Tijdschrift van het
**Nederlands Radiogenootschap**
DEEL 22 No. 6 NOVEMBER 1957

## Introduction

The task of the Dr. Neher Laboratory as central laboratory of the Netherlands Postal and Telecommunications Services is to perform the development work on behalf of the various branches of PTT, and to do research work in some limited fields.

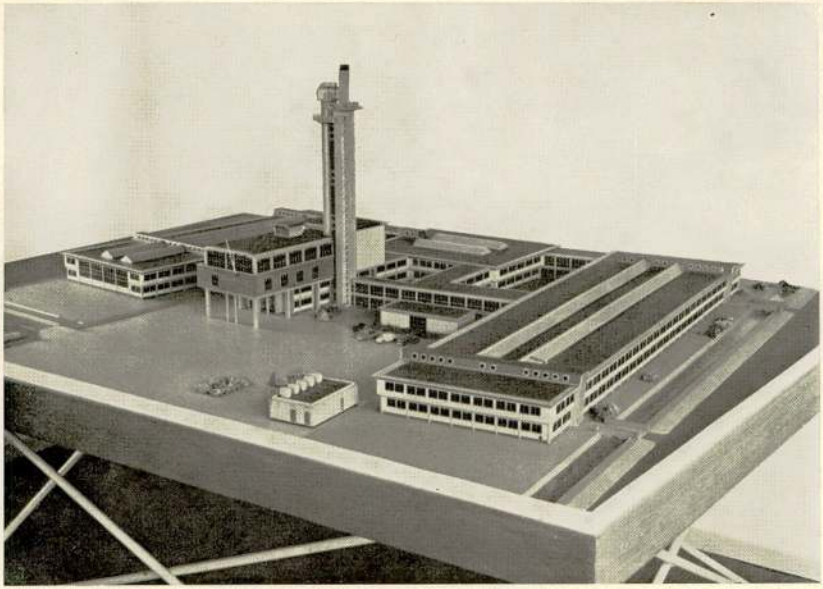
In the present age of revolutionary progress in technics and science, we should of course endeavour in the interest of the rapidly expanding services, to recognize the future trend of technical and scientific development.

Therefore it is an important feature of the laboratory's task to follow and study these new developments and to keep in view the possibility of applying new working-methods in the Postal and Telecommunications Services with its many aspects. Many and various indeed, for they comprise i.a. telephony, telegraphy and telex, both with regard to transmission via carrier cables or via radio, and with regard to switching techniques and equipment. The many problems to be dealt with refer to equipment for radio broadcasting and television, to postal mechanisation, to the introduction of automation in the Postal Cheque and Clearing Service, and to the use of computers for data processing. It is clear that system engineering, too, forms a field of action for the Laboratory.

With a limited scientific staff the Laboratory tries to accomplish all these tasks to the best of their abilities and it is gratifying to note that the present issue of the "Tijdschrift van het Nederlands Radiogenootschap" has entirely been devoted to contributions of some of the Laboratory's staff. I hope that these articles will give the readers an insight into the work done at the Dr. Neher Laboratory and make them interested in it.

The Hague, October 1957.

Director General of the Netherlands PTT.



Maquette PTT laboratories at Leidschendam.

The interior of the right-hand hall showing the work rooms.

# Some problems concerning the choice of cable-circuits for television transmission

by A. P. Bolle *)


## Summary

In this article some problems concerning the choice of cable-circuits for television transmission are discussed. Typical advantages and disadvantages of the symmetrical and asymmetrical circuits in relation to television transmission are shown.

The results of these considerations are applied to the different possibilities of television transmission systems.


## 1. Introduction.

It can be shown, that for the television system (625 lines, 25 pictures per second) applied in the Netherlands and also in many other countries, the highest frequency amounts to about 5 Mc/s. The lowest frequency is 25 c/s, apart from the dc-component in the signal, which component can be re-added to the signal at an arbitrary point in the chain between camera and receiver.

Due to this large relative bandwidth (25 c/s—5 Mc/s) high demands are made on the transmission medium, especially if a cable is chosen as such. For the frequency band which can be used on cables usually lies under 12 Mc/s, (except in the case of very short connections, which are not dealt with here), because of the fact that the attenuation of a cable expressed in dB increases at least with the square root of the frequency. This means that on cables a television signal can be sent either as a video-signal (i.e. unmodulated 25 c/s—5 Mc/s), or amplitude-modulated on a carrier the frequency of which has been chosen in such a way that both sidebands or one complete sideband and one vestigial sideband are under 12 Mc/s. Even in the second case the relative bandwidth is large. With radio-transmission there are fewer difficulties in this respect, because of the fact that higher frequency-bands can be chosen. As a result, the

*) Transmission Branch.

relative bandwidth is far smaller than in the case of cable-transmission. In some cases a higher frequency-band is chosen even on cables, in order to obtain a simple modulating apparatus and a less large relative bandwidth.

## 2. Transmission requirements.

The transmission requirements which must be met can be derived from the fact that for a good picture-quality it is necessary that the time-dependency of the output-signal should approximate the time-dependency of the input-signal of the transmission-medium. From this it follows that the phase-frequency characteristic plays as important a role as the attenuation-frequency characteristic. This is in contrast with telephone-transmission practice. Instead of describing the transmission properties with the aid of the above-mentioned two characteristics, it is also possible to do so by means of the so-called pulse-response or wave-form response. Both methods are identical, as is well known, but since the television-signal itself consists of a great number of successive voltage-steps, corresponding to variations in lucidity, the last method is advantageous in that it gives more direct information about the behaviour of the transmission-medium. The C.C.I.T.T. has formulated a number of provisional requirements for the transmission characteristics as well as for the wave-form-response [1].

It is important that the cable used should have a good impedance regularity. The result of too large an impedance irregularity is, that after reception of the wanted (primary) signal a more or less great number of secondary signals (German: Mitfluss, Dutch: nasleep) are received which cause the so annoying multiple pictures (ghosts). The C.C.I.T.T. basing itself on the work of Kaden [2] and Fuchs [3] has formulated the requirements for the impedance regularity which can be checked either by steady-state measurements or by pulse measurements [1]. Based on the same phenomenon there is also a requirement for the minimal admissible return loss between the cable-impedance and the input and output impedance of repeaters.

It is evident that there is also a requirement for the signal-to-noise ratio; as well for continuous random noise as for periodic noise [1].

All the C.C.I.T.T. provisional requirements are given for a 2500 km international hypothetical reference circuit for television

transmission. Till now only little experience has been gained as regards the contributions made by each of the constituent parts of the reference circuit to the total admissible deviation from the ideal case.

## 3. Possible cable circuits

There are two sorts of circuits; namely the symmetrical and the asymmetrical circuits. A circuit is symmetrical when both conductors have the same electrical position with respect to earth. In the many forms of telephone cables a more or less great number of symmetrical pairs are present. The phantom circuit which can be made from two symmetrical pairs (the side circuits) when these pairs together form a quad-structure, is also a possible symmetrical circuit for the transmission of television signals. In most cases these circuits are paper-insulated. Due to the losses in the paper, the rate at which the attenuation of the circuit increases is more than proportional to the square root of the frequency. Of late a cable type with a better sort of insulation material (e.g. poly-ethylene) has come into use. In such cables the attenuation increases proportional to the square root of the frequency as a result of the fact that the losses occur practically exclusively in the conductor material. This is a great advantage, especially when these cables are used for high frequencies (e.g. television transmission). There is little to be said about the impedance regularity of the symmetrical circuit. It is reasonable to suppose that with cables having plastic insulation the impedance regularity is greater than with those having paper-insulation, because of the better geometrical construction. For short distances the requirements become less stringent. It is possible that even the older paper-insulated circuits give satisfactory results in those cases.

The coaxial circuit is the most used asymmetrical circuit. There are two types, one with solid insulation (poly-ethylene) and one air-insulated. For both types the attenuation increases proportional to the square root of the frequency. The impedance regularity of the latter type is better than that of the first. If a coaxial circuit with solid dielectricum for very long distances is used for television transmission, some difficulties may arise. As a rule several coaxial pairs are covered by one sheath just as with symmetrical circuits.

Each of the two circuits (symmetric and coaxial) has its own specific advantages and disadvantages.

The advantage of the symmetrical circuit is that it is more insensitive to asymmetrical interference from outside (mains hum, strong transmitters). Small disturbances in the symmetry give rise to the penetration of the interference into the wanted signal. A disadvantage of symmetrical circuits is that they influence each other (crosstalk) especially at high frequencies when there is no screen. As a rule the crosstalk increases with the frequency. There are compensation methods, which are, however, never perfect. As a matter of fact the crosstalk determines the highest admissible frequency on symmetrical circuits. As a rule, for the successful application of symmetrical circuits at high frequencies it will be necessary to make a sufficiently good screen around each symmetrical pair. Naturally, this is not without influence on the attenuation and the price.

With the phantom-circuits the conditions are more favourable. The construction of the quad being ideal, the external field of the phantom-circuit is very small. As a result the crosstalk between phantom-circuits will be less than with the side-circuits. Under these ideal conditions the screening can be omitted. The quad configuration, however, is never perfect. The deviation from the ideal construction is larger as the conductor diameters are smaller and it is larger in paper-insulated cables than in poly-ethylene-insulated cables. The advantage of the phantom-circuits is complete, when no use is made of the side-circuits. The crosstalk between the side-circuits and "their" phantom-circuit is fairly large, whereas the combined use of the side-circuits and phantom-circuits makes the presence of complicated provisions necessary, especially at high frequencies.

The advantage of the coaxial pairs is that the crosstalk decreases with the frequency because of the fact that the outer conductor acts as a screen. At lower frequencies (below say, 100 kc/s) the screening becomes insufficient as a rule. The use of a coaxial pair for those frequencies may be dangerous. The same holds with respect to the penetration of asymmetrical interferences. These phenomena can be reduced to a more or less large extent by making an additional screen or by preventing the flowing of currents, induced from the outside, into the outer conductor. These currents cause a voltage drop on the inside of the outer conductor, especially at low frequencies. As a result a current arises in the coaxial circuit, which current cannot be separated from the wanted signal. The flowing of the above-mentioned currents can be prevented to a more or

less large extent by inserting a high impedance into the outer conductor. However, this high impedance should not be present for the wanted currents. This can be done by winding the inner and outer conductor together on a core. In this way the so-called coaxial coil comes into existence, which coil introduces into the coaxial pair some additional attenuation. The interruption of the outer conductor can also be reached by inserting a transformer. This can only be done if the relative bandwidth of the wanted signal allows the construction of a transformer with good transmission properties.

The attenuation of the different circuits must also be considered. For all circuits it is possible to choose the dimensions in such a way that the attenuation has a minimum value. Assuming that the attenuation arises from losses in the conductor only and that both conductors are made from copper, the results are as follows [4]:

*a.* The symmetrical pair (with screen) has a diameter which is 1.63 times as large as the diameter of a coaxial pair for the same attenuation. The quantity of copper is 1.3 times as large.

*b.* The phantom-circuit (without screen) with the same outside diameter as the C.C.I.T.T. coaxial pair ($\approx$ 10.5 mm) has an attenuation which is 1.3 times the attenuation of that coaxial pair and needs only half the quantity of copper. It is true that both side-circuits have a lower attenuation than the phantom-circuit (about 10 % lower), but they need a screen in order to make the use at high frequencies possible. As a result the attenuation increases considerably ($\approx$ 50 % higher than C.C.I.T.T. coaxial pair). The quad structure being not ideal, a screen is necessary and consequently the attenuation and the quantity of copper increase.

From the foregoing it follows that with respect to the insensitivity for low-frequency interferences the symmetrical circuits (even without screen) are by far superior. The ideal phantom-circuit also has a good insensitivity for high-frequency interferences. The deviation from the ideal case being too large, the presence of a screen is necessary and as a result the phantom-circuit is surpassed by each of the side-circuits (lower attenuation, less copper). The increase in attenuation is considerable, however, and therefore the coaxial circuit is superior in this respect.

For a complete comparison of the different possibilities,

many other factors must be considered: price and quantity of the dielectricum, necessary amount of screening, manufacture difficulties etc. etc. In each case it is necessary to consider which type of circuit must be chosen.

## 4. Transmission systems

If television signals are to be transmitted over short distances, they can best be transmitted unmodulated, in order to avoid the use of expensive modulating equipment. Because of the frequency-band involved the symmetrical circuit is preferred. If a great insensitivity against high-frequency interferences is desired, a screen is necessary. An additional advantage of the symmetrical circuit is that completely balanced amplifiers can be used because it is nearly impossible to construct a transformer with good transmission properties. With balanced amplifiers it is very easy to avoid the coupling via the supply-voltage sources. With unbalanced amplifiers this coupling cannot be avoided completely in view of the lowest frequencies involved. With the balanced amplifiers it is also possible to avoid the cathode and screen grid decoupling condensors.

Because of the very large relative bandwidth the equalisation of the attenuation and phase is fairly complicated.

If television signals are to be transmitted over large distances, it is evident that they must be modulated. The C.C.I.T.T. has recommended a vestigial sideband system with a carrier frequency of 1056 kc/s. The modulation apparatus is complicated. The lowest frequency involved is such that a coaxial cable can be used.

In Germany another system has come into use: a double sideband system with a carrier frequency of 21 Mc/s [5]. Therefore the relative bandwidth (16 Mc/s—26 Mc/s) is rather small. The modulation equipment seems to be fairly simple. As a result, this system can be used for short distances on the C.C.I.T.T. coaxial cables with a repeater spacing of about 4.7 km. The equalisation of the attenuation and phase is easy.

## 5. Bandwidth-reduction

All transmission problems become less complicated when the bandwidth involved can be reduced.

With very good DC-restorers it seems to be possible to

omit from the video-signal the frequencyband below, say, 10 kc/s [6]. The design of a video-transformer becomes easier in that case. The advantages of such a transformer are evident: impedance-transformation at the input and output stages of amplifiers, a change from balanced to unbalanced circuits etc. etc.

With some forms of coding it is also possible to omit a great deal of the high frequencyband of the video-signal [7]. This is, from the point of view of the transmission engineer, also very attractive. As far as the present author knows the two reduction methods are based only on theoretical considerations and neither has been applied in practice.

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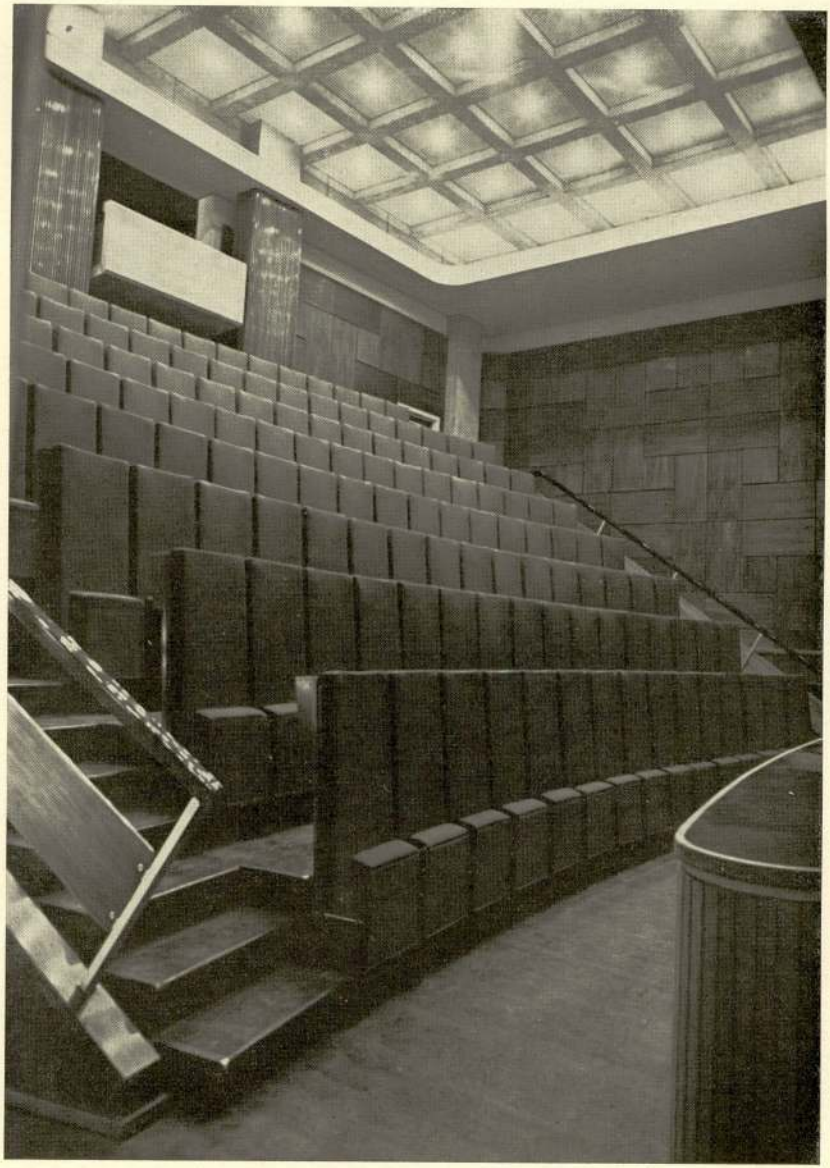

The auditory.

# Some audio problems in modern telephone systems

by H. Mol *)

**Summary**

The present paper gives a survey of the audio problems pertaining to the improvement of the transmission quality of a telephone circuit. Besides the improvement of elements the study of the fundamental problems of speech turns out to be an important factor.

## 1. Introduction.

In general the average radio engineer lends a very unwilling ear to audio transmission problems. Perhaps the following, admittedly very brief survey will convince him of the existence of interesting problems in this field, worthy of the attention of engineers.

Of course the various subjects cannot be presented in the round: some subjects are treated more elaborately because more is known about them at the moment. This does not mean that less time is or should be devoted to other subjects that are dealt with in a few lines. In general the treatment of the problems is admittedly streamlined to the $n^{th}$ degree but we tried to alleviate that deficiency by giving a modest number of references to literature.

## 2. The elements of a telephone-set.

In commercial telephony the classic device for transforming sound into electrical signals still is, and practically always has been, the carbon microphone. It has been able to stand its ground that long because it is an amplifier as well, providing a power gain of some 30 dB. It admittedly has a bad reputation for distortion, instability and noise. Consequently it is despised by studio people who discarded it as soon as they could invent types of microphones better suited to their purpose.

Since 1945 the carbon microphone has been studied at laborious length at the Transmission Laboratories with the aim of finding the limiting factors of its performance. This study was necessitated by the reality that after the end of World War II Germany ceased to be the main source of carbon microphones for the Netherlands PTT. One started looking for an appro-

*) Transmission Department

priate substitute for the almost classic Siemens-microphone. Apart from some untimely improvisations this search turned out to be fruitless. It was in that situation that the Netherlands PTT undertook to compile a specification for the construction of a new type of carbon microphone that could be used throughout the Netherlands telephone network. We at the laboratories seized the opportunity for incorporating the latest theoretical and practical findings in the design which also had to have normalized outer dimensions.

Moreover the microphone was expected to function satisfactorily in the multitude of different feeding systems this country could boast of at that time.

Needless to say that a hoard of thought-provoking problems was thrown up, both of a practical and a theoretical nature. It stands to reason that the ultimate design turned out to be a compromise, fitted as it were by least squares to the conflicting requirements. In order to test our theoretical pronouncements several prototypes were produced by the laboratory and demonstrated to the industry in order to facilitate the process of putting ideas into mass-production. Though in general PTT took the initiative in these matters we would do a grave injustice to the industry, including that of the Netherlands, by not mentioning and acknowledging the painstaking activities of the contracting factories.

The basic principles underlying the final design can be briefly summarized as follows. (1)

The carbon chamber of the microphone is nothing but a miniature coal-scuttle, the walls of which are insulated except for the two electrodes. It is next to a miracle that this collection of loose contacts can be stable at all. In a stable microphone the carbon granules remain in the same geometric configuration throughout the duration of a telephone call. Now, like a heap of sand, the granular mass has the tendency to assume a smaller volume. In the microphone we call this the phenomenon of packing. Among the many factors that have a stake in the problem of packing dry friction between the granules plays an important role. Dry friction keeps the packed granules in their frozen condition, against the restoring forces furnished by the polarizing direct current. The influence of the feeding current on the stability can be either detremental or beneficial. When the granules are mechanically displaced along the streamlines of the feeding current the electro-static forces

between the granules will support the packing action of dry friction. If, on the other hand, the granules are displaced perpendicularly to the streamlines, the electro-static forces will drive them back to their original positions, in that way providing an elasticity of electro-static origin. This is the so called transverse-current principle. Unfortunately the transverse-current microphones suffer from a low output. On the other hand the microphones of obsolete design with flat electrodes that are exceedingly inviting to packing have a high output. The modern compromise is the microphone with dome-shaped electrodes, said electrodes being parts of spheres. Here the transverse-current principle is active only in certain parts of the carbon chamber. The stabilized pathways as it were "keep up" adjacent streamlines that guarantee a higher output but run the risk of packing. For stability it is essential to keep the tolerances of the distance between the electrodes within narrow limits in mass-production.

The frequency response of the carbon microphone is governed by the resonance of the diaphragm and its clamping and by the resonance of the carbon chamber. The mechanical impedance of the carbon chamber is non-linear and depends on the magnitude of the displacement of the moving electrode. Consequently the shape of the frequency response depends on the acoustic pressure of the sound field in which the microphone is tested. Also packing makes itself felt in the frequency response because packing increases the stiffness of the carbon chamber so that consequently the lower end of the frequency range is attenuated.

Though, clearly, the frequency response of the carbon microphone is a rather vague affair, we must nevertheless smooth out disturbing mechanical resonances. This can be done either by introducing acoustical or mechanical friction or, more elegantly, by adding additional resonances, in other words by introducing more degrees of freedom. In this respect we meet the same type of problems in electro-dynamic, electro-magnetic, piezo-electric or semi-conductor microphones.

The harmonic distortion, even in a modern carbon microphone can easily reach a value of some 20%. The seat of this distortion is the non-linear mechanical impedance of the carbon-chamber, as "seen" from the moving electrode. Interesting enough, one can prove, both theoretically and experimentally, that the relation between resistance-variation of the carbon chamber and displacement of the moving electrode is surprisingly linear. It

is the non-linear carbon impedance that causes the non-linear relation between sound pressure and resistance-variation. Many people pin the deficiences in intelligibility on the non-linear distortion of the carbon microphone. Our preliminary measurements, however, seem to indicate that the limited frequency range is to blame.

The only reason for discarding the carbon microphone in future is that it will not be stable on the low line currents electronic telephone exchanges will be able to furnish. In the USA a semi-conductor microphone (2) has been announced that requires only one tenth of the current of a carbon microphone. As it is a modulator it also provides amplification, just like the carbon microphone. It is not, however, afflicted by packing.

Since the invention of the telephone in the early seventies the construction of the telephone receiver has not fundamentally changed. It still contains pole-pieces with coils, a permanent magnet and a flat diaphragm. The major improvement was the gradual reduction in size of the permanent magnet. Whereas in the early days the magnet was used as a generously sized handle, quite often in modern receivers the magnet can hardly be found at all. About 20 years ago it appeared that the frequency response of the receiver could be made flat (so called equalized receiver) by placing a hard surface very close behind the diaphragm, a textile-covered hole in this surface providing the necessary friction. Since several years this principle is incorporated in the new receivers that are introduced into the Netherlands telephone network. It is interesting to know, however, that said principle was already present in Ader's receiver (3), dating back to around 1880, though it was not recognized as such at that time.

Though the present type of receiver guarantees a good service there are several reasons for replacing it by a different type in future. It is difficult, for instance, to adjust and keep stable the air-gap between the diaphragm and the pole pieces. The diaphragm, being continuously exposed to the attracting force of the permanent magnet, has the tendency to sag. Also, there is the danger of saturating the diaphragm and many variants to the classic configuration have the same aim: letting the lines of force pass the membrane at right angles and preventing their return through the thin vibrating plate. In the balanced magnetic systems there is no constant force of mag-

netic origin on the armature, which is separated from the cone-shaped membrane as such. In general, there is a limit to screwing up the magnetic field because, apart from saturation effects, the field acts as a transformer ratio matching the mechanical system to the electric source. As such it has an optimum value (3). The comparatively high acoustic impedance (as seen from the earpiece) of the present receiver with its hard, flat membrane is felt as a drawback because the receiver does not short-circuit the room noise which is produced by a low-impedance source, namely the air. Moreover the receiver is a bad radiator so that it cannot be used as a small loudspeaker to replace the classic ringer with its gongs. The above-mentioned and other problems are still under consideration for the time being.

## 3. The improvement of transmission quality

Though recently developed types of telephone receiver such as the ring-armature (4) type and the rocking armature type give a substantially higher output than the present type, there is a limit to increasing the sensitivity of the receiver. The same argument holds for the microphone as will appear from the following.

When the total air-to-air sensitivity of a telephone channel is too high there will be uncomfortable listening provided an automatic volume regulation of some sort takes place which has many technical consequences in the network. We think to have reached that comfortable limit in the PTT network, considering that, in the complete absence of noise, one can carry on a conversation via two modern telephone sets even with an attenuation of some 50 dB between them. When a white noise of 0,5 V, as measured with the CCITT-psophometer at the terminals of the set, is present this attenuation reduces to some 35 dB.*) Only in noiseless networks a less sensitive microphone can be compensated for by an oversensitive receiver. When there is noise present the microphone output should be well above the noise level. So-called sound powered systems may also cover a "distance" of 50 dB in the noiseless case, but under the prevailing noise conditions in the present telephone systems they never have a chance.

In the modern telephone set the microphone and the receiver

*) These figures may not be regarded as hard-and-fast figures. They merely indicate the order of magnitude.

are coupled to the line by means of a balancing transformer which reduces the so-called side-tone. The expression side-tone is technical jargon for hearing one's own voice coming from the receiver which can be most unpleasant. The more the sensitivity of both the microphone and the telephone is screwed up the more difficult it becomes to combat side-tone. One of the many solutions for the balancing transformer is the so-called Ohnesorge-system, among other things characterized by the fact that the balancing resistance acts as a DC shunt across the microphone (which is beneficial to the latter's stability) and the property that the receiver is connected to a separate winding on the transformer. The separate winding permits the insertion of a transitorized amplifier for the hard of hearing without the need for an additional transformer, providing a gain of some 25 dB.

Before the war the Ohnesorge-system was used practically throughout the Netherlands telephone network.

It was maintained in the new telephone sets developed after 1945 though the ratio of the windings was slightly changed in order to permit the transformer to receive 100 $\Omega$ microphones as well as 200 $\Omega$ microphones. The 200 $\Omega$ type is necessitated by the $2 \times 400$ $\Omega$, 24 V feeding coils of the Rotterdam exchange.

When the attenuation between two telephone sets is reduced by means of amplifiers, echo-problems arise. In this country the so-called 4 wire-system is in general use. As the classic amplifiers are unilateral there are separate go- and return paths. The coupling of these paths to the subscriber's lines is realized by balance transformers that also have the task of preventing coupling between the go- and return paths. As the balance conditions can never be ideal there always remains a certain amount of coupling. Consequently the talker who sends speech waves along the go-path hears his own voice coming back via the return-path in the form of an echo. The higher its level and the longer its time-delay the more disturbing the talker's echo will be. (5)

A detailed description of the laboratory's engagement with the echo problem lies outside the limited scope of this paper. We shall confine ourselves to stating the interesting fact that for this country it is the improvement of the echo conditions on international lines that calls for expensive measures in the national network.

As the improvement of transmission quality marches on more

problems are posed to the transmission engineer. For instance, the possibilities of reducing the attenuation in the peripheral ramifications of the telephone network are certainly worth exploring. The telephone networks of factories, the buildings of which are spread over large areas, are nevertheless without further ado connected to the PTT-network via often already long subscriber's lines. It is certainly necessary to bring these networks within the scope of the general improvement of transmission quality though at the moment is not yet quite clear what will be the most economical method. One might contemplate loading*), 2 wire amplifiers, negistors**) etc.

In the long run no modern transmission laboratory can be blind to the fundamental problems of speech.

Especially in (carrier) systems where the same channel is used for speech and signalling the signal receivers must be safeguarded against signal imitation by speech. In this respect again and again one comes to the conclusion that our knowledge of the statistical properties of speech waves is still far from mature.

Also several problems of automation call for voice operated devices.

Since the realization of the transatlantic cable the attention of the world is re-focussed on the vocoder problem (6). The vocoder aims at bandwith-reduction (or better: capacity-reduction) so that more than one telephone conversation can be transmitted over a "straight" channel that normally carries only one conversation.

The construction of a successful vocoder must, in our opinion, root in a sound knowledge of the articulatory and auditory processes. We must never forget that Nature, with its vast experience in telecommunication is our best teacher and that it has sound reasons for not applying certain techniques.

Needless to say that the development of a vocoder is only worth-while for application to expensive circuits.

For the above mentioned reasons a small group at the dr Neher Laboratories is engaged in fundamental studies on speech.

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**) These are negative impedance amplifiers.

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# Non-linear properties of carbon resistors

by C. E. Mulders *)


## Summary

The application of the conventional method of determining the non-linear effects in carbon resistors, as outlined in many test specifications for these components, encounters difficulties in case of good specimens. An alternative method is proposed, based on the measurement of the production of harmonics in the resistor, when a pure sinusoidal current is passing through it. A description of the experimental procedure is given and some results are shown. A comparison is made between both methods in some cases where either method is applicable.


## 1. Introduction.

Wire-wound resistors generally obey Ohm's law

$$V = i\,R, \tag{1}$$

where $R$, the resistance value, is a function of temperature.

Equation (1) does not entirely hold for carbon resistors. A better approximation of the voltage-current relation for this type of resistor is given by

$$V = i\,R_a \pm a\,|\,i^n\,|\,. \tag{2}$$

(The notation $\pm$ in this equation must be interpreted in such a way that for positive $i$ the $+$ sign must be used and for negative $i$ the $-$ sign. In this way $V$ is a symmetrical function of $i$, irrespective of the value of $n$, as in the case with carbon resistors).

If we define $R$, the resistance value, as the quotient of $V$ and $i$, (2) yields

$$\frac{V}{i} = R = R_a + a\,|\,i^{n-1}\,|\,. \tag{3}$$

*) Physical department.

For carbon resistors $a\,|\,i^{n-1}\,|$ is small compared to $R_a$ and negative. Both $R_a$ and $a$ are temperature dependent, though the temperature dependence of the small term $a\,|\,i^{n-1}\,|$ will be generally neglected.

For $n = 2$, (3) becomes

$$R = R_a + a\,|\,i\,| \approx R_a + a\,\frac{|V|}{R_a}\ ,$$

$$\text{or} \qquad R - R_a = \triangle R \approx \frac{a\,|V|}{R_a}\ . \tag{4}$$

In this case $\triangle R$ is proportional to the applied voltage and a "voltage coefficient" can be defined i.e. the percentage resistance variation per volt. So

$$\text{voltage coefficient} = \gamma = \frac{100 \times \triangle R}{R_a \times |V|} \quad \text{or}$$

$$\triangle R = \frac{\gamma \cdot R_a \cdot |V|}{100}\ . \tag{5}$$

Comparing (4) to (5) gives

$$a = \frac{\gamma \cdot R_a^2}{100} \tag{6}$$

It is common practice to give an allowable upper limit for $a$, as it results from testing procedures for carbon resistors. It will be shown, that the exponent in (3) is not necessarily 2, so that the "voltage coefficient" concept cannot always be used.

## 2. Conventional method for the determination of $a$ or $\gamma$.

A straightforward method to determine $a$ or $\gamma$ is to measure $R$ as a function of $i$ or $V$. By substituting the measured values in (3) $a$ and $\gamma$ can be calculated.

A difficulty arising when the method is applied is that the measuring current will cause the resistor to rise in temperature, which means that $R_a$ changes and a wrong value of $a\,|\,i^{n-1}\,|$ results. An obvious possibility to overcome this difficulty is to measure $R$ so rapidly that no appreciable rise in temperature will take place. In e.g. the procedure as described by the Radio Industry Council in the specifications RIC 112 and 113 for "resistors, fixed, composition", such rapid measuring is pre-

scribed. "The samples shall be dried for 2 hours at not less than 70°C. The voltage coefficient shall be determined by measuring the resistance value at a voltage corresponding to the normal maximum rating and again at 1/10th of that voltage. The voltage shall be applied for not more than 1 second every 10 seconds to avoid overheating".

For good carbon resistors, however, where $\alpha$ and $\gamma$ are small, the said value of 1 second proved to be unacceptably long. An experimental study of the rate of heating of resistors of different kind was made and the influence on resistance measurements could be estimated (see Section 8).

A description of a proposed new method for the determination of $\alpha$ and $\gamma$ will be given in the next Section.

## 3. Determination of $\alpha$ and $\gamma$ by measuring the harmonic production.

When a pure sinusoidal current $i = i_m \sin \omega t$ passes through a carbon resistor, harmonics will be present in the voltage across the resistor, due to the non-linearity of (2) or

$$V = R_a\, i_m \sin \omega t \pm \alpha \,|\, i_m^n \sin^n \omega t \,| \,. \tag{7}$$

The Fourier expansion of this equation is

$$V = R_a\, i_m \sin \omega t + E_1 \sin \omega t + E_3 \sin 3\omega t + E_5 \sin 5\omega t \,. + \ldots . \tag{8}$$

Only uneven harmonics appear, due to the symmetry of (2) and (7).

The most important factor of (8) is $E_3$. Application of Fourier analysis to (7) and (8) gives

$$\int_0^\pi E_3 \sin^2 3\omega t \, d\omega t = \int_0^\pi \alpha\, i_m^n \sin^n \omega t \sin 3\omega t \, d\omega t \,,$$

so that

$$\frac{E_3 \times \pi}{2} = \alpha\, i_m^n \int_0^\pi \sin^n \omega t \sin 3\omega t \, d\omega t \,. \tag{9}$$

The integral in (9) which will be called $f(n)$ has the value

$$f(n) = \int_0^{\pi} sin^n \omega t \, sin \, 3\omega t \, d\omega t = -\frac{n-1}{n+3} \frac{\frac{n}{2}!}{\frac{n+1}{2}!} \sqrt{\pi}, \quad (10)$$

$$\text{for } n > -\frac{1}{2}$$

so that (9) can be written in the form

$$E_3 = \frac{2\, a\, i_m^n f(n)}{\pi} = a\, i_m^n \times Q\,, \quad (11)$$

where $Q = 2 f(n)/\pi$.

The following table gives values of $f(n)$ and $Q$ for a number of values of $n$

| $n$ | $f(n)$ | $Q$ |
|---|---|---|
| 1.0 | 0 | 0 |
| 1.5 | 0.11 | 0.07 |
| 2.0 | 0.27 | 0.17 |
| 2.5 | 0.34 | 0.215 |
| 3.0 | 0.39 | 0.25 |
| 3.5 | 0.43 | 0.275 |

## 4. Experimental procedure.

*a*. A very simple realization of the principle laid down in Section 3 would be the following.

A generator supplies a pure sinusoidal audio frequency current to the resistor under test in series with a high-value wire-wound resistor. The amplitude of the harmonic voltage across the resistor is measured and $a$ and $\gamma$ can be calculated with the formulae of Section 3 and 1.

A numerical example, however, shows the difficulties encountered in this procedure. Take, for example, a 1 watt 1000 $\Omega$ resistor for which $n = 2$, so that the voltage coefficient definition can be applied. An accepted maximum value for good carbon resistors for $\gamma$ is 0.002 % per volt, which, for the resistor considered here, means that under full load ($V_{eff} = 31.6$ volt), according to (5) and (6),

$$\triangle R = 0.63\ \Omega \text{ and } a = 20\ \Omega \,.\, A^{-1}\,.$$

The amplitude of the fundamental current at full load follows from the relation $\frac{1}{2} i_m^2 \times R = W$ or $\frac{1}{2} i_m^2 \times 1000 = 1$ so that $i_m^2 = 0.002$ and according to (11)

$$E_3 = 20 \times 0.002 \times 0.17 = 6.8 \times 10^{-3}\ \mathrm{V} = 6.8\ \mathrm{mV},$$

the amplitude of the fundamental voltage at the same time being $31.6\sqrt{2} = 45$ volt.

It is clear that for the execution of the measurement indicated here, the harmonic content of the generator must be considerably less than 6.8 mV, the harmonic voltage generated in the resistor. This means that, the fundamental voltage being 45 V, the harmonic content of the generator must be less than $1.5 \times 10^{-2}\ \%$. This is, of course, difficult to realize. It is also difficult to measure the 6 mV harmonic voltage in the presence of 45 V fundamental voltage, so that the wave analyzer for the measurement of $E_3$ will be complicated.

*b.* An alternative method, in which both the fundamental and the harmonic voltage of the supply are balanced in a bridge, will be discussed now.

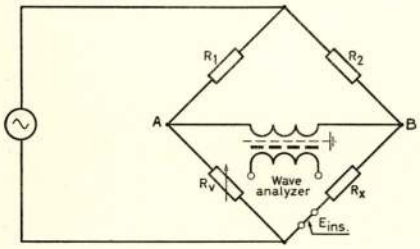


Fig. 1.
Wheatstone bridge arrangement for the measurement of harmonic production in carbon resistors.

The resistor under investigation $R_x$ is introduced as one of the four arms in a Wheatstone bridge, together with three wire-wound resistors, one of which is adjustable. All four resistors have approximately the same ohmic value. Audio-frequency voltage is supplied to the bridge in such a way that $R_x$ is loaded to its rated wattage. A wave analyzer that can be tuned to the fundamental as well as to the third harmonic voltage of the generator frequency, is connected to the bridge diagonal. Bridge equilibrium for the fundamental voltage is established by means of the adjustable resistor. In this situation both the fundamental and the third harmonic frequency in so far as they are produced by the generator are not present in the bridge diagonal, so that measurement of the third harmonic voltage in the diagonal now

gives an undisturbed indication of the harmonics produced in $R_x$. Let the measured voltage of the third harmonic be $E_3'$. The relation between $E_3'$ and the value of $E_3$ as it appears in (8), can be calculated for a given bridge configuration. It is, however, simpler to determine this relation by calibration. This can easily be done by inserting a given small voltage of any frequency in series with $R_x$ ($E_{ins}$ in fig. 1) and measuring the voltage produced by it in the wave analyzer in the bridge diagonal. The measured relation between these two voltages is the same as that between $E_3$ and $E_3'$, provided the bridge configuration is the same in both cases.

## 5. Some details of the method.

The measurements can be carried out with standard equipment i.e. no excessive requirements are made on the wave analyzer or the audio frequency oscillator. The oscillator must be able to supply the rated wattage, which means that when e.g. a 1 watt resistor is under test, 4 watts must be supplied to the bridge.

The transformer in the bridge diagonal must have a carefully wound symmetrical primary winding, so that points $A$ and $B$ (fig. 1) present the same parasitic impedance to earth. An electrostatic screen between primary and secondary winding must be provided in order to prevent false bridge equilibrium. A simple method to verify whether these conditions are satisfied is, to take for $R_x$ a wire-wound resistor and to measure whether the fundamental and harmonic voltages in the bridge diagonal cancel out simultaneously.

## 6. Some measurements of non-linear behaviour of carbon resistors.

A number of resistors of different values and constructions was measured according to the method outlined in Sections 4 and 5. The applied audio-frequency was 250 c/s, so that measurements of harmonics were made at 750 c/s. Harmonics were measured at a number of applied voltages up to the rated voltage. Fig. 2a and 2b give the values of $E_3$ as a function of the peak value of the applied fundamental voltage, in a double logarithmic plot. Fig. 2a gives values for cracked-carbon resistors, 2b for composition-type resistors.

It appears from the straight lines in figure 2a, that for cracked-carbon resistors a simple exponential relation as indicated by (2) and (7) is valid. For composition-type resistors no such simple relation holds, as is shown by the curved lines in fig. 2b.

The value of the exponent $n$ in (2) can be determined from the graphs by means of (11), from which we derive

$$log\, E_3 = n\, log\, i_m + log\, a + log\, Q\,. \tag{12}$$

$n$ turns out to be the tangent of the lines in fig. 2a, assuming

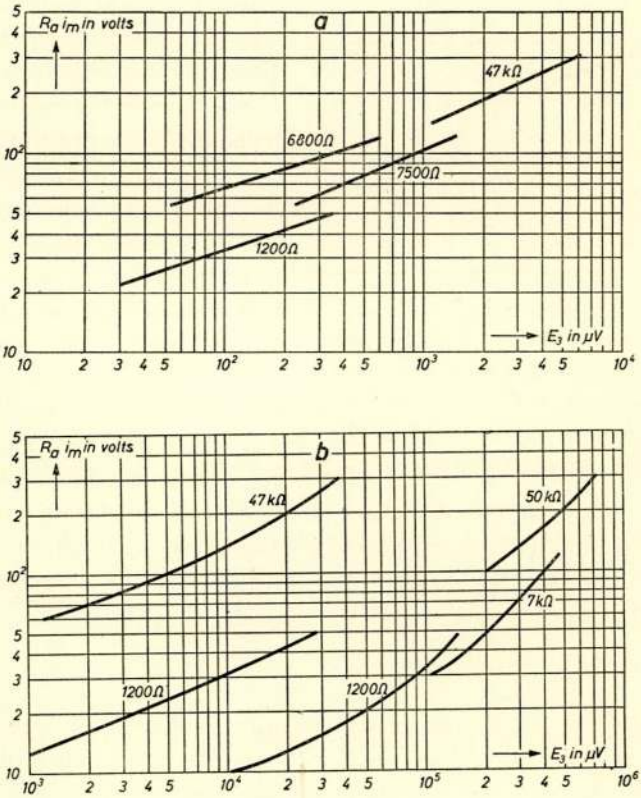


Fig. 2.

Peak value of third harmonic voltage produced by the passage of a current $i_m \sin \omega t$ through a resistor $R_a$ as a function of $i_m R_a$.

Fig. 2a values for cracked-carbon resistors

Fig. 2b values for composition-type resistors.

that $a$ (and so $\log a$) is independent of temperature. Since the points of the lines of fig. 2 are measured at different voltages, which means different temperatures of the resistor, $\log a$ in (12) would not be a constant if $a$ were a function of temperature. In Section 9 it will be shown by separate measurements that $a$ is indeed only slightly dependent on temperature, so that $\log a$ can be taken as a constant.

## 7. Verification of the method.

For the composition-type resistors, where the non-linear effects are important, the conventional method of Section 2, viz. a direct current method, gives reliable results, so that in these cases a comparison can be made with the new method proposed in the present article.

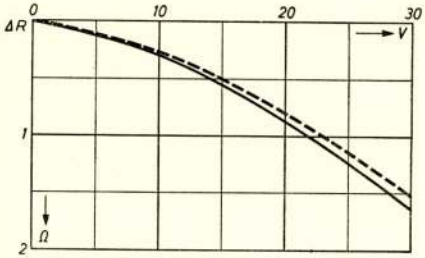


Fig. 3.
$\triangle R$ as a function of the voltahe across a resistor, determined by two methods,
——— by a DC method
– – – – – by harmonic production measurement.

One of the resistors of fig. 2b, the 1200 $\Omega$ one, was chosen for this purpose and its resistance for a number of different voltages in a Wheatstone bridge was measured by means of direct current. The voltage was applied for the shortest time necessary to make reading possible. Fig. 3 gives $\triangle R$, the deviation of the actual resistance from that measured with a very small voltage, as a function of the applied voltage. It proves to be possible to represent the function by

$$\triangle R = 7.0 \times 10^{-3} \times V^{1.60} \tag{13}$$

On the other hand, the measurements of harmonic production of this resistor as shown in fig. 2b enable us to determine $\triangle R$. From the tangent of the slope of the function, it appears that $n$ is 2.35, so that $Q$ in (10) = 0.22. For 30 volt we read from the graph $E_n = 9 \times 10^{-3}$ V and $R_a$ being 1140 $\Omega$, substitution in (11) yields

$$a = \frac{E_3}{i_m^n \times Q} = \frac{E_3 \times R_a^n}{V^n \times Q} = \frac{9 \times 10^{-3} \times 1140^n}{30^n \times 0.22} .$$

Substitution in (3) gives

$$\Delta R = a\, i^{n-1} = a \times \frac{V^{n-1}}{R_a^{n-1}} = \frac{9 \times 10^{-3} \times 1140^n}{30^n \times 0.22} \cdot \frac{V^{n-1}}{1140^{n-1}} =$$

$$= \frac{9 \times 10^{-3} \times 1140}{30^{2.55} \times 0.22} \times V^{1.55} = 7.8 \times 10^{-3} \times V^{1.55} \quad (14)$$

The dotted line in fig. 3 represents this function. It will be seen that, although the value of the exponent in (13) and (14) is slightly different the over-all representation of the results of both methods gives consistent values.

It would be very interesting to compare the results of both methods for the cracked-carbon resistor in fig. 2a. The difficulty is obvious from the following. The calculation of this Section applied to the 1200 $\Omega$ resistor in fig. 2a would give a $\Delta R$ value at full load of 0.02 $\Omega$. In Section 2 it was mentioned that the measurement of such small resistance variations is seriously hampered by the heating of the resistor by the measuring current even for a very short measuring time. The next Section will give some details about this heating-up effect.

## 8. Heating by the measuring current.

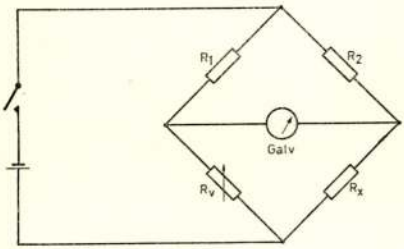


Fig. 4.
Wheatstone bridge arrangement for the measurement of the resistance change $\delta R$ due to the heating effect of a current.

$R_x$, the resistor to be tested, is again part of a Wheatstone bridge, which is approximately brought into equilibrium by means of the variable resistance $R_v$. At a moment $t = 0$ the voltage is applied to the bridge circuit and the galvanometer is read as a function of time. From this the value of $R_x$ as a function of time after the switching on of the current can be calculated or determined by calibration.

In fig. 5 the resistance is plotted as a function of time in such a way that $\delta R$, the difference of the resistance value at time $t$ compared with the initial value, is given as a function of time. The three examples give three different types of heating effect. In fig. 5a, representing the behaviour of the cracked-

carbon resistor of 1200 $\Omega$ in fig. 2a, $\delta R$ can be accurately described by an exponential function with one time-constant

$$\delta R = -18\,(1 - e^{-t/40}) \ ,$$

where $t$ is expressed in seconds.

For the composition-type resistor in fig. 5b, we find

$$\delta R = -10\,(e^{-t/200} - 1) - 20\,(e^{-t/25} - 1) \ .$$

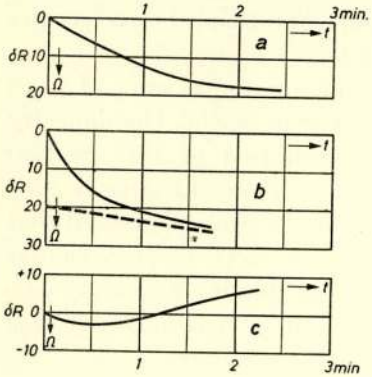


Fig. 5.
Resistance chance $\delta R$ (fig. 4) as a function of time for three resistors of different construction.

Fig. 5c represents a resistor for which the resistance is not a linear function of temperature so that an irregular heating curve results.

From the functions shown in fig. 5a and 5b for which a simple mathematical expression can be found, an estimation can be made of the resistance change immediately after the switching on of the current. In the case of fig. 5a, $\delta R$ is 0.045 and 0.45 $\Omega$ after 0.1 and 1 sec respectively. This value of 0.045 $\Omega$ compares very unfavourably with the 0.02 $\Omega$ calculated for the non-linear effect of the same resistor at full load (see end of preceding Section). This means that even after so short a time as 0.1 sec the resistance change due to heating is already more than double the one to be expected from the non-linear effect.

## 9. Temperature dependence of $\alpha$.

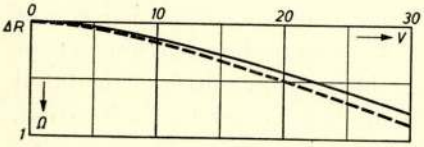


Fig. 6.
$\triangle R$ as a function of $V$ at two temperatures.
——— at 70° C
– – – – – at 90° C

It was mentioned in Section 6 that the calculations made are there valid only if $\alpha$ is independent of temperature. That the latter is indeed the case can be proved by using a resistor with pronounced non-linear effects.

Figure 6 gives the measure-

ment of $\triangle R$ as a function of $V$ with short pulses of DC for two temperatures of 20° and 70° C respectively. It will be seen that for this rather large temperature difference $\triangle R$ does not change very much so that in view of the limited accuracy sufficient for this kind of measurements it will be allowed to neglect this second-order effect.

## 10. Conclusions.

*a.* The conventional method of determining the non-linear properties of carbon resistors is only applicable to resistors where this effect is appreciable e.g. composition-type resistors.

*b.* The concept of voltage coefficient is only applicable to cases where in equation (2) $n = 2$. This is true only in a limited number of cases.

*c.* The conventional method cannot be applied to good resistors because even measuring times as short as 0.1 sec. give heating-up effects that can be several times greater than the effect we want to measure.

*d.* A method is proposed that is based on the measurement of the production of harmonics and does not present the difficulties indicated in *c*. This method has the additional advantage that the measured quantity, viz. the amplitude of the harmonics produced, is in fact, in the majority of cases, the reason why we are interested in the non-linear effects of resistors.

*e.* For resistors with pronounced non-linear behaviour, where both methods could be used, the agreement between the results turned out to be quite satisfactory.

Acknowledgement.

I am indebted to Mr. W. E. Knip for his valuable assistance in this investigation.

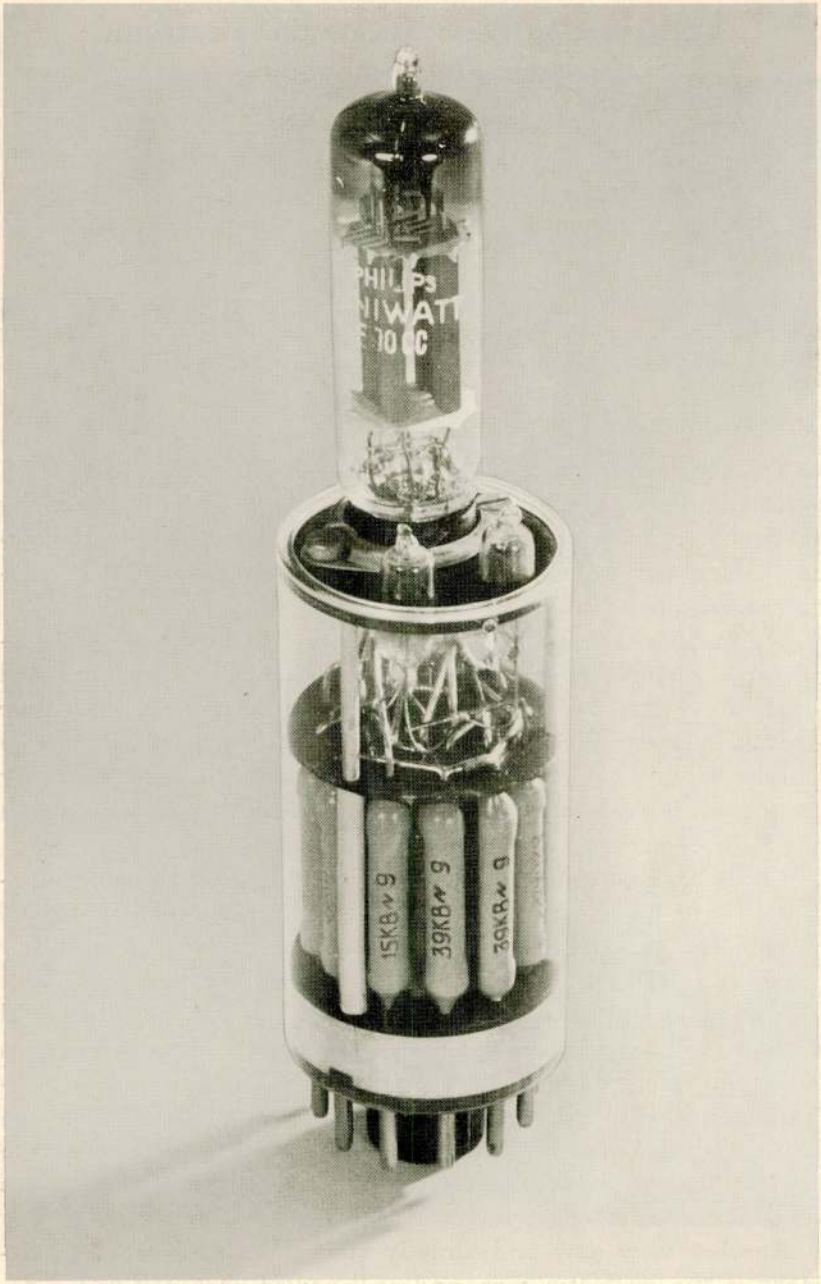


Fig. 1
Early trigger circuit

stage or not, depends on many factors, some of which will be reviewed hereafter.

Radio and switching have now a common base in the form of a group of elements which are used in both techniques. The design methods in radio and switching, however, are completely different. Radio is an art of transmission of information, while switching concerns processing of information. The information to be processed is generally in digital or coded form. A widely used form is coding in the form of binary numbers, which consist of the digits 0 and 1 only. Binary numbers can be stored on bistable switching elements. A relay is a typical example of such an electro-mechanical element. Many types with different windings and spring sets have been developed. In the last few years it has proved possible, by applying systematic circuit design methods, to standardise on a few types, which can be machine-made.

## 2. Electronic building stones

In the beginning of the development of electronic switching circuits a building stone analogus to a relay was missing. It has been one of the most important projects of the Switching Laboratory to develop such an electronic building stone. In 1949 this laboratory started to develop fully electronic switching techniques with diodes and triodes. The design of 5 building stones, of which a trigger was the most important one, has been finished in the spring of 1951. This building stone is shown in fig. 1, its circuit in fig. 3. It consists of a double triode (E90CC Philips) on top of a cylindrical aluminium cover (which has been replaced in the photograph of fig. 1 by a perspex cover) containing the resistors of the trigger circuit. A 12-pin tube socket closes the bottom of the aluminium cover. Two small neon tubes are placed on top of the cover. They indicate the position of the trigger.

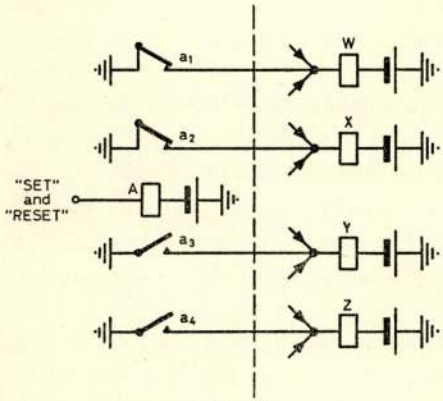


Fig. 2
A relay controlling four other relays

A trigger circuit can be compared in many cases with a relay

with one make and one break contact. It is possible in electronic switching to control a much larger number of other circuits by means of diode gates. These diode gates are the equivalents of the contacts in relay circuits. Fig. 2 shows a relay circuit in which a relay controls four other relays *W, X, Y,* and *Z*. Fig 3

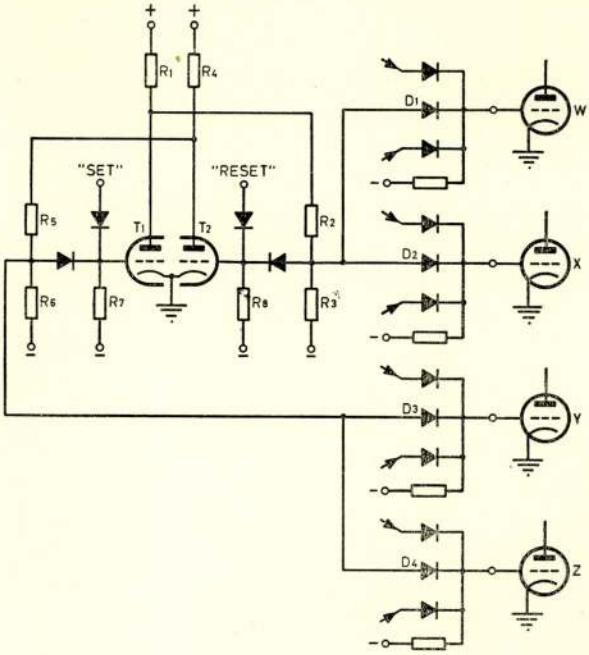


Fig. 3
Equivalent electronic circuit of fig. 2

shows the equivalent electronic circuit in which a trigger *A* controls four tubes *W, X, Y,* and *Z*. The diodes $D_1$ up to $D_4$ function in a way analogous to the contacts $a_1$ up to $a_4$. The relays can be controlled by a plurality of contacts as is indicated by the multiple arrows. The electronic elements *W, X, Y,* and *Z* can be controlled by a plurality of diodes, connected to other trigger circuits.

The resistors of the trigger shown in fig. 1 have an accuracy of 5%. The development of this trigger was ready in the spring of 1951. The nominal working voltage is 220 volts dc, but the trigger operates on voltages from 130 volts up to 300 volts.

A special problem in designing these building stones with high vacuum tubes forms the heat dissipation of the filaments (about 2.5 watts per tube) and of the anode current (about

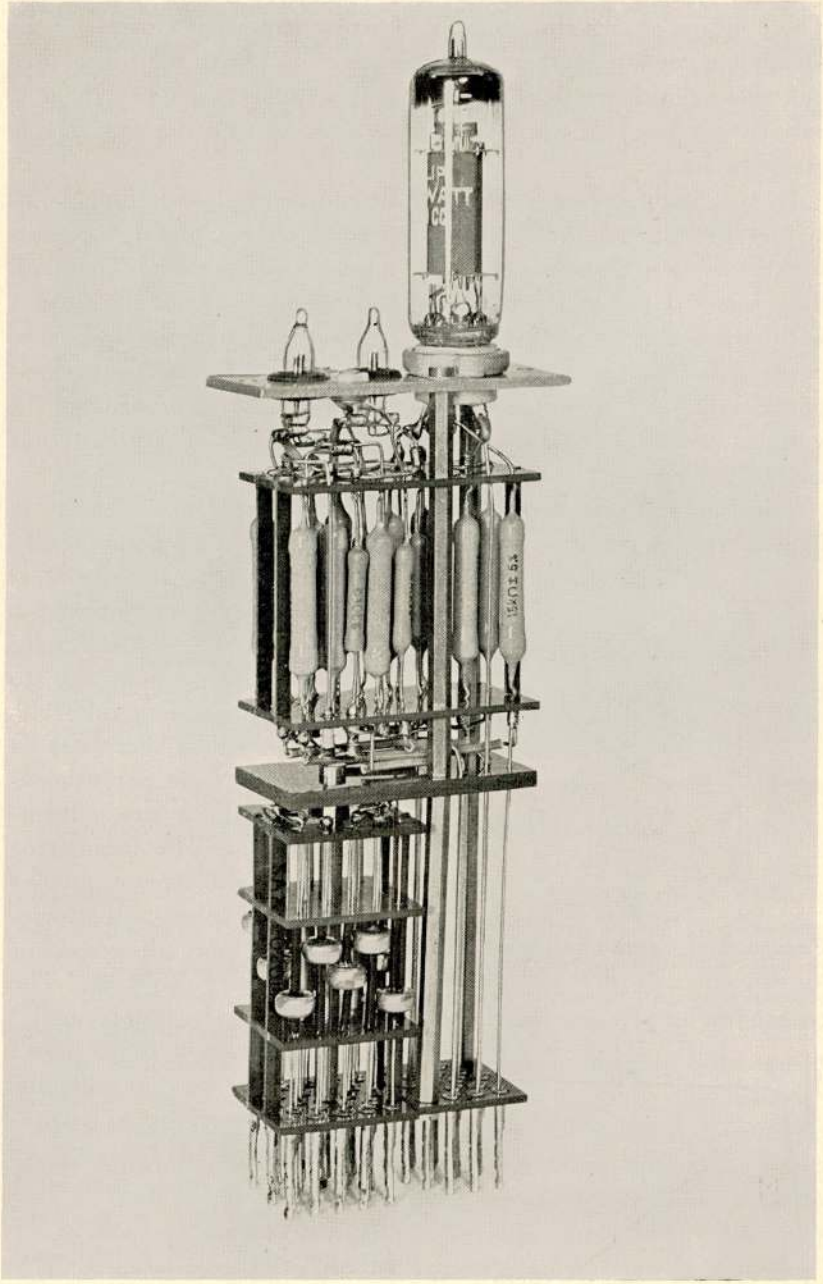

Fig. 4
Open trigger circuit with built-in diodes.

of elements. The plug contacts which form a part of the copper-plating of the card are gold-plated. These cards have the great advantage that wiring faults are impossible in the reproduction. Moreover each card is an exact copy of the original circuit, which ensures a great stability in the final product. These cards are well suited for machine insertion of components and for dip soldering.

Very complex automatic electronic circuits can be designed with these components. It does not belong to the scope of this article to review all the possibilities. There are other components which can be used in switching, and which have not been described here. The most important one is the magnetic core, which has two stable magnetised conditions. Other components as e.g. the so-called ferro-electric materials are under development, but have not reached the practical stage. It would go too far to describe all the future possibilities in this article which has been confined to the description of some practical electronic building stones for switching purposes.

# A portable instrument for measurements on intermediate-frequency level on frequency modulated microwave radio links

by J. W. A. van der Scheer *)


## Summary

A description is given of a portable instrument used for measuring characteristics of frequency modulated microwave radio links. The measurements are carried out in the i.f. band (frequency-ranges 60—80 Mc/s or 95—115 Mc/s) and involve amplitude and group delay variations versus frequency (with an accuracy better than 0,1 dB and 1 *n*sec respectively), the linearity of the discriminator (within 1 %) and the reflection-coefficient of the input and output of i.f. units up to 0.01. The measured characteristics are made visible on a c.r. tube.

Separate i.f. components can be measured as well as the overall behaviour of a long-distance circuit. Moreover one-way measurements can be made.

The instrument is developed both for the maintenance of microwave equipment and for laboratory tests.


## 1. Introduction.

In transmitting wide-band information e.g. multichannel telephony or television (50 c/s—5 Mc/s) over a radio-path, use is frequently made of microwave links on centimeter waves with frequencies between 4,000 and 11,000 Mc/s, frequency modulation being applied.

For such a transmission high demands are usually made on the quality and the stability of the link. On these frequencies atmospherics are not important but the radio signals are subject to fading. Besides the quality of the transmission is dependent on the thermal noise level and the distortion caused by the epuipment. In what follows special attention will be paid to distortion.

Distortion, which in television e.g. gives rise to a less good gradation, and in multichannel telephony to crosstalk and intermodulation noise is to a large extent introduced in the i.f. part of the microwave radio equipment. [1] [2] [3]

*) Radio laboratory.

If, taking into account the technical possibilities, the maximally admissible distortion contributions are divided somewhat equally, certain requirements can be drawn up for the separate components such as amplifier, limiter, discriminator and modulator.

Specification of the contributions of the various components has the advantage that parts can be replaced without re-adjustment and without the admissible distortion being exceeded. This procedure will promote the reliability of the equipment, which is of course most important in the PTT-service.

For the development, the adjustment and the maintenance of the microwave equipment and for the checking of the equipment before being accepted from the manufacturers, it was necessary to have an easily transportable measuring set available, by means of which the properties of the various components could be determined.

Hereafter a brief description will be given of the measuring set which was developed in the PTT Dr Neher laboratory, and which is suitable for the making of those measurements on intermediate frequency level which are of importance for the reliability of a microwave radio link. In this connection there may be mentioned: Amplitude and group delay variations as function of the frequency, linearity of the demodulator and the reflection coefficient at the terminals.

## 2. Principle of the measurements.

The block schematic diagram (fig. 3) gives the main essentials of the measuring set, consisting of two parts: the sending and receiving unit. The frequency range on i.f. level is centred round 70 or 105 Mc/s. Characteristics, which are displayed on a cathode-ray tube, are:

a. the amplitude versus frequency
b. the group delay versus frequency
c. the discriminator-curve
d. the reflection coefficient at the connection points.

The scanning of the frequency range to be measured is effectuated by means of a so-called Megasweep, being part of the sending unit. The Megasweep is essentially an oscillator varying relatively slowly in frequency, (iterative frequency 50 c/s) and consists in principle of two klystrons (*a* and *b*), operating on a frequency of approximately 9,000 Mc/s with a 70 or 105 Mc/s mean beat-frequency. The klystron *a* is frequency mo-

dulated by the horizontal deflection voltage of the oscilloscope (maximum deviation of $\pm 15$ Mc/s).

The output voltage of the crystal mixer is very small and moreover shows amplitude modulation, which makes amplifying by a wide-band amplifier and limiting necessary (Group-delay variations in the measuring equipment are corrected by a phase-equalizing network). For some measurements it is necessary to have superposed on the slow frequency variation of the Mega-sweep a more rapid variation with a constant, small deviation. This is effectuated by means of frequency modulation of the klystron *b*.

### 2.1. *Measurement of the amplitude frequency characteristic.*

A signal with a double frequency modulation passes through the i.f. under test. The frequency modulation consists of a slow one over the entire band to be measured, controlled by the horizontal deflection voltage of the oscilloscope and one with high frequency and small deviation. The index of the last-mentioned modulation is chosen smaller than 1, so that the frequency-spectrum consists mainly of a carrier and two sidebands.

Dependent on the slope of the amplitude frequency-characteristic the frequency modulated signal will vary in amplitude. After amplitude detection at the output of the network and selective amplification of the high modulation frequency the last signal is supplied to the oscilloscope, the slope being displayed on the c.r. tube as function of the frequency.

The detector (fig. 1) has been developed as a special coaxial detector with a flat amplitude and phase characteristic up to 1,000 Mc/s. This detector can serve as an element in a circuit or if terminated with its characteristic impedance of 75 ohm as a load impedance at the output of an i.f. amplifier. The detector consists of a crystal diode and a low-pass filter contained in a piece of coaxial conductor having a somewhat larger diameter than the standard type. The output is attainable at the upper side. A second plug serves for the injection of the signal of the frequency marker oscillator.

### 2.2. *Measurement of group-delay variations.*

To determine group-delay variations as function of the frequency a method is used which was applied by Nyquist in 1930. [4]

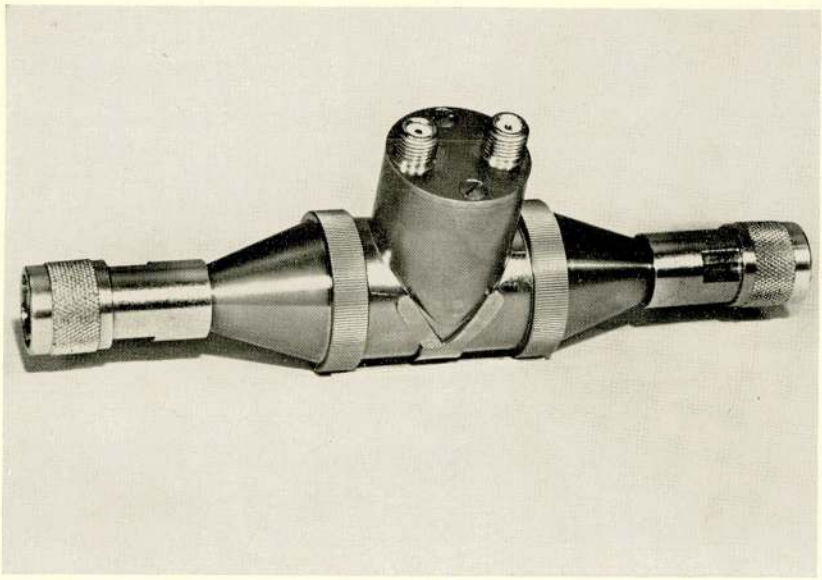

Fig. 1.
Coaxial detector.

The principle of the measurement is shown in fig. 2 Just as in 2.1. a double frequency modulation of the Megasweep is effectuated.

The output of the network under test is supplied to the amplitude detector and then amplification of the modulation frequency is effectuated. Dependent on the derivative of the phase characteristic of the network $\frac{da}{d\omega} = \tau$ ($a$ = phase angle) the phase of the demodulated signal will be shifted with respect

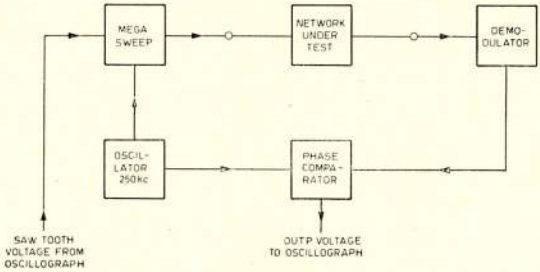


Fig. 2.
Simplified block schematic diagram
of the measurement of group delay.

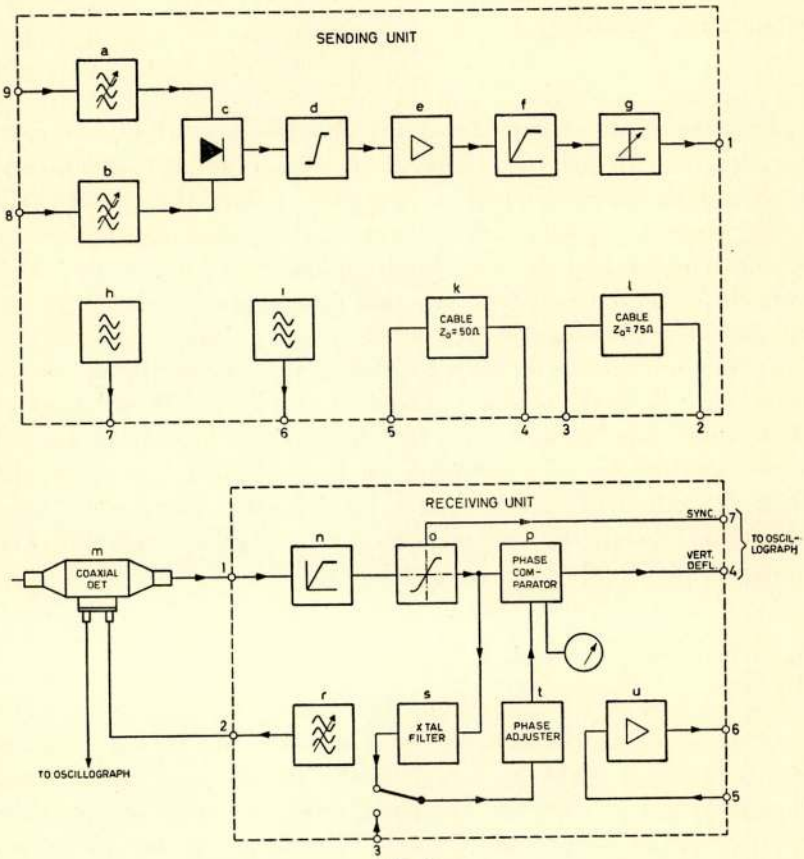


Fig. 3.

Block schematic diagram of the whole equipment.

a. klystron oscillator
b. ,, ,,
c. mixer
d. phase equalizer
e. amplifier (70 or 105 Mc/s)
f. limiter (70 or 105 Mc/s)
g. attenuator
h. xtal oscillator 250 kc/s
i. ,, ,, 60 kc/s
k. cable $Z_0 = 50$ ohm (20 meter)
l. ,, $Z_0 = 75$ ohm (12 meter)
m. coaxial detector
n. limiter
o. discriminator
p. phase comparator
r. frequency marking oscillator
s. xtal filter 250 kc/s
t. phase adjuster
u. amplifier 60 kc/s

to the phase of the modulating signal. This phase difference is now measured by means of the phase comparator. It is proportional to the group-delay $\tau$ and to the frequency $p$ of the modulating signal

$$\varphi = 2\pi p\tau$$

Because of noise considerations and the desirability of measuring group delay variations of 1 *n*sec ($10^{-9}$ sec) accurately, $p$ should be given the highest possible value, the upper limit being given by a sufficient amount of the resolving power. So a compromise had to be found and a frequency of 250 kc/s was ultimately decided on. In this case a group-delay of 1 *n*sec results in a phase shift of $\varphi = 0.09°$.

In this connection it may be stated that a measuring equipment for television receivers and developed by van Weel operates with amplitude modulation. However, this equipment is not very suitable for measurements on f.m. systems, because of the limiters contained in the latter. [5] Several instruments have been constructed for the measurement of group delay in f.m. systems. [6] [7] [8]

## 3. Measuring method.

Figure 4 is a front view of the measuring equipment. The arrangement can be built up by means of short pieces of cable. By interchanging the parts *d*, *e* and *f* of the sending unit and the parts *n* and *o* of the receiving unit, i.f. amplifiers with a central frequency of both 70 and 105 Mc/s can be measured. See block schematic diagram fig. 3.

### 3.1. *Measurement of the amplitude frequency characteristic of an i.f. amplifier.*

The amplifier to be tested is connected to the output of the sending unit, if required via a matching network. The output of the amplifier is fed to the coaxial detector terminated by 75 ohm.

The input 9 of klystron *a* is connected to the horizontal deflection voltage of the oscilloscope and the detector output to the vertical deflection input. The amplitude frequency characteristic becomes visible on the screen. By inserting a 1 dB attenuator and by making use of the d.c. input of a Tektronix

Fig. 4.
General appearance of sending unit (below) and the receiving unit (above).

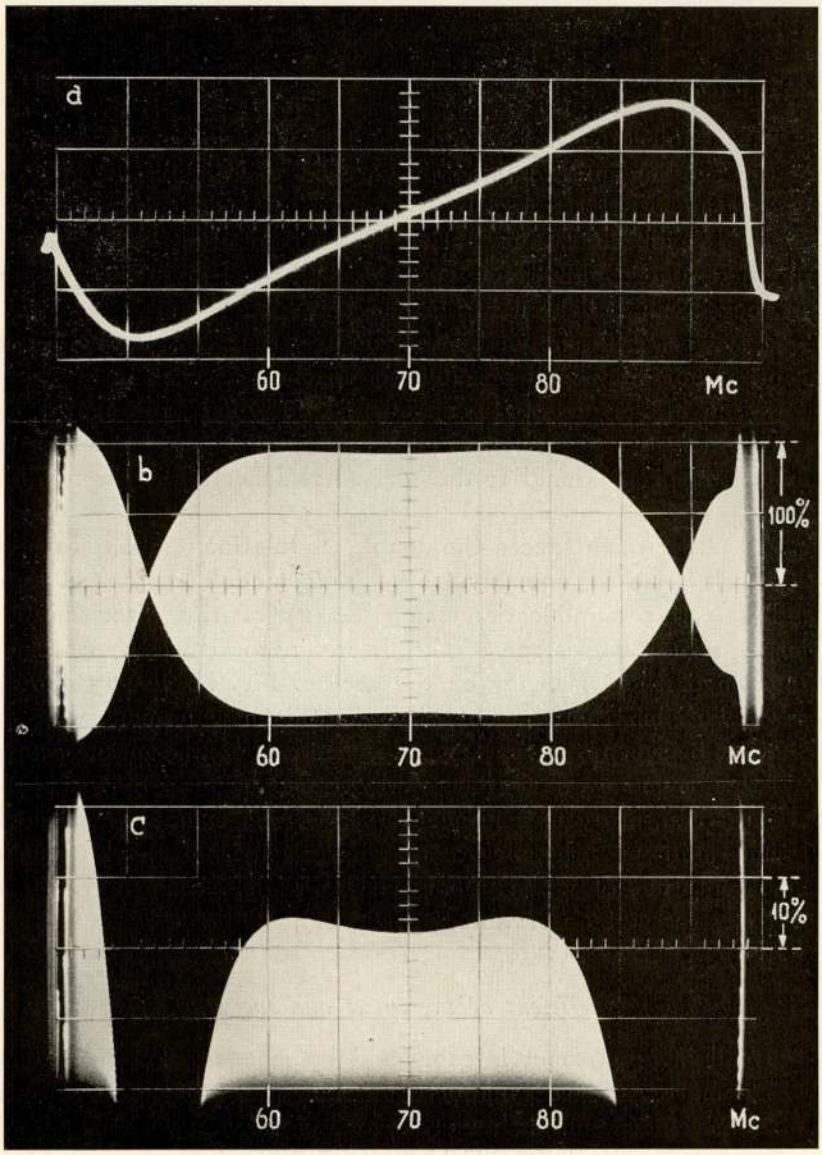


Fig. 6.

a. discriminator output as function of the frequency

b. slope of the discriminator

c. top of figure 6b five times amplified.

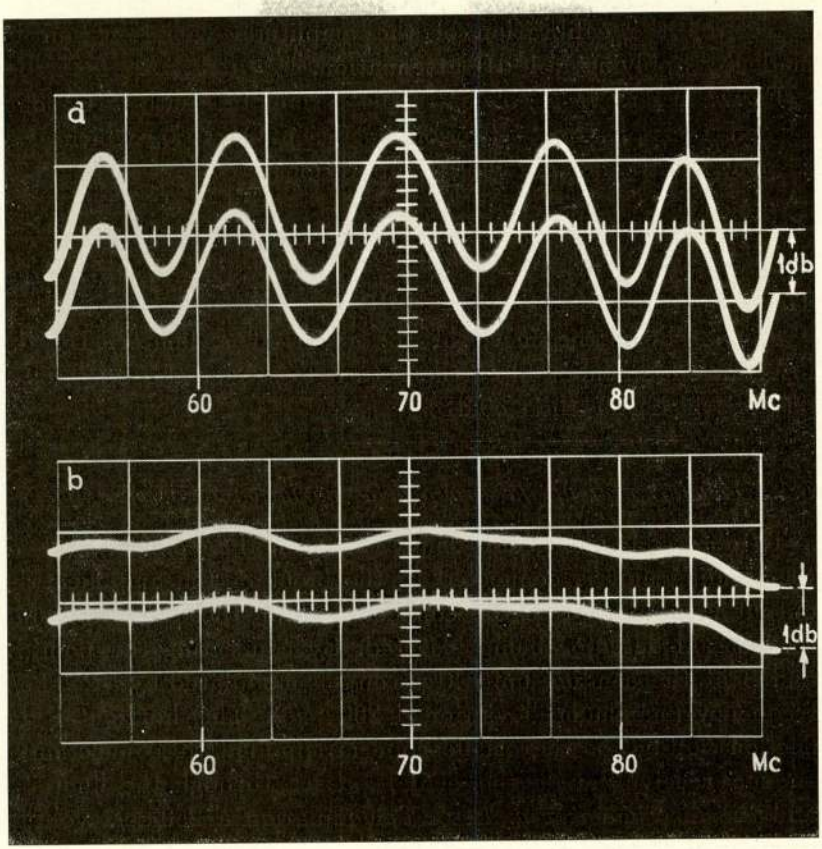


Fig. 7.

a. amplitude variations at the beginning of a 12 m long cable with appreciable mismatch at the end ($k = 0.09$)

b. amplitude variations at the beginning of the same cable terminated with the input of a limiter ($k < 0.003$).

In order to be able to determine the reflection coefficient of an input or output, the amplifier is connected via a long cable 1 (present in the sending unit) to the output 1, the coaxial detector being inserted.

At a 60 to 80 Mc/s deviation there will be voltage variations as function of the frequency (fig. 7). A simple deviation shows that the reflection coefficient

$$K = \frac{e_{max} - e_{min}}{e_{max} + e_{min}}$$

In this formula $e_{max}$ = the maximum voltage on the detector
$e_{min}$ = the minimum voltage on the detector.

Here too the calibration of the amplitude variation can be made by applying a 1 dB attenuation.

Fig. 7a refers to a 75 ohm cable terminated with a 62 ohm carbon resistor. The reflection coefficient is 0.09, which corresponds with the following computation:

$$K = \frac{75 - 62}{75 + 62} = 0.09$$

The return loss is $20 \log K = 21$ dB.

Fig. 7b shows the curve of the input of a limiter. By the insertion of an input filter the return loss over the band 60—80 Mc/s is better than 30 dB.

### 3.4. *Measurement of the group delay variation of an i.f. amplifier.*

To determine the group delay variation the connection 1 of the sending unit is connected to the input of the amplifier; the output is connected to the connection 1 of the receiving unit, the coaxial detector being inserted. By connecting the points 7 and 8 of the sending unit, klystron *b* is modulated by a 250 kc/s frequency (maximum deviation ± 200 kc/s). The horizontal deflection voltage of the oscilloscope is supplied to connection 9. In the receiving unit the signal passes through the limiter *n* and the discriminator *o*. After demodulation a 250 kc/s voltage varying in phase is obtained. In the phase comparator *p* this voltage is compared with a 250 kc/s reference voltage, which can be taken from the sending unit (connection 7). The reference voltage can also be obtained from the receiving unit via the crystal filter *s*; this filter being so narrow that already the first sidebands of the 250 kc/s phase-modulated signal are eliminated (bandwidth 10 c/s). The mean phase difference can be adjusted by means of the phase adjuster *t*. When the adjustment is correct (phase difference 90°), the comparator indicates zero. The output of the phase comparator (connection 4) is connected to the oscilloscope, synchronised by a signal originating from the discriminator in which the sawtooth voltage is still present (connection 7).

When the crystal filter is used, it is possible to measure a microwave radio link in one direction without the necessity of transmitting the 250 kc/s reference signal. The amplitude characteristic can be made visible together with the group delay curve by connecting the coaxial detector with the oscilloscope.

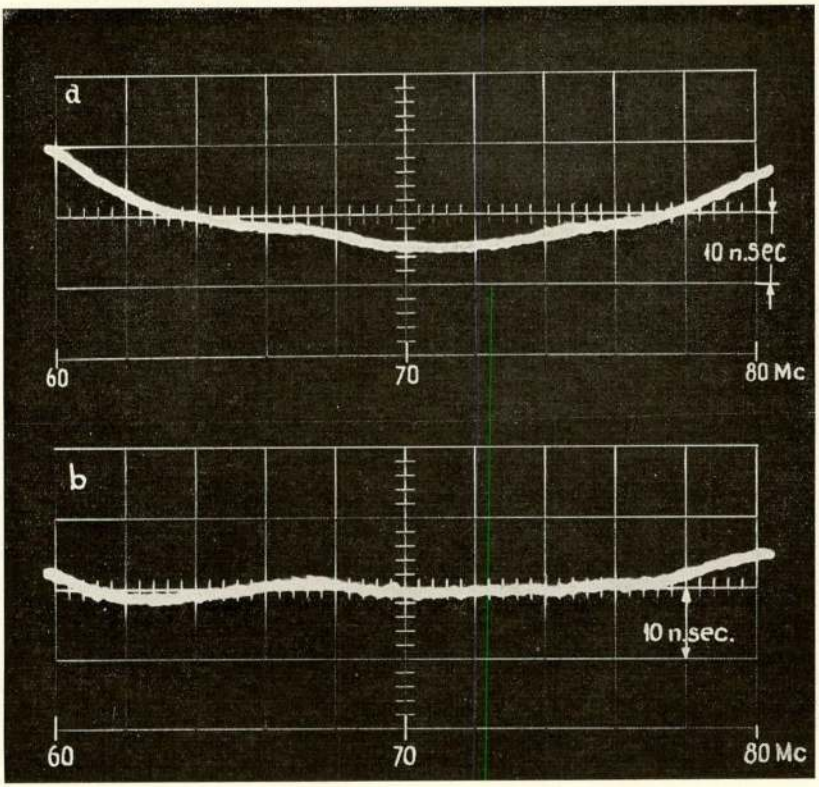


Fig. 8.
Group-delay vs frequency curve on an i.f. amplifier (gain 60 dB)
a. without phase equalization
b. with an equalization filter inserted

For the calibration of the group delay meter the sending unit has been provided with a cable *k* with a characteristic impedance of 50 ohm, having a fluctuating group delay variation of exactly 10 *n*sec.

The figures 8a and 9a show the results of group delay measurements on radio relay equipment without phase correction. By means of phase equalizing networks group delay variations can be reduced as can be seen from fig. 8b and 9b. [9] [10]

## 4. Final remarks.

By making some provisions it is possible to measure waveguide filters lying in the frequency range of the klystron *a*. For

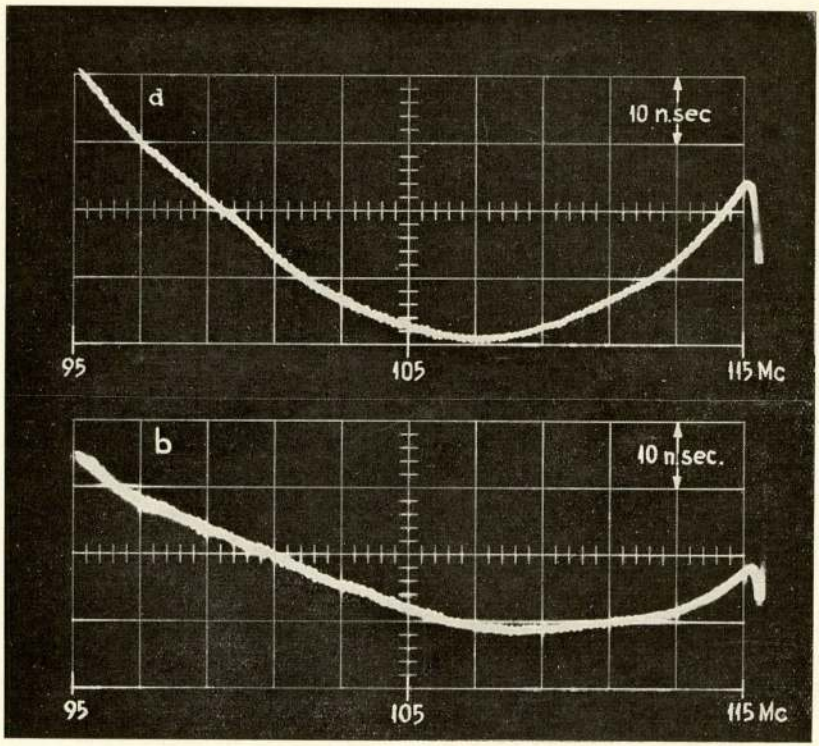


Fig. 9.
Group-delay vs frequency curve of a one-way radio link (40 km)
a. without phase equalization
b. with an equalization filter inserted.

this purpose the filter is inserted between the klystron and the crystal detector *c*, the filter being separated from the klystron by a ferrite isolator.
Klystron *a* is modulated both by the horizontal deflection voltage of the c.r. tube and by the 250 kc/s signal (mean frequency = = centre frequency of the filter); the frequency of klystron *b* is adjusted to give a frequency difference of 70 (or 105) Mc/s.

Furthermore connection 1 of the sending unit is connected to the coaxial detector. For the measuring procedure 3.4. is referred to.

Attention should be drawn to the fact that an amplitude variation of the output voltage of the network under test may cause a group delay variation in the meter (amplitude phase conversion) in consequence of which the amplitude characteristic will affect the group delay curve. This difficulty was over-

come by an effective construction and adjustment of the limiters and by eliminating any feedback in the amplifiers.

The author wishes to mention the contributions of Mr. A. S. van den Bosch and Mr. M. Gerritsen for the general lay-out and of Mr. M. van Sliedregt for the construction of the Mega-sweep.

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# A waveguide filter theory

by M. van Sliedregt *)


## Summary

Following the work of Goubau [2], [3] a general treatment of $\lambda_g/4$ coupled bandpass waveguide filters is given. Formulae for the insertion loss, the envelope delay and in particular the insertion loss at center frequency due to dissipation are derived for waveguide filters consisting of $n$ stages having as parameters the ratios of the bandwidth of the stages to that of the total filter.

By simply modifying these ratios, special filter designs are possible. Examples of a filter consisting of three stages are given.


## 1. Introduction.

With regard to the design of bandpass waveguide filters two methods are available. The first method, which is commonly used, involves the equivalence of the discontinuity in the waveguide with a lumped circuit. The elements of the second method, used by Simon and Broussaud [1] and Goubau [2] [3], are the reflection and transmission coefficients of the discontinuity.

Goubau introduces in his theory a wave matrix, the elements of which can be expressed in these coefficients; the theory is given for two equivalent types of microwave filters, viz. the direct and the $\lambda_g/4$ coupled filters and in particular for filters consisting of identical stages. In special cases, however, on account of this restriction certain requirements cannot be met. In this paper therefore a general treatment, using the wave matrix notation of $\lambda_g/4$ coupled waveguide filters, consisting of different stages, is given. The bandwidth and the center frequency of such a stage, comprising a piece of waveguide and two identical discontinuities, are related to the wavelength in the guide and to the reflection and transmission coefficients of the discontinuities. Assuming that the bandwidth and the center frequency of a single stage can be chosen arbitrarily a variety of filter designs with specific characteristics is obtained by taking specific bandwidth ratios for the filterstages, and keeping the center frequency of all stages the same.

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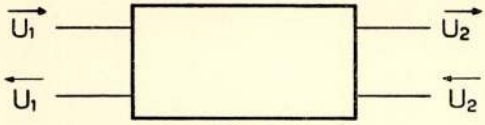


Fig. 1.
Incoming and outgoing waves.

Now the electromagnetic states of a filterstage can be characterised by incoming and outgoing waves (fig. 1). $\vec{u}_1$ and $\overleftarrow{u}_1$ are the (complex) amplitudes of the incoming and outgoing waves on one side and $\vec{u}_2$ and $\overleftarrow{u}_2$ the amplitudes of the corresponding waves at the other side of the filterstage. As the relations between these waves are linear, they can be written in the form of a wave matrix equation

$$\left\| \begin{matrix} \vec{u}_1 \\ \overleftarrow{u}_1 \end{matrix} \right\| = \left\| \begin{matrix} N_{11} & N_{12} \\ N_{21} & N_{22} \end{matrix} \right\| \left\| \begin{matrix} \vec{u}_2 \\ \overleftarrow{u}_2 \end{matrix} \right\| \tag{1}$$

The matrix elements $N$ are closely related to the reflection and transmission coefficients of the discontinuities, quantities that are assumed to be determined directly by measurements or by calculation. The calculation can be carried out by means of an intermediate step involving the classic elements of the equivalent circuit method (based on the lumped element theory).

In practical cases the filters are loaded with a non-reflecting system, so $\overleftarrow{u}_2 = 0$ and consequently only the first element $N_{11}$ of the matrix is of importance. As will be shown in the subsequent chapters, the properties of a composite filter can easily be derived by matrix multiplication of the matrices of the separate links.

## 2. The wave matrix of a single stage.

Mr. Goubau [2] [3] starts with the introduction of the wave matrix of a waveguide section with the electrical length

$$\Phi = \psi - \psi_v \tag{2}$$

and containing a single discontinuity. The electrical length is related to the material length $l$ by the expression

$$\Phi = 2\pi\, l/\lambda_g \tag{3}$$

$\lambda_g$ being the wavelength in the guide,

$$\lambda_g = \frac{c/f}{\sqrt{1 - \left(\frac{f_c}{f}\right)^2}} \tag{4}$$

where $f_c$ = cut-off frequency of the guide.

The reflection coefficient of the discontinuity can be written as $\Gamma = -\cos\varphi\, \varepsilon^{-j\psi_v}$ and the transmission coefficient as $T = j \sin\varphi\, \varepsilon^{-j\psi_v}$.

A characteristic reflection factor is derived, i.e. that reflection factor measured on the output side of the transducer, that yields a reflection factor, appearing on the inputside of the transducer, of the same value.

The corresponding transmission constant $g$ proves to be given by

$$\varepsilon^g = q + j\sqrt{1 - q^2} \quad \text{with } q = \frac{\sin\psi}{\sin\varphi} \tag{5}$$

The transmission constant will be purely imaginary (passband) for $|q| < 1$ $(\pi - |\varphi| < \psi < \pi + |\varphi|)$. At center frequency $f_o$, $\psi = \pi$.

In general a filter is used between non-reflecting loads and in order to have the characteristic reflection factor at the terminals of the filterstage, matching transformers at input and output are necessary. These transformers are asymmetrical stages and must have the following properties.

*a*. No reflection at the input and output of the total filter at center frequency.

*b*. The output and input transformer connected in series must have the same frequency response as one filterstage.

In this way an original stage with input and output transformers is equivalent to two identical stages connected in series, both consisting of an input and an output transformer. Such a stage is shown in fig. 2.

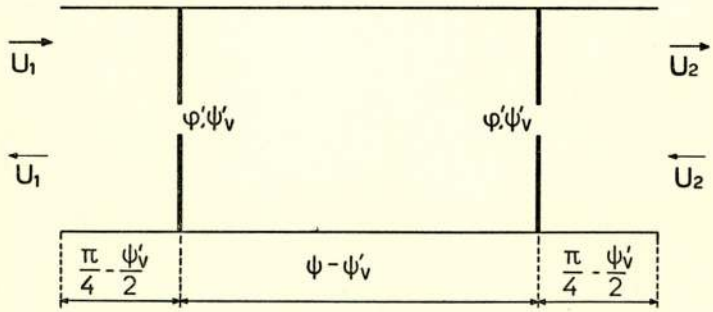


Fig. 2.
Filterstage consisting of two identical discontinuities.

The reflection coefficient of the discontinuity of this new type can be written as $\Gamma = -\cos\varphi'\,\varepsilon^{-j\psi' v}$ and the transmission coefficient as $T = j\sin\varphi'\,\varepsilon^{-j\psi' v}$. The relation between $\varphi$ and $\varphi'$ is given by

$$\cos\varphi' = \frac{1}{\cos\varphi} - \sqrt{\frac{1}{\cos^2\varphi} - 1} \tag{6}$$

and the wave matrix $(N)$ of the filterstage of fig. 2 by

$$(N) = -\frac{1}{\sin^2\varphi'}\begin{Vmatrix} j(\varepsilon^{j\psi} - \cos^2\varphi'\,\varepsilon^{-j\psi}) & \cos\varphi'(\varepsilon^{j\psi} - \varepsilon^{-j\psi}) \\ -\cos\varphi'(\varepsilon^{j\psi} - \varepsilon^{-j\psi}) & -j(\varepsilon^{-j\psi} - \cos^2\varphi'\,\varepsilon^{j\psi}) \end{Vmatrix} \tag{7}$$

Using the equations (5) and (6) we can express the wave matrix (7) as follows

$$(N) = \begin{Vmatrix} q - j\cos\psi & -jq\cos\varphi \\ jq\cos\varphi & q + j\cos\psi \end{Vmatrix} \tag{8}$$

With the approximations $\cos\psi = -1$ around the center frequency and $\cos\varphi \approx 1$, it is found that

$$\begin{Vmatrix} \vec{u}_1 \\ \overleftarrow{u}_1 \end{Vmatrix} = \begin{Vmatrix} q + j & -jq \\ jq & q - j \end{Vmatrix} \begin{Vmatrix} \vec{u}_2 \\ \overleftarrow{u}_2 \end{Vmatrix} \tag{9}$$

If $\overleftarrow{u}_2 = 0$, the insertion loss $L$ will be

$$L = \frac{\left|\vec{u}_1\right|^2}{\left|\vec{u}_2\right|^2} = 1 + q^2 \tag{10}$$

It is readily seen, that for the 3 dB bandwidth $B$, $L = 2$ and $q = 1$. In order to find the relation between $\varphi$ and the bandwidth $B$ we differentiate $\Phi$ (2, 3 and 4) with respect to the frequency $f$.

$$\frac{d\Phi}{df} = \frac{2\pi l}{c} f \sqrt{1 - \left(\frac{f_c}{f}\right)^2}\,\frac{1/f}{1 - \left(\frac{f_c}{f}\right)^2} \tag{11}$$

At the center frequency $f_0$ the equation (11) takes the form

$$\frac{d\Phi}{df} = |\psi|_{f=f_0}\,\frac{1/f_0}{1 - \left(\frac{f_c}{f_0}\right)^2} \tag{12}$$

Hence by putting $sin\,\varphi \approx \varphi$ ($\varphi$ being substantially constant) and using (2) and (3), we obtain

$$sin\,\psi = -\,|\psi|_{f=f_o} \frac{1}{1-\left(\frac{f_c}{f_o}\right)^2} \frac{f-f_o}{f_o}$$

and
$$q = \frac{sin\,\psi}{sin\,\varphi} = \frac{|\psi|_{f=f_o}}{\varphi} \frac{1}{1-\left(\frac{f_c}{f_o}\right)^2} \frac{f-f_o}{f_o} \tag{13}$$

For the 3 dB bandwidth $B$ we find for

$$q = 1 = \frac{|\psi|_{f=f_o}}{\varphi} \frac{1}{1-\left(\frac{f_c}{f_o}\right)^2} \frac{B/2}{f_o} \tag{14}$$

$$\text{and for } \varphi = |\psi|_{f=f_o} \frac{1}{1-\left(\frac{f_c}{f_o}\right)^2} \frac{B}{2f_o} \tag{15}$$

$$\text{and } q = -\frac{2\pi(f-f_o)}{\pi B} = -\frac{\Omega-\Omega_o}{\pi B} \tag{16}$$

$q = \frac{2\,\Delta f}{B}$, being a normalised frequency variable.

The reflection coefficient of the discontinuity required for a given waveguide, a center frequency $f_o$ and a bandwidth $B$, can be obtained by means of the equations (6) and (15).

Till now we have neglected the dissipation loss; it can be taken into account by introducing complex values for $\varphi$ and $\psi$ viz. $\psi = \psi_r + j\psi_i$ and $\varphi = \varphi_r + j\varphi_i$

$$\text{whence } q = \frac{sin\,\psi}{sin\,\varphi} = \frac{sin\,(\psi_r + j\psi_i)}{sin\,(\varphi_r + j\varphi_i)} \approx \frac{sin\,\psi_r}{sin\,\varphi_r} + j\sigma \tag{17}$$

where $\sigma = \frac{\psi_i + \varphi_i}{\varphi_r}$

similar to [4] $\sigma = \frac{Q_E}{Q_u}$

($Q_E = \frac{f_o}{B}$ being the quality factor without dissipation and $Q_u$

the quality factor of the cavity with the dissipation and no coupling with the waveguide).

Once the $Q_u$ for a given waveguide and frequency range is measured or calculated, the ratio $\sigma = \frac{Q_E}{Q_u}$ can be determined for any practical case.

## 3. *n* filterstages with different bandwidths and equal center frequencies

In this chapter we will derive general equations for the insertion loss $L$ without dissipation, the envelope delay $\tau$ and the insertion loss $L_0$ at center frequency due to dissipation, each equation being a function of the ratios of the bandwidths of the single stages to the bandwidth $B$ of the total filter.

Supposing that for the $r^{th}$ stage the bandwidth $B_r = B/a_r$ the frequency variable $q_r$ for this stage will be

$$q_r = \frac{2\,\Delta f}{B_r} = a_r q \tag{18}$$

$q$ referring to the total filter.
According to (9) the wave matrix of the $r^{th}$ filterstage can be written as

$$\left\| \begin{matrix} a_r q + j & -j a_r q \\ j a_r q & a_r q - j \end{matrix} \right\| = a_r q\,(A) + (B) \tag{19}$$

where

$$(A) = \left\| \begin{matrix} 1 & -j \\ j & 1 \end{matrix} \right\| \text{ and } (B) = \left\| \begin{matrix} j & 0 \\ 0 & -j \end{matrix} \right\|$$

The wave matrix $N$ of the total filter is obtained by matrix multiplication

$$(N) = \prod_{r=1}^{n} \left[ a_r q\,(A) + (B) \right] \tag{20}$$

We have indicated in chapter 1 that only the relation $\vec{u}_1 = N_{11}\,\vec{u}_{n+1}$ is of importance.

On performing the matrix multiplication, the multiplication being not commutative, we get several products of $(A)$ and $(B)$, like $(A)(B)$, $(B)(A)$, $(B)(A)(B)$, $(A)(B)(A)$, $(A)(B)(A)(B)$ and

so on. All these products can be simplified to one of the following forms $(A)^k$; $(B)^k$; $(A)^k(B)^l$; $(B)^l(A)^k$; $(B)^l(A)^k(B)^m$ and $(A)^k(B)^l(A)^m$. It can easily be shown that

$$
\begin{aligned}
(A)^k &= 2^{k-1}\,(A) \qquad (21)\\
(B)^k &= j^k \begin{Vmatrix} 1 & 0 \\ 0 & (-1)^k \end{Vmatrix}\\
(A)^k(B)^l &= 2^{k-1} j^l \begin{Vmatrix} 1 & -(-1)^l j \\ j & (-1)^l \end{Vmatrix}\\
(B)^l(A)^k &= 2^{k-1} j^l \begin{Vmatrix} 1 & -j \\ (-1)^l j & (-1)^l \end{Vmatrix}\\
(B)^l(A)^k(B^m) &= 2^{k-1} j^{l+m} \begin{Vmatrix} 1 & -(-1)^m j \\ (-1)^l j & (-1)^{l+m} \end{Vmatrix}\\
(A)^k(B)^l(A)^m &= \begin{cases} 2^{k+m-1} j^l\,(A) & (l = \text{even}) \\ 0 & (l = \text{odd}) \end{cases}
\end{aligned}
$$

From the latter equation one can deduce that if the combinations of the form $a_p\, a_q$ - - - - - $a_v\, a_w$ are arranged in such a way that $p < q \ldots\ldots < v < w$, those combinations of which the difference between any two adjacent indices is even, must be omitted. The first element $N_{11}$ of the matrix $(N)$ can be written

$$
\begin{aligned}
N_{11} &= b_n\, q^n + j\, b_{n-1}\, q^{n-1} + j^2\, b_{n-2}\, q^{n-2} + j^3\, b_{n-3}\, q^{n-3} + \ldots\ldots\\
N_{11} &= b_n\, q^n - b_{n-2}\, q^{n-2} + b_{n-4}\, q^{n-4} - \ldots\ldots\\
&\quad + j\left[\, b_{n-1}\, q^{n-1} - b_{n-3}\, q^{n-3} + b_{n-5}\, q^{n-5} - \ldots\ldots \right] + j^n \qquad (22)
\end{aligned}
$$

Noticing that the first elements of the matrices of the equations (21) are all unity, we obtain for the coefficients $b$ the equations

$$
\begin{aligned}
b_n &= 2^{n-1} \prod_{k=1}^{n} a_k \qquad (23)\\
b_{n-1} &= 2^{n-2} \left[ \prod_{k=1}^{n-1} a_k + \prod_{l=2}^{n} a_l \right]\\
b_3 &= 2^2 \sum_{k=1}^{n-2} \sum_{l=2}^{n-1} \sum_{m=3}^{n} a_k\, a_l\, a_m \qquad k < l < m\\
b_2 &= 2 \sum_{k=1}^{n-1} \sum_{l=2}^{n} a_k\, a_l \qquad k < l\\
b_1 &= \sum_{k=1}^{n} a_k
\end{aligned}
$$

In general $N_{11} = Re\, N_{11} + j\, Im\, N_{11}$, so the insertion loss $L$ of the filter will be

$$L = \frac{Pi}{Po} = \frac{\left|\vec{u}_1\right|^2}{\left|\vec{u}_{n+1}\right|^2} = (Re\, N_{11})^2 + (Im\, N_{11})^2 \tag{24}$$

Applying (22) gives

$$L = s_{2n}\, q^{2n} + s_{2n-2}\, q^{2n-2} + \ldots . s_4\, q^4 + s_2\, q^2 + 1 \tag{25}$$

with $s_{2p} = b_p^2 - 2b_{p-1}\, b_{p+1} + 2b_{p-2}\, b_{p+2} - 2b_{p-3}\, b_{p+3} + \ldots .$ (26)

and $b_0 = 1$

The envelope delay $\tau = -\, d\Theta/d\Omega$ can be obtained from the phaseshift $\Theta$ between $\vec{u}_1$ and $\vec{u}_{n+1}$ given by $tg\Theta = Im\, N_{11}/Re\, N_{11}$.

Noticing that $\dfrac{d\Theta}{d\Omega} = \dfrac{d\Theta}{d\,(tg\,\Theta)} \cdot \dfrac{d\,(tg\,\Theta)}{dq}\; \dfrac{dq}{d\Omega}$

with $\dfrac{dq}{d\Omega} = -\dfrac{1}{\pi B}$ (from equation (16)), we find, applying equation (22), the following expression for $\tau$

$$\tau = \frac{t_{2n-2}\, q^{2n-2} + \ldots . t_6\, q^6 + t_4\, q^4 + t_2\, q^2 + t_0}{s_{2n}\, q^{2n} \quad + \ldots . s_6\, q^6 + s_4\, q^4 + s_2\, q^2 + 1}\; \frac{1}{\pi B} \tag{27}$$

with $t_{2p} = b_p\, b_{p+1} - 3b_{p-1}\, b_{p+2} + 5b_{p-2}\, b_{p+3} - 7b_{p-3}\, b_{p+4} + \ldots .$

and $b_0 = 1$ (28)

It may be observed that the denominator of equation (27) has the same form as the right-hand expression in equation (25). The envelope delay can be written as a power series

$$\tau = \tau_0 + \frac{1}{2!}\left[\frac{d^2\tau}{d\Omega^2}\right]_{\Omega=\Omega_0} (\Omega-\Omega_0)^2 + \frac{1}{4!}\left[\frac{d^4\tau}{d\Omega^4}\right]_{\Omega=\Omega_0} (\Omega-\Omega_0)^4 + \tag{29}$$

$$+ \frac{1}{6!}\left[\frac{d^6\tau}{d\Omega^6}\right]_{\Omega=\Omega_0} (\Omega-\Omega_0)^6 + \ldots .$$

The odd higher derivatives are zero since the envelope delay characteristic is symmetrical with respect to $\Omega = \Omega_0$ (the bandwidth being relatively small with respect to $\Omega_0$). Repeated differentiation of equation (27) shows that for $\Omega - \Omega_0 = 0\ (q = 0)$

$$\frac{d^2\tau}{d\Omega^2} = 2!\left(t_2 - t_o s_2\right)\left(\frac{1}{\pi B}\right)^3 = 2!\, T^{(2)}\left(\frac{1}{\pi B}\right)^3 \tag{30}$$

$$\frac{d^4\tau}{d\Omega^4} = 4!\left(t_4 - t_o s_4 - s_2 T^{(2)}\right)\left(\frac{1}{\pi B}\right)^5 = 4!\, T^{(4)}\left(\frac{1}{\pi B}\right)^5$$

$$\frac{d^n\tau}{d\Omega^n} = n!\left(t_n - t_o s_n - s_{n-2} T^{(2)} - s_{n-4} T^{(4)} - s_{n-6} T^{(6)} \ldots\right)\left(\frac{1}{\pi B}\right)^{n+1}$$

The insertion loss $L_o$ at center frequency due to dissipation is obtained by substituting $q = j\sigma = j\, Q_E/Q_u$ in the equation (22)

$$N_{11} = b_n (j\sigma)^n - b_{n-2}(j\sigma)^{n-2} + \ldots j\left[b_{n-1}(j\sigma)^{n-1} - b_{n-3}(j\sigma)^{n-3} + \ldots\right] + j^n$$

$$= j^n\left[1 + b_1\sigma + b_2\sigma^2 + b_3\sigma^3 + b_4\sigma^4 + \ldots b_n\sigma^n\right]$$

Therefore $$L_o = \frac{\left|\vec{u}_1\right|^2}{\left|\vec{u}_{n+1}\right|^2} = (1 + b_1\sigma + b_2\sigma^2 + b_3\sigma^3 + \ldots b_n\sigma^n)^2 \; (f = f_o) \tag{31}$$

By means of this equation a calculation of the insertion loss at center frequency is possible, if the $Q_u$ is known.

## 4. Waveguide filters comprising three sections.

From the equations (23) we see that

$$b_1 = a_1 + a_2 + a_3 \tag{32}$$

$$b_2 = 2(a_1a_2 + a_2a_3)$$

$$b_3 = 4a_1a_2a_3$$

From the equations (25) and (27) we find the insertion loss

$$L = s_6 q^6 + s_4 q^4 + s_2 q^2 + 1 \tag{33}$$

and for the envelope delay

$$\tau = \frac{t_4 q^4 + t_2 q^2 + t_o}{s_6 q^6 + s_4 q^4 + s_2 q^2 + 1}\;\frac{1}{\pi B} \tag{34}$$

From equation (31) follows that the insertion loss at center frequency is

$$L_o = (1 + b_1\sigma + b_2\sigma^2 + b_3\sigma^3)^2 \quad (\sigma = Q_E/Q_u) \tag{35}$$

The values for the coefficients $s$ and $t$ are obtained from the equations (26) and (28)

$$s_2 = b_1^2 - 2b_2 \tag{36}$$
$$s_4 = b_2^2 - 2b_1 b_3$$
$$s_6 = b_3^2$$
$$t_0 = b_1$$
$$t_2 = b_1 b_2 - 3b_3$$
$$t_4 = b_2 b_3$$

4.1. *The filterstages are identical.*

This is the case treated in ref [2], [3] (Goubau) by introducing the Tschebeyscheff's polynomials. In our approach this means that $a_1 = a_2 = a_3 = a$.

From the equations (32) follows that

$$b_1 = 3a;\ b_2 = 4a^2 \text{ and } b_3 = 4a^3$$

Hence $s_2 = a^2$; $s_4 = -8a^4$; $s_6 = 16a^6$ and

$$t_0 = 3a;\ t_2 = 0;\ t_4 = 16a^5$$

For the insertion loss $L$ we obtain, by using (33)

$$L = 16a^6 q^6 - 8a^4 q^4 + a^2 q^2 + 1 \quad \text{(see fig. 3).}$$

Solution of this equation for $L = 2$ i.e. 3 dB and $q = 1$ gives the value of $a = 0.761$.

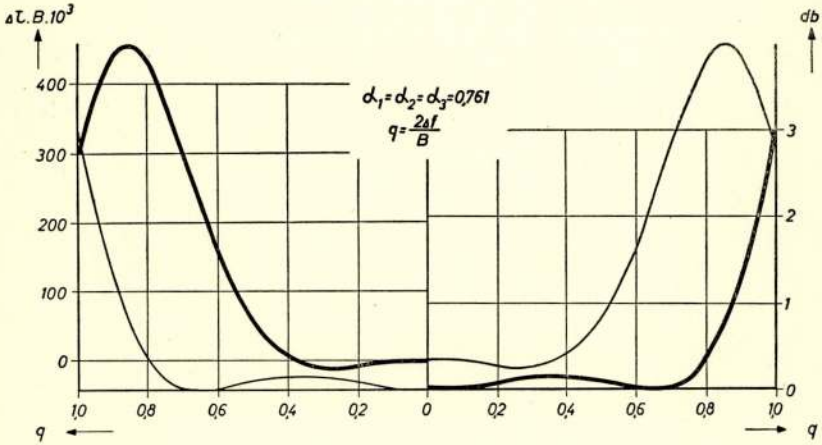


Fig. 3.
The envelope delay and insertion loss of a three-cavity filter with identical stages.

The expression for the envelope delay $\tau$ follows from (34)

$$\tau = \frac{16a^5q^4 + 3a}{16a^6q^6 - 8a^4q^4 + a^2q^2 + 1} \frac{1}{\pi B} \quad \text{(see fig. 3)}$$

and for the insertion loss $L_o$ at center frequency due to dissipation we have, according to (35)

$$L_o = (1 + 3a\sigma + 4a^2\sigma^2 + 4a^3\sigma^3)^2$$

4.2. *The amplitude characteristic as flat as possible.* ("Butterworth": [5])

An amplitude characteristic as flat as possible will be obtained, if $L = 1 + q^6$ (see fig. 4).

Whence in equation (33) $s_2 = s_4 = 0$ and $s_6 = 1$, using (34) we find that

$$\left.\begin{aligned} b_1^2 - 2b_2 &= 0 \\ b_2^2 - 2b_1b_3 &= 0 \\ b_3^2 &= 1 \end{aligned}\right\} \quad \text{and } b_1 = 2;\ b_2 = 2 \text{ and } b_3 = 1$$

From (32) we obtain $a_1 = a_3 = \frac{1}{2}$ and $a_2 = 1$ and after substitution of these values in (36) $t_o = 2$; $t_2 = 1$ and $t_4 = 2$. Therefore the equation (34) for the envelope delay $\tau$ becomes

$$\tau = \frac{2q^4 + q^2 + 2}{q^6 + 1} \frac{1}{\pi B} \quad \text{(see fig. 4)}$$

The insertion loss $L_o$ at center frequency due to dissipation, will be $L_o = (1 + 2\sigma + 2\sigma^2 + \sigma^3)^2$.

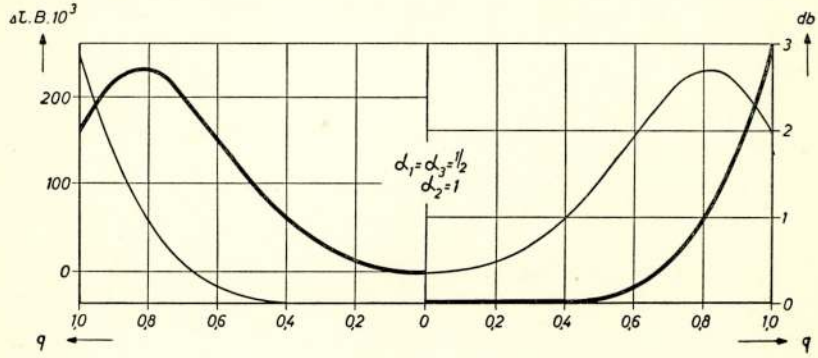


Fig. 4.
The envelope delay and insertion loss of a three-cavity filter with maximally flat amplitude.

### 4.3. *The envelope delay characteristic as flat as possible* ("Thomson": [6])

The condition that should be met implies zero values of $(n-1)$ higher derivatives of the envelope delay. (In our case 2 higher derivatives).

From the equations (30) therefore follows that $t_2 - t_0 s_2 = 0$ and $t_4 - t_0 s_4 = 0$ and from (36) that $5b_2 = 2b_1^2$ and $15b_3 = b_1^3$.

Substituting these values in the equations (33) and (34) we find for

$$L = \frac{1}{225}(b_1 q)^6 + \frac{2}{75}(b_1 q)^4 + \tfrac{1}{5}(b_1 q)^2 + 1 \quad \text{(see fig. 5)}$$

$$\text{and } \tau = \frac{6(b_1 q)^4 + 45(b_1 q)^2 + 225}{(b_1 q)^6 + 6(b_1 q)^4 + 45(b_1 q)^2 + 225}\,\frac{b_1}{\pi B} \quad \text{(see fig. 5)}$$

Solving the equation for $L = 2$ i.e. 3 dB and for $q = 1$ we find for $b_1 = 1.7557$. With this value we can by means of the relations between $b_1$, $b_2$ and $b_3$ calculate the values $a_1$, $a_2$ and $a_3$, yielding $a_1 = 1.1017$; $a_2 = 0.4853$ and $a_3 = 0.1687$. The insertion loss $L_0$ at center frequency due to dissipation will be $L_0 = (1 + b_1\sigma + b_2\sigma^2 + b_3\sigma^3)^2$.

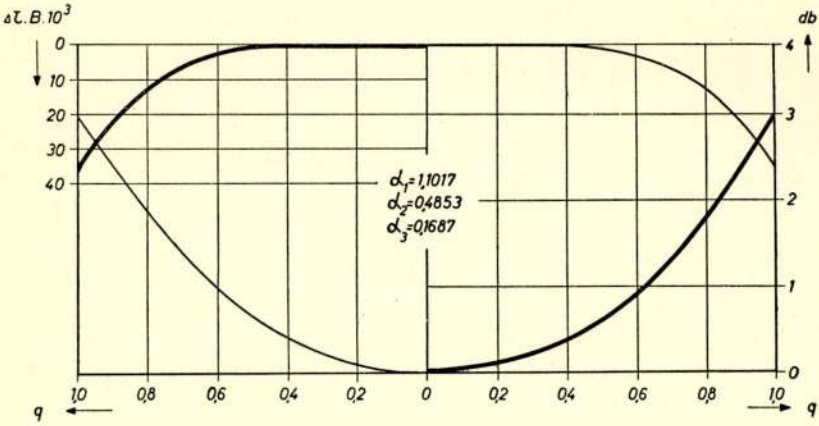


Fig. 5.
The envelope delay and insertion loss of a three-cavity filter with maximally flat envelope delay.

### 4.4. *Special cases.*

Apart from the types treated in 4.1, 4.2 and 4.3 still other variants may be of importance.

Due to phase non-linearity intermodulation noise is introduced into an FM multi-channel telephone system. The noise can be expressed in terms of the derivatives of the envelope delay, as can be seen from an equation given by Starr and Walker [7]. It may be expected that by using an effective frequency deviation, small with respect to the filter bandwidth and by making only one higher derivative of the envelope delay equal to zero, the intermodulation noise can be reduced in comparison with the Butterworth design. In this way an amplitude characteristic can be obtained which in comparison with the Thomson design is more suitable. Proceeding in this way we have only one condition for the bandwidth ratios; for simplicity two stages are made identical. This means that two arrangements of bandwidths are possible, viz. $a_1 a_2 a_1$ and $a_1 a_1 a_2$.

The equations (32) will be in

| case I | case II |
|---|---|
| $b_1 = 2a_1 + a_2$ | $b_1 = 2a_1 + a_2$ |
| $b_2 = 4a_1 a_2$ | $b_2 = 2(a_1^2 + a_1 a_2)$ |
| $b_3 = 4a_1^2 a_2$ | $b_3 = 4a_1^2 a_2$ |

From the condition $t_2 - t_0 s_2 = 0$ and from the equations (36) we find $3b_1 b_2 - 3b_3 - b_1^3 = 0$.

So we get for case I and case II resp.

$$8a_1^3 + a_2^3 - 6a_1 a_2^2 = 0 \text{ and } 4a_1^3 - a_2^3 - 6a_1^2 a_2 = 0.$$

We can express $a_1$ in $a_2$ or v.v.

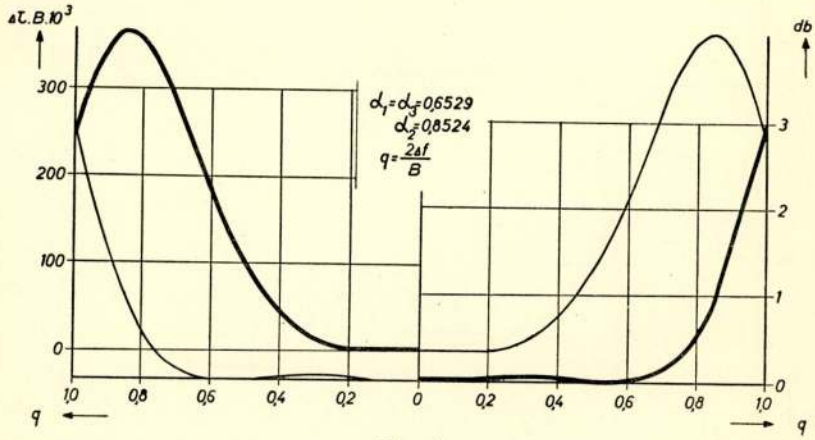


Fig. 6.

The envelope delay and insertion loss of a three-cavity filter, one higher derivative of the envelope delay being zero.

The first equation yields two solutions. Substituting the values for $s$ from (36) in the equation (33) we find for the insertion loss $L$ with $L = 2$ and $q = 1$, $a_1$ or $a_2$

case I $\quad a_1 = a_3 = 0.6529$ and $a_2 = 0.8524$

$\quad a_1 = a_3 = 0.2096$ and $a_2 = 1.2071$

and case II $\quad a_1 = a_2 = 0.8194$ and $a_3 = 0.5128$

The characteristics of the insertion loss $L$ and the envelope delay $\tau$ are shown in figures 6, 7 and 8.

These filter designs can be considered as a variation between the Butterworth and the Thomson design which are similar to the transitional Butterworth-Thomson filter in [8].

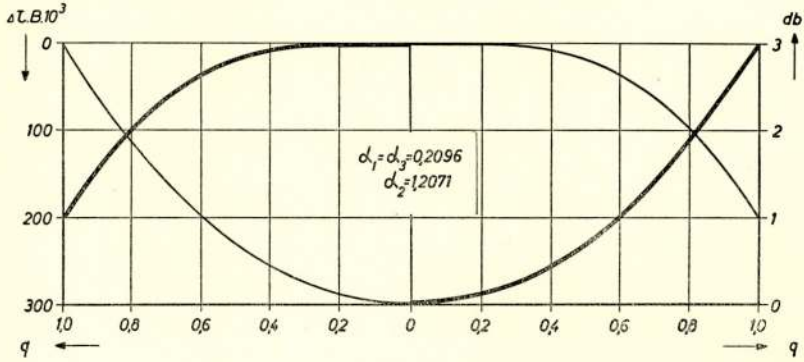


Fig. 7.

The envelope delay and insertion loss of a three-cavity filter. one higher derivative of the envelope delay being zero.

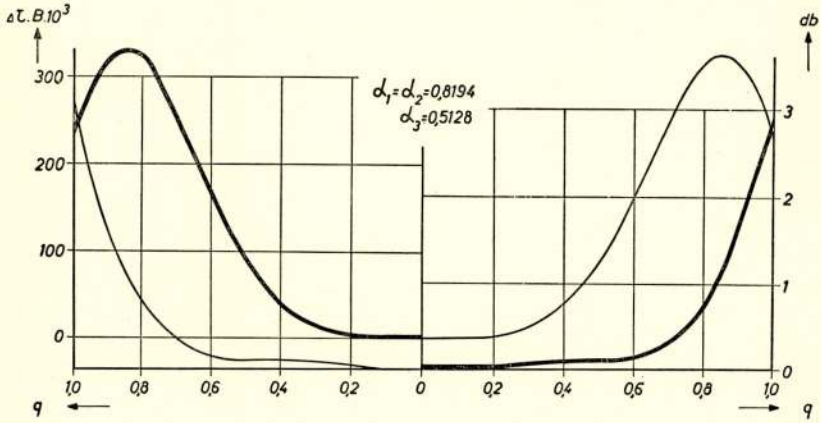


Fig. 8.

The envelope delay and insertion loss of a three-cavity filter, one higher derivative of the envelope delay being zero.

**Acknowledgement.**

The author is indebted to Mr. L. Krul for many helpful and valuable discussions, and to the Mathematical Branch of the Dr. Neher Laboratory for the computational work.

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**AVONDCOLLEGES IN DE STERRENKUNDE**

Het Pedagogisch Instituut, afdeling Leraarsopleiding aan de Rijksuniversiteit te Utrecht, Lucas Bolwerk 11, tel. 21741, verzocht ons de volgende mededeling op te nemen.

Op de donderdagen 16, 23, 30 januari en 6 februari 1958 zullen aan de Sterrenwacht der Rijksuniversiteit te Utrecht avondcolleges gegeven worden voor afgestudeerden. Het onderwerp van deze colleges is Radiosterrenkunde.

De aanvang der colleges is 19,15 precies. Einde 21.00 uur.

Aan het volgen van deze colleges zijn geen kosten verbonden.

Leraren kunnen hun reiskosten (geen verblijfkosten) vergoed krijgen voorzover deze kosten hoger zijn dan *f* 2,50.

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**FLUGNAVIGATION UND FLUGSICHERUNG**

Van 19 tot 24 mei 1958 wordt te Berlijn een congres gehouden met als onderwerp luchtvaartnavigatie en luchtverkeersbeveiliging.

Nadere inlichtingen en aanmeldingen bij: Ausschuss für Funkortung, Am Wehrhahn 94, Düsseldorp.

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# Boekbesprekingen

*Basic mathematics for radio and electronics*, door F. M. Colebrook, B.Sc., D.I.C., A.C.G.I. Derde druk, juli 1957, bewerkt door J. W. Head, M.A. (Cantab). Voor Wireless World uitgegeven door Iliffe and Sons Ltd, London; 359 bladzijden, 90 figuren, 7½" x 5". Prijs 17 s 6 d.

Het boek is een door J. W. Head bewerkte uitgave van "Basic mathematics for radio students" en is bestemd voor hen die, reeds vertrouwd zijnde met de radiotechniek en de elektronica als zodanig, het gevoel hebben in hun wiskundige kennis te kort te schieten.

In de eerste vier hoofdstukken worden de elementaire algebraïsche bewerkingen alsmede de begrippen logaritme, complex-getal, limiet en reeks besproken. In hoofdstuk 5 wordt na de behandeling van enige stellingen uit de vlakke meetkunde en de goniometrie het begrip vector ingevoerd waarbij gebruik wordt gemaakt van de complexe schrijfwijze. Hoofdstuk 6 bespreekt in vogelvlucht de differentiaal- en integraalrekening.

In hoofdstuk 7 worden vervolgens enige op het vakgebied liggende toepassingen gegeven (lopende en staande golven, voorbeelden van de complexe rekenwijze).

Tenslotte volgen dan nog 2 door Head aan het oorspronkelijke werk toegevoegde hoofdstukken, waarvan het eerste de operatorenrekening van Heaviside en het tweede enige begrippen uit de matrixalgebra, het numerieke rekenen en de waarschijnlijkheidsrekening behandelt. De operatorenrekening wordt gegeven in de vorm van een rekenrecept, hetgeen de argeloze lezer voor onverwachte moeilijkheden kan plaatsen.

Gezien deze grote verscheidenheid kan men niet verwachten dat alle onderwerpen die plaatsruimte kregen waarop ze krachtens hun belangrijkheid recht hebben; zo worden de reeksen van Fourier b.v. wel erg summier behandeld. Dit neemt echter niet weg dat de radiotechnicus groot nut van dit werk kan hebben, terwijl wellicht ook de beginnende docent er zijn voordeel mee kan doen.

Het boek is vlot geschreven en typografisch uitstekend verzorgd. Een honderdtal opgaven met antwoorden bevordert de zelfwerkzaamheid van de lezer.

L. K.

*Electronisch Jaarboekje 1958,* 11de uitgave, 160 bladzijden tekst en illustraties, 9½ x 14 cm, uitgegeven door „de Muiderkring" te Bussum. Prijs *f* 2,95.

Het electronisch jaarboekje heeft zich zo langzamerhand een bijzondere plaats veroverd bij allen die zich met de practische radiotechniek bezig houden. In overzichtelijke vorm zijn tabellen, gegevens betreffende televisie, gegevens van de meest gebruikte electronenbuizen opgenomen. Veel aandacht is ook besteed aan transistoren en geluidsregistratie. Diverse schema's van in „Radio Bulletin" gepubliceerde ontwerpen zijn eveneens opgenomen. Een korte verklaring hoe de opgenomen nomogrammen betreffende resonantiekringen moeten worden gebruikt achten wij gewenst aangezien vele amateurs hiermede niet op de hoogte zijn. Het boekje bevat tevens een agenda, en kan dus als normale zakagenda gebruikt worden. De uitvoering is zeer verzorgd.

H.

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# Uit het Nederlands Radiogenootschap

## PERSONALIA

### PROF. Ir. W. H. VAN ZOEST

Met ingang van 1 juli 1957 werd Ir. W. H. van Zoest benoemd tot buitengewoon hoogleraar in de Afdeling der Elektrotechniek van de Technische Hogeschool te Delft, om onderwijs te geven in de lijntransmissie.

De Heer Van Zoest begon zijn carrière, na enige jaren als assistent bij Professor Bähler te hebben gewerkt, op 16 maart 1934 bij het Staatsbedrijf der P.T.T. bij het Laboratorium van de dienst Kabels en Versterkers. Hij bekleedde daar vele rangen en werd op 1 januari 1954 benoemd tot Hoofdingenieur der P.T.T. bijzondere diensten. Hij vervulde de functie van Chef van het transmissie laboratorium en werd aangewezen als plaatsvervangend directeur der laboratoria der P.T.T. Op 1 januari 1955 volgde zijn benoeming tot Hoofdingenieur algemene dienst der P.T.T.

Vele publicaties over uiteenlopende onderwerpen op het gebied van kabels, transmissie, distributie, verschenen van zijn hand.

Gaarne wensen wij Ir. Van Zoest veel geluk bij het aanvaarden van de taak van Buitengewoon Hoogleraar aan de Technische Hogeschool.

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**CORRECTIE OP No. 5**

De foto's op de blz. 307 en 312 van het artikel van M. Stanesby dienen verwisseld te worden.

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