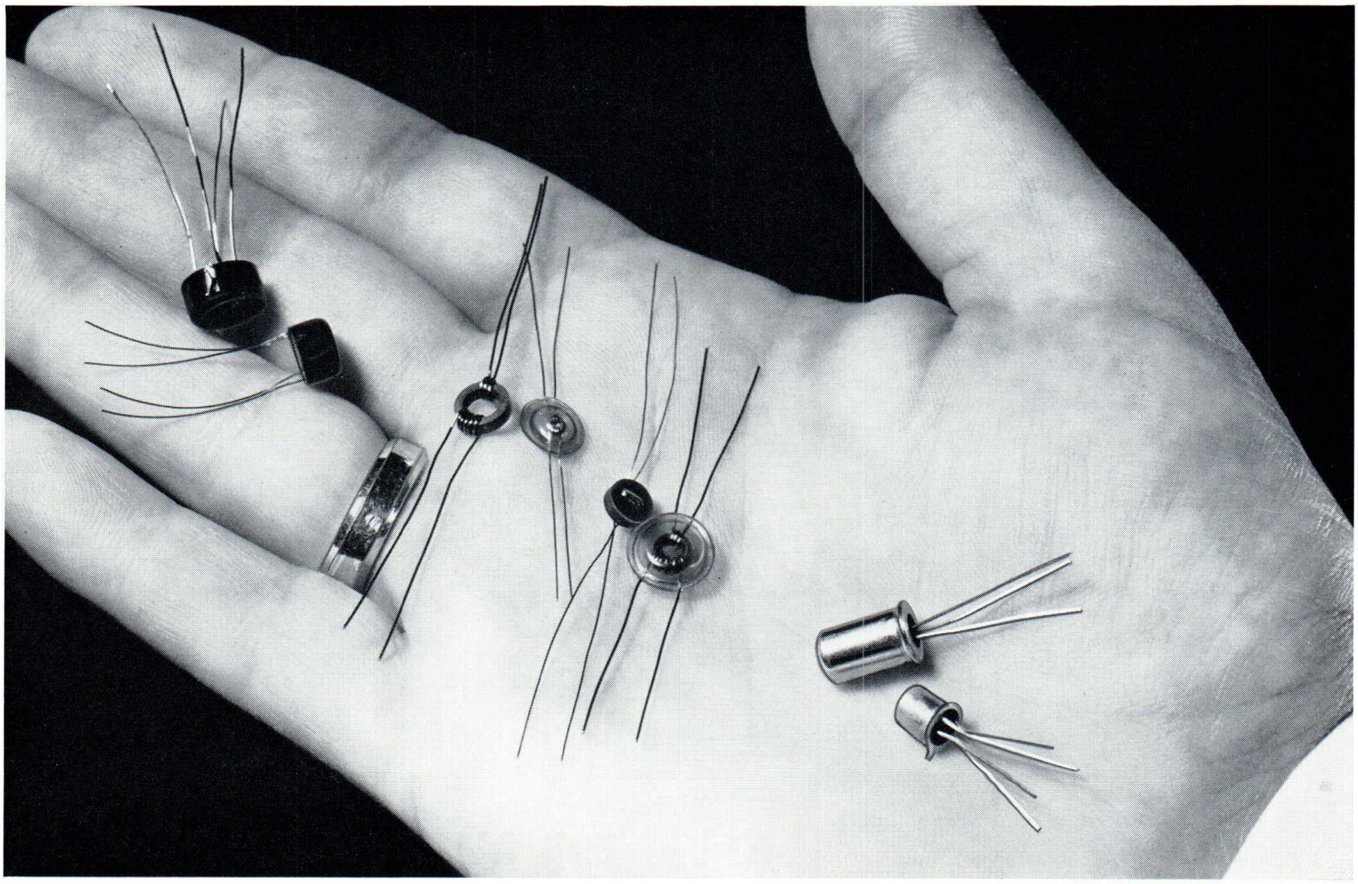
In the field of ferrite miniature transformers, new technical approaches have become possible by the development of high-permeability ferrites. Miniature transformers are, for instance, needed for various tasks in nanosecond-pulse techniques: e.g. impedance transformation for matching lines of different characteristic impedance, lines and transistor stages, or transistor stages with widely different input and output impedances; for polarity reversal of pulses, or galvanic separation of circuits. The advantages afforded in this area by ferrites of very high permeability are manifold. With increasing permeability, propagation factor and bandwidth increase also; the required number of turns and hence the rise time inherent in the transformer itself are reduced.

Cores of very high permeability have almost ideal transmission properties. Voltage, current and impedance transformation becomes possible for pulses up to a length of 1 μsec even at high pulse frequencies.

Ferrite miniature transformers can be manufactured in various shapes. The outer dimensions of ferrite cores range from 2 to 10 mm. To the simple wound toroidal core and the core with two holes, a new component has been added: the complete miniature transformer, sealed into a chunk of plastic or installed in a transistor case.

These few examples show that ferrites of particularly high initial permeability can be used for manufacturing new components which permit promising applications in many more different ways.

Fig. 8 Ferrite core miniature transformers can be manufactured in various shapes. To the simple wound toroidal core and the core with two holes, a new component has been added: the complete miniature transformer installed in a transistor case



Superhets for Cars — Old and New Models

By Walter Hirschmann and Friedrich Seibt †

Superhets for cars have ceased to be luxuries. On the contrary: Each car should be fitted with a radio in the interest of central traffic control—an absolute must in today's Sunday and holiday traffic. The development of the car superhet, which has to meet special requirements over and above those to be filled by a normal broadcast receiver, is shown in retrospect, and the advantages of an all-transistor set are illustrated.

To give the car driver a better view, the stylists have put the upper edge of the dashboard as low as possible. As the driver's leg-room must not be restricted, less space is available for indicating instruments and controls. Moreover, an ever-increasing number of technical devices have to be mounted in the dashboard. All this limits the space for a car receiver and leads to the necessity of constructing extremely flat sets. As the temperatures encountered in a car may be very high, especially after long exposure to sunshine, all the components in the receiver must be of rugged design and highly immune to temperature changes. In addition, such sets must have a pronounced resistance to mechanical vibrations.

As a car travels along, the built-in receiving system picks up a signal of continuously varying field strength. The input voltages to the set are between some tenths of a microvolt and some volts. They may thus be of the order of the noise voltage, but may also be strong enough to overload the receiver. If vhf broadcasts are to be received, even more exacting requirements have to be met. Only a superheterodyne receiver with outstanding regulation characteristics can cope efficiently with such conditions.

Conventional sets

Early car superhets had a complement of 4 to 5 tubes. The receiver quality could have been improved by using more tubes, but the space available set a limit, for one. Another reason for keeping down the number of tubes was the current drain: the heater current was already as high as about 1.5 to 2 amps at a battery voltage of 6 v, and the total plate current 40 to 60 ma at an anode voltage of 200 v. The latter had to be derived from the battery voltage through a separate electromechanical vibrator with complex interference suppressing facilities. If we take the vibrator efficiency of about 70% into account, about 25 watts were consumed in all.

Hybrid sets

As semiconductor development progressed, the tubes in the output stages could first be replaced by Siemens power germanium transistors. To secure a low-distortion drive of the transistorized pushpull final stage, Siemens developed a special vlf tube, the ECF 86. These hybrid sets worked with no more than 3 to 5 tubes, but had already 3 to 5 transistors, because a transistor circuit could be substituted for the electromechanical vibrator, which resulted in simpler interference suppression on account of the higher "vibrator" frequency. In keeping with the relatively low signal at the input stages, the anode voltage of the tubes left to provide for r-f amplification and of those required in the vlf input and driver stages, saw a decrease to 60 to 120 v.

Conventional sets took a rather high percentage of the dashboard space in a car and consumed as much as about 25 watts if equipped with 4 or 5 tubes *(Blaupunkt photo)*

This went hand in hand with a reduction of input power to 12 to 14 watts. The hybrid sets dominated the market for car superhets for several years.

Transistor sets

With the advent of the high-frequency transistor came the all-transistor car receiver. The sets could be built to take up less space and be lighter than tube sets. It was not only the small dimensions of the transistors that made this design possible, but also the fact that these components did not need sockets. Immunity to vibration and a considerably longer life were further advantages of the transistors.

The current drain could be slashed. When the set was adjusted for medium volume, it was as little as 0.1 to 0.3 amp, i.e. lower than that of the smallest lamp in a car. The input power to a transistorized superhet, averaging about 1 watt, was thus less than one twentieth that of a conventional set.

All-transistor design required the development of new circuitry techniques, which also brought some improvements. For instance, the electromechanical relay in the automatic tuning system can now be controlled by a transistor circuit.

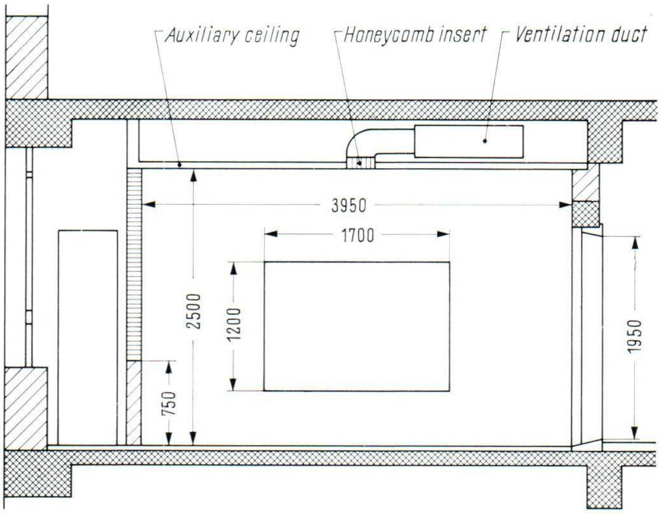
Because of the strongly varying signals picked up by a car aerial, very exacting requirements as to stability and noise level had to be met by the vhf input stages. Germanium mesa transistors—incorporated of late in a-m/f-m i-f stages as well—have shown a satisfactory performance. Initially there was some doubt because of the temperature limit of the germanium transistors, but efficient cooling and circuit techniques were able to keep the increase of the collector currents with temperature within such narrow tolerances that even an ambient 60 °C is well below the temperature where the stability of the circuit is affected. Transistors have shown extremely good results in car radios, and all-transistors sets are dominating the market everywhere. Moreover, transistors are being used in the general electrical system of a car, and some considerable modifications in their favor are in the offing.

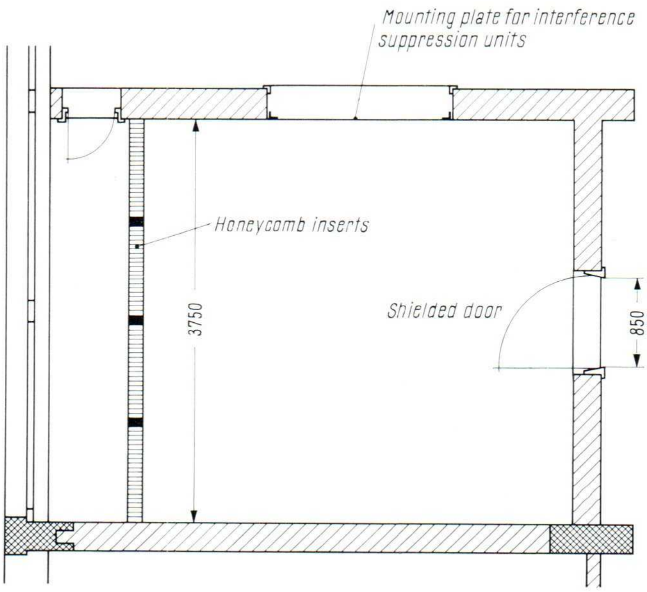
Transistor sets may be of space-saving design. The input power to a modern all-transistor car superhet is as low as about 1 watt *(Blaupunkt photo)*

Shielded Room for Calibration and Measurement

An air traffic control center required a shielded room for calibration and measurement. The shielding was chosen to be of the permanently installed type in order to make the best use of the available space. Since a ventilation duct ran through the room, a second ceiling had to be constructed on top of which a 0.1 mm copper screening foil was laid. A honeycomb insert was mounted over the duct opening. A masonry base of 70 cm height built 80 cm away from the glass windows carried a solid wall with honeycomb inserts which reached up to the second ceiling.

Interference suppression for the voltage supply required four radio interference suppression units and one telephone interference suppression unit (as well as three h-f plug connections) to be arranged on a mounting plate. This plate was built into a wall between this and an adjacent room; it can be closed off by two sheet metal doors. *Klaus Kowalkowski*





Publications About Components

The Siemens Traveling-Wave Tube RW 21, a modern amplifier tube for educational tv and f-m microwave radio systems operating in the frequency band of 2.4 to 2.8 Gc.
Technical report from the Tube Factory, Siemens AG, Order No. 2-6220-015-01

This new 32-page technical report on the traveling-wave tube RW 21 represents a continuation of publications on practical, comprehensive application data of products from the Siemens Tube Factory. The RW 21 has been designed for use in microwave transmission networks, particularly those carrying special programs such as educational tv, and in conventional broadband microwave radio systems.

The RW 21 covers the frequency band of 2.4 to 2.8 Gc, delivers an average pulse saturation power of 32 watts and has an average small signal gain of 42 db. With concurrent visual and aural amplification a peak sync power of 10 watts may be obtained at a three-tone intermodulation ratio of 30 db. In normal f-m microwave radio systems the tube may be operated up to 20 watts output power.

The tube is a periodic permanent magnet focused traveling-wave tube and is a plug-in match in its associated magnet system. The magnet system has a particularly small leakage field. The tube is designed to operate with depressed collector. Cooling may be effected by conduction or convection.

The r-f input and output ports are designed for connection to coaxial cables. The magnet system including the tube and the connections is provided with r-f shielding. This technical report covers the design and principle of operation of the tube and magnet system, and furthermore includes a comprehensive discussion on the various types of distortion with their definitions which have considerable influence on the amplification of tv signals. These are amplitude distortion dependent on drive level, such as compression, expansion, pulse compression, differential gain, linearity, and phase distortion dependent on drive level such as phase difference, a-m/p-m conversion and differential phase, and finally intermodulation products with measuring methods and their relationship to the black-white transition of a tv signal.

The information is not only applicable to the RW 21 alone, but also on a more general basis to the amplification of tv signals.

The actual performance of the RW 21 is equally well covered with the aid of 48 pictures and graphs, its properties under different operating conditions are discussed, and the power supply with hints for tube switch on and switch off described. Finally methods of cooling are mentioned.

Siemens ELECTRONIC COMPONENTS Bulletin